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EDITOR, H. P. Westman

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This Issue in Brief

Application of Pulse-Code Modulation to an Integrated Telephone Network—Part 1 describes the advantages of applying pulse-code modulation to telephone networks. In pulse-code modulation, a speech wave is sampled periodically at a rate at least twice its highest frequency. The amplitude of each sample is defined by a number that is converted into a code group of pulses. After transmission as on-or-off pulses, they are reconverted to the corresponding amplitude values at the receiver.

Distortion in transmission occurs only if noise exceeds half the amplitude of a pulse in a position where no pulse should be or if more than half of the amplitude of a pulse is lost. Pulses are so brief that reflections from impedance mismatch are unimportant. Noise, crosstalk, and attenuation are independent of the number of repeaters and of the length of the transmission path.

Many trains of pulses, each for a different channel, can be interleaved to reduce the number of transmission paths. Transmission is practicable over ordinary telephone pairs in cable with repeaters at 2-kilometer (1.25-mile) intervals for 24 voice communication channels, using one pair for each direction in a 4-wire arrangement. The pairs for 2 conventional channels then accommodate 24 channels.

Initially, pulse-code modulation may be used to serve up to, say, 200 subscribers through a concentrator that may be connected to the exchange either by pulse-code or conventional connections. This will reduce greatly the number of space-path crosspoint switches as each crosspoint will be able to carry 24 channels.

A local exchange will serve a group of concentrators that need not be physically in the central office but each may be near its own group of subscribers. It will, of course, also provide paths to other exchanges. All connections will be over pulse-code multiplex highways. A transit exchange will operate on similar principles. There are only 2 basic switching stages in such a system, connections to the subscribers through concentrators and interconnections between pulse-code multiplex highways.

The timing of the interleaved series of pulses of each channel is critical for proper demodulation and switching. To reserve this particular time interval in all the exchange equipment traversed by that signal would require much duplication of equipment. Speech memories are therefore provided at the incoming and outgoing highways and are interconnected by a crosspoint, all being under control of switches that connect them at proper time intervals to permit the incoming signal to be stored in an incoming memory, transferred at a convenient time through the crosspoint to an outgoing memory, from which it is read at the proper time to transmit it to the next exchange or to the other subscriber.

Despite the flexibility of speech memories, blocking can still occur in the exchange and connections may be broken down and reconnected by a rearrangement method.

Pulse-code modulation uses signals similar to those for data transmission and will inherently provide wide frequency bands also suited to data transmission. Although signal distortion and losses of pulses are of no significance in telephony, some redundancy will be needed for adequate accuracy for data signals.

Part 2 is on transmission and encoding. Because in pulse-code modulation the amplitude, duration, and precise position of pulses are of no significance, the pulses can be regenerated at any point and transmission quality will be completely independent of transmission distance and of the number of repeaters. Consequently, the transmission-line equivalent remains absolutely constant regardless of variations in the transmission medium.

Use of digital techniques alone is adequate for all functions, such as modulation, demodulation, regeneration, control and supervisory operations, and speech transmission. Semiconductors fit readily into such a system. Speech transmission and switching information use the same

kind of pulses and are handled over the same paths with about the same bandwidth requirements.

In coding, the amplitude of the sample is quantized to 1 of a prescribed number of levels. As the number of levels is increased, the uncertainty of exactly where the signal amplitude falls is decreased as each band represented by a code number becomes narrower.

To provide minimum and equal percentage widths for these levels as a function of the amplitude, nonlinear quantization is used. This is equivalent to amplitude compression followed by linear quantizing. The coder selected is completely digital in operation. It uses a compressor characteristic that can be reduced to a group of 5 segments of straight lines with different slopes producing 128 quantization steps.

In the decoder, the digits for each code group of pulses are accumulated in a digital memory and then decoded to give the resultant value of the sample.

Part 3 treats switching. There are only 2 switching stages required, one for connecting subscribers' lines to a pulse-code-modulation multiplex highway and the other for interconnecting such highways. The timing of the individual pulse trains to permit multiplex operation and the synchronization and coordination of these timing intervals for switching as a call passes through an exchange and from one exchange to another are important elements in the switching design.

For average traffic, a concentrator will serve 150 subscribers' lines and for light traffic 200 lines. It may be located near the subscribers and remote from the exchange to which it is connected. In an experimental system constructed to test the design, 7-digit code groups to designate the instantaneous amplitude of each sample of speech are transmitted at 10 kilohertz to provide 24 channels on a 4-wire basis using 2 telephone pairs for each conversation.

The subscribers' lines are connected to a concentrator, which assigns to each of 24 active lines a time interval for multiplexing and pulse-code modulates the speech samples during 4-microsecond intervals repeated every 100 microseconds. The pulse-code multiplex output of the concentrator goes to the exchange via 2 telephone pairs, 1 for each direction, called a highway.

At the exchange, the highway from a concentrator terminates in 2 speech memories, 1 for each direction. These 2 memories each have 24 lines in which speech code groups may be written for the 24 conversations it must handle. A free line, equivalent to a multiplexing time interval, is assigned to each new call by a time-switching memory.

As the outgoing highway must use for each new call a multiplexing time interval that is free to the concentrator or exchange to which the call is directed, a similar pair of memories under control of another time-switching memory is provided. These 2 memories will then be interconnected to transfer their pulse groups to each other through an electronic crosspoint switch under control of a space-switching memory.

All 3 switching memories record in identically numbered memory lines the addresses of the memory lines in the 2 speech memories so as to connect these 2 addresses together through the crosspoint for a 4-microsecond interval each multiplex cycle. Thus, through these memories, a speech sample arriving at the exchange at a randomly assigned interval in the multiplex cycle is transferred to another memory for transmission elsewhere at another randomly assigned multiplex time interval.

Multiplex switching is controlled by a clock in each exchange and variations among exchange clocks or variations in propagation time over lines or through repeaters are corrected by introducing under control of a synchronization circuit the required number of flip-flop stages to delay a "fast" pulse train.

If an incoming multiplex highway offers a call to an exchange at a time interval that is not available on the outgoing multiplex highway

to which it is directed, internal blocking, which is a characteristic of all time-division-switching systems, occurs. A rearrangement of existing connections can then be made to accommodate the new call.

Experimental Pulse-Code-Modulation Transmission for Local-Area Telephony—A field experiment was made of pulse-code modulation for interexchange junctions in Madrid on 18 telephone pairs of 0·64-millimetre wire in existing 900- and 600-pair cables over a distance of 6·2 kilometres with unattended repeaters installed at 1-kilometre spacings in manholes.

The 24-channel system provided 23 speech paths and 1 synchronizing channel with signalling facilities suited to the rotary switching in the exchanges. Speech sampling was at 8 kilohertz with 70 quantization levels and about 25 decibels of signal compression. A unit-disparity unit-distance code requires 7 binary digits for speech and an 8th digit was added for signalling, making $1 \cdot 536 \times 10^6$ digits per second. By further time division, the 8th pulse provides 4 signalling conditions.

A unit-disparity code limiting to 1 the difference in the number of marks and spaces gives each code group an average density of pulses and transitions that aids in extracting the 1·536-megahertz timing wave from the received signals, reduces crosstalk, and permits codes for levels near 0, which are most frequently encountered, to avoid extreme numbers of transitions. A unit-distance code, in which codes for adjacent levels differ by only 1 digit, reduces errors that may occur in a parallel coder if several digits are simultaneously in a state of indecision.

Each speech input, after filtering and amplitude limiting, is sampled for 2·6 microseconds every 125 microseconds to produce amplitude-modulated pulses that are then multiplexed. After compression, the amplitude of each pulse is quantized by an array of comparators biased to operate at different levels. Each comparator is designed to define 2 levels. Through a diode matrix, the 7 digits of each code group are produced in parallel by the output of the comparators. The receiver is complementary to the transmitter. It uses magnetic cores to store the serially received digits for parallel readout. Transistor amplifiers and pulse-shape regenerators are used.

The compressor uses the approximately logarithmic characteristic of diodes, which are temperature controlled for stability. The expander is given the complementary characteristic by using a compressor network in a reverse feedback path.

Speech quality was good at optimum volume, tolerable for a 20-to-25-decibel range, and intelligible for a range of 40 to 45 decibels. A moderate extension in volume range over that provided by the test equipment would be desirable. Digit rates of 1·5 megahertz are feasible over paper-insulated multi-pair telephone cable and repeater spacings may exceed 1 kilometre.

United Kingdom–Faroes–Iceland (Scotice) Submarine-Cable Telephone System—The South section between Gairloch, Scotland, and Torshavn, Faroe Islands, is 290 nautical miles (537 kilometres) and includes 10 submerged repeaters. The North section linking Torshavn and Vestmannaeyjar, Iceland, is 408 nautical miles (756 kilometres) and is equipped with 15 repeaters and an equaliser. The polythene coaxial cable is 0·46 inch (11·7 millimetres) in diameter and has a characteristic impedance of 52 ohms. In one direction in the band between 24 and 96 and in the other direction between 120 and 192 kilocycles per second, there are 24 telephone channels, a third of which may be used for voice-frequency telegraphy. Engineering communication uses the bands 18–24 and 192–198 kilocycles per second and supervisory signals occupy bands from 96 to 100 and from 192 to 200 kilocycles per second.

Each repeater consists of a parallel pair of 3-stage vacuum-tube amplifiers whose gain rises from 22 decibels at 24 kilocycles per

second to 65 decibels at 196 kilocycles per second. The amplifier being used for both directions of transmission, the required stability margin of attenuation is provided by filters. Power is supplied equally from both terminals over the cable with a drop of 120 volts for each repeater. Repeaters are mounted in rigid cylindrical steel housings and the electrical connections passing through the bulkheads are tested at 5 tons per square inch (787 kilogrammes per square centimetre) for a month before being accepted for actual use.

All the cable for a link and its repeaters were connected together and stowed aboard ship. As the cable is laid, insertion-loss-versus-frequency measurements were made to the shore terminal to predict the required characteristics of the equaliser, which was assembled to these specifications just before being jointed to the cable and laid.

Shipboard-Adjustable Submerged Equaliser—Equalisers are required on long submarine telephone cables to adjust for failure of repeater gain and cable loss to compensate, say, within 3 decibels, for each other over the entire frequency range of transmission. The attenuation characteristic of the cable determines the design of the repeaters. In addition to design limitations, manufacturing tolerances for both cable and repeaters, and variations in cable characteristic resulting from laying and sea-bottom conditions affect equalisation.

Part of the equalisation can be predicted but part must be designed, built, and inserted in the cable on shipboard as a result of measurements made during the laying operations and extrapolated to the completed length. The laying of the cable must not be interrupted while the equaliser is completed, installed in the pressure housing, and jointed to the cable, which requires about 16 hours.

That part of the equaliser to compensate for the predictable errors is built and hermetically sealed on shore in an environment of "dairy" cleanliness. To insure the same quality and reliability for the part that must be added on shipboard, several units of each component of resistance, inductance, and capacitance are also hermetically sealed in cans on shore for assembly on shipboard. A wide range of canned components is carried although only a few cans are used for each equaliser.

Equalisers are mounted in the same pressure housings as are used for repeaters. Made of high-tensile steel, the main body is a cylindrical tube or casing closed by a bulkhead at each end. Each bulkhead is sealed by two *O* seals carried in grooves round its periphery. Petroleum jelly is forced into the space between the seals to a pressure comparable to that at sea bottom and when submerged is raised slightly above sea pressure by a pressure-intensifying arrangement of pistons of different diameters. This pressure can be measured as frequently as desired without loss of pressure or jelly and provides a ready means for checking the sealing of the housing before laying.

A lead gasket between the inner face of the bulkhead and a retaining shoulder on the casing provides a diffusion seal against possible entry of water vapour passing the *O* seals. This lead seal is pre-stressed to insure that the lead has flowed even though in shallow water the sea pressure would not be sufficient.

A cable centre conductor is moulded in polythene through a threaded hole in the bulkhead and the polythene also covers a ridged stem that extends the hole on the high-pressure side. Water pressure seals the polythene to the stem, and the threading in the entry hole prevents the pressure from extruding the cable into the apparatus compartment.

The cable is firmly anchored within the outer end of the housing and is secured against rotating in the housing. The anchorage, bulkhead, and apparatus frame may be removed and replaced aboard ship with a few simple tools, and the sealing can be verified to be adequate to withstand sea pressure before the housing is layed.

Variable Reactance Frequency Multipliers—A nonlinear reactance, although free from dissipative losses, accepts real power from a source and releases it at other frequencies. It thus appears as a power-consuming device to the source but, since no power is retained, the over-all power input equals zero.

In the case of a lossless nonlinear resistance, the maximum efficiency as a frequency doubler would be 25 percent, the remaining 75 percent would be converted to direct current. The lossless nonlinear reactance does not rectify and so releases its power at one or more harmonic frequencies; the efficiency of conversion to a single frequency can reach the theoretical limit of 100 percent.

The characteristics of a voltage-sensitive semiconductor capacitor are given and some features affecting efficiency, power, circuit complexity, and tuning, especially for large frequency multiplication over wide bands, are treated. An example given starts with a band of frequencies in the 500-megacycle-per-second region and multiplies them to an 8000-megacycle-per-second band.

Supporting theoretical data appears in an appendix on the frequency doubler, the simplest and most efficient of the varactor frequency multipliers.

Contributions to Studies of Overflow Traffic—Consideration is given first to a simple primary group of trunks overflowing to a secondary group provided only for that purpose. Both rigorous and approximate equations are given for the occupation probabilities of the overflow group.

Next the case of 2 primary groups overflowing into a common secondary group is considered for various conditions of offered traffic and number of lines but with the same mean and variation of overflowing traffic. Approximate equations are given for the blocking probabilities or losses in the traffic offered to each primary group. These results are compared with those obtained using the method of Wilkinson; only slight discrepancies are evident, the latter being pessimistic for high-congestion conditions of overflow and optimistic for low congestion values. Finally, the results are generalized for any number of primary groups.

Interregister Multifrequency-Code Signalling for Telephone Switching in Europe—A description is given of a multifrequency-code signalling system that has been accepted by many of the leading telephone administrations in Europe. Based on the simultaneous transmission of 2 out of 6 voice frequencies between 1380 and 1980 hertz in the forward direction and between 500 to 1140 hertz in the backward direction, combinations are provided for 15 codes in each direction. By use of a "shift" signal, the number of useful codes is increased. Tables of these codes are given and their application in local, toll, national, and international networks is listed in detail.

The availability of an independent frequency band in each direction permits the use of compelled signalling. The duration of a signal is not important. The forward signal, for instance, is maintained until a proper reply signal is returned to the source to confirm its receipt and the performance of the required action.

End-to-end signalling is envisioned. In it a register in the office nearest the subscriber, or in a toll center acting as a pivotal point for a number of local areas, maintains control of the switching through all later offices, providing only the information necessary at each switching point to advance the connection to the next office and finally supplying the last digits of the called number to the exchange to which the called subscriber is connected.

Definitions and functions of line signals and definitions of general terminology are given in two appendixes.

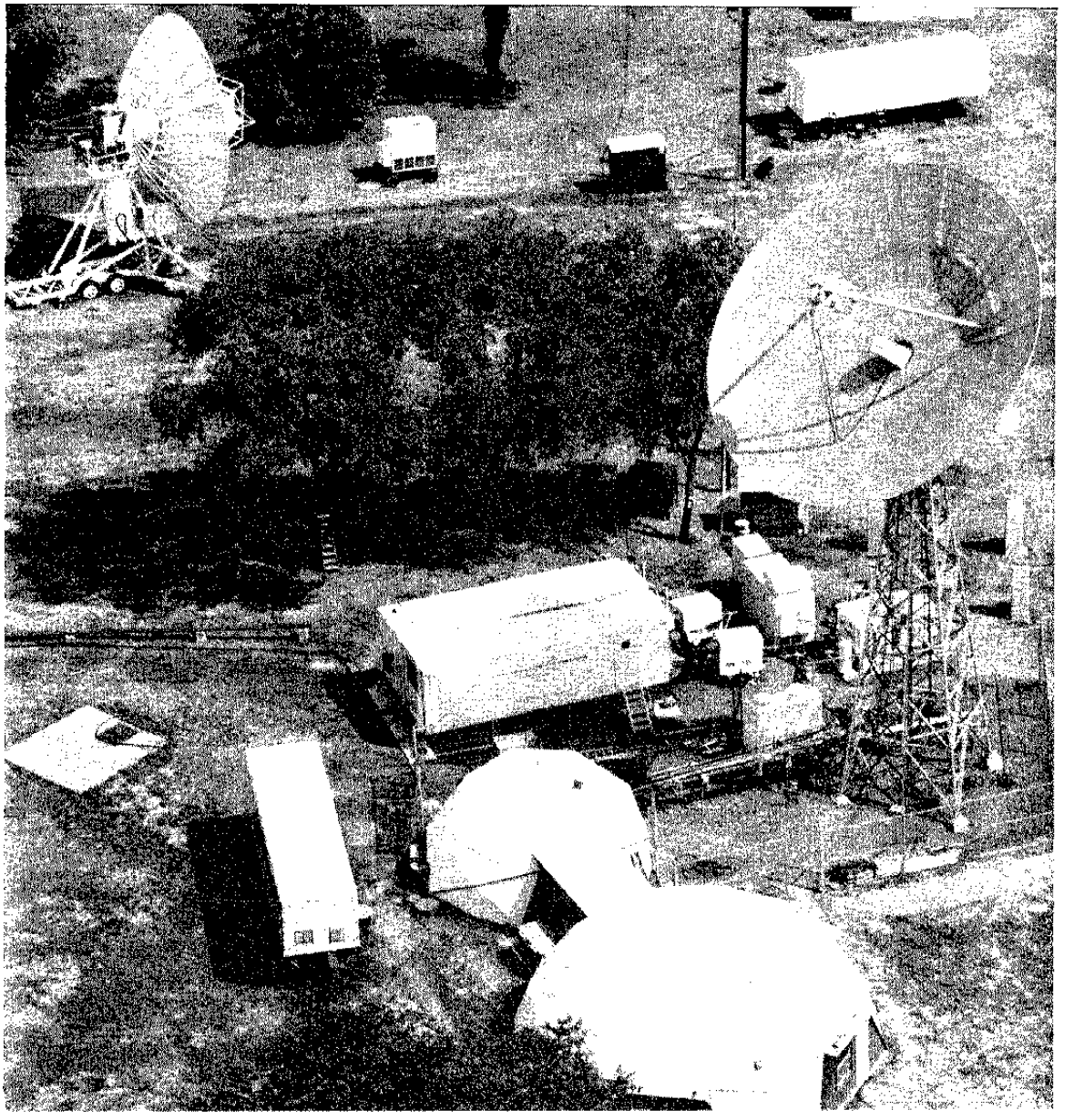


Figure 1—Project Relay satellite ground equipment. The permanent station in the foreground is at Nutley, New Jersey, and the transportable station in the background was sent to Rio de Janeiro for the linking of North and South America by radio via the satellite.

Project Relay—As part of its plans for Project Relay, the National Aeronautics and Space Administration provided for tests to link North and South America via the satellite over a 5000-mile (8000-kilometer) radio circuit.

In the foreground of Figure 1, is the permanent ground terminal near New York City that, built two years ago, is the first private station licensed by the Federal Communications Commission for earth-space communication

experiments. It uses a 40-foot (12-meter) paraboloidal antenna. The terminal apparatus is housed in two igloo-type structures and includes a transmitter, tracking and communication receivers, antenna controls, monitoring consoles, and means of combining several voice channels into one signal for ease in handling.

In the background of the picture are the antenna and control van of a transportable station to be set up near Rio de Janeiro, Brazil, for operation by Companhia Radio Internacional do Brasil. This station travels in a van and three trailers and can be shipped by sea, air, or road. It can be erected by four men in as little as 16 hours. The 30-foot (9-meter) paraboloidal antenna can be dismantled into pie-shaped pieces. The station can handle 12 simultaneous two-way telephone conversations; 12 teleprinter or data circuits can replace each voice channel.

ITT Federal Laboratories
United States of America

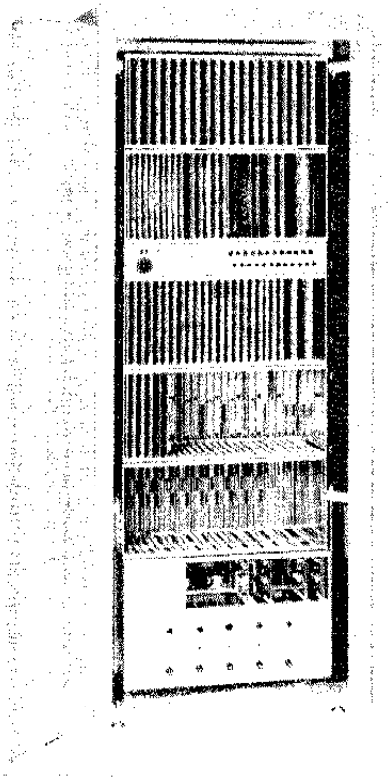
Electronic Telephone Switching with *PNPN* Diodes—A recently demonstrated 100-line private automatic telephone switching system provides a separate physical path for each two-way speech channel. The connections are established by a single equipment on a time-sharing basis.

When points on the input and output rows and columns of a 3-stage network of 4-layer *PNPN* semiconductor diodes are marked by potentials, a series of the diodes will fire sequentially across the network to form a conducting path between the marked points. A path will always be found between any two terminal points, but this will not always be the same path. Defective diodes will not disrupt traffic but will simply be avoided in setting up a path.

Lifting of the subscriber's handset will mark one side of the network and actuate a link allotter, causing a free link to mark the other side of the network. Diodes across the network then fire, connecting the subscriber to the link and initiating dial tone. Responding to the dialled digits, the link will mark the network input terminal of the called line. A second path will then fire through the network and connect the called line to the link and through it to the calling line. Ringing tone and then speech current are connected.

A printed-circuit board accommodates each 5 subscribers' line circuits, which give access to the link circuits. In the reverse direction, a link is connected to the line circuit of the called subscriber's line. The link circuit provides dial tone, stores the dialled digits, and establishes a route to the called line. It also provides ringing or busy tone. Fail-safe performance is provided by link circuits. The link allotter serves all calls on a one-at-a time basis. On failure, transfer to a stand-by unit is automatic.

Figure 2—A 100-line all-electronic private telephone exchange using *PNPN* semiconductor diodes for switching.



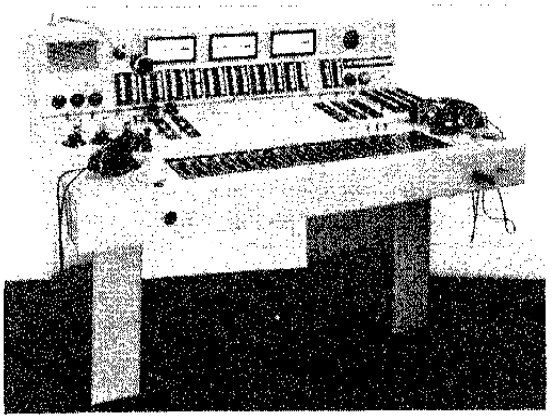
Maintenance may be by nontechnical personnel. An audible tone signals a malfunction and a lamp indicates the defective circuit board, which is simply withdrawn and a spare plugged in to replace it. A 100-line equipment is shown in Figure 2.

ITT Kellogg
United States of America

Stereophonic Studio Equipment—Two broadcast studios especially designed for stereophonic pickup have been installed for the Belgian Radio and Television service. They may be used for monaural operation also.

For stereophonic pickup, 6 input channels may be mixed and transmitted over the required 2 output channels. Each input channel has 2 connections for microphones and 7 for magnetophones, turntables, and other external sources. The attenuators for controlling and mixing inputs can be coupled in pairs. They are equipped with contacts for signalling purposes and to connect any of the additional inputs to monitoring loudspeakers via cue circuits. The output signals, at a level of 6 decibels referred to 1 milliwatt, are monitored with loudspeakers and light-spot modulation meters.

Figure 3—Belgium Radio and Television stereophonic broadcast studio equipment.



In addition to the required complement of amplifiers, power supplies, and control equipment, each studio is provided with 3 magnetophones and 3 turntables. One of the control consoles is shown in Figure 3.

Bell Telephone Manufacturing Company
Belgium

Telex Automatic Exchanges—Automatic teleprinter switching systems have been inaugurated in the New York and San Francisco offices of the American Cable & Radio Corporation. They permit its telex subscribers to set up international and intercontinental connections. All incoming traffic is handled on a fully automatic basis as is outgoing cable traffic. Outgoing radio traffic is on a semiautomatic basis.

Charges are recorded on perforated tape and reported to the calling subscriber by teleprinter after each call. The tape is also used for billing.

The New York exchange has a capacity of 1300 subscribers and 180 trunks connected to the most important switching centers of South America and Europe, to the domestic Western Union and telex systems, and to the Canadian telex network. The San Francisco installation will accommodate 200 subscribers and 37 trunks giving access to switching centers for Japan, Philippines, and Hawaii as well as to the domestic Western Union and telex systems. Both permit ordinary cablegrams to be exchanged via teleprinter.

Bell Telephone Manufacturing Company
Belgium

American Cable & Radio Corporation
United States of America

Telex Experiments Via Telstar—Following several successful exchanges between the United States and France via Telstar, the communication satellite of the American Telephone and Telegraph Company, the first direct connection between telex subscribers over this relay between London and New York occurred on 23 August 1962 and on the following day between Paris and New York.



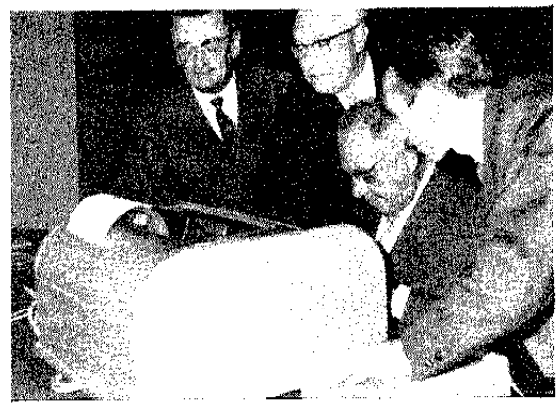


Figure 4—First experimental Telex connections between subscriber stations in New York and Europe via the Telstar communication satellite occurred on its 406th orbit of the earth. At left is the group in Croydon, England. From left to right are: Group Captain F. C. Daubney, Chairman of the Board and Managing Director of Creed & Company; Mr. Richard Thompson, Member of Parliament; Mr. D. L. B. Lithgow, General Commercial Manager of Creed; and Councillor John L. Aston, Mayor of Croydon. In Paris, at the right, also from left to right, are: Mr. E. Delpoux, Commercial Manager of Compagnie Générale de Constructions Téléphoniques; Mr. P. Lizon, General Manager of Le Matériel Téléphonique; and Mr. M. Pérès, President and Managing Director of Les Téléimprimeurs, at the keyboard.

Greetings were sent by Mr. H. S. Geneen, President of International Telephone and Telegraph Corporation, through the facilities of the American Cable & Radio Corporation in New York, and were received simultaneously in London by Standard Telephones and Cables and in Croydon by Creed & Company. On the following day, similar transmissions were made to Laboratoire Central de Télécommunications, Le Matériel Téléphonique, and Compagnie Générale de Constructions Téléphoniques in Paris. Suitable replies were returned and members of the international press also exchanged greetings via this new and promising radio relay.

In the United States, the Andover, Maine, Telstar terminal was used. The British General Post Office terminal was employed in England as was that of the Centre National d'Etudes des Télécommunications in France.

Northern Ireland Factory—Lord Brookeborough, prime minister of Northern Ireland, proclaimed "the beginnings of an important new industry" in Ulster when he welcomed the arrival of Standard Telephones and Cables (Northern Ireland) Limited to Monkstown with the opening of its first plant, shown in Figure 5. An additional main factory is under construction nearby and both will produce telephone exchange equipment.

Standard Telephones and Cables
United Kingdom

Pompes Salmson Joins Le Matériel Téléphonique—The Société des Pompes Salmson has recently merged with and will assume the name of Le Matériel Téléphonique.

Founded in 1917 as a family business, Société des Pompes Salmson has become a major supplier of centrifugal pumps in France. It manufactures a wide range of pumps for both industrial and domestic heating systems. A great variety of electric-driven pumps also go to the mechanical and chemical industries.

Its engineering, development, and test laboratories are at Argenteuil as are its service and repair shops. Manufacture is at a modern plant in Laval, where Le Matériel Téléphonique al-

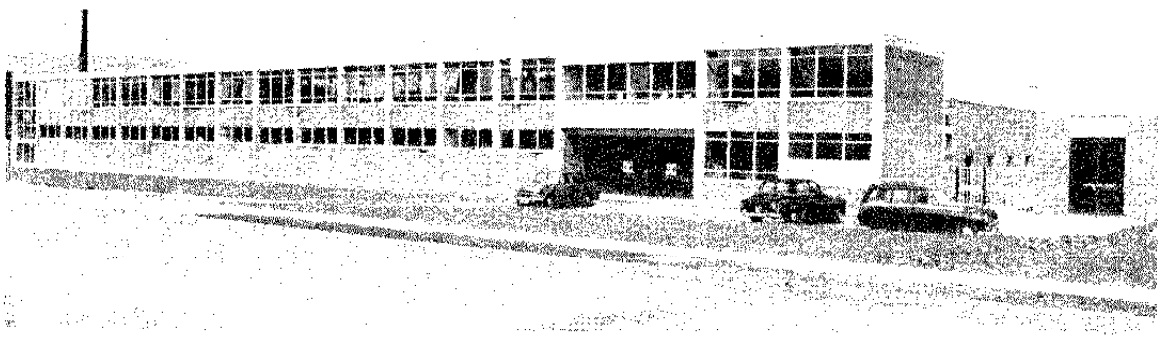


Figure 5—First factory of Standard Telephones and Cables (Northern Ireland) Limited in Monkstown, Newtownabbey, Northern Ireland.

ready has two factories. It employs over 600 persons.

Le Matériel Téléphonique
France

Australia Opens Its First Pentaconta Exchange—The first crossbar Pentaconta public telephone exchange in Australia was cut over in September 1962 at Melbourne. Present at the official acceptance of the Kew Exchange in July were: Messrs. Brian F. Jones, deputy director general of the Postmaster-General's Department; Glen N. Smith, director, and Arthur Stephens, assistant director of Posts and Telegraphs (Victoria); and T. H. Skelton, controller of stores and contracts of the Postmaster-General's Department.

With 3900 subscribers connected, 34 thousand calls were handled during the first 4 hours of service. Provision is made for 5200 lines, 5000 numbers, 238 incoming trunks, and 245 outgoing trunks. An important feature of the exchange is automatic alternative routing within the network to which it belongs. A second or a third alternative routing will be sought if the preferred route is busy. This may require a modified number of digits to be transmitted to a transit exchange.

Compagnie Générale de Constructions Téléphoniques
France

Standard Telephones and Cables Pty. Limited
Australia

Communication Between Austria and Italy—The map, Figure 6, shows the coaxial-cable and microwave radio network of Austria and its extension into Italy. The recently inaugurated coaxial system between Klagenfurt and Tarvisio will increase the 159 present channels between the two countries to an ultimate of 960 channels.

From Klagenfurt, two 4-gigacycle-per-second radio systems provide for multichannel telephone and a 625-line television signal to Graz and to Sonnwendstein, near Vienna. A 6-megacycle-per-second cable will connect the Vienna studios with the transmitter at Kahlenburg.

The coaxial cable for the Klagenfurt-Tarvisio system was manufactured in Austria. Part of the equipment at its Italian end and the coaxial system extending to Udine were provided by Fabbrica Appariccchiature per Comunicazione

Elettriche Standard of Italy. The remaining major part of the system was provided by Standard Telephones and Cables.

Standard Telephones and Cables
United Kingdom

Receiver-Indicator for Loran C—A new receiver-indicator for low-frequency long-range loran is shown in Figure 7. It is self tracking, once a signal has been acquired, and will provide continuous precise navigation data out to 1500 miles (2778 kilometers) from the transmitter. Previous systems were forced to sacrifice accuracy to obtain such range.

The equipment requires only 150 watts, permitting emergency operation from batteries, occupies 1.75 cubic feet (0.05 cubic meter) of space, and weighs 75 pounds (34 kilograms).

ITT Federal Laboratories
United States of America

Pneumatic Tube System for Toll Tickets—Despite the extensive automatization of the international toll system, with its attendant increase in traffic, many manual toll boards are still in service. A fully automatic toll-ticket pneumatic tube system has been installed to improve manual operations at the Zurich toll exchange.

One of two distributing centers is shown in Figure 8. The incoming tubes, which concentrate into a single tube, are above the outgoing

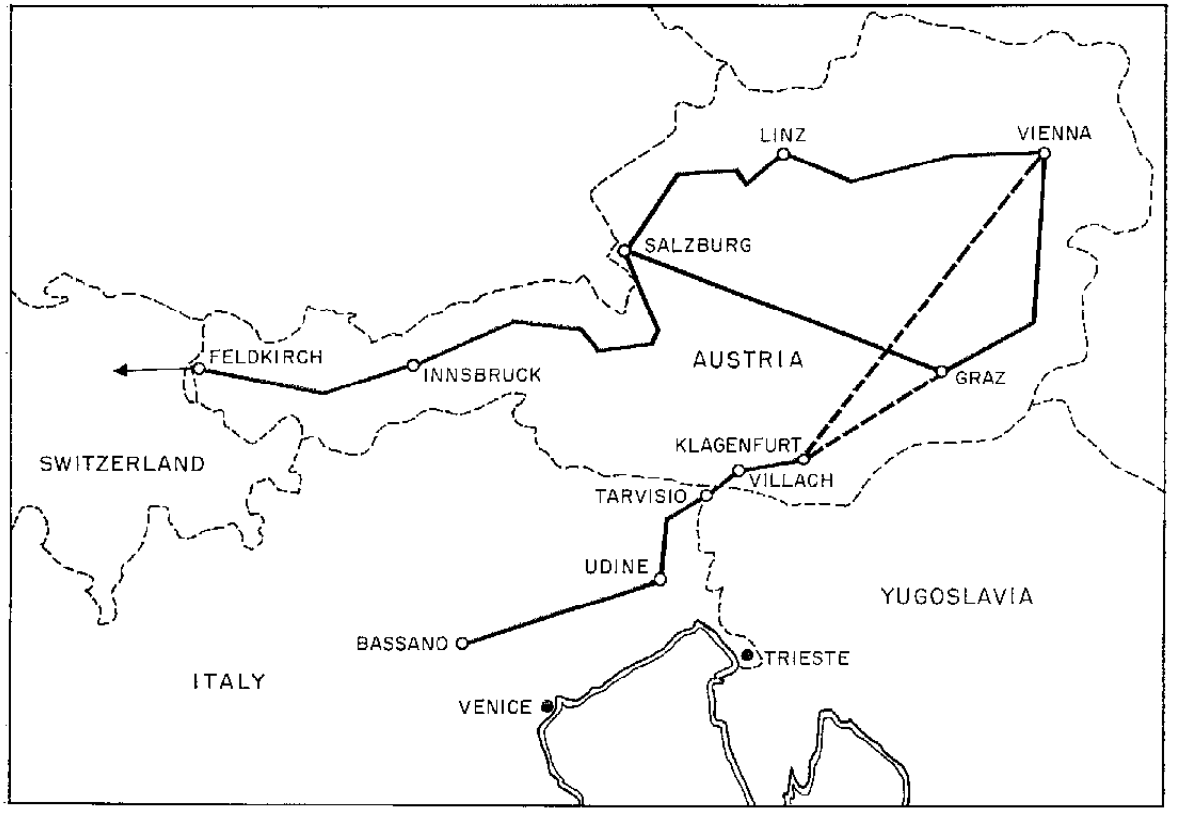


Figure 6—Austrian trunk system and its extension into Italy. Solid lines are coaxial cables and broken lines are microwave radio systems. Bassano and Udine are connected by small-diameter coaxial cable.

tubes with their remotely controlled routing valves. Supervisory photoelectric devices are not visible in the photograph. Capable of serving 208 stations, the tubes not yet needed have not been installed.

Control is from a central register utilizing conventional telephone components including a Pentaconta selector. The tickets are routed by transmission of digits from the key set at the toll position.

Standard Telephone et Radio
Switzerland

Gyroscope Within a Gyroscope—The precision to which a gyroscope will maintain its orientation in space is importantly related to the amount and variation of friction between the rotating mass and its support. Figure 9 shows a gyroscope consisting of an optically smooth fused-quartz sphere mounted within a metal sphere that rotates on a shaft.

Initially, both spheres are rotated as a unit by an air-driven turbine. When operating speed is achieved, several high-pressure air jets float the quartz sphere within its enclosure, providing a nearly frictionless bearing and a highly flexible coupling between the two spheres that rotate at the same speed.

Deviations of the axis of rotation of the outer sphere with respect to that of the inner sphere

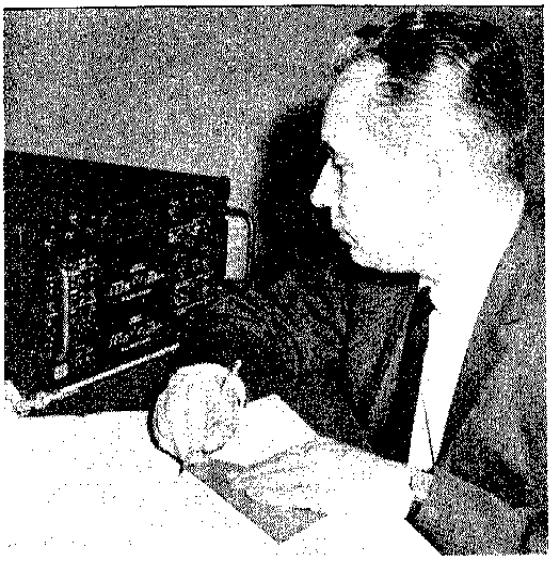


Figure 7—Self-tracking low-frequency loran receiver providing continuous precise position data out to 1500 miles (2778 kilometers).

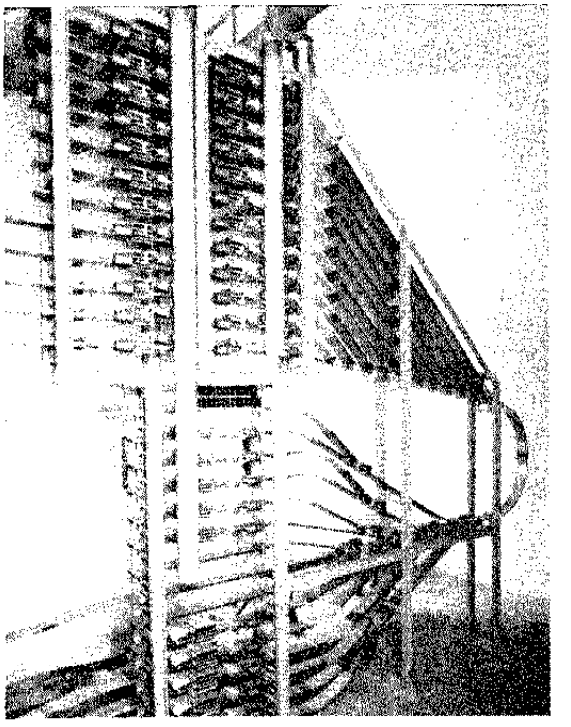


Figure 8—Distributing center for toll-ticket pneumatic tube system in the Zurich toll exchange.

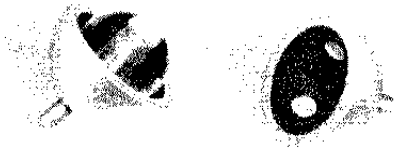


Figure 9—The rotating parts of the gyroscope within a gyroscope consist of a fused-quartz sphere that floats within an outer turbine-driven metal sphere.

are sensed by a beam of light reflected from a mirror on the inner sphere and cause the platform supporting the outer sphere to be adjusted to return the system to its proper orientation.

ITT Federal Laboratories
United States

Data-Processor Control of Switching—A stored-program data processor has been combined with a switching center to supervise its operation. Messages may be switched automatically to alternative routes in event of blockage or loss of the preferred path, transmission accuracy is continuously checked and retransmission ordered to correct errors, and message accounting is performed to insure receipt of all data at the proper destination.

The programming may be changed as needed, providing great flexibility of operation. The equipment shown in Figure 10 has been in continuous operation as part of a major communication complex since 1960.

ITT Federal Laboratories
United States of America

Destination Area Recording in Long-Distance Dialling—Information on traffic flow over telephone toll-exchange trunk groups is most useful for network design if it is in terms of the areas to which the calls are directed. The recorder, shown in Figure 11, to do this is connected to an originating register in the toll exchange to punch on tape the destination area code for each call.

The tape is scanned by automatic equipment that may extract any desired information such as the proportion of total calls to each area, their value in traffic units, or the sum of events recorded on several tapes.

Standard Elektrik Lorenz
Germany

Deep-Water Acoustic Monitor—Capable of probing to an ocean depth of 7000 feet (2129 meters), a new hydrophone, low-noise solid-state amplifier, high-speed constant-tension winch, and specially developed communication cable make up a new light-weight acoustic monitoring system. The new cable consists of two silver-plated copperweld conductors covered with an extremely tough insulating material that is completely free of pin holes that usually degrade the performance of cables operating at deep-sea pressures. Less than 0.1 inch (2.5 millimeters) thick and weighing about 25 pounds per mile (6.1 kilograms per kilometer), the cable is capable of carrying wide-band television signals.

ITT Federal Laboratories
Surprenant Manufacturing Company
United States of America

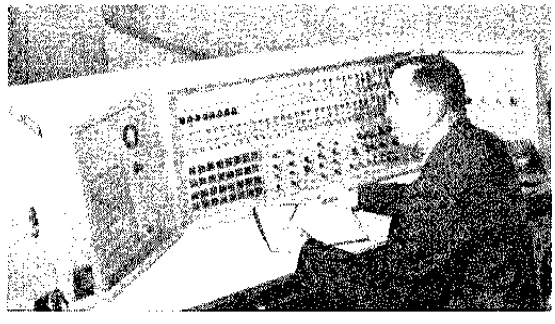


Figure 10—A combined switching center, computer, and data processor that can handle 75 000 words per minute.

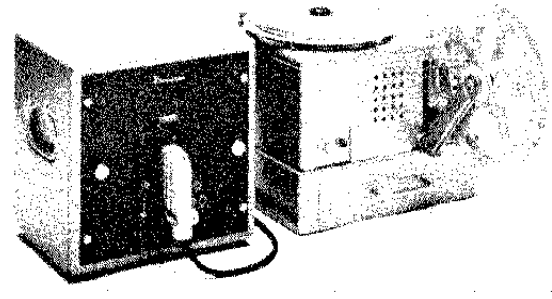


Figure 11—Equipment for recording, for toll traffic calculations, the area codes to which long-distance calls are directed.

Single-Sideboard Radio Equipment—Transmitters and receivers for single-sideband radiocommunication have been developed for use in uninhabited tropical regions. Simplicity, low cost, and ease of maintenance and operation have been primary features of the design. Vacuum tubes have been used throughout.

The fixed equipment includes a receiver, a 300-watt transmitter, and a 75-watt transmitter-receiver. The latter, shown in Figure 12, is available in both simplex and duplex versions.

The mobile transmitter-receiver units are of 30 and 75 watts capacity. A separate control unit mounts on the vehicle dashboard. The antenna unit permits efficient loading over the band from 3 to 15 megacycles per second of antennas designed for passenger cars and for trucks.

Accessory units permit remote control, use of teleprinter circuits, and interconnection with telephone exchanges.

Standard Electrica
Portugal

Marine Radio Receiver—The triple-conversion radio receiver shown in Figure 13 covers from 70 kilocycles per second to 30 megacycles per second. It has the exceptional stability, sensitivity, and selectivity required for operation aboard ship.

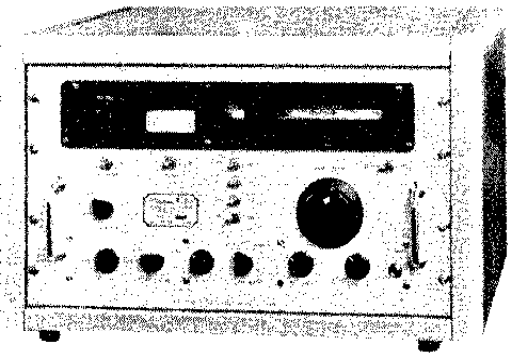


Figure 12—Single-sideband 75-watt transmitter-receiver for fixed station.

Image rejection is better than 80 decibels over the entire tuning range. The bandwidths at 6 and 60 decibels down are 6 and 20 kilocycles per second, respectively, for the broad conditions and 3.1 and 6.5 for the medium selectivity condition. Narrow-band selectivity is provided by a mechanical filter. The inclusion of a 100-kilocycle-per-second crystal calibrator assures dial settings to within 2 kilocycles per second of any specified frequency.

Mackay Radio and Telegraph Company
United States of America

Radar Indicator with Voltage Readout of Marker Positions—A 9-inch (23-centimeter) plan polar indicator for a short-range precision radar is shown in Figure 14. The console includes deflection and video amplifiers and the extra-high-voltage generator. The accompanying power supply also contains the main sweep generator and circuits for producing a marker

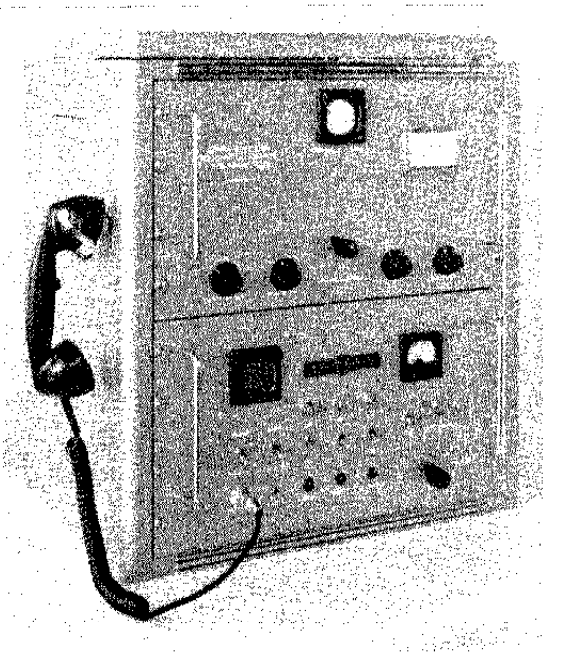


Figure 13—Marine radio receiver, which covers from 70 kilocycles per second to 30 megacycles per second.

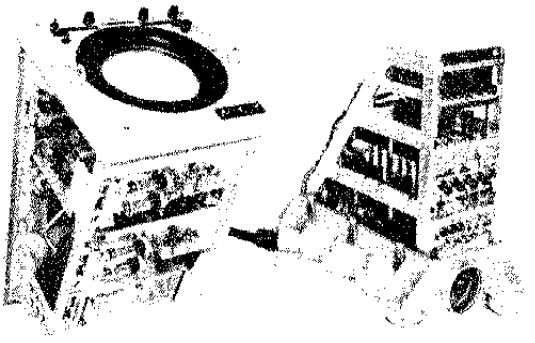


Figure 14—Plan polar radar indicator. The position of a marker is under the control of the operator. Coordinate voltages corresponding to the marker position are produced for use in other equipment.

and controlling its position on the screen of the cathode-ray tube. The position of the marker is converted to voltages corresponding to x and y coordinates originating at the center of the polar plan indicator. The coordinate voltages are available for use in other equipment.

The equipment will operate over a range from −40 to +65 degrees centigrade and will withstand mechanical shocks up to 50 gravitational units. Each weighs about 40 kilograms (88 pounds).

Standard Radio & Telefon
Sweden

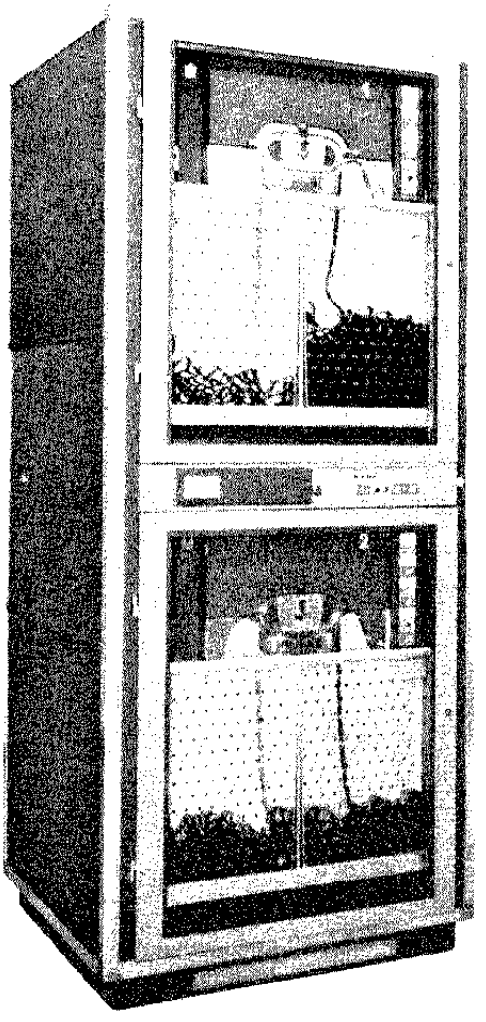


Figure 15—Magnetic tape unit with 5 transports. Two are visible at the front, 2 are at the rear side, and the remaining transport is within the framework.

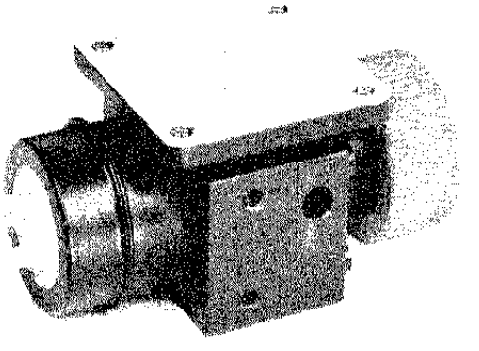


Figure 16—Reflex klystron for the 11-gigahertz region producing 0.5 watt with 600 volts on the accelerator electrode.

Compact Precision Timer—For demolition work, arming programs, and similar functions requiring extreme accuracy, a compact timer has been developed. A quartz crystal, transistor driven, provides the time base. The timing interval may be preset to several hours in increments of minutes and seconds. A 100-millisecond output pulse of 2 amperes is produced. The timing accuracy is within 0.1 percent for a supply voltage that varies between 4 and 8; it can be improved by providing a constant supply voltage. A typical unit occupies 15 cubic inches (246 cubic centimeters).

ITT Federal Laboratories
United States of America

Precision High-Resolution Oscilloscope—A 14-inch rectangular-screen cathode-ray tube provides resolution of 25 lines per centimeter. Deflection linearity is 1 percent of full-scale value along its major axes. The maximum plotting error anywhere on the screen is 0.6 centimeter. Drift is less than 0.5 centimeter per hour after warmup.

This high-resolution tube employs electrostatic focusing and magnetic deflection. It is operated in a circuit that will accept either single-ended or balanced inputs. For the latter, the sensitivity is 10 millivolts per centimeter. Response is flat to within 3 decibels from direct current to 50 kilocycles per second for full deflections, decreasing linearly at higher frequencies.

ITT Industrial Products Division
United States of America

High-speed Magnetic Tape Units—Displayed at exhibitions in Brussels, Munich, and Paris, were magnetic tape units of a new type shown in Figure 15. The tape is driven by pneumatic capstans and stored loosely in a bin of modular design to meet various dimensional requirements.

The tape speed may be between 1 and 5 meters per second (40 and 200 inches per second), its width either 12.55 or 25 millimeters (0.5 or 1 inch) and the starting and stopping times are 3 milliseconds. The recording density is 20 bits per millimeter (500 bits per inch) and 100 000 alphanumeric characters can be transferred per second on 8 adjacent tracks at this density and the highest speed. Also at this density, the capacity is 6 million 7-bit characters. Only 1 nonpermanent reading error will occur in 100 million bits.

Bell Telephone Manufacturing Company
Belgium

Reflex Klystron—A reflex klystron, VLR 222, is built of metal and ceramic for operation between 10.7 and 11.7 gigahertz. It is suitable for transmission by radio of 300 or 600 telephone channels or a television signal.

The reentrant resonant cavity that determines the operating frequency is tuned mechanically at a rate of 340 megahertz per turn of a screw acting through a diaphragm.

A modified Heil type electron gun is used. Perveance is 4.4 microamperes per volt$^{3/2}$ at an accelerator voltage of 600. The cathode current is 65 milliamperes.

Operated in the 2¾ mode with 600 volts on the resonator, an output of at least 0.5 watt is delivered to a matched load. Dependent on the mechanically set resonator frequency, the electronic tuning range varies between 60 and 110 megahertz with a modulation sensitivity between 0.5 and 1.4 megahertz per volt. The modulation characteristic is linear within 2 percent for frequency deviations of ±5 megahertz.

As a local oscillator in the 3¾ mode at 400 volts a power of 80 milliwatts is delivered to a matched load. As evident in Figure 16, output is delivered through a circular ceramic window in a *UG-39U* flange for *R100* waveguide.

Standard Elektrik Lorenz
Germany

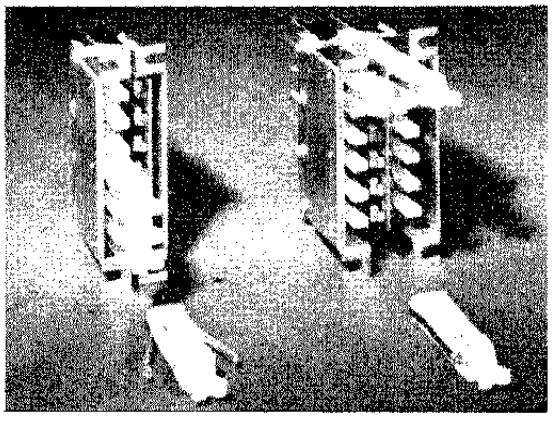
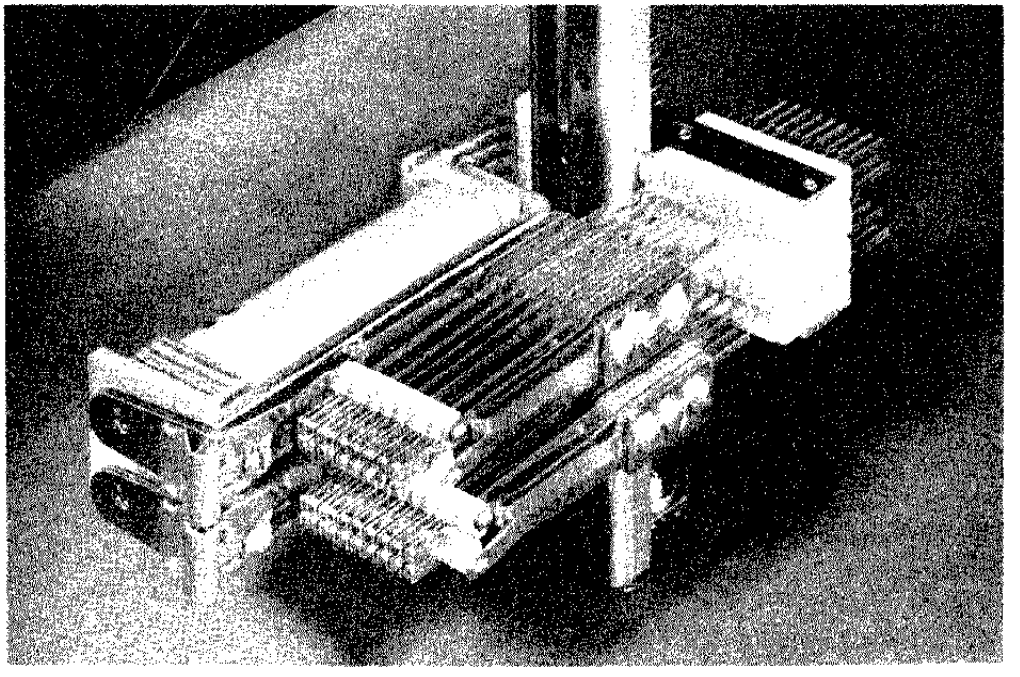
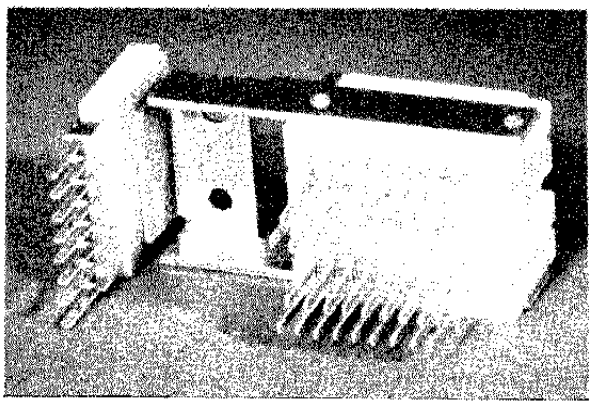
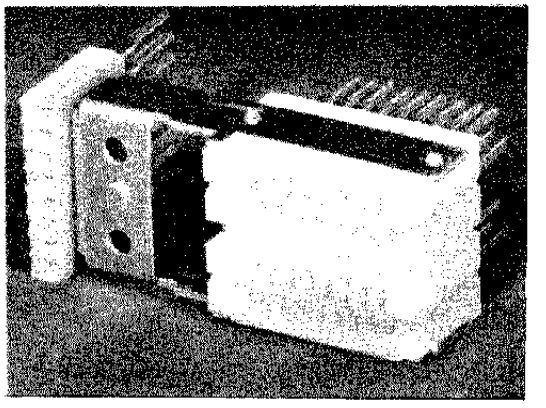


Figure 17—Cases for 6 and 12 telephone fuses. The fuse wire is soldered between the ends of the flat springs, thus holding them close to the molded frame, as in the fuse at the right.

Figure 18—Contact base into which the standard Pentaconta relay may be plugged.







Protection for Silicon Rectifiers—Silicon rectifiers, including controlled types, have limited thermal capacity and can be destroyed by rapid overheating due to voltage transients. Selenium rectifiers will absorb brief overloads and special units called Safe-T-staCs have been developed to protect silicon units. Their use generally allows employment of silicon rectifiers with reduced ratings.

Standard Telephones and Cables
United Kingdom

Telephone Fuses and Panels—For use in telephone exchange power supply leads, the compact fuses and panels shown in Figure 17 have been developed. The fuse element is soldered between the ends of 2 nickel-silver springs that also provide electrical contact to the panel. They are mounted in 2 polyamide molded pieces.

The maximum current rating is 2 amperes and the tension exerted on the fuse wire is low enough to permit a minimum rating of 0.2 ampere.

The fuse receptacles provide signalling contacts insulated from the fuses. The dimensions of the panel mounting areas are $3\frac{5}{32}$ inches (80 millimeters) by $\frac{31}{32}$ inch (25 millimeters) for the 6-fuse unit and $1\frac{33}{64}$ inches (39 millimeters) for the 12-fuse model.

Compagnie Générale de Constructions Téléphoniques
France

Plug-in Connectors for Pentaconta Relays—A recently developed connector, which may be fastened to a support by 2 screws, permits the standard Pentaconta relay to be plugged into a circuit. As evident in Figure 18, a maximum of 18 connecting points are provided for each of 2 spring pile-ups and 8 more are available for the operating coils.

The connector springs are of nickel-silver with 2 contact points in parallel, each exerting a pressure of 300 grams. Solderless wrapped wiring connections may be used.

Compagnie Générale de Constructions Téléphoniques
France

Television Receiver for 405 and 625 Lines—Production has been started on a broadcast television receiver designed for operation on the British 405-line very-high-frequency transmissions and on the 625-line ultra-high-frequency bands that will also be used in the United Kingdom. Only one additional valve is required over the conventional 405-line design.

Kolster Brandes
United Kingdom

Deflection Yoke for Line-Free Television—The television picture-tube deflection assembly shown in Figure 19 has a winding, in addition to the horizontal and vertical windings, that when activated by a crystal-controlled 13.56-megacycle-per-second current wobbles the spot to produce a line-free picture. To correct for both cushion and keystone distortions, four adjustable magnetic shoes are arranged at 90-degree intervals around the yoke. A negative-temperature-coefficient resistor compensates for changes in the resistance of the vertical deflection coils.

Standard Elektric Lorenz
Germany

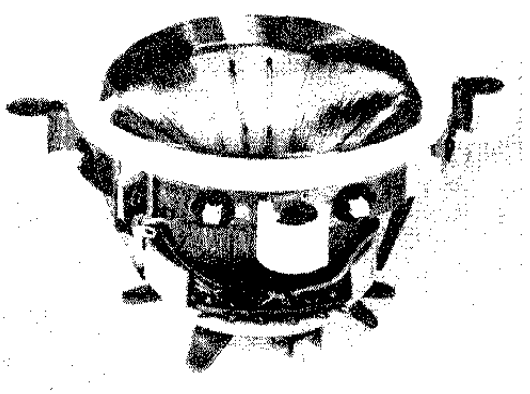


Figure 19—Deflection system incorporating spot wobble to produce a line-free television picture.

Local Interexchange Telephone Carrier—The use of carrier systems to increase the capacity of existing multipair cables linking telephone central offices is primarily an economic matter: whether it is cheaper to lay new cable or provide carrier equipment to work with existing cable. In a test arranged with the General Post Office in England, carrier equipment has been

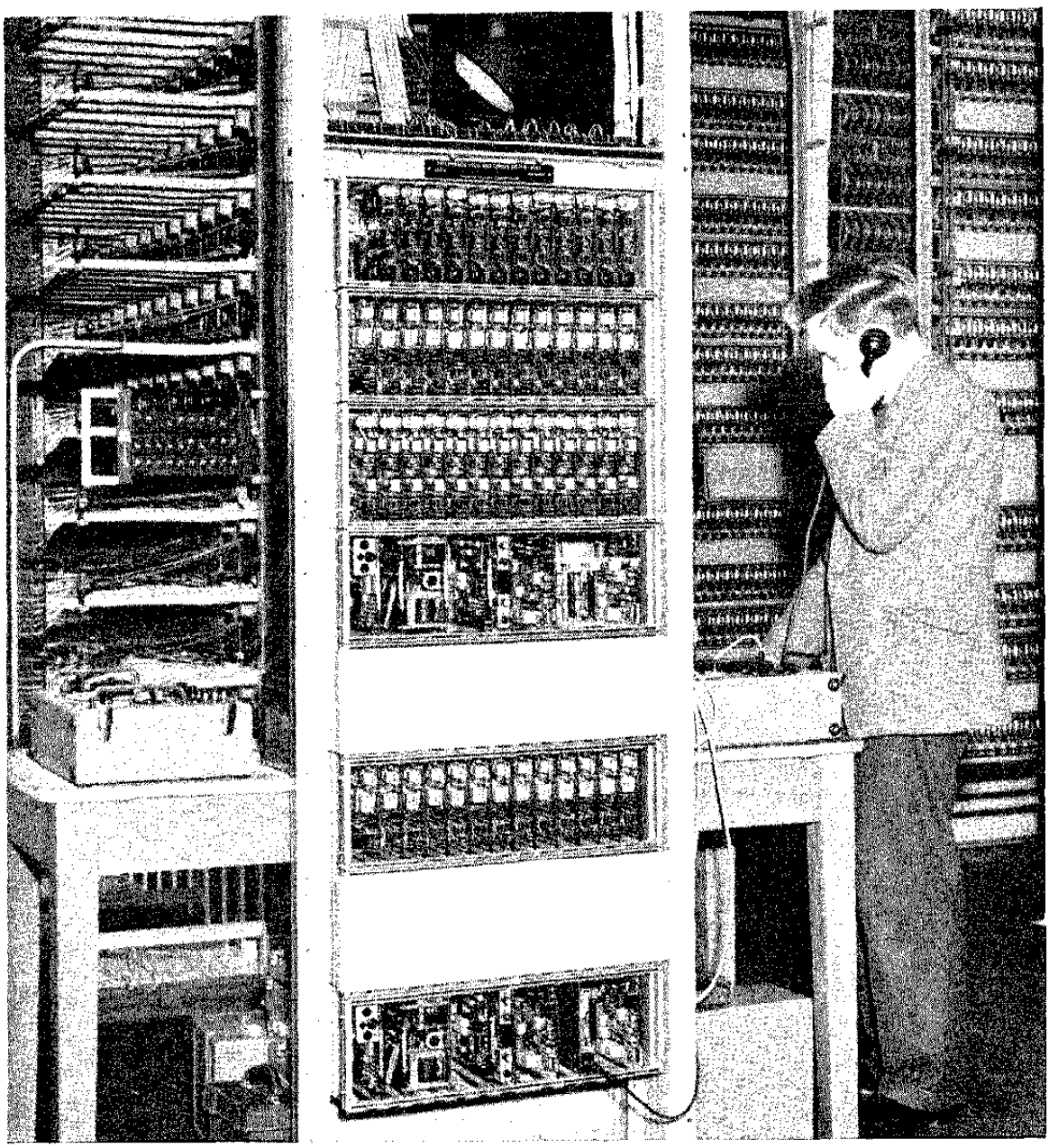


Figure 20—Carrier and double sideband is being tested for interexchange telephone carrier over existing multipair cable.

installed at London and Kent and at 9 points along this cable route.

Using new developments in transistors and ferrites and avoiding expensive filters by employing carrier and both sidebands, two pairs in existing cables carry 12 both-way voice channels and a 13th channel that is utilized for dialling and other signalling. Some of the terminal equipment in the Crayford exchange in Kent is shown in Figure 20.

Standard Telephones and Cables
United Kingdom

Barrier-Grid Storage Cathode-Ray Tube—In Figure 21 is a barrier-grid storage cathode-ray tube usable as a moving-target-radar indicator, as a random-access binary store, and as an image-storage device. The 5-inch (127-millimeter) tube contains a precision electron gun and a sandwiched target made up of a metal backplate, a dielectric layer, and a barrier grid having square apertures. The scanning electron beam passes through the grid to the dielectric layer. Secondary electrons are emitted from the dielectric and pass through the barrier grid to a collector, producing an electric signal.

Depending on the charges left on the elemental dielectric areas and the potential on the barrier grid, a state of equilibrium can be reached. No secondary electrons then pass through the grid to the collector and there is no signal output.

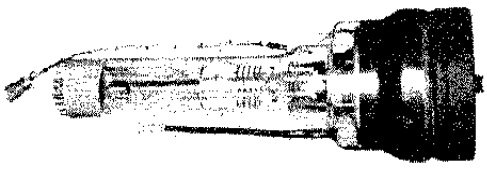


Figure 21—A barrier-grid storage cathode-ray tube usable as a moving-target-radar indicator, a random-access binary store, and as an image-storage device.

As a radar moving-target indicator, the radar video signal is applied to the backplate in synchronism with the scanning beam. After a few scanning cycles, the repetitive input signal from a stationary or clutter target causes the tube to reach equilibrium and output is produced only for moving targets. A cancellation rate of 26 decibels is achieved for clutter.

As a random-access binary store, the tube can store about 58 000 bits in a 240-by-240-bit raster. The resolution is 1200 lines per diameter and 15 writing levels are distinguishable in the output.

ITT Industrial Laboratories
United States of America

Reciprocal-Type Ferrite Phase Shifter—A microwave phase shifter of the reciprocal type, useful for phase modulating microwave signals and for simulating Doppler signals, consists of a ferrite body in a short rectangular plastic X-band waveguide. An exciting current at 10 kilocycles per second produces a phase shift of 1.3 degrees per milliampere. The insertion loss is less than 1 decibel and the voltage standing-wave ratio is less than 1.2. A power supply and meter calibrated in degrees of shift from 0 to 450 form a convenient operating aid.

ITT Federal Laboratories
United States of America

Surbond and Surock Wire Insulation—Surbond is a Teflon insulated wire with a coating that assures adequate adhesion in modern potting compounds, such as polysulfide, polyurethane, epoxy, and silicone types as well as improved surface resistance and aging over uncoated Teflon.

Surock is a specially coated thin-wall Teflon insulated wire that is two to three times as good as the uncoated product for abrasion resistance, cut-through resistance, and hot-soldering-iron endurance. Electrical properties and resistance

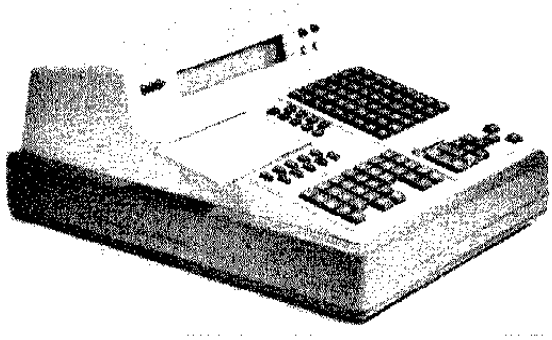


Figure 22—Load-control set used in the weight and balance system installed at Copenhagen-Kastrup airport for Scandinavian Airlines.

to solvents are unaffected by the coating. Adhesion to potting compounds is increased.

Surprenant Manufacturing Company
United States of America

Aircraft Weight and Balance System—More than a year ago, there was installed at the Copenhagen-Kastrup airport for Scandinavian Airlines a system for reporting the weight of passengers, luggage, and freight to be put aboard each aircraft and its distribution as an aid in balancing the load.

A Stantec Zebra computer is associated through a central electronics system with 40 weight-input and 3 load-control sets. The weight-input sets are used in all bookings of passengers and luggage to record flight class, passenger category, reservation status, number of passengers, and destination airport. The central equipment acknowledges an acceptable booking operation with a green light. A lack of seats or weight overload for a particular aircraft will cause a booking to be refused, which is signalled by a red light and the pertinent information is recorded on a page printer.

The load-control sets, shown in Figure 22, are in the central office and permit the limits within which each particular aircraft may be loaded to be recorded in the store of the computer. A visual display indicates the bookings already completed and permits changes to be made to accommodate different types of loads. Shortly before departure, a final statement of loading is provided by the load-control set. It includes the addresses of both senders and receivers of all freight, which are transmitted by page printer to the destination airport.

The system handles about 15 thousand input items per day in serving 90 to 100 flights. Engineers of Standard Telephones and Cables, United Kingdom, collaborated in the design.

Standard Elektrik Lorenz
Germany

Clavier Receives Award

The Achievement Award, the highest issued by the Professional Group on Communication Systems of the Institute of Radio Engineers, was presented on 3 October 1962 to Andre G. Clavier "for his outstanding contributions to the science and technique of communication systems."

A member of the research staff of the International System from 1925 to his retirement in 1962, he published extensively on his continuous activities on microwave equipment design and propagation studies.

Application of Pulse-Code Modulation to an Integrated Telephone Network

Part 1—Advantages of Pulse-Code Modulation

P. MORNET
Laboratoire Central de Télécommunications; Paris, France

1. Introduction of Electronics into Telephone Switching

The essential advantages expected from the introduction of electronics into telephone switching are a reduction in the volume of equipment, greater reliability, and increased flexibility of operation.

The first two advantages reduce housing and installation costs and maintenance. The third can increase operating profits by offering additional services to subscribers.

Electronic switching methods differ in the way the principles of space or time division are applied to the speech circuit and to the control circuit.

The distinction between these two types of circuits—the former actually providing the speech signals between the subscribers, and the latter controlling the speech connections to be established or released—appeared more and more clearly as the evolution of electromechanical exchanges progressed and, of course, remains valid for exchanges introducing electronic elements of any type whatever.

Both space and time division of switching functions have developed progressively in the past.

In time division, the frequency of use of any component or element is a direct function of its operating speed. If its response time is very short compared to the time allotted to perform its particular operation, the whole exchange may need only one such element, which will handle all calls and will usually be installed in a central position.

Several experimental exchanges have been made having a control circuit operating on this principle and a speech circuit of the space-division type in which the crosspoint, which actually provides the continuity of the speech path, consists either of metallic contacts or electronic elements.

Among these experimental exchanges, two may be mentioned; one was installed and run by Bell Telephone Laboratories at Morris, Illinois, in the United States, and the other is a private 240-line electronic exchange built by Laboratoire Central de Télécommunications, which has been in operation in Paris since January 1960.

However, as none of the solutions for the problem of the electronic crosspoint has been considered completely satisfactory, attention has been focussed on semielectronic systems as the immediate successor to electromechanical systems.

2. Time-Division Speech Network

While such semielectronic systems are of great interest, there seems to be a controversy over whether they are only a stage toward techniques in which the speed of the electronic elements can be used to better advantage.

In fact, in a space-division speech network, the number of crosspoints is always very large, even using the most elaborate methods for their number. Moreover, the time during which a given crosspoint is used always is short compared to the time during which it is waiting.

As has long been known, it is not necessary to transmit speech as a continuous signal, but it is sufficient to transmit brief samples at a rate that is at least twice the highest frequency of the speech spectrum.

Several transmission systems based on this principle have been tested or operated, all using of course the fundamental advantage of interleaving the samples from several conversations to obtain multiplex operation.

With fast electronic elements, the principle of space division can also be applied to the speech circuit. Crosspoints may still be required but through multiplexing their operating rate is

much higher than in space division; as a result their number may be considerably reduced and thus affect the volume of the exchange equipment.

3. Time-Division Transmission Systems

Transmission systems based on time division differ essentially from one another in the method used for the transmission of the information contained in the sample of the speech wave. The following 3 types of pulse modulation have been developed.

A. Pulse-amplitude modulation in which the amplitude of the pulse itself varies with that of the sample.

B. Pulse-width modulation in which the amplitude remains constant but the width of the pulse is a function of the modulation.

C. Pulse-time modulation in which the amplitude and width of the pulse remain constant but the time of the pulse with respect to a reference time varies as a function of the modulation.

4. Pulse-Code Modulation

These various types of modulation are all more-or-less vulnerable to noise and crosstalk. Much more effective protection can be obtained with the use of pulse-code modulation disclosed in French patent 852 183 issued to Reeves in 1938.

According to this process, the amplitude of each sample is measured and its value is transmitted in the form of a series of pulses. Only the presence or the absence of a pulse is significant.

Thus, only the disappearance of a pulse from the signal or the appearance in it of an additional pulse can cause a deterioration of the information transmitted.

In fact, if the maximum amplitude of an extraneous pulse is less than half of that of the coded signal pulses, the original signal can be reconstituted without error at the receiver. A mere alteration of the characteristics of one or several pulses caused by crosstalk, distortion, or noise has no effect provided the above limit is not exceeded.

Generally speaking, pulse-code modulation provides great reliability of transmission, and noise, crosstalk, and attenuation of the speech are independent of the transmission length and of the number of repeater points.

4.1 Increase in Switching Reliability

The increase in reliability of pulse-code modulation is however not limited to the transmission itself. It is a well-known and evident fact that, a priori, elements operating under ON and OFF conditions guarantee good operation superior to that from elements that are required to have precise characteristics independent of time and the conditions of use, such as temperature, power supply, et cetera.

Moreover, the ON-OR-OFF conditions are particularly convenient for the operation of transistors, which are thus especially suited to the field of switching. In both these conditions, the power dissipated is, in fact, negligible. In the ON condition, the difference of potential between emitter and collector is very low, of the order of a tenth of a volt for germanium and of the order of a volt for silicon, whereas in the OFF condition, the current flowing through the transistor is only a leakage current, with an intensity of a few microamperes.

The adoption of this type of modulation permits applying such operating conditions to all the circuits dealing with speech signals. Such is the case for transmission equipment and for the speech circuit in switching.

4.2 Problem of the Crosspoint

For the speech circuit in particular, the problem of the crosspoint is greatly simplified. In space switching, besides a low price, an impedance ratio of at least 10 000 between the closed and open (off) conditions is required, this ratio determining the crosstalk. In time-division switching, such as in pulse-amplitude modulation, one

imperative requirement of the crosspoint is that it forgets quickly the information that it has just transmitted so as not to disturb the following information or, in other words, not to introduce crosstalk. This means that all transistors used as gates must possess extremely short switching times.

Pulse-code modulation is, on the contrary, perfectly compatible with closed-to-open impedance ratios of the order of 10 and requires no particular speed of operation. These features together with the greatly reduced number of crosspoints for time division make this type of modulation ideal for switching.

For an objective assessment of the merits of pulse-code modulation, however, it is necessary to consider also the conversion of analog signals to digital signals and vice versa. The coder and decoder units that perform these operations are the only elements of the speech circuit that do not work completely in this ON-OR-OFF mode. Thus, their quality of communication depends only on these units provided the minimum conditions required of transmission and of the switching circuits are fulfilled.

5. Problem of The Telephone Network

The advantages of pulse-code modulation can best be used in an integrated network, that is, one in which both transmission and switching are done with the same type of modulation.

Experience has shown that pulse-code modulation, because of its resistance to interference, can be transmitted on ordinary telephone pairs whether or not they form part of ordinary telephone cable. The attenuation of the pulse is of course great, but it is sufficient to introduce a simple repeater at about every 2 kilometers, to restore the whole of the information without loss. In practice, these repeaters could occupy the place of the Pupin coils used at present, thus involving only small installation expenses.

In actual practice, a 24-channel pulse-code-modulation system requires two telephone pairs, one for each transmission direction. Nevertheless, 24 communications are handled by 2 ordinary pairs which, in conventional systems, carry only 2 communications.

Moreover, the principle of the 4-wire connections imposed by pulse-code modulation might perhaps be advantageously extended to the subscribers' lines. The subscriber's set *and line circuit*, elements whose price has a large influence on the total cost of a system, can indeed be simplified under certain conditions.

The possibility of multiplying transmission capacity by a large factor can be exploited in different ways, but it appears, especially if one considers the local telephone network where connections are short, that the switching circuits also should be adapted to pulse-code modulation to avoid costly equipment for changing the type of modulation. It is also preferable to extend the multiplex transmission as close as possible to the subscribers and this leads to considering the generalized use of a structure of the "expanded" type for the local network. Figure 1 gives an example of this.

The subscribers in such a network are connected to concentrators. Each concentrator is served by a pulse-code-modulation highway of, for example, 24 channels. The local exchange itself then essentially comprises only the switching circuits between pulse-code-modulation multiplex highways.

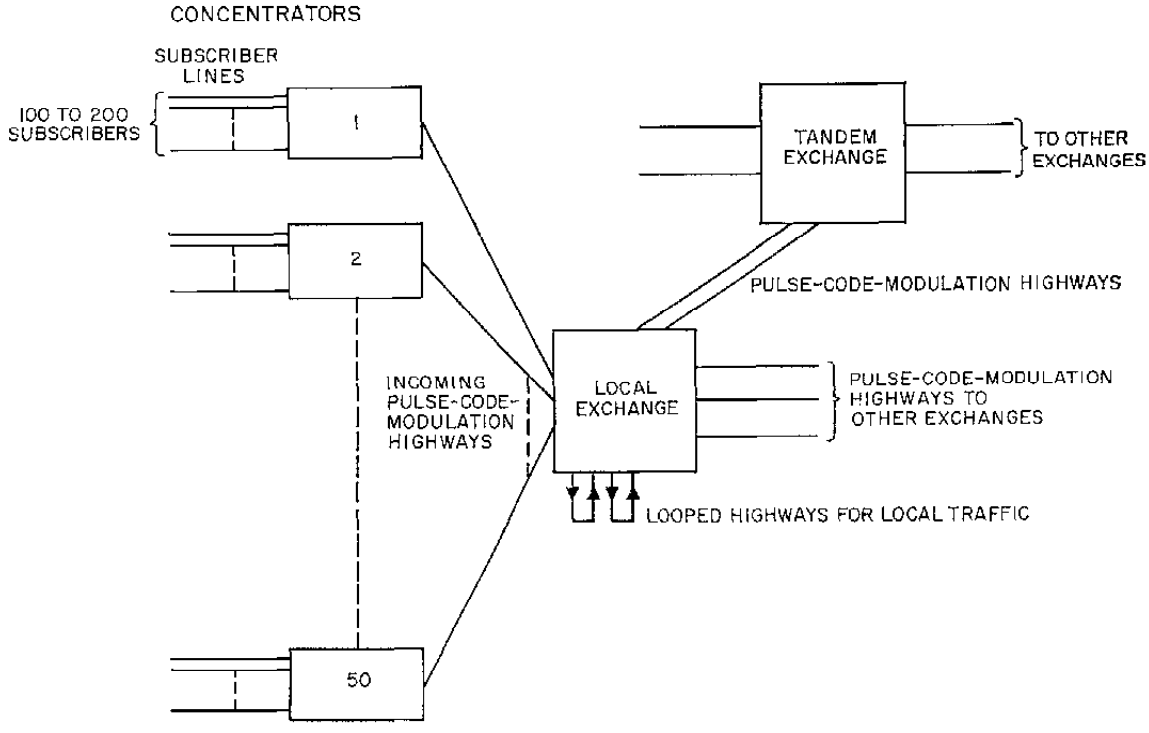
These highways are of 3 types: those providing connection to concentrators, those allowing interconnection with other exchanges, and a few to handle local traffic.

Lastly, the transit exchanges ensure interconnections between local exchanges and only need perform switching operations between pulse-code-modulation multiplex highways.

5.1 CONCENTRATOR

A concentrator, shown in Figure 2, consists essentially of the switching stage between the subscribers' lines and the pulse-code-modulation highway allotted to this concentrator. It can be

Figure 1—Local telephone network using pulse-code modulation.



placed in the same building as the local exchange but, to make the best possible use of the expanded type of exchange, it will preferably be put near the subscribers who are connected to it, and thus minimize the length of their lines. It includes the sampling circuits as well as the coding and decoding devices. Its very small size permits easy installation.

5.2 Local Exchange

Consider as a typical example a local network of 8000 subscribers having individually a traffic of 0.1 erlang at the busy hour, a local traffic of 20 percent of the total traffic, and a distribution of the lines over 50 concentrators with each one serving 160 subscribers.

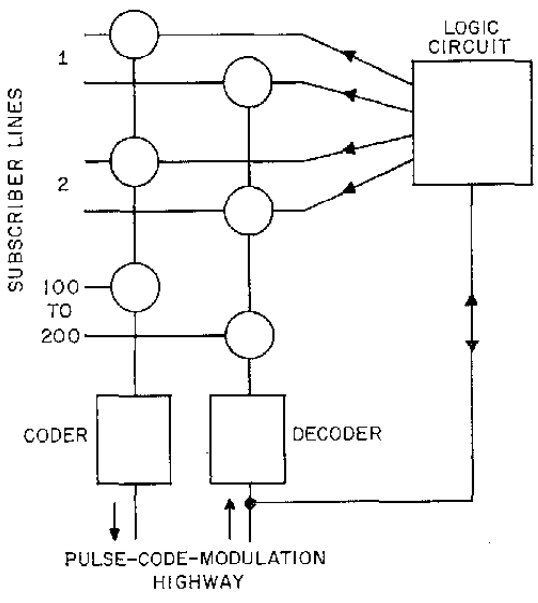
The local exchange must thus ensure both the interconnections between the 50 pulse-code-modulation multiplex highways connecting it to the concentrators and the highways connecting it to other local or transit exchanges.

In principle, a transit exchange will then be no different from a local exchange since, in the two cases, the switching operations to be carried out are made only on pulse-code-modulation multiplex highways.

There are therefore, in such a network, only two basic switching stages, the first assuring connections between subscribers' lines and a pulse-code-modulation junction, which are the concentrators, and the second, which is the basic element of the local and transit exchanges, assures interconnections between the various pulse-code-modulation multiplex highways.

Both types of switching stages have been constructed and tested experimentally.

Figure 2—Concentrator.



A preliminary study of the speech network of a complete exchange has furthermore shown that, in general, one switching stage could be sufficient. The adoption of a more-complex connection network with several stages does not seem justifiable. Even if it permits some savings in the number of crosspoints, this number is already much reduced by multiplex operation and the gain in absolute value can only be small. The network with one stage therefore seems to be the best solution, even for very large exchanges.

5.3 Network Operation in Time Division

The digital nature of the pulse-code-modulation signals makes it possible to record them in a binary-cell memory. This possibility is made use of as soon as the signals arrive at the exchange. Without this input memory, the establishment of a connection between two subscribers would present the problem of finding a time position simultaneously free on all the multiplex highways involved in the connection between the concentrator of the calling subscriber and the concentrator of the called subscriber. The choice of the time position would then necessitate the passing of information among all the interested exchanges. Moreover, an alignment of the time positions would have to be made on a variable number of multiplex highways according to the geographic positions of the two subscribers. The traffic possible on these highways would then be very small and this would lead to a poor use of the transmission and switching circuits.

5.3.1. *Alignment of Time Positions in Each Exchange*

The situation described above can fortunately be remedied by recording the signals on their arrival in the exchange in what is called a speech memory.

With this speech memory is associated another, the so-called time-switching memory, which contains the identity of the various lines of the speech memory. A cyclical reading of the time-switching memory allows the reading or writing of the speech-memory lines in any order.

This order is chosen by the control circuit of the exchange so as to satisfy the alignment of the time positions inside the exchange.

The unit formed by these two memories, of speech and time switching, makes it possible to carry out a switching in time of the channels of an incoming highway. They serve therefore to make the time at which a signal is handled in the exchange independent of its arrival time.

An example of this is given in Figure 3. A speech signal arrives on channel 12 of highway J_i and is written on line 12 of *MP1*, under control of the synchronization circuits. The identity of this line 12 is written on line 15 of *MCT1* under the control of the logic circuit. At time interval 15, line 15 of *MCT1* is read, line 12 of *MP1* is read and its contents transferred to line 17 of *MP2* connected to highway J_o. The identity of this line is supplied by *MCT2* and read at time 15. The transfer from *MP1* to *MP2* is made by the crosspoint connecting J_i to J_o,

which is conductive at the time of the reading of line 15 of *MCS* at time 15. The speech signal will be transmitted to J_o at the reading of line 17 of *MP2* at time 17.

5.3.2 *Rearrangement*

Despite the great flexibility provided by speech memories associated with switching memories, blocking can occur inside an exchange, which reduces efficiency in the order of 50 percent for 24-channel highways. A novel method of connection break-down and rerouting, referred to as the "rearrangement" method, overcomes this difficulty and eliminates blocking due to the alignment of the time positions by taking advantage of the flexibility provided by the speech memory.

The principle of the method is explained in part 3 of this series of articles. At this point it suffices to mention that this method greatly increases the highway efficiency at the cost of only a slight increase in the control equipment of the exchange, and that the rearrangement operations are performed with negligible or zero loss of information.

In the application of this method, it is feasible either to limit the transpositions to those that can be made within an exchange or to make transpositions also on the time positions in the transmission circuits between exchanges.

The complexity of the signalling problems involved in the latter solution leads us to think that at present the solution that limits the rearrangements to the inside of each exchange is preferable, despite its smaller savings in equipment.

The flexibility provided by the purely digital nature of all the signals makes it possible to realize exchanges of absolute operational reliability that have a speech circuit of very-reduced dimensions and that permit best utilization of the transmission circuits resulting from increased capacity and very-high efficiency.

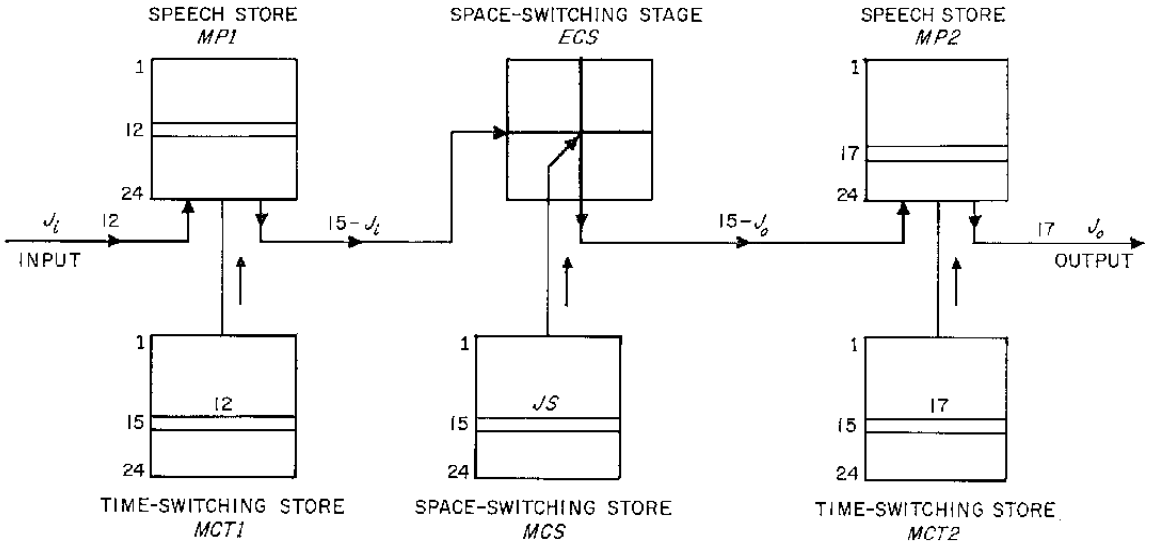


Figure 3—Switching stage between pulse-code-modulation highways.

5.4 Synchronization

A very-great problem is that signals arriving over the highways with any phase whatever must, under control of a local clock at the exchange, be identified, separated, and correctly routed.

This problem is easily resolved for the concentrator and would not be very difficult for a star-shaped network, but it becomes more difficult in a mesh network because priorities must be assigned to the different exchanges to properly control the whole network and because the transmission parameters are variable.

There is, however, a very sure solution through use of the digital memory that records the signals from each highway on their arrival in the exchange.

Writing in this memory is controlled by a synchronization circuit that introduces time delays that may be as large as the time between adjacent bits to compensate for slow shifts or rapid jitter in propagation time. It places all bits in positions corresponding to their original periodicity.

The synchronization circuit also identifies each channel by its code and routes its incoming bits to a designated channel line in a memory. Because the timing of the bits has already been corrected, this memory can operate normally and the reading can then take place at any time determined by the local clock of the exchange. This memory thus plays the role of a buffer and allows asynchronous operation by eliminating those long slow variations of propagation time that are too large to be handled by the vernier time-delay circuit.

Obviously, this process involves in certain cases a loss of information, which is manifested only by a noise that is absolutely negligible in telephony. A relative variation of frequency of 10^{-6} between the incoming information and the local time causes a loss of the order of 1 pulse out of 10^6.

6. Introduction of Pulse-Code Modulation into Existing Networks

The question arises of course as to how such a network could gradually replace the existing network or at least be introduced into it.

It is first necessary to mention that there will probably be an intermediate stage during which the telephone exchanges will have centralized electronic control circuits with a programming flexibility such that the introduction of new network operating techniques will be possible. This intermediate stage should permit the later introduction of the pulse-code-modulation integrated network.

But, independent of this consideration, a certain number of suppositions can be made. Pulse-code-modulation might first be introduced into connections between exchanges to avoid installing new cables. Some pairs in saturated cables could thus be used to constitute pulse-code-modulation highways. Satisfactory tests have already been made along this line.

Another possibility could be that in areas of rapid population expansion existing subscribers' pairs could be grouped in two's to constitute pulse-code-modulation highways serving concentrators.

The efficiency of the lines would thus be multiplied by a factor of the order of 100 since only 2 pairs would permit connecting up to 200 subscribers to the network.

Equipment for changing the type of modulation as well as the type of synchronization would then have to be provided at the points of connection with the existing local exchanges while waiting for the number of concentrators to become high enough to justify transformation of the exchange itself.

It is, of course, possible to imagine other processes, but only a methodical study of the extension of the network and of its economy will show how the problem may best be approached.

In the case of a network that is to be built completely, only the interconnections with other

networks would present problems. But these being few in number, the solutions, even costly ones, can then be considered without difficulty.

7. Problem of The Military Network

One particular case, however, the military network, merits special attention.

One of the most important considerations in this case is the setting up of the network. The network can cover a rather large area and have connections of several hundreds of kilometers established by varied means, some of which do not always present great uniformity of transmission qualities. These means can include, in fact, multiconductor cables, coaxial cables, radio links, and others that must in certain cases be connected end to end.

Questions of impedance and level matching then arise, which, with conventional types of modulation, can be resolved satisfactorily only by using a 4-wire connection from one end to the other of the transmission. It would even be preferable to handle the switching on a 4-wire basis to avoid many conversions between 2- and 4-wire circuits.

With pulse-code modulation, these problems remain but are simplified. As has been stated above, this modulation is effective with a useful signal-to-stray-signal ratio of 6 decibels and makes it possible to neglect the reflections that would be inadmissible with voice-frequency signals.

The switching problems are particularly easy to handle with pulse-code modulation and it is no longer necessary to consider conversions between 2- and 4-wire systems, which are always delicate.

Pulse-code modulation not only offers the possibility of an integrated network, but solves the complex problem with respect to conventional systems assuring a quality of speech practically independent of the length and quality of the connections.

8. Transmission of Digital Information

It is now certain that in the future telephone network, signals to transmit information other than speech will take on more and more importance. Already devices such as the Dataphone, put on the American market by Western Electric, permit such exchanges at low speed. Pulse-code modulation itself consists of information of this type transmitted at a very-high speed, 70 kilobauds for a telephone channel sampled at a rate of 10 000 times per second and for a 128-level quantization corresponding to a 7-digit binary code.

An integrated telephone network using pulse-code modulation over a 24-channel multiplex highway must provide for the transmission of digital data at a rate of 2 megabauds.

As was mentioned previously, the necessity of synchronizing the signals on arrival in the exchange and of making rearrangements from time to time may possibly involve loss of information. These losses are absolutely negligible for telephony but require, nevertheless, the application of a certain redundancy for the transmission of digital information.

The required minimum redundancy will be determined by a systematic study of the information losses in the transmission circuits as well as in the switching circuits.

Nevertheless, without appearing to be too optimistic, rates of the order of 35 kilobauds for a single channel or of 1 megabaud for the whole of a highway can be envisaged. These may appear larger than needed at the moment but they can be completely used for example in the case of long-distance centralized handling of numerical data in a computer. This possibility could be offered to subscribers.

9. Conclusion

Pulse-code modulation makes it possible to draw the maximum benefit from the introduction of electronic elements into telephone systems because it gives a homogeneous solution

to the problem of the telephone network taken as a whole and not under only one particular aspect. Application of this type of modulation to telephone switching requires new techniques that have many points in common with those of digital computers, but it also presents problems that are peculiar to telephony, such as signalling, operation, and in a more-general manner those of the organization of a network.

Our theoretical and experimental work has illustrated the possibility of a practical application of this modulation to the public telephone network in a not-too-distant future.

The following articles dealing in more detail with two aspects of this problem, transmission and switching, will give a more-precise idea of the present state of our developments.

Pierre Mornet was born in Châteauroux, France, on 21 March 1919. He graduated as a radio engineer from the Ecole Supérieure d'Electricité of Paris.

He was a mechanical engineer in the Air Force before joining Laboratoire Central de Télécommunications in 1943. In 1945 he was recalled by the Air Force Army, Transmission Service. Mr. Mornet returned to Laboratoire Central de Télécommunications in 1952 and since 1956 has been in charge of the switching and transmission laboratory.

Application of Pulse-Code Modulation to an Integrated Telephone Network

Part 2—Transmission and Encoding

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1. Introduction

Scarcely 15 years elapsed between the famous invention of Alexander Graham Bell in 1876 and the description of the first multiplex telephone system by Maurice Hutin and Maurice Leblanc in French patent 215 091, filed in 1891. Nevertheless, the first commercial equipment, Western Electric, type *A*, was not built and cut over before 1918. No telephone multiplex system was in operation in France before 1932, when the *D* system of the French Post Office was inaugurated.

These first multiplex telephone systems, using aerial lines or underground cables, as well as the first multiplex radio link tested in 1931 by A. G. Clavier and L. Gallant between Saint-Inglevert, France, and Lympne, England, were based on the frequency-division multiplex principle.

At the present time, the great majority of multiplex telephone systems operated in the world are of this type. There are tens of thousands of them and the whole length of their circuits amounts to tens of millions of kilometers.

In a frequency-division multiplex system, a given frequency range corresponds to each telephone channel. The electric filters necessary for separating each voice-frequency band from 300 to 3400 hertz must differ according to the exact position in the transmission band to which that speech channel was transposed. Moreover, these filters are delicate and expensive elements.

These reasons are sufficient to explain why, very soon after the first frequency-multiplex systems were put into operation, many researchers tried to develop multiplex systems that were completely different.

It is odd to note that time-division multiplexing was not only discovered but used before the invention of the telephone: in fact, the Baudot telegraph multiplex system was put in operation in 1874.

The idea of replacing the low-frequency electric wave from the microphone by pulse signals dates from 1924; it is due to R. A. Heising. The considerations that led Heising to modulate the width of the pulses were related to improving the efficiency of a radio transmitter.

The idea of multiplexing telephone channels by interleaving pulses obtained by the cyclic sampling of the various speech signals corresponding to these channels, was conceived after 1930.

The first experimental transmissions utilizing this type of modulation were conducted in France, Great Britain, and the United States from 1937 to 1941, but they did not result in commercial production of equipment.

Other systems of pulse modulation perfectly adapted to time-division multiplexing have been proposed in French patent 846 849 issued to E. M. Deloraine and A. H. Reeves and have even led to factory production of devices, the first model of which was the multiplex type 10 of the British army, manufactured as far back as 1942.

However, in all these systems, the information transmitted was directly related to a characteristic of the electric signal that is continuously variable such as amplitude, width, or position of the pulse. The sources of distortion, crosstalk, and noise inherent in these various systems of transmission, could still considerably affect the received signal.

The credit of having invented a system of pulse-code modulation that makes it possible, theoretically, to get completely rid of the disturbances usually introduced during transmission, is attributed to A. H. Reeves, to whom French patent 852 183 was issued in 1938.

In pulse-code modulation, numbers are transmitted, represented by signals similar to telegraph signals but at a speed much higher than that usually utilized in telegraphy.

These numbers are the amplitude measurements of samples taken regularly from the speech signals. The information then lies only in the presence or absence of the code pulses and is to a certain extent independent of their amplitude, width, and phase.

The principle is simple; however, the necessity of multiplying the number of the transmitted pulses by at least 5 per sample to obtain an acceptable definition increases the bandwidth for transmission in corresponding proportion. Besides, the complexity of the encoding and decoding elements common to all channels of a multiplex pulse-code-modulation telephone system sets a minimum for the number of channels that may be advantageously multiplexed by this process.

In 1938, it was not easy to measure accurately the pulse amplitudes at the rate of several tens of thousands per second. The number of active elements used, instead of the numerous filters utilized in the frequency multiplex systems, was such that the use of vacuum tubes involved considerable difficulties.

As in all kinds of digital techniques, the introduction of semiconductors has simplified and considerably hastened the resolution of the problems raised.

2. Advantages of Pulse-Code Modulation in Transmission

It is of interest to note some of the most-important advantages inherent in the use of pulse-code modulation.

2.1. Transmission Quality Independent of Distance

For transmission, the most-spectacular advantage of pulse-code modulation is certainly the possibility of regenerating signals in an ideal manner and thus avoid, theoretically, any loss of information, so as to obtain a quality of transmission that is independent of distance. Experiments have shown that this is indeed the case, indicating that the decrease of transmission quality with distance is absolutely negligible, which is contrary to what happens in all other known systems.

2.2 Absolute Stability of Transmission-Line-Equivalent

It is also obvious that the transmission-line equivalent remains absolutely constant, however the quality of the transmission medium may vary. The variations of this transmission-line equivalent in the numerous systems using amplitude modulation, particularly in the great majority of the carrier systems on cables, are well enough known to make further explanation unnecessary.

2.3 Homogeneity

The use of a single telecommunication technique and of a coded form for the information being transmitted makes possible a more-homogeneous transmission system than was previously available.

2.3.1. *Homogeneity of Techniques*

In whatever is concerned: modulators or demodulators, control and supervision devices, regenerators, or channel-dropping equipment, to mention just transmission elements, only the digital technique is used.

In pulse-code multiplex transmission, it is possible to dispense with the use of channel filters, which are as varied as they are fragile and which are used in such great numbers in frequency-division multiplex telephone equipments.

Moreover, semiconductor logic elements are still in a state of rapid development as distinct from filters, which it seems now improve very slowly. The time is probably not far off when telecommunication engineers will tend to ignore

the flip-flop mode of operation and will be concerned with only the bistable logic element in the form of a hermetically sealed multipole.

2.3.2. Homogeneity of Transmitted Information

The homogeneity of the pulse-code-modulation information becomes clearly apparent in the increasing resemblance between transmission and switching. The fundamental advantage of this is that information can be switched and transmitted in the same form. This homogeneity appears also in various points of the transmission system.

Thus, the speech information that appears as a series of binary digits can take advantage of all the facilities offered by the digital-information-handling techniques: analysis, memorizing, ciphering, correlation, et cetera.

The choice of bandwidth to provide for a given type of information, such as, telephony, high-quality radio, television, high-speed data transmission, or radar display transmission, can be made automatically by digital means.

The dropping, inserting, and combining of channels are also carried out in a very flexible manner.

3. Transmission in an Integrated Telephone Network

The main purpose of multiplexing is to reduce the number of transmission paths, permitting saving on part of the price of cables to be laid, or avoiding the necessity of laying new cables if those already in service are overloaded.

In the system to be described, conversion from voice-frequency signals to pulse-code-modulated multiplex signals is performed through pulse-amplitude modulation. This solution seems to be the most economical.

It should be noted in Figure 1 that the signals are in the form of pulse-amplitude modulation only as an intermediate modulation inside the concentrator. This avoids any deterioration of the information. Besides, as subscribers' lines are generally very short, variations of the transmission-line equivalent between subscribers will be minimized. It may also make possible the use of a 4-wire subset. Otherwise, the conversion from 2-wire to 4-wire operation is made in the concentrator. The pulse-code modulation in the network is handled entirely on a 4-wire basis.

At the concentration stage, the subscribers' lines are cyclically scanned by pulses supplied

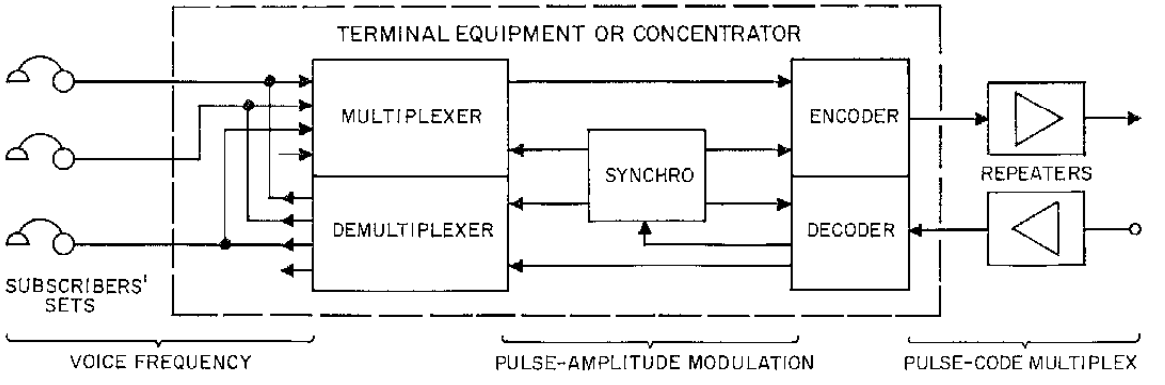


Figure 1—The voice-frequency signals from the subscribers sets are sampled in the multiplexer to produce pulse-amplitude modulation that is then converted to pulse-code modulation in the encoder.

from a timing device, controlled by signals from the exchange to which the concentrator is connected. This control uses the same path over which speech information is sent by the exchange.

The multiplex signal made up of interleaved amplitude-modulated pulses obtained by scanning various subscribers' lines goes to a coder. The coder converts from pulse-amplitude to pulse-code modulation and the multiplex signal is sent to the exchange via a junction equipped with regenerative repeaters. The spacing between repeaters is determined by the quality of the transmission path. It is these pulse-code-modulation junctions and channels that are switched at the exchange. Moreover, the exchange adds to the speech signals other information that enables the concentrator, firstly, to receive the information from the exchange with perfect synchronization, and, secondly, to send the speech signals that are continually coming from the exchange to only the subscribers' lines for which they are intended after they have been demodulated.

The timing device performs certain other operations which will be described in the following article.

4. Multiplexing and Sampling

In an integrated telephone network, the role of the multiplexing element is to concentrate the calls in process at any one time and to sample the speech signals of these active subscribers.

At the present time, multiplexing by interleaving amplitude-modulated samples of the signals to be transmitted, is certainly the simplest and the least-expensive method.

It is sufficient to sample a signal at a rate that is at least twice the highest fundamental frequency contained in the spectrum of the signal. Demodulation consists in sending the pulse-amplitude-modulated samples through a low-pass filter having a cut-off frequency equal to the maximum frequency in the original signal.

For the frequency band from 300 to 3400 hertz, it is theoretically sufficient to sample at 6800 times per second. As low-pass filters become proportionately more costly as they approach the ideal, the sampling frequencies usually chosen for time-division multiplex telephone systems are from 8000 to 10 000 hertz.

The choice of the sampling frequency is governed by considerations of the total cost of the system.

In ordinary multiplex transmission, the number of filters depends only on the number of channels being multiplexed. The tendency is to choose a sampling frequency approaching twice the maximum frequency in the signal to be transmitted, since the bandwidth of the multiplex transmission is proportional to the sampling frequency.

For an integrated telephone network, the number of filters is related to the number of subscribers connected to the concentrator. This number can be 10 times higher than the number of channels being multiplexed. In this latter case, the cost of the low-frequency filtering is an important element in the price of the whole equipment, and the tendency will be toward a high sampling frequency to reduce the cost of the filters.

It is important to note that the cost of the devices that are common to all subscribers in the concentrators or terminals, such as, the coder, decoder, and power supply, is practically independent of sampling frequencies between 8 and 10 kilohertz.

5. Coding and Decoding

The conversion from pulse-amplitude modulation to pulse-code modulation is done in an element called the coder. The convertor for the reverse operation is called the decoder.

The fundamental operation of coding is the quantization of the amplitude-modulated pulses derived from the speech wave. This quantizing operation is fundamentally the measurement of the amplitude of each sample taken from the

speech channels. For economy, this measurement must be made during the time between 2 consecutive samplings, which requires that one coder must be able to handle all the samples coming from all the channels that are being multiplexed. For example, if the sampling frequency is 10 kilohertz and 25 channels are to be multiplexed, the coder has 4 microseconds to carry out this operation. It is necessary that this operation be sufficiently precise to permit the faithful reconstitution at the receiving end from the pulse-code-modulated signal of, firstly, a pulse-amplitude-modulated signal and then the original speech signal.

Nevertheless, it must not be too precise, for this will increase the bandwidth and the equipment will be delicate or costly. In fact, the numbers provided by the measurement require more digits, binary or other, as the measurement becomes more precise.

There is no structural difference between a coder for a single channel and one for a multiplexed group of speech channels. The latter coder should simply operate p times faster than the former, p being the number of channels being multiplexed.

It may also be pointed out that the coding-decoding operation produces an inherent type of disturbance in the signal. This results simply from the uncertainty that is inherent in every measuring operation, no matter how precise it may be. The fact that this disturbance is called "quantization noise," "quantization distortion," or even "quantization error" shows clearly that it concerns a phenomenon new to telephony. Fortunately this type of disturbance is normally the only one that affects the system. Moreover, it can be reduced as much as desired, but at the price of increasing the complexity of the equipment. The essential problem is to compromise between an inexpensive transmission system and giving the highest possible quality of service.

The coding of the signals is called linear when all the quantization steps are equal. Let

U = maximum amplitude of the unclipped speech signals

u = amplitude of the weakest speech signal to be transmitted

q = quantization step.

To clarify this idea, assume that the maximum level of the signals U exceeds by 40 decibels that of the minimum transmitted signals u. Assume, also, that the signal-to-noise ratio of quantization must remain higher than 20 decibels. The quantization step q must be smaller than $u/10$ and the necessity of coding the signals of amplitude U requires that the total number of steps be higher than 2000.

For linear coding, the coding precision of weak signals is just sufficient, whereas that of the strong signals is largely redundant. The equipment used is exaggeratedly complicated and very bad use is made of transmission bandwidth.

The advantage of nonlinear coding is that the amplitude of quantization steps may be made to increase with the amplitude of the signal, thus maintaining a constant precision of coding for all amplitudes. This results in a compression of the signal.

5.1 Structure of Coders

The various speech coding devices, proposed and experimented with, can be divided into two categories, those in which the actual coding elements are completely distinct from the compressing elements and those in which these elements are inseparable.

5.1.1 *Compressor Plus Linear Coder*

In Figure 2, the compressor is of the instantaneous type. It looks like a quadripole, the transfer function of which is a decreasing function of the input signal amplitude. The output voltage must depend only on the input voltage at the same moment, otherwise crosstalk will be introduced between the samples that come from different channels and are applied successively to the compressor.

At the receiving end, an expander having a transfer function the inverse of the compressor must, of course, restore the relative amplitudes of the original signals.

The major difficulty of the compression-expansion operation is to be able to reproduce in practice and maintain for a long time the transfer functions of the compressors and the expanders, otherwise distortion will be introduced in these processes.

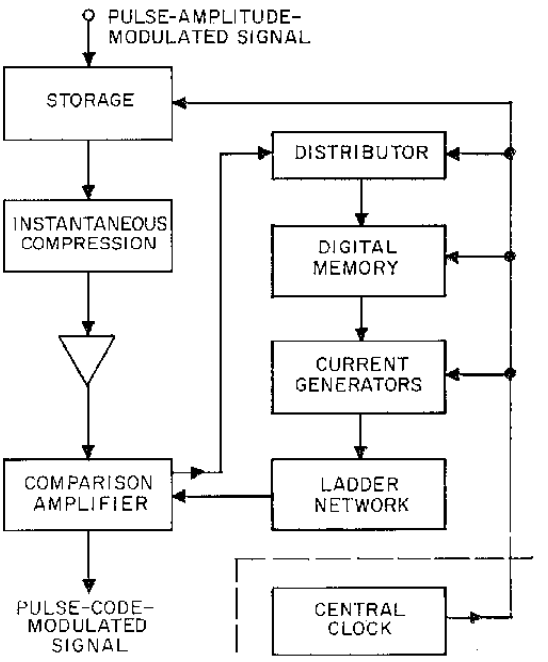
The important role played by compression is already known. The problem arises of determining what transfer function of the compressor will limit the number of digits, which are generally binary, that define the quantization value of a given sample and will also maintain a signal-to-noise ratio that is acceptable over the whole dynamic range. The complexity of the equipment will usually affect its cost and reliability and must, of course, not be neglected in the choice of the transfer function.

Numerous functions: logarithmic, hyperbolic, and exponential, either continuous or discontinuous, have been proposed for compression. The most attractive is the logarithmic function which may be readily obtained from the curvature of the characteristic $v = f(i)$ of a diode. Nevertheless, these devices, which theoretically present the advantage of great simplicity, are very handicapped by variations in the characteristics of individual diodes and by further variations with changes of temperature and finally by their aging.

Another type of compressor uses the abrupt threshold variations in the gain of an amplifier when the amplitude of the input signal exceeds certain predetermined values.

The linear coders associated with such compressors are of the parallel type when the measurement of the pulse-amplitude-modulated sample coming from the compressor is made in one single step. When this measurement is made by successive approximations the coder is called the series type. The number of successive tests in such a method is obviously related to the

Figure 2—Means of producing pulse-code modulation by using a signal compressor and linear encoding.



precision of the measurement. The time allotted to each measurement is fixed: it is the interval that separates 2 successive samplings.

The series-type coders thus have elements that work at higher speed than those of the parallel-type coders, but they are fewer in number.

In the linear coder of the series type shown in Figure 2, the signals from the compressor are quantized in the following manner. The signal to be measured is compared successively with the reference signals derived by the coder itself from the results of the preceding comparisons. The results of the successive comparisons appear at the output of the comparator in the form of electric signals, which can take only 1 of 2 values. These results are transmitted in pulse-code modulation and also stored in a digital memory after appropriate switching by a distributor that is controlled by a local clock. The p binary digit obtained at the end of the p first comparison of a measurement cycle initiates a change in the reference signal to make it

suitable for the $(p + 1)$th comparison. There are 2^p possible values.

The reference signals are obtained by addition in a ladder network of the currents supplied by the generators controlled by the cells of the digital memory.

The average difference between the signal to be measured and the reference signal is halved for each successive comparison. The accuracy of the measurement is thus defined by the number of successive comparisons, which are limited by the number of binary digits that may be transmitted in each code group.

All these operations are synchronized by the local clock, which also defines the cadence of the sampling and multiplexing operations.

5.1.2 *Coder-Compressor*

Only the coder-compressors that use a special cathode-ray tube in which the path of the electron beam is intercepted by a specially cut mask will be cited. Even though this solution is very flexible as far as the type of code is concerned, the disadvantages, due to the use of nonhomogeneous techniques and to the fragility of their electron optics, become all the more prohibitive as the required fineness of the coding of weak signals increases.

Coder-compressors may also use a voltage divider in which the cells are increased with the number of quantization steps. However, this design is quickly limited by the quantity of material to be used. It should probably be reserved for special applications where the norms imposed on the transmission are less severe.

The coder-compressor developed for our use is of a series type. Its structure resembles that of the linear coder described previously. However, its compression characteristic can be reduced to a group of segments of straight lines with different slopes. Such a characteristic can be obtained by purely digital means.

Before describing the coder-compressor, we shall see how it is possible to determine the optimum compression characteristic. In our particular case, this is closely related to the structure of the coder-compressor, which is why this subject has not been discussed previously.

To simplify this study, we shall start from the following hypotheses.

A. The code used is a pure binary code with $n + 1$ binary digits and 2^{n+1} steps.

B. Only a certain range of signals will be considered, those that have the same polarity with respect to the potential corresponding to the absence of speech.

C. The compression characteristic will be represented by using, as abscissa, the value x of the signals to be compressed and encoded, the unit chosen being the maximum value of the useful signal, and as ordinate, the ratio y of the measurement given by the coder, to the maximum value.

The signal-to-quantization-noise ratio remains appreciably constant for any signal level, if the ratio of the amplitude of the largest step used to measure a sample, to that of the sample itself, is constant.

For this, it is necessary that the increase of signal Δx corresponding to the increase of a step Δy be proportional to the signal x, that is

$$\frac{\Delta y}{\Delta x} = \frac{a}{x}.$$

The points that correspond to the passage from one quantization step to the following step must then be found on the curve of equation

$$y = 1 + \alpha \log x.$$

The essential differences between such a compression characteristic and the one proposed by H. Holzwarth and studied in detail by Bernard Smith are as follows.

A. The curve described by Holzwarth, which has the following equation in our system of reference,

$$y = \frac{\log (1 + \mu x)}{\log (1 + \mu)}$$

passes through the origin ($x = 0$, $y = 0$). This is not necessary if the dynamic range of the useful signals is taken into account. This range is obviously limited: in particular, the signals of amplitude lower than the first quantum cannot be transmitted.

B. The characteristic of $y = 1 + \alpha \log x$ presents the advantage over the preceding equation of corresponding to a signal-to-quantization-noise ratio that is constant over the entire useful dynamic range, whereas this ratio deteriorates about 6 decibels for weak signals by the second type of logarithmic compression.

If we now consider the ratio of the slopes of the chords that subtend 2 adjacent arcs having respectively, as ends, the points of ordinates

$$y_i + (k/2^n), \quad y_i, \quad \text{and} \quad y_i - (k/2^n)$$

that is, corresponding to successive quantization step groups that have the same number of steps, it is seen that

$$\rho = \frac{x_{i+k} - x_i}{x_i - x_{i-k}} = 10^{k/2n\alpha}.$$

This ratio is independent of i, that is, if we choose a compression characteristic formed of segments of a straight line subtending, as previously, the logarithmic characteristic of equation

$$y = 1 + \alpha \log x$$

the signal-to-quantization-noise ratio at the corresponding points of these segments does not vary over the entire useful dynamic range. It is nevertheless obvious that this ratio varies along each segment and increasingly so as the number of segments becomes smaller.

We shall now establish the relations that exist between the factor α, which determines the rate of the logarithmic compression, and the parameters of the pulse-code-modulation transmission, that is:

A. The number of digits used for the coding, $n + 1$.

B. The ratio of the amplitude of the signal to that of the largest step used to make the measurement p.

C. Finally, the useful dynamic range, which is defined as the ratio A of the amplitude of the unclipped maximum signal to that of the smallest signal judged useful for coding with the precision that corresponds to the ratio p.

The ratio p must remain constant over the entire useful dynamic range.

The smallest signal of this useful dynamic range, which must not be confused with the smallest useful signal, has in our system the amplitude

$$x = 1/A.$$

It is natural to consider that there is no compression for the signals that are weaker than those of the useful dynamic range. This is another way of saying that the coding of the very weak signals is linear.

Actually, this is the case with most of the existing systems. The ordinate that corresponds to the abscissa point $x = 1/A$ is then $y = p/2^n$.

Hence the first relation

$$p/2^n = 1 - \alpha \log A.$$

The first step of the useful dynamic range is defined by

$$\Delta x_1 = x_2 - x_1 = x_1 \quad (10^{1/\alpha 2^n} - 1)$$

$$\Delta y_1 = 1/2^n$$

the ratio $x_1/\Delta x_1$ being equal to p, we obtain the second relation

$$1/p = 10^{1/\alpha 2^n} - 1$$

or even

$$\frac{1}{\alpha} = 2^n \log \left(1 + \frac{1}{p}\right).$$

Finally, we can deduce the relation between the useful dynamic range A, the number of digits $n + 1$, and the ratio of signal-to-quantization noise p

$$2^n = p + \frac{\log A}{\log (1 + 1/p)}$$

which can also be written

$$2^n \approx p(1 + 2.3 \log A). \qquad (1)$$

It is seen for example that the use of natural binary codes with 7 digits ($n = 6$) makes it possible to obtain, if logarithmic compression is used, a signal-to-quantization-noise ratio that is constant and equal to about 22 decibels inside a useful dynamic range of 36 decibels. Under these conditions, the ratio of the strongest unclipped signal to the weakest coded signal is about 60 decibels.

We shall now examine the deterioration in signal-to-quantization-noise ratio when the logarithmic characteristic is replaced by a variable number of chords that subtend its arcs.

The deterioration observed over the range of a knee of this new characteristic is given by the value of the ratio of the slopes of the segments that form the knee. This ratio is, as we have seen,

$$\rho = 10^{k/2^{n\alpha}}$$

or even

$$\rho = 10^{k \log (1+(1/p))}.$$

The deterioration of the signal-to-quantization-noise ratio that occurs over a knee is then

$$20 \log \rho \approx 8.7 \ (k/p), \text{ decibels.} \qquad (2)$$

The order of magnitude of this deterioration is therefore $k/2$ decibels. This signifies that the use, for a given polarity, of a characteristic having 4 straight-line segments subtending a logarithmic characteristic leads to a slightly greater deterioration than that which corresponds to the loss of the smallest binary digit. On the other hand, the use of a characteristic having 8 segments leads to a lesser deterioration than that which corresponds to the loss of this digit.

The coder-compressor that was used in our first experiments gives a compression characteristic that is comprised of 5 segments in all.

The central segment has 64 quantization steps and corresponds to an improvement of 25 decibels of the signal-to-quantization-noise ratio for weak signals. The 4 other segments each have 16 quantization steps.

A schematic diagram of the device used is shown in Figure 3. It does not include an analog-type of instantaneous compressor ahead of the linear coder. The operation is completely digital, and is in all ways like that of an ordinary linear coder. The current generators are naturally more numerous since it must be possible to perform the coding of the last digits according to 3 different laws depending on the amplitude of the sample measured.

The greatest difference results from the introduction of a decoding matrix between the code store and the current generators. This matrix makes it possible at the time of each comparison and according to the results of the previous comparisons to determine which reference signal range, supplied by the current generators, should be used for the rest of the coding.

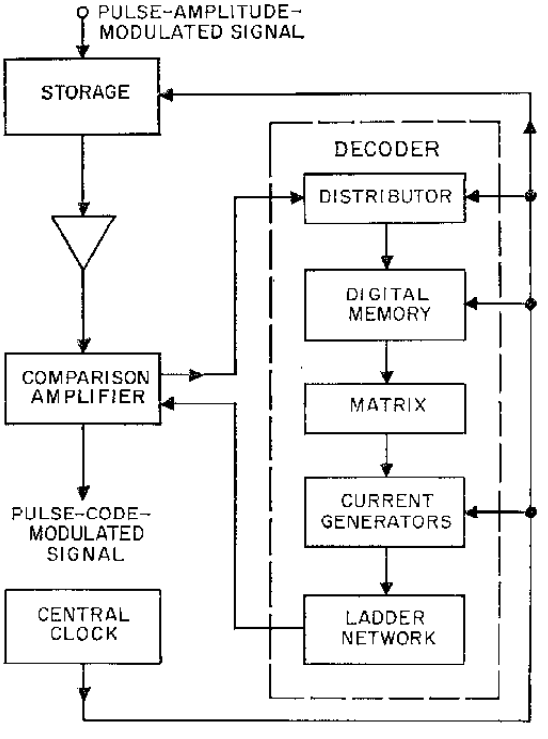


Figure 3—Digital method of producing pulse-code modulation with nonlinear quantization.

From the technological point of view, the only new difficulty concerns increasing the sensitivity of the comparison amplifier. This increase in sensitivity is, moreover, equal to the improvement in the signal-to-quantization-noise ratio for the weak signals.

5.2 Decoding

The decoder is the element that permits the reconstitution of multiplex signals from the codes.

A noticeable symmetry exists between the decoding and coding elements. In fact, the very device that permits the derivation of the reference signals for the comparisons, constitutes a decoder. This device is represented in Figure 3.

The decoding proper does not present any particular difficulty.

The information is almost always transmitted in series; in other words, the different digits of each code appear successively at the input of the decoder. The highest-valued digits generally precede the lowest and, in linear coding using the pure binary code, each digit has a value that is double that of the following digit and half that of the preceding digit.

There are two different ways of using the information that arrives continually at the input of the coder. Either accumulate the information in a digital memory and do the decoding all in one step or translate separately each one of the elements of information as it arrives. In the latter case, the accumulation or integration of the complete code information is made in an analog memory. This analog memory can either be at the level of the common pulse-amplitude-modulation amplifier that precedes the demultiplexer, or it can belong to each one of the channels. The filter that permits the reconstitution of the low-frequency information from the demultiplexed amplitude-modulated pulses, can be used for this.

There is a major disadvantage to using an analog memory for decoding. The energy sent into this analog memory is proportional to the time during which the memory is connected to the decoder; that is, any modulation of the width of the control pulses of each partial integration results in a stray modulation of the demodulated signal.

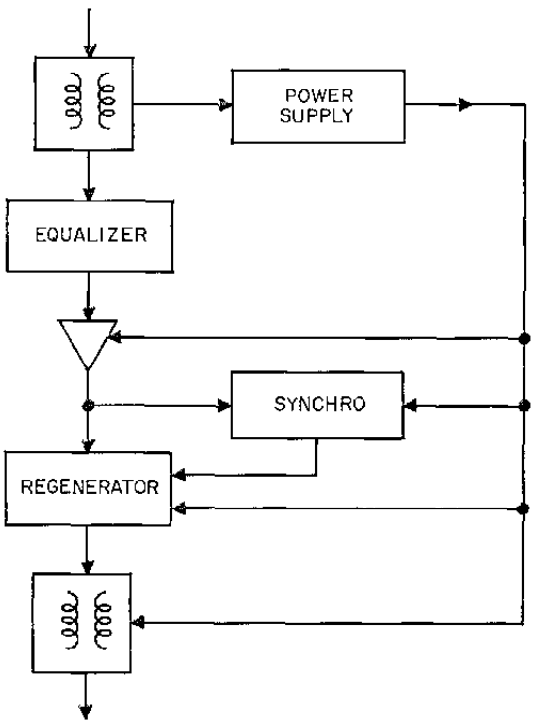
The situation is the reverse if decoding is effected as a whole. Then the energy transferred depends on the width of a signal control pulse and it is even possible to conceive of devices where to a certain extent the width and the shape of this control pulse have only a secondary influence on the demultiplexing.

In our decoding equipment, accumulation is in a digital memory identical to the one in the coder.

6. Amplification and Demultiplexing

In a time-division multiplex telephone equipment that is an integral part of a network, it is not possible to consider separately the different operations that follow the decoding, namely: pulse-amplitude-modulation, demultiplexing, and filtering, combined or not with a low-frequency amplification. In fact, more than anywhere else, the economic problems are closely connected to the technical problems. The fluctuations of the prices of the different electronic components are such that any solution can rapidly become obsolete. It must not be forgotten that in an integrated telephone network a single coder can serve more than 200 subscribers. The total amplification necessary for reception can be obtained through an amplifier common to 200 subscribers. Considerations of the over-all price of demultiplexing gates and filters suitable for each subscriber may make desirable the addition of a rudimentary low-frequency amplifier to each line equipment. Moreover, protection may be thus provided to the speech signal against crosstalk introduced into the part of the equipment that handles pulse-amplitude modulation.

Figure 4—Regenerative repeaters for pulse-code-modulation signals.



Each one of the line equipments used in our system has a low-frequency amplifier. A regenerated repeater for pulse-code-modulation signals is shown in Figure 4.

7. Signalling

The signalling problems in an integrated telephone network using pulse-code modulation are closely related to the switching problems. Neglecting speech signals, the information that passes between the subscriber and the exchange through the concentrator, such as: dialling, supervision, ringing, and of course various other tones, is routed in the same way as the speech signals. As far as the coding, decoding, multiplexing, and demultiplexing are concerned, these signals are not separated from the speech signals. Let us recall that the subset used is a 4-wire set with a transistor amplifier in the ringer.

8. Synchronization

To recognize the beginning of each group of pulses, it is necessary to transmit synchronization codes of the same nature as the speech codes but which have a constant structure and are repeated cyclically. For example, the repetition cadence of these codes can be equal to the sampling frequency of the speech signal. At the receiving end, a digital discriminator makes it possible to distinguish between the synchronization code and other codes, and a correlating device, also digital, assures that it is indeed the synchronizing signal and not a signal formed, for example, by the end of one speech code and the beginning of the following code. The separation of the synchronization code from the range of speech codes is not an a priori requirement.

The synchronization makes it possible to recognize the beginning of each code group, that is, the rank of each unit of binary information received, as well as to determine the number of the multiplexed channel corresponding to this code. The synchronization device possesses a certain inertia so that the accidental destruction of a single unit of information of a synchronization code would not cause the loss of useful information transmitted during the following cycles. This inertia, supplied by a digital memory, is closely associated with the correlation used in searching for synchronization at the start of operation or after accidental loss of synchronization.

It is also to be noted that, on the scale of binary information, a finer synchronization is necessary. The propagation-time variations, the crosstalk between the different multiplex highways or between these and other transmission systems, the fluctuations of the supply sources, and noises of various origins, have repercussions on the phase of the pulses at the input of the exchanges or the concentrators, as well as at the input of the repeaters. The problem of synchronization in an integrated and mesh telephone network using pulse-code modulation has been solved by using digital memories. This

solution will be developed in the paper dealing with switching.

9. Conclusion

The use of pulse-code modulation in an integrated telephone network offers the considerable advantage of making the transmission quality independent of the distance and of the network configuration. The transmission proper introduces practically no additional crosstalk, distortion, or noise. The transmission-line equivalent remains absolutely stable.

The introduction of pulse-code-modulation into telephony offers, moreover, other advantages that, even though unknown to the subscriber, are none the less real. Thus, the elimination of the numerous, costly, and delicate filters used in frequency-division multiplex telephone systems involves substantial savings in volume, weight and price of equipment. Moreover, the use of a single technique for both transmission and switching should lead to the disappearance of the radical differences that exist at present between these two fields of telecommunications.

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Application of Pulse-Code Modulation to an Integrated Telephone Network

Part 3–Switching

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1. General

The application of pulse-code modulation to telephone switching leads to an examination of the 3 main points.

A. Design of a switching stage connecting voice-frequency subscribers' lines to a pulse-code-modulation multiplex highway.

B. Design of a switching stage for interconnecting pulse-code-modulation multiplex highways.

C. Synchronization of the coded and multiplexed speech signals at the input of each exchange to which they are transmitted.

When these 3 problems are solved, the development of the speech circuits in the various types of exchanges through which a call may be routed is straightforward.

The control circuit of a pulse-code-modulation exchange does not raise any particular problem. The interconnections with the speech circuit will be facilitated by the multiplex operation of the latter.

The experimental system, developed to test pulse-code multiplex operation of a telephone system that will be described, used a 7-digit code with a sampling rate of 10 kilocycles per second to provide 24 channels on a 4-wire basis.

2. Speech Circuit

Two types of switching stages and the synchronization circuits will be examined.

2.1 Subscribers' Lines and Concentrator

The switching stage connecting the subscribers' lines to a pulse-code multiplex highway may be in a concentrator near the subscribers and at a distance from the exchange. The 24 channels of the multiplex highway allow 24 simultaneous conversations. The number of lines that can be served depends on their traffic; for average traffic, about 150 subscribers' lines can be connected to a concentrator, 200 or more for light traffic.

2.1.1 *Speech Circuit*

Figure 1 is a simplified representation of the concentrator, which is connected to the ex-

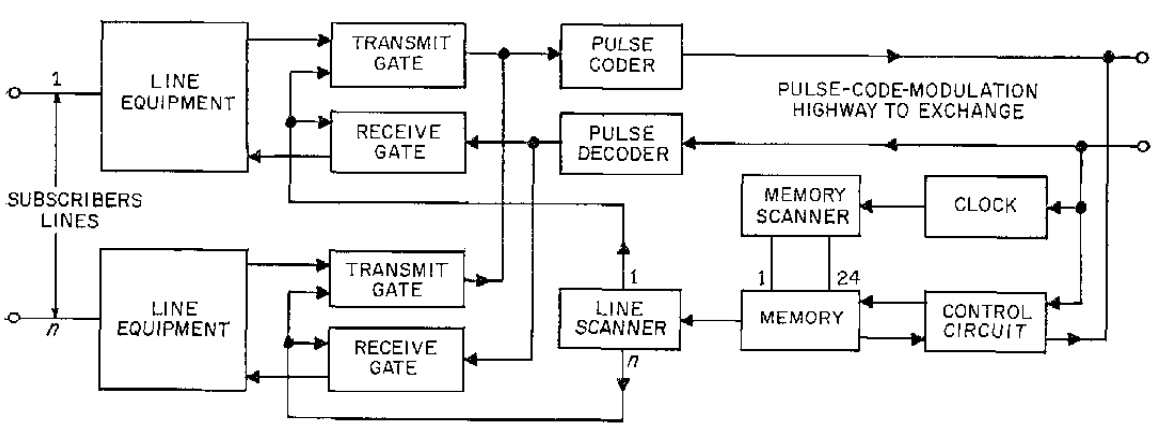


Figure 1—Concentrator for subscribers' lines.

change on a 4-wire basis over 2 telephone pairs. These pairs terminate in a coder and a decoder.

Under control of the line scanner, each active subscribers' line, up to 24 in number, will be connected for 4-microsecond time intervals to the coder and the decoder, giving that subscriber access to the exchange.

The identities of the subscribers in conversation are recorded in a memory having a line for each of the 24 possible connections. Each memory line will record 8 digits to identify in coded form any one of 255 subscribers.

Every 100 microseconds, at time t_n corresponding to connection n, row n of the memory is read. The identity of the subscriber to be connected is transmitted to a line scanner, called the decoding matrix, each output of which controls the 2 gates associated with a subscriber.

The continuous signals corresponding to the state of the line (receiver on hook or unhooked), as well as the dialling signals, are transmitted through the same path as the speech signals.

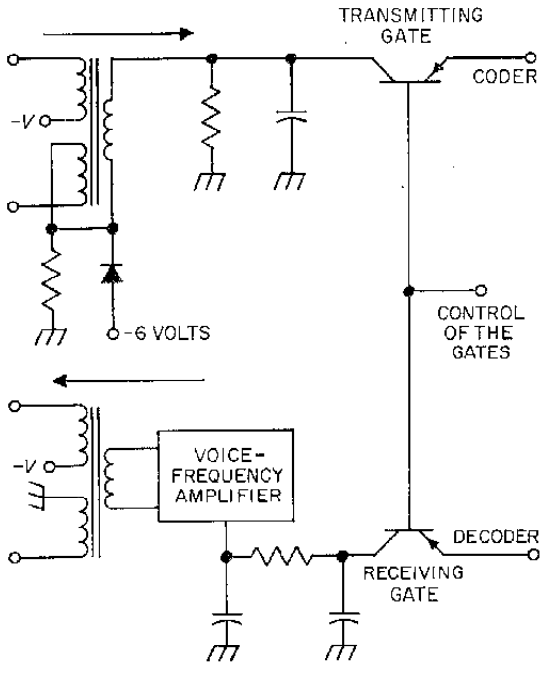


Figure 2—The 4-wire solution of the subscribers' line equipment.

2.1.2 *Control Circuit*

It is economical to perform in the concentrator only the minimum of logic functions, since they can be fulfilled in the exchange by circuits common to several concentrators.

The only functions required from the concentrator are the establishment and release of a connection between a subscriber and a given channel. These controls are accomplished in the concentrator by writing or erasing information on a memory line.

For the passage of information between the concentrator control circuit and the exchange, use is made of the same channel as that for speech.

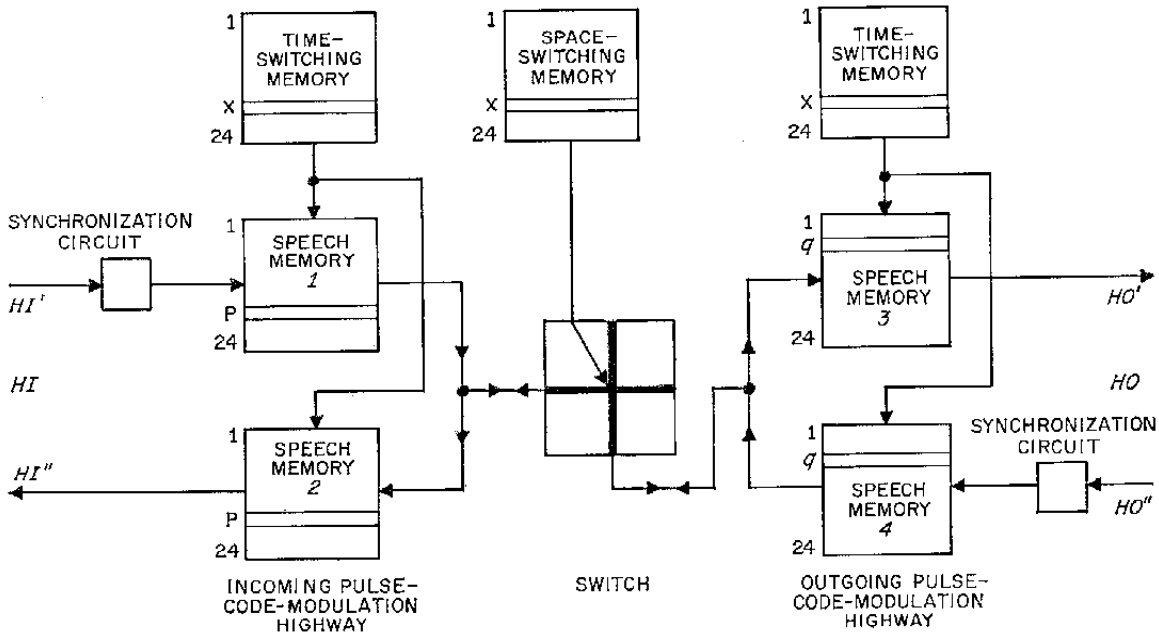
To connect a subscriber to a channel, the exchange sends the code of this subscriber to the concentrator, at the scanning times assigned to the channel. The logic circuit of the concentrator tests the subscriber's line (free or busy) by ensuring that the code of the subscriber is not already written on one of the memory lines, sends the corresponding information to the exchange, and establishes the connection if the called subscriber is free.

For the release of a connection, the exchange sends a release code to the concentrator at the scanning time assigned to the connection to be released.

As digital information can be transmitted between concentrator and exchange at a rate of 70 000 binary digits per second for each channel, the transmission of orders can be performed with great redundancy ensuring very reliable remote control of the concentrator.

The detection of a calling subscriber is controlled from the exchange. Through the control circuit and memory of the concentrator, the line scanner is made to connect each subscriber to the exchange through a free channel, one after

Figure 3—Switching stage between incoming highways through coupled outgoing highways.



the other, thus assuring the connection of all the subscribers to a call-detection circuit, which then knows the state of each line and can recognize a new calling subscriber. All lines are tested about 4 times per second.

2.1.3 *Line Equipment*

The equipment proper to each subscriber's line deserves particular attention. Its influence on both cost and volume of the whole equipment is very important; the other devices of the concentrator are common to all subscribers. One solution, a 4-wire arrangement of line equipment, is given in Figure 2.

Since the use of concentrators reduces considerably the length of the subscribers' lines, the utilization of 4-wire lines may become possible, thus permitting a simplification of the subscriber's set (antisidetone circuit) and of the subscriber's line equipment.

As the sampling is made through 2 different gates, 1 for each direction, these gates are of a simple design employing only 1 transistor of alloyed type per gate.

The transmitting gate operates at a low level compatible with a correct level of crosstalk. A single-transistor voice-frequency amplifier following the receiving gate allows it to operate at a level equivalent to that of the transmitting gate. The control of these gates is thus simplified.

The use of a relatively high sampling frequency of 10 kilohertz permits a simple design for the filtering circuits.

The 2 pairs of the subscriber's line terminate in 2 transformers permitting the transfer from 2-wire circuits on the subset side to 1-wire circuits on the concentrator side.

The average voltage of the signals sent to the coder is −6 volts when the receiver of the subset is unhooked. In this condition, the diode connected to −6 volts is conductive. The voltage drops to about 0 when the receiver is hooked.

Numerous other solutions are possible for the type of line equipment described. In particular, the replacement of each transformer by a symmetrical circuit having 2 voice-frequency transistors may be considered. These transistors, which are already inexpensive, make it possible to reduce the volume of the equipment.

In the case of 2-wire subscribers' lines, the use of a differential transformer for the transfer from 2 to 4 wires seems to be the most economical solution.

The sampling of the conversation is then made by 2 different gates as in the case of a 4-wire line.

A voice-frequency ringer with amplification in the subscribers' set is required so that ringing signals can be transmitted through the speech circuit.

2.2 Switching Stage Between Pulse-Code Multiplex Highways

Any pulse-code-modulation exchange will be connected to other exchanges or concentrators by multiplex highways. The connections between the various channels of these highways are made by a switching stage connecting the so-called incoming highways to the so-called outgoing highways. A description of this switching stage was given in Part 1 of this article.

Figure 3 shows, in a simplified manner, the elements acting in a connection between an incoming multiplex highway *HI* and an outgoing highway *HO*.

2.2.1 *Space Switching*

A space-switching stage having as many inputs and outputs as there are, respectively, incoming and outgoing highways, operates in time division. During any one of the 24 time intervals of 4 microseconds each repeated every 100 microseconds, it can connect any input to any output and the connections made can have any configuration whatever.

In the 4 microseconds during which an input is connected to an output, 2 coded speech samples relating to one conversation are transmitted, 1 in each direction.

The 7 digits of a sample are transmitted successively in time. The speed of the switch makes it possible, during the 0.5 microsecond allotted to a digit, to transmit 1 bit of information in each direction. The control of the connections in the switch is made from a space-switching memory. One of these memories having 24 rows is allotted to each incoming highway. During 1 of the 24 time intervals, t_x for example, row x of this memory is read. On this line is written the coded identity of the output to be connected to the incoming highway; a decoding matrix controls the crosspoint corresponding to the output.

2.2.2 *Time Switching*

Each highway, incoming *HI* and outgoing *HO*, is equipped with 2 speech memories, one for transmit and one for receive, and with a time-switching memory.

For example, the coded speech signals arriving at the exchange on *HI'* are written on the speech memory 1 through the synchronization circuit. The 7 digits of channel 1 are written on line 1 of the memory, those of channel 2 on line 2, et cetera.

The synchronization circuits needed for this operation will be described later.

The coded speech signals leaving the exchange on *HI"* are extracted from speech memory 2. At time t_1, the information corresponding to channel 1 is read in row 1; at t_2, of channel 2; et cetera.

The reading of speech memory 1 and the writing in speech memory 2 are controlled by a time-switching memory similar in operation to the space-switching memory. If, for example, at time t_x, channel p of the incoming highway is handled through the switching stage, the coded identity of row p is written in row x of the incoming time-switching memory.

The readout at time t_x of this row of the incoming time-switching memory causes the readout of row p of speech memory 1, the speech codes there registered are transmitted through the switch toward the outgoing highway, and also causes the writing of the information coming from the switch into row p of speech memory 2.

The outgoing highway equipment functions in the same way as that of the incoming highway.

This type of switching stage thus makes it possible to handle a conversation inside the exchange at a time that is completely independent of its timing on the incoming and outgoing highways. In principle, this can be applied to any type of pulse-code multiplex exchange of the local network. But, as will be seen below, appreciable simplification of the equipment allotted to a highway is possible in many cases.

The establishment of a conversation through the switching stage involves the alignment of the time positions on the highways concerned. Consider, for example, a connection that is to be established between channel p of incoming highway HI and channel q of outgoing highway HO. The marker circuit of the switching stage begins the search for a time position that is simultaneously free on the input of the switching stage III and on the output HO.

Let t_x be a free time. The connection is then established by writing, on row x of the 3 switching memories, the following information:

Code for channel p in the incoming time-switching memory.

Code for channel q in the outgoing time-switching memory.

Code for the outgoing highway in the space-switching memory.

The release of the connection is made by erasing this information.

In certain types of exchanges, it is necessary to make connections between incoming highways or between outgoing highways. This is particularly the case for a local exchange, the incoming highways of which serve concentrators. The local traffic between subscribers necessitates connections between incoming highways. These connections are made by local highways HL, each 1 comprised of 2 coupled outgoing highways HOA and HOB as shown in Figure 4.

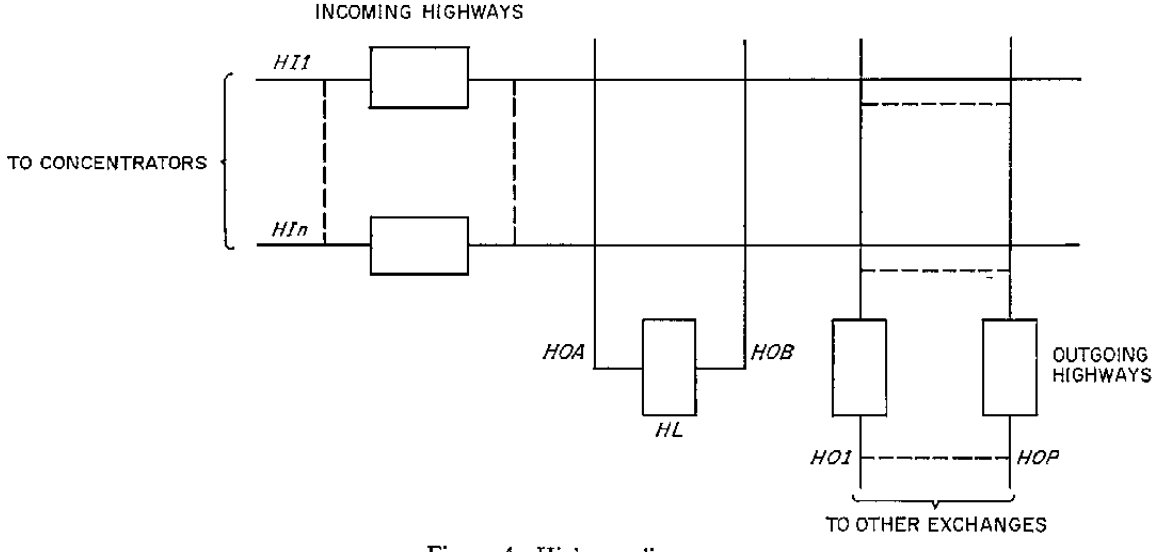


Figure 4—Highway diagram.

2.2.3 *Simplified Highway Equipment*

The memory equipment associated with each pulse-code multiplex highway at the input in the exchange can be simplified in many cases.

If subscribers' lines are connected to the exchange directly, the general organization of the switching stages interconnecting these subscribers to pulse-code multiplex highways remains the same as that of the concentrator shown in Figure 1.

For a pulse-code highway serving a group of subscribers, it is then possible to directly connect the outputs of the pulse coder and decoder to the input of the switch. No speech memory is necessary; the identity memory of the subscribers in conversation serves the same purpose with respect to the control circuit of the exchange as the incoming time-switching memory of Figure 3.

The equipment of a multiplex highway serving a remote concentrator can also be reduced. In comparison with the preceding case, where the stage of connection to the subscribers was in the exchange, it is sufficient to compensate for the propagation times between the exchange and the concentrator. For this, a supplementary delay is added at the input of the exchange to the information coming from the concentrator so that the total delay will be a multiple of 100 microseconds. This additional delay can be provided by a matrix memory of the same type as the speech memories or by a circulation memory such as a delay line.

Simplification is also possible for local highways such as shown in Figure 4. The 2 pulse-code multiplex highways constituting this local highway require only 1 time-switching equipment (2 speech memories and 1 time-switching memory).

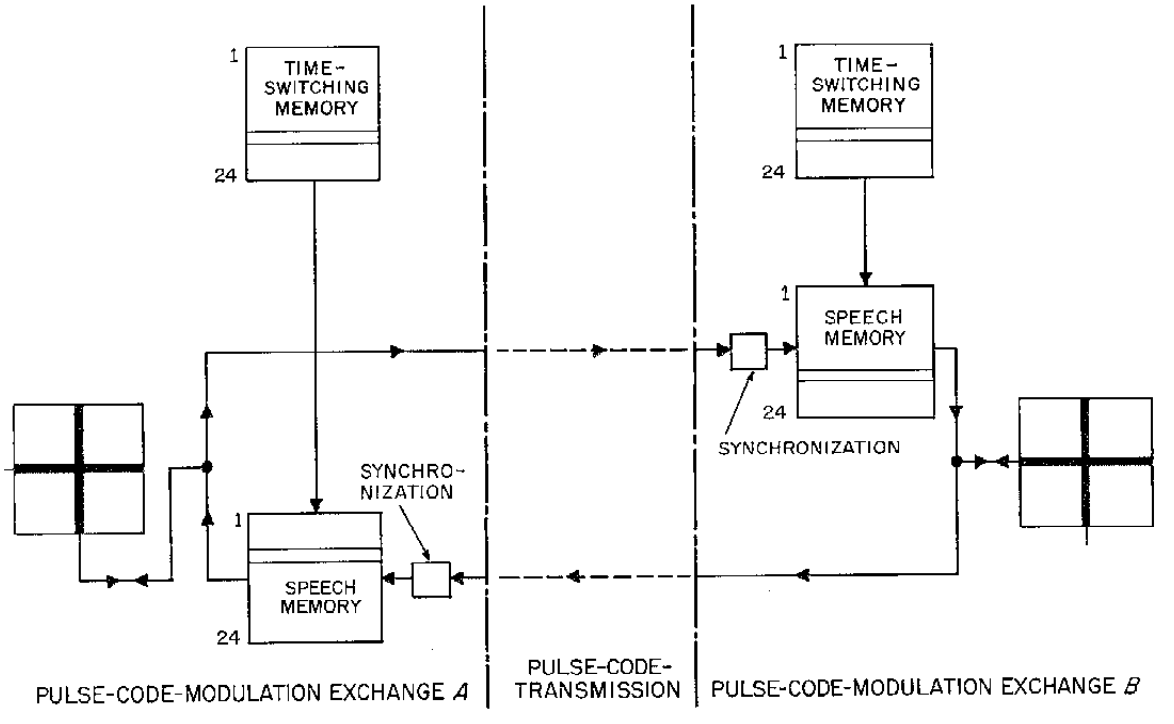


Figure 5—Highway connecting 2 pulse-code multiplex exchanges.

It may be noted that a pulse-code multiplex highway connecting 2 pulse-code exchanges of the type shown in Figure 3 has in each direction of transmission 2 speech memories, 1 in the sender and the other in the receiver exchange. As has been shown, this design permits complete independence of the 2 exchanges and of the transmission circuits.

The deletion of 1 memory is possible provided certain conditions be fulfilled between the control circuits of the exchanges.

Figure 5 shows one possible design. A single speech memory is placed in each exchange at the input of the information, which provides for ready synchronization.

Let us consider a call passing from exchange A to exchange B. The time position chosen by exchange A to handle the call will be used for the transmission direction from A toward B. If exchange B must use a different time for this same call, it must inform exchange A of the change so that A can proceed to the desired time shift for the information arriving from exchange B.

If, in the rest of the routing of the call, connection modifications are necessary in exchange B as at the end of dialling, for routing problems, et cetera, exchange A must be informed of these modifications.

Thus the saving on speech memories requires supplementary exchanges of information between central offices.

2.3 Synchronization Equipment

The principle of the synchronization circuits was described in the first article.

The problem consists in writing the coded speech signals of the 24 channels of a pulse-code multiplex highway in their correct place in the speech memory associated with the highway at the input of the exchange.

These signals may undergo time fluctuations because of imperfections of the transmission circuits and differences in the frequencies of the exchange clocks.

Figure 6 shows the circuits providing for synchronization of the information.

The writing and reading of information in the speech memory are done successively. The time allotted to a digit of 0.5 microsecond is divided into 4 elementary times, the first being

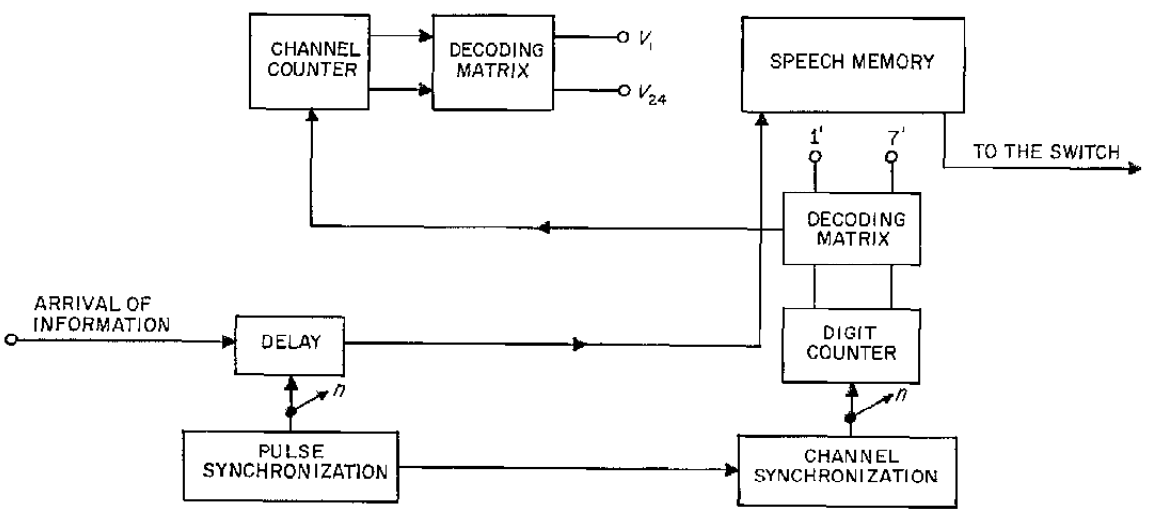


Figure 6—Sychronization circuits.

allotted to the writing in the memory, the third to the readout, the two others being used to select the points to be written or read.

The bits of information arrive in the exchange at any time; it is first necessary to individually fix each one of them in time before writing them in their place in the memory. This leads to dividing the synchronization circuits into two parts.

2.3.1 *Pulse Synchronization Circuit*

The incoming information is delayed by a variable time of less than $\tau = 0.5$ microsecond so as to arrive at the input of the memory at the intended elementary time.

The delay, which is produced by flip-flops, varies in a discontinuous manner and can take the 4 following values: 0, $\tau/4$, $\tau/2$, and $3\tau/4$.

The adjustment of delay is made by a pulse synchronization circuit common to several multiplex highways. This circuit continually checks the time delay of each highway. To do this, it compares the arrival time of the information on a highway to that of the time pulses delivered by the exchange clock and then estimates from this the necessary delay on this highway.

When the delay on a given highway increases from $3\tau/4$ to τ, the addressing circuits for writing the information in the memory are modified and the delay introduced in the highway is returned to 0.

In the same way, when the delay is 0 and the information arrives too early, the addressing circuits of the memory are modified and a delay of $3\tau/4$ is introduced into the highway.

2.3.2 *Channel Synchronization Circuits*

The address of the memory point where the information must be written, at a given time, is derived from 2 counters; 1 counts by 8 and gives the identity of the column of the memory and the other counts by 25 and gives the identity of the row of the memory.

These counters operate with a certain time shift with respect to the exchange clock. This corresponds to the shift of the synchronization channel with respect to the local time t_{25}.

A channel synchronization circuit common to several multiplex highways also works in time division. It is successively connected to all the multiplex highways and serves two purposes.

It checks for correct synchronization of the 2 counters in the following manner. When the channel counter reaches the count of 25, it checks that the incoming information conforms with the synchronization code. A decision is taken only after 3 successive checks to avoid the possibility of an error due to the transmission circuits. The counters are resynchronized when their phases are wrong.

It modifies the addressing of the information in the memory when the pulse synchronization circuit has to change the delay from $3\tau/4$ to 0 or vice versa. This modification consists of putting the digit counter of the highway concerned forward or backward by 1 time unit.

The synchronization problems are simpler when the clocks of the 2 switching equipments connected by a multiplex highway are synchronized. The connections between concentrators and exchanges are of this type. The synchronization of the concentrator consists of correctly setting the counters with respect to the synchronization channel, and then no delay is necessary.

3. Evaluation of Possible Traffic Improvement In Highway Efficiency

3.1 Traffic Possible by Time Position Alignment

The establishment of a connection through an exchange necessitates, as shown in 2.2.2, the existence of a time position that is simultaneously free on the input and on the output to be connected. Therefore, a certain internal blocking results in all time-division switching systems.

Computations have been made of the traffic possible on time-division multiplex highways [3–5].

In the case of 24-channel multiplex highways, the possible traffic on a highway is about 12 erlangs, admitting 1 lost call out of 100. The corresponding efficiency of the highways is 50 percent.

As an example, let us consider an average traffic per subscriber at the busy hour of 0.1 erlang. Each concentrator can then serve 120 subscribers. A local exchange of 9000 subscribers will interconnect 75 incoming highways arriving from the concentrators and 75 outgoing highways going to other exchanges or serving local traffic.

Elimination of internal blocking due to time-position misalignment is possible by applying a rearrangement operation [6, 7].

3.2 Principle of Rearrangement

If we consider, in a switching stage between pulse-code multiplex highways, a connection to be established between an incoming highway HI having only the time position t_x free and an outgoing highway HO having only the time position t_y free, the problem cannot be solved without rearrangement.

A rearrangement is then made by releasing the connection that occupies the time position t_x of HO, which makes it possible to establish the desired connection, then by re-establishing the connection that was just released at time position t_y. This re-establishment may perhaps require releasing another connection, which will be itself re-established at another time position, and so forth until a free path is found for each connection.

Take, as an example, a 5-by-5 switch interconnecting inputs 1, 2, 3, 4, 5 to outputs *a, b, c, d, e*. Each input and each output serve 1 pulse-code multiplex highway (see Figure 3).

Column *A* of Table 1 shows the connections between the inputs and outputs during t_x and

TABLE 1
Successive Arrangements

A			B			C			D			E		
	t_x	t_y		t_x	t_y		t_x	t_y		t_x	t_y		t_x	t_y
1	b	d	1	b	d	1	b	d	1	b	*d*	1	**d**	b
2		a	2		a	2		a	2		a	2		a
3	d	e	3	d	c	3	d	*e*	3	**e**	**d**	3	e	d
4	a		4	a	**c**	4	a	c	4	a	c	4	a	c
5	e	c	5	e	*c*	5	**c**	e	5	c	e	5	c	e

t_y, 2 of the 24 time positions. At time t_x, input 1 is connected to output *b*, input 2 is free, input 3 is connected to output *d*, et cetera; at time t_y, input 1 is connected to output *d*, et cetera.

Let us assume that a new connection is to be established between input 4 free at time t_y and output *c* free at time t_x, this output not being connected to any input at this time.

The connection 4-*c* cannot be made without modification of the connections already established.

The connection 5-*c* existing at time t_y is released. This change makes connection 4-*c* possible at time t_y. Column B shows the connections that are then made.

The connection 5-*c* is then re-established at time t_x by releasing the connection 5-*c*, which is itself re-established at time t_y by releasing the connection 3-*e* (Table 1, Column *C*).

The same process, consisting in permuting the time positions t_x and t_y during which 2 connections are established, is continued.

The connections 3-*e* and 3-*d* are permuted in time, the connection 1-*d* being released (Column *D*).

The connections 1-*b* and 1-*d* are permuted (Column *E*).

The re-establishment of the connection 1-*b* at time t_y does not necessitate any release, the output *b* being free at time t_y. The rearrangement of the connections is finished.

The process just described can be generalized for any size exchange. It is easily shown that the number of connection permutations always remains lower than the number of inputs or of outputs of the exchange.

Calculations have been made of the average number of connections that must be modified in order to establish a new one for various sizes of multiplex switching stages. This number is relatively low; in the most pessimistic cases, it is necessary to modify on the average one conversation to establish a new one.

Application of the rearrangement method does not introduce any additional equipment into the speech circuits of a pulse-code-modulation exchange; only a slight increase is required in marker equipments.

When the connections between pulse-code-modulation exchanges are designed as in Figure 5 (1 speech memory in each exchange; see paragraph 2.2.3), any permutation of time on a highway required by 1 exchange for a rearrangement must be made known to the corresponding exchange.

On the other hand, if the highways are equipped with 2 speech memories, as in Figure 3, the rearrangements are made independently in each exchange.

3.3 Traffic Possible by Rearrangement

Since the rearrangement method eliminates internal blocking in switching stages between pulse-code multiplex highways, the efficiency of these highways depends only on the amount of traffic handled in 1 direction.

A highway connecting a concentrator to an exchange constitutes a trunk of 24 channels, which can handle a traffic of about 15 erlangs, permitting 1 lost call out of 100.

Assuming traffic of 0.1 erlang per subscriber at the busy hour, the concentrator could serve 150 subscribers. A local exchange of 9000 lines would have under these conditions 60 incoming highways.

The efficiency of the outgoing highways handling the total traffic, that is, 900 erlangs, depends on the number of directions in which this traffic is handled. Efficiencies of the order of 80 percent can be expected, which leads to 47 outgoing pulse-code multiplex highways.

It will be recalled that in the case of an exchange operating by time-position alignment (paragraph 3.1) having the same number of lines and handling the same traffic, 75 incoming highways and 75 outgoing highways were necessary.

The rearrangement method therefore allows an important saving in equipment, as much from the point of view of transmission as from the point of view of switching.

Improvement in the efficiency of the highways is possible by other methods than that described. Notably, the increase in speed of electronic elements makes it possible to envisage multiplexing a greater number of channels on the inputs and outputs of the space-switching stage; for example, 48 instead of 24. The blocking due to the time-position alignment is then decreased.

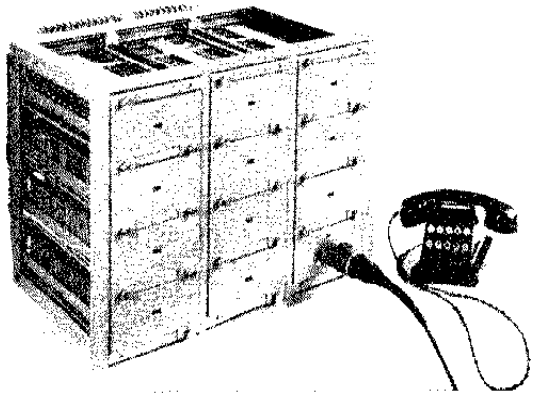
4. Switching Models

Theoretical and experimental work has led to the construction of test models of a concentrator and of a pulse-code-modulation exchange.

The model of the concentrator is illustrated in Figure 7. The equipment common to the subscribers; coder, decoder, memory, and control circuits, was constructed as for a real concentrator, and 28 subscribers' lines were equipped. A concentrator serving 150 subscribers' lines would have only the slightly greater volume of 0.15 cubic meter (5.6 cubic feet).

The model of the pulse-code multiplex exchange has 2 incoming pulse-code multiplex highways, one able to be connected to the concentrator, the second serving 3 registers and an operator's position, and 3 outgoing pulse-code multiplex highways, 2 of which are associated to form a local highway.

Figure 7—Experimental concentrator equipped for 28 subscribers' lines.



These highways are fully equipped with speech and switching memories. The design of the switching stage conforms to Figure 3.

The control circuit would be suitable for an exchange of several thousand subscribers' lines so far as its essential elements are concerned.

The control circuit for a pulse-code-modulation exchange is simpler than that for an exchange having a space-division speech circuit. In fact, all the information written in the switching memories of the speech circuit of a pulse-code exchange is accessible to the control circuit. The latter then requires only very few memories. Moreover, the transfer of information between speech and control circuits is accomplished with only a few interconnecting elements due to the multiplex operation of the speech circuit and to the digital nature of the signals that it routes.

The model required elementary logic circuits having relatively high-speed performance: response time < 0.1 microsecond. Fast transistors and diodes were used.

The storage points in the speech memories use a capacitor associated with 2 diodes, this capacitor being charged or not according to whether the information to be written is 1 or 0.

5. Prospects

The switching models that were built have made it possible to verify the validity of the solutions of the problems met in applying pulse-code modulation to telephone switching.

From the point of view of switching, the advantages are very important decreases in volume, weight, and power consumption of the equipment; and a very great flexibility in operation.

The use of the same type of modulation for both transmission and switching presents, among numerous advantages, a very easy exchange of supervisory information between exchanges that makes possible the locating of subscribers' line concentrators conveniently near the subscribers and remote from the exchange.

The very great progress made in recent years in both performance and cost of fast digital techniques, now makes it possible to consider pulse-code-modulation solutions that will be competitive with conventional designs. It is very likely that even the near future will show still further improvement in the situation.

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Jean Le Corre was born in Guiler-sur-Goyen, France, on 24 August 1927. He graduated as a radio-electrical engineer from the Ecole Supérieure d'Electricité of Paris in 1950.

In 1954, he joined Laboratoire Central de Télécommunications and was engaged in work on electronic telephone switching. He is now head of the electronic switching department.

Experimental Pulse-Code-Modulation Transmission for Local-Area Telephony

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1. General Description

1.1 Purpose of Development

Pulse-code modulation, being extremely resistant to perturbations in the transmission path, is commonly postulated for use on obviously poor transmission media. Another application, potentially no less important, is to extend the transmission capacity of conventional paths such as multi-conductor cable. Paired cable intended for voice-frequency traffic is a poor medium for multiplex or other wide-band signals, on which it superimposes large and variable amounts of crosstalk and noise. These degradations are readily withstood by a binary signal, in which only the presence or absence of a pulse has to be detected, and which can in consequence be regenerated without accumulation of impairment.

The increase in capacity offered by this technique is most likely to be needed on junction routes between exchanges in large cities or conurbations,* which combine a substantial investment in multi-conductor cables with a large and growing traffic. Looking ahead a little further, it is also in these areas that the combined problems of transmission and switching may best be solved by digital techniques. Consequently, this equipment was designed for junction working; specifically, for interconnecting rotary exchanges in a designated experimental location, although there is nothing to preclude its use elsewhere.

1.2 Facilities

The system basically provides 24 channels, but, since one channel is allotted to synchronizing, 23 speech paths are provided, each with associated signalling facilities. Each speech waveform is sampled at 8 kilohertz, a high enough rate to permit transmission of the normal 0·3 to 3·4 kilohertz pass band. Each sample amplitude is quantized into one of 70 levels. Since this number is inadequate for transmission of speech signals having a wide range of mean volume, amplitude compression is used to extend the volume range by about 25 decibels, the ratio of clipping level to the smallest quantum interval being then about 56 decibels. Two similar paths, one in each direction, are combined by voice-frequency hybrid coils.

The coded speech requires 7 binary digits for each sample. An 8th digit is appended to carry signalling information. This could convey 2-condition signalling directly, but since the rotary exchanges require 3 or 4 conditions a further stage of time-division is used to break down this digit into 2 binary digits at 4 kilohertz each. The resulting signalling capacity, namely 2 high-speed channels in each direction associated with each speech channel, is adequate for any known type of switching plant.

1.3 Line Transmission

The total number of binary digits generated per second is the product of the sampling rate (8 kilohertz), number of digits per sample (8), and number of channels (24), namely, $1{\cdot}536 \times 10^6$. These are assembled sequentially for transmission on a single pair in each direction.

The attenuation of 0·64-millimetre diameter paired cable at 1·536 megahertz is about 25 decibels per kilometre; crosstalk margins vary considerably but in the worst cases either near end or far end can be less than 40 decibels on a kilometre length. Repeaters are therefore

* A great aggregate of urban communities, con+urban.

necessary. The mean spacing used is 1 kilometre with some scatter to permit the use of existing manholes. This interval, decided on early in the project, is unnecessarily short. The sending level is 2 volts peak to peak into 100 ohms, and a typical receiving level (after equalization) is about 100 millivolts. The repeaters will tolerate fluctuations of input level within the range ±6 decibels at the expense of some deterioration of noise rejection.

The repeaters are supplied with power over the signal pair. The whole chain is supplied in series from the transmitting terminal, each repeater dropping 9 volts with a current supply in the range of 25 to 40 milliamperes.

Power separation, as well as matching and balancing, requires the repeaters to have line transformers on input and output. These transformers do not pass direct-current or voice-frequency components of the code signal; the resulting distortion of long sequences of marks or spaces would make their regeneration uncertain unless some special form of signal were used. The code used in this equipment avoids long sequences, so making possible the regeneration of signals in which each mark is a unidirectional pulse.

1.4 Terminal Equipment

A block diagram of the transmitter is shown in Figure 1. Each speech input, after filtering and

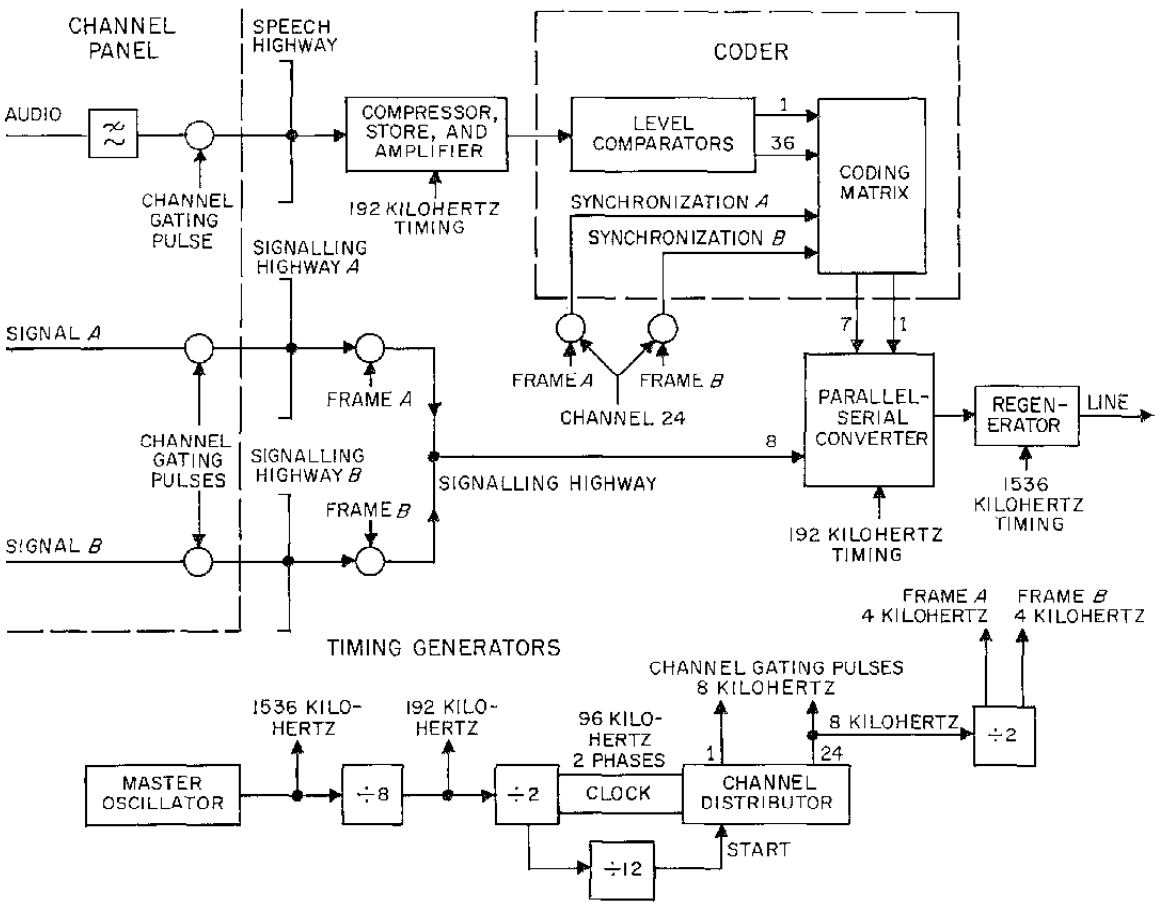


Figure 1—Transmitter.

amplitude limiting, is sampled for 2·6 microseconds in every 125 microseconds. The conversion efficiency of the storage-type modulator [11] is high and no channel amplifier is needed. The channels are combined on the speech highway, where the signal takes the form of multiplex amplitude-modulated pulses with a guard space between channels approximately equal to the sample width. The amplitude range of the pulses is reduced by the compressor. The compressor output is sampled again, and the sample amplitudes are stored as a charge on a capacitor to provide a constant-amplitude input to the coder throughout the coding operation.

The coder contains an array of amplitude comparators, biased to operate at different levels, to define the quantizing intervals. By an artifice explained in Section 3.2 each comparator serves to define 2 levels. The outputs of the comparator array define the sample level; they are connected to a matrix of diodes that energizes the appropriate digit outputs to generate a 7-digit code character. The digits are applied simultaneously to 7 of the 8 distinct inputs of a parallel-to-serial converter. They appear in sequence at the single output, and after regeneration are launched into the line.

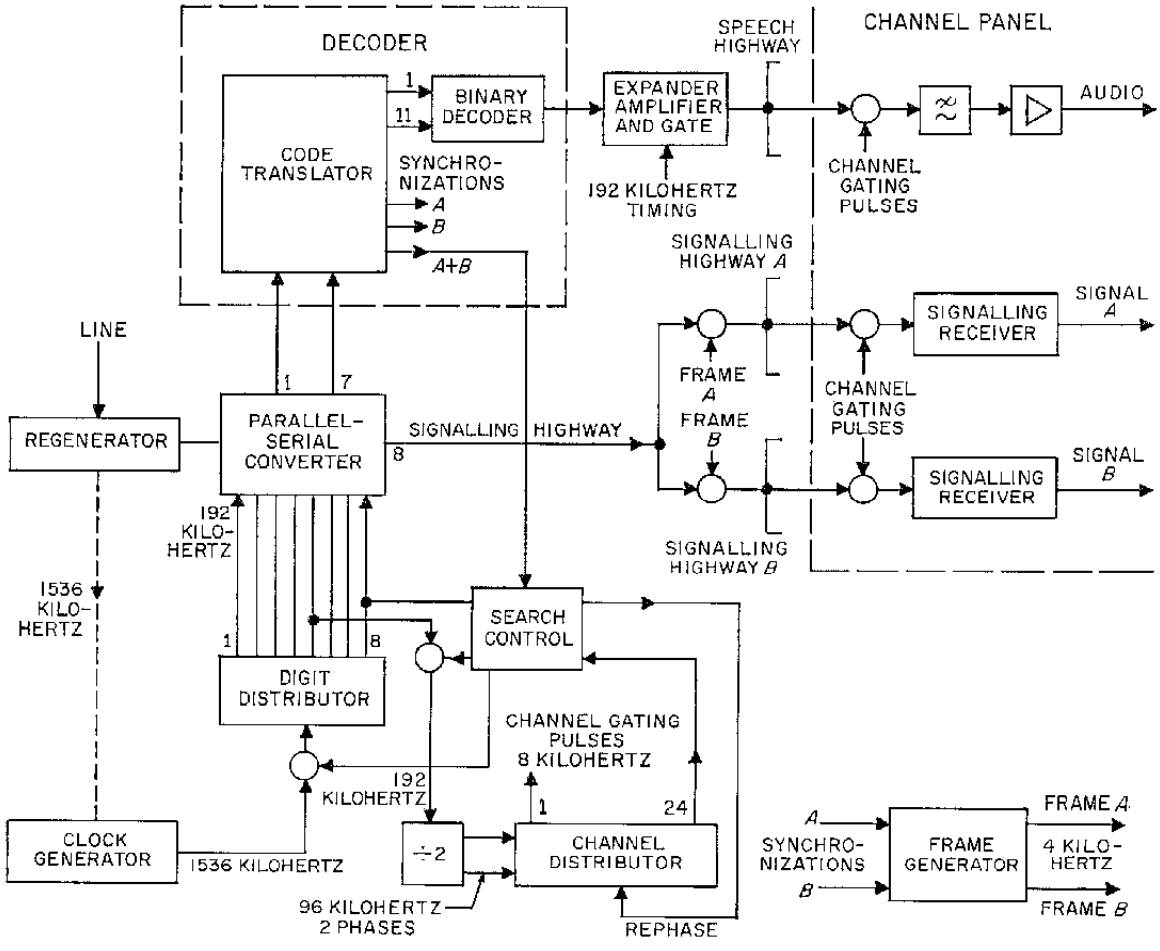


Figure 2—Receiver.

The number of signalling conditions necessitates the use of two binary signals, A and B, for each speech channel. Each signal input is sampled, and the signals from the various channels are assembled on signalling highways A and B. The latter are connected through alternately energized gates to the common signalling highway, which carries a complete frame (125 microseconds) of A signals followed by a complete frame of B signals, and so on. This combined signal is applied to the remaining input of the parallel-to-serial converter and appears as the 8th digit in the sequential output.

To identify the channels, so that they may be correctly separated at the receiver, distinctive code characters are transmitted in channel 24. To identify the frames so that signalling digits A and B may be separated, these synchronizing characters differ in alternate frames producing synchronizations A and B. Each character is generated by energizing an appropriate line in the diode matrix during the synchronizing channel period; the coder is inhibited at this time.

A block diagram of the receiver is shown in Figure 2. Much of this is obviously complementary to corresponding parts of the transmitter. Each incoming code character is arranged in parallel form, translated to a binary number by an array of saturable magnetic cores, and then converted by the decoder into an amplitude-modulated pulse. This, after amplitude expansion complementary to the initial compression, represents the quantized approximation to the input sample. The amplitude-modulated pulses are distributed to the individual channels; each channel audio signal is recovered by filtering, again with high conversion efficiency, and amplified to the required output level.

The signalling digit is separated and distributed via frame and channel gates to signalling receivers, each of which provides as output a make contact for actuation of the relay sets.

The process of searching for and maintaining synchronism with the transmitter requires some apparatus with no counterpart at the other terminal. The code translator array includes elements responsive to each synchronizing character. During normal running, a synchronizing character is received, and recognized, during the period allotted by the receiver channel distributor to channel 24. If the receiver is out of step with the transmitter, this recognition does not occur, and its absence initiates a process of searching. The counting rate of the digit distributor is changed, by inhibiting some of its clock pulses, with the effect that every possible group of 8 adjacent digits is offered in turn to the synchronizing detector. Eventually the synchronizing character, if present, is recognized; this stops the search, and starts the receive distributor running normally in synchronism with the incoming characters and frames.

2. Regeneration

2.1 Regenerable Signals

The process of regeneration has two aspects. One is the standardization of the individual pulses, which must be indistinguishable by amplitude, shape, or duration. The second aspect is the allocation of pulses to discrete positions on a regular time scale. This offers no especial difficulty at the transmitter, where a timing wave may be taken from the master oscillator. At the repeaters and the receiver, the only timing information available is that which can be recovered from the pulse-code-modulated signal. That is to say, we must extract a timing wave at the mean digit frequency and of constant phase in relation to the mean digit position despite the facts that some digits are mark and others space, more or less at random, and that any digit may be perturbed during its passage through the line.

If a signal is to be regenerable, its information content per unit time must not exceed the capacity of the medium. This is necessary but not, however, sufficient; since there are various practical hazards that may be overcome only by some form of restriction on the signals to be transmitted. We may, for instance, refrain from

fully utilising the available bandwidth, signal-to-noise ratio, or repertoire of code sequences. Any of these procedures amounts to the provision of excess capacity as the price of some other benefit. The point may be clarified by discussion of two specific benefits desired.

As noted in Section 1.3, the exigencies of line transmission include the generation of some form of signal that will not suffer unduly from the inability of line transformers to pass direct current and very-low-frequency alternating-current components. Possible solutions may be classified according to the period of time over which the average (loosely the direct-current level) is reduced to negligibility. In order of increasing time period the possibilities are:

A. Period of 1 digit: use of dipole impulses, either amplitude-modulated (Figure 3B) or phase-modulated (Figure 3C).

B. Period of a few digits (from 2 up to character length): alternate mark inversion (Figure 3D).

C. Period of 1 character: use of a low-disparity code, in which limited numbers of marks and spaces occur (Figure 3E).

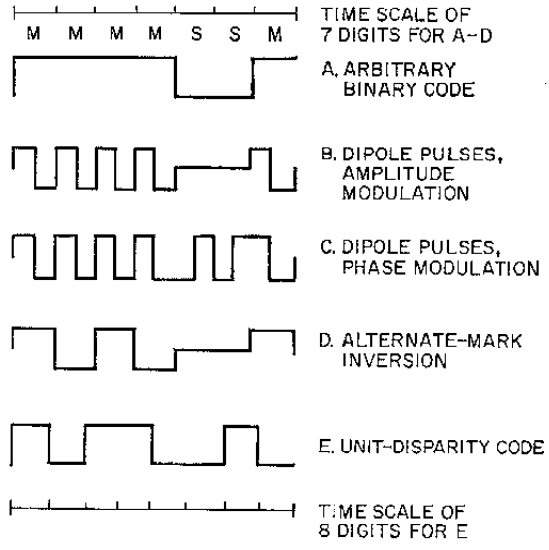


Figure 3—Alternative forms of coded signal.

D. Period of 1 cycle of modulation: ensure that any constant conditions that occur use characters of low disparity, reserving the high-disparity characters of the code for levels excited intermittently.

Method *D* alone is inadequate, since the time period is inconveniently great; although it is a useful adjunct to other methods. Method *A* tolerates the greatest low-frequency cut-off, but occupies by far the greatest bandwidth. Method *B*, requiring 3 distinguishable levels, is less resistant to crosstalk or noise. Of the practicable methods, *C* appears to be the most efficient in its use of informational capacity. It required more development, since no suitable codes or coding methods were known as previous pulse-code-modulation work invariably used binary-number or reflected binary codes. A variety of codes and coding methods has been devised [1–4] and the one used in this equipment will be described in Section 3.

If we define disparity as the excess of marks over spaces in any one character, then the disparity of a binary number of n digits (or any other code having the full repertoire of characters, 2^n in number) ranges from $+n$ (all marks) to $-n$ (all spaces); and the number of consecutive marks or spaces that can occur is indefinitely large. At the other extreme, it is possible for all the characters of the code to have the same disparity. A compromise seems admissible, and the one chosen is that the disparity of a speech character shall be ± 1. That is to say each character in a code of $2n + 1$ digits has either n marks and $n + 1$ spaces, or vice versa. The number of levels pertaining to

TABLE 1
NUMBER OF LEVELS FOR VARIOUS NUMBERS OF DIGITS

Number of Digits	Number of Levels	$\log_2$ (Number of Levels)
5	20	4·33
7	70	6·12
9	252	7·97
$2n + 1$	$2(2n + 1)!/n!(n + 1)!$	

various numbers of digits is shown in Table 1. The redundancy is equivalent to about 1 binary digit.

With speech conveyed by a 7-digit unit-disparity code, and signalling by an independent digit, each 8-digit character has 3, 4, or 5 marks. The longest run of marks or spaces that can occur in an arbitrary sequence of characters is 9.

Timing information may be derived only from transitions of the signal waveform. According to the form of the signal, these may depend on the presence of marks or of transitions between mark and space. (Phase-modulated dipole pulses are an exception.) In a code using the full range of disparity, the density of marks or transitions can fluctuate over a wide range:

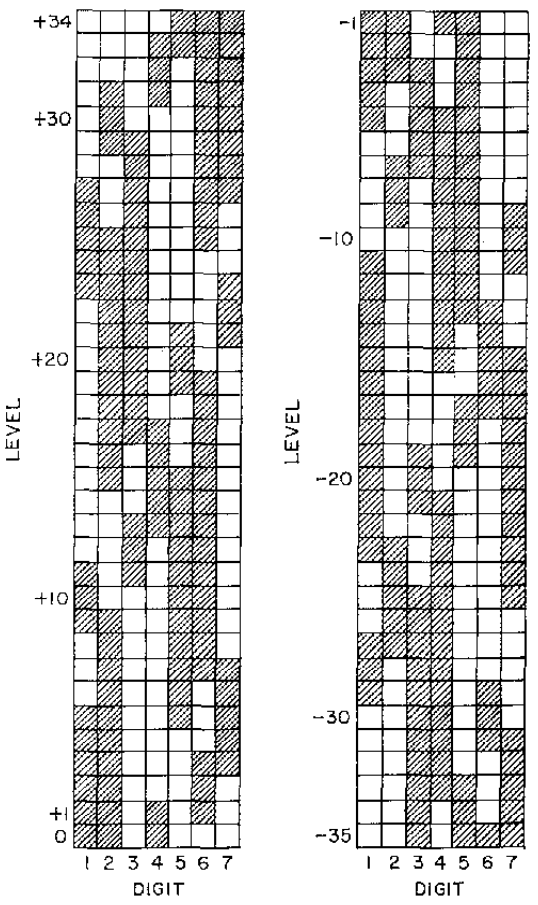


Figure 4—The code.

as an extreme, there may be none at all. Even if we dismiss complete absence as unlikely to be sustained, fluctuations of say 8 to 1 are undesirable for two reasons. Firstly, temporal fluctuations in the density of any single pulse train increase the difficulty of extracting a sufficiently regular timing wave. Secondly, differences of mean density among the various signals in a cable reduce the crosstalk margin proportionately: a high-density signal perturbs the timing of a low-density signal more than 2 equal signals perturb each other.

A low-disparity code, by definition, constrains the density of marks. The density of transitions is also constrained, though not so closely; moreover, it is easy to ensure that characters belonging to levels near 0 (which occur most frequently) avoid extreme transition densities. The code actually used (Figure 4) is a reasonable solution to each of the problems discussed.

To make the repeaters as simple as possible, pulses occupying only half the digit period were used in the first trial. This served its immediate purpose, but transmission of such pulses is wasteful of bandwidth. Later work, not described here, postulates a bandwidth approximately half that required in the original system.

2.2 Repeaters

Use has been made of 2 types of self-timing repeaters, whose block diagrams are shown in Figure 5. The essential difference is that the timing waveform is derived from the output or the input signal, respectively. Each has its own advantages, but on the whole input timing is preferred; it allows a wider tolerance on the mean frequency, and it is applicable to more-ideal regenerators with narrow sampling.

Several functions are, of course, common to all types of repeater. The line attenuation is equalized to within 1 decibel for a line section of average length, from low frequencies up to almost the digit rate, by an equalizer having 2 resonant sections. The signal level is raised by

a 2-transistor preamplifier to about 2 volts peak, enough to operate a discriminator or gate.

The design of the timing filter is a compromise. To obtain a reasonably pure unmodulated sine wave from an input signal with virtually random modulation, a very narrow pass band is desired. However, the phase shift of the extracted sine wave must be confined within perhaps 10 degrees for any likely frequency drift of oscillator or filters; so frequency tolerances set a lower limit to the bandwidth. The practical solution was a quartz-crystal filter with a bandwidth of about 3·5 kilohertz. A 10-degree phase shift occurs at ±300 hertz, which is substantially wider than a practicable crystal tolerance.

It has been shown possible to use linear addition of the signal and timing wave [5, 6] but the use of a combined discriminator and timing gate has been preferred, even in an avowedly simple repeater, since the gate determines pulse width as well as time of occurrence, so permitting the use of a simple output stage; the gate isolates the controlled signal from the timing, on which it has little or no influence; and variations in timing-wave amplitude have less effect. A quite-satisfactory version used a saturated-transistor switch as a timing gate, but a second model using a current switch exhibited less intersymbol interference and will be described here.

In Figure 6A, transistors *T1*, *T2*, and *T3* are emitter-coupled, although with the low supply voltage used in the repeater it is not possible to define the emitter current precisely. The output is taken from the collector of *T3*, which can draw the full current only when *T1* and *T2* are cut off by positive excursions of their bases. The timing wave applied to *T1* ensures that it diverts current from the other transistors for half of each digit period. During the other half, current flows in *T2* or *T3* depending on whether the base of *T2*, driven by the amplified

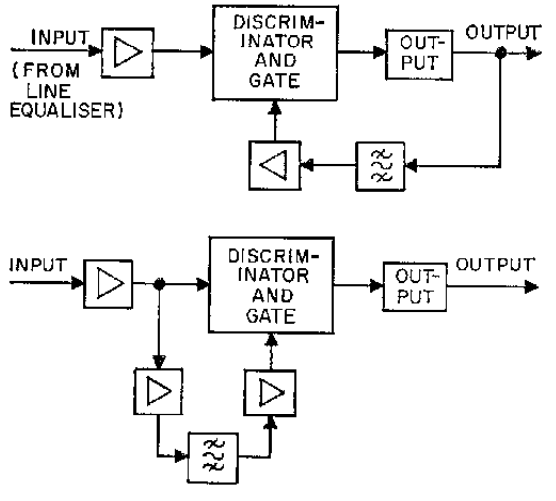


Figure 5—Repeater diagrams.

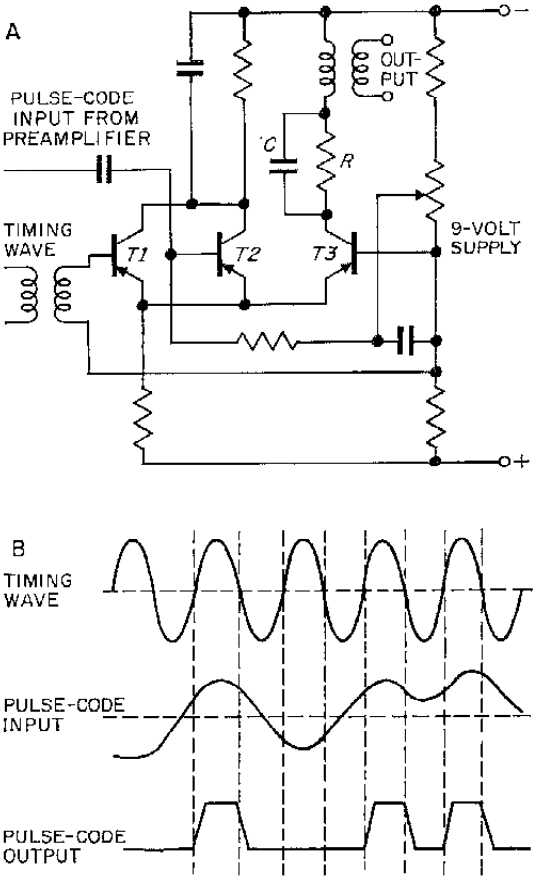


Figure 6—Combined discriminator and timing gate.

pulse-code-modulated signal, is negative or positive. The effect is that the circuit acts as a combined timing gate and amplitude discriminator, as illustrated by the waveforms of Figure 6B. Since a substantial output power is drawn from the collector of *T3*, this transistor is on the verge of saturation at pulse peaks; so better waveforms are obtained by building out the load to a constant resistance, using the capacitance-resistance network shown in series with the effective inductance and resistance of the transformer-coupled line.

It will be noted that, while the gate can restrict the duration of output pulses below that of the input, it cannot augment the width of narrow pulses. Consequently, it is dependent on a bandwidth restriction in the combination of line, equalizers, and preamplifier, which produces roughly the pulse shape shown in the figure.

All repeaters tested were capable of rejecting symmetrical interfering signals whose peak-to-peak amplitude was 6 decibels below signal under favourable conditions and 12 decibels below over a range of input levels and patterns.

The repeaters are supplied with power over the the signal pair. The consumption of each one, namely 20 milliamperes at 9 volts, is sufficiently low for several to be supplied in series over the lightest gauges of cable in service.

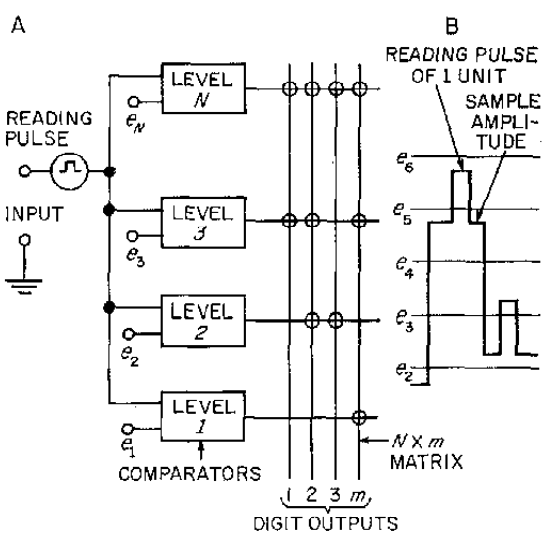


Figure 7—Coder for arbitrary code.

3. Coding and Decoding

3.1 Code

To provide an easily regenerable binary signal, as discussed in Section 2.1, a unit-disparity code was desired.

The various digits of a parallel coder are independently generated or regenerated, and large errors can occur if several digits are simultaneously in a state of indecision. It is, therefore, an advantage to employ a unit-distance code, in which characters belonging to adjacent levels differ in only 1 digit. Moreover, as will appear in Section 3.2, the structure of a parallel coder is simplified by use of a unit-distance unit-disparity code. The 7-digit code shown in Figure 4 has both these properties. It is also cyclic: the 2 extreme characters are suitable for juxta-position, so that the assignment of an origin is quite arbitrary. Also, any digit column in the raster is equivalent to any other when translated by some multiple of 10 places, so the assignment of digit positions is arbitrary. Ignoring obvious permutations there are 12 essentially different codes of this type. The choice of code and origin was based mainly on 2 criteria: (A) the characters belonging to levels near 0 which occur most frequently, should have an average number of transitions between mark and space and (B) these same characters should have the minimum probability of simulating suitable synchronizing characters.

3.2 Principle of Coder

We first refer to a perfectly general method for generating an arbitrary code, and then show how it may be simplified in the case of a unit-disparity unit-distance code. In Figure 7, consider an array of level comparators with a series of equally spaced bias levels e_1, e_2 . . . e_N. Each comparator gives an output pulse when the

common input exceeds its threshold. For example, with the input waveform shown in Figure 7B, the first sample surpasses all the thresholds up to e_4, and so actuates several comparators when it is applied. To actuate only 1 comparator, a small pulse known as a reading pulse is superimposed on the input signal to be coded, giving an increment equal to the spacing between levels and therefore surpassing 1 more threshold (e_5 in the example). Thus an output from level r initiated by the reading pulse indicates that the input amplitude lies between e_{r-1} and e_r.

Any desired m-digit code may be obtained from a coding matrix with N inputs and m outputs and connections at the crosspoints so that each level energizes the digit outputs that are to be marks in the corresponding character of the code. The crosspoint connections are made by means of any elements capable of constituting an OR gate, for example, diodes, transistors, or windings on saturable magnetic cores. The comparators may employ semiconductor elements as in this equipment or saturable magnetic cores [1, 7].

This arrangement as it stands is rather lavish with comparators and crosspoints. However, the structure of a unit-distance unit-disparity code permits it to be simplified. Consider any 4-mark character of such a 7-digit code. Each adjacent character may differ from it in only 1 digit, and must contain 3 or 4 marks; it must therefore be a 3-mark character containing 3 out of the 4 marks specified. Also, the two 3-mark characters bracketing any 4-mark character must have different combinations of 3 out of the 4 marks. We may therefore deduce that 3- and 4-mark characters alternate in the code raster, and that each pair of 3-mark characters bracketing any 4-mark character contains between them all of its 4 marks.

These properties may be verified for the code used by reference to Figure 4. They lead to the coder [1] shown in Figure 8. This is similar to the previous arrangement, except that:

A. Comparators are provided only for odd levels, from 1 to $N + 1$ (N is assumed even).

B. The reading pulse amplitude is 3 units (that is 1·5 times the spacing of the comparator threshold).

C. For a $(2n + 1)$-digit code, there are n crosspoint connections in the matrix for each comparator. For example, 3 for a 7-digit code, corresponding to all the 3-mark characters.

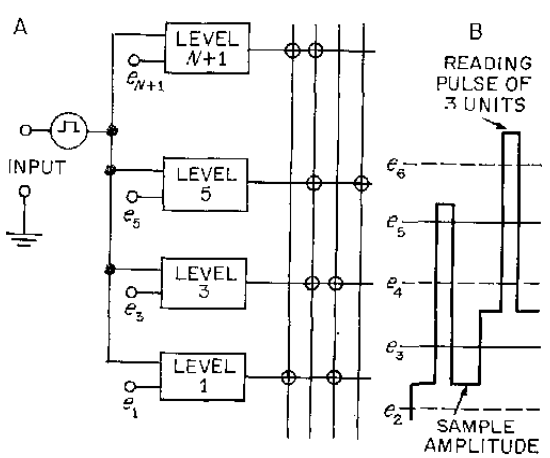


Figure 8—Coder for unit-distance unit-disparity code.

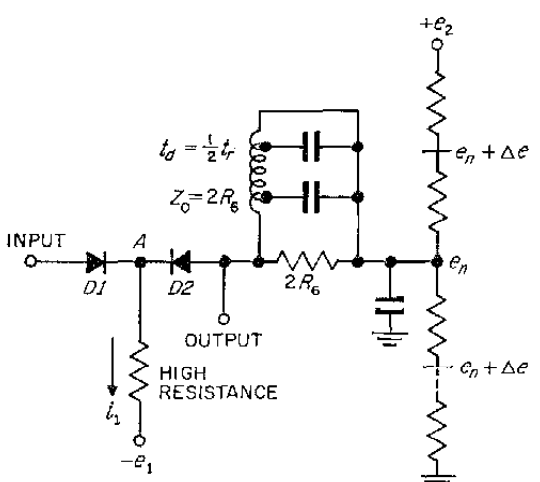


Figure 9—Comparator.

The result of these changes may be seen in Figure 8B. Here the full horizontal lines indicate the actual thresholds e_1, e_3, e_5, et cetera, and broken lines are equally spaced virtual thresholds e_2, e_4, et cetera. Any input lying above a real but below a virtual threshold (for example, the right-hand sample in the figure) causes the reading pulse to actuate 1 comparator and so generate an n-mark character. Any input lying above a virtual but below a real threshold as the left-hand sample in the figure, causes the reading pulse to actuate 2 comparators and so generate the $(n + 1)$-mark character that is bracketed by the 2 n-mark characters pertaining to the comparators.

This specialization has not only reduced the number of comparators to just over half and the number of crosspoints to fewer than half; it also widens the tolerances on threshold levels and reading-pulse amplitude. Any likely variation in these magnitudes leads only to irregular spacing of quantum levels, whereas in the basic arrangement small errors can cause complete failure. In consequence, although the general parallel coder is rather unattractive the specific version for a unit-disparity unit-distance code is entirely practical.

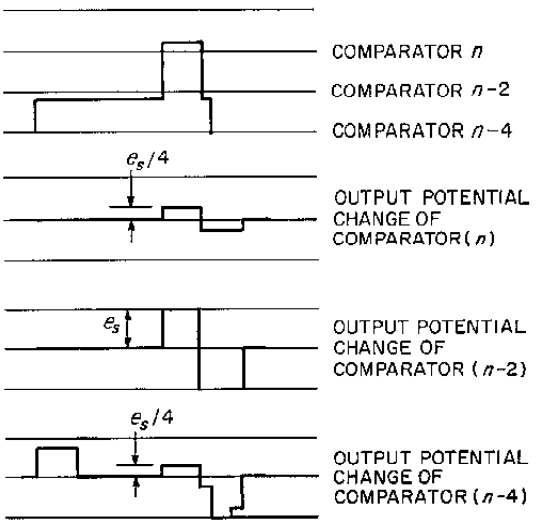


Figure 10—Waveforms of the comparator.

3.3 Practical Coder

A practical comparator is shown in Figure 9. The coder contains 36 such circuits, with their inputs connected in parallel and their reference potentials forming an arithmetic progression with a common difference Δe volts. If the input be sufficiently negative with respect to the reference potential e_n, diode D1 is cut off and a current i_1 flows through D2 and the short-circuited delay line. A sufficiently positive excursion cuts off D2 and diverts the current through D1 and the signal source, so generating a change of potential $e_s = i_1 R_6$ at the output terminal. The full output occurs only if the input excursion includes the whole interval from $(e_n - V_f)$ to $(e_n + V_f + e_s)$, where V_f is the potential drop across a diode passing a forward current i_1. Partial coverage of this comparator window diverts only part of the current and produces a proportional output. Reflection from the short-circuited line cancels any output after a time t_r.

A typical set of waveforms is shown in Figure 10. Only those outputs coincident with the reading pulse are significant; comparators below n-4, which would be energized only by the edges of the pulse-amplitude-modulated signal, are not shown.

Each comparator output drives 1 row of the matrix via an amplifier, which is represented by a current generator in Figure 11. The terminals a are connected to digit gates, which conduct only during the period of the reading pulse. Figure 12 shows the set of gated signals corresponding to the waveforms of Figure 10. A full output e_s from any comparator yields an amplitude e_a at this point; the total amplitude for any digit is proportional to the sum of the outputs of the comparators having a matrix element for that digit.

The fact that the comparators have a variable output as the input traverses the window, rather than a binary output dependent on a sharp threshold, would appear to invalidate the explanation given in Section 3.2. However, by

making a binary decision after passage through the matrix it is possible to arrive at the same result. The decision threshold is set to $e_d/2$ volts (see Figure 12).

The effect of the subsequent binary decision is that only if it is driven through more than half its range can a comparator contribute to the output a mark peculiar to itself. Any input level causes either 1 or 2 comparators to be adequately driven by the reading pulse, so generating the 3- and 4-mark characters respectively. Critical levels occur at the centre of every interval defined by the comparator bias voltages. Thus the example shown yields the 3-mark character 1101000 corresponding to level *n*-2.

3.4 Principle of Decoder

Just as the coder generates digits in parallel, so the decoder requires parallel inputs. The first process towards decoding is therefore to separate the incoming serial digits. The digit distributor generates 192-kilohertz pulse trains in 8 phases, each corresponding to a digit. For each of the 7 speech digits a gate, operated by a pulse train of appropriate phase, separates that digit from the incoming pulse-code-modulated signal and admits it to a magnetic storage element. The 8th digit period, which in the transmission path is used for signalling, is occupied in the speech path by the decoding process. Early in this period, therefore, the stored digits are read out simultaneously and pass after amplification to the parallel decoder.

The decoder, like the coder, is most easily explained (as, indeed, it was devised) by starting with a general method of decoding or translating and then seeing how it can be simplified when we take into account the structural properties of the unit-disparity unit-distance code. Consider an array of detectors, all supplied with the pulse-code-modulated signal, but each responsive to only 1 character from the code (Figure 13). Binary output from these detectors may be used to drive the rows of a ma-

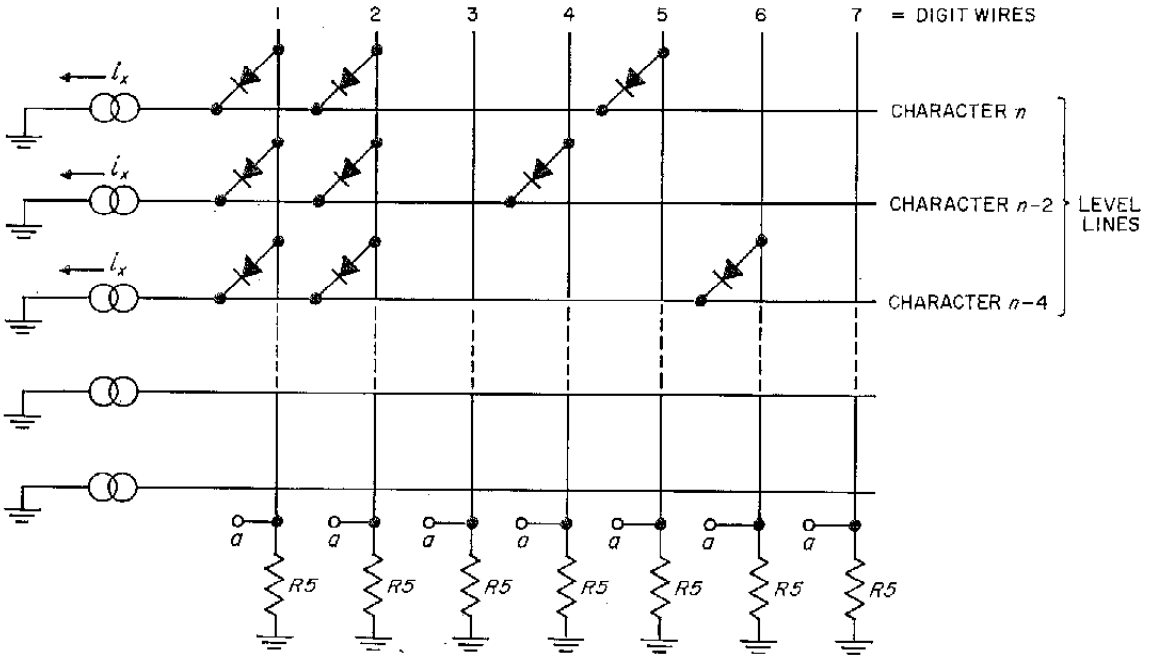


Figure 11—Coding matrix.

trix, similar in principle to a coding matrix (Figure 7), whose crosspoints may be connected to digit columns so as to yield any desired output code, in particular a binary number that is easily decoded.

The simplification possible in a decoder or translator for unit-disparity unit-distance code depends on 2 properties: adjacent characters differ in only 1 digit, and characters with positive and negative disparity alternate so that in a 7-digit code there are both 4- and 3-mark characters respectively.

These properties lead to the decoder or translator [8] shown in Figure 14. There are half as many detectors as in Figure 13, each responsive to either of 2 adjacent characters. If we can make a detector responsive to one character, we can equally well make a detector responsive to a pair of characters by omitting any discrimination of the digit in which they differ and operating solely on the remaining digits, in this case 6, which they have in common. Now the binary-number translation of such adjacent pairs also differs in only 1 digit: the least significant. Consequently, all but 1 of the digits of the translation can be obtained from a reduced matrix of $N/2$ rows and m-1 columns, for a code with N levels and m digits. The least significant digit, the only one differing in the translation, may be found by checking the property in which the characters of a pair differ, their disparity. Positive or negative disparity in the input corresponds to a final 1 or 0 in the output. This requires an auxiliary detector, of about the same degree of complexity as the others.

The construction of the detectors has not so far been specified. Given parallel inputs, the array of detectors can be built as a matrix rather like the output matrix. In this case, specialization to the unit-distance unit-disparity code halves the number of rows in the matrixes and reduces the number of crosspoints in each row by 1·5 on the average.

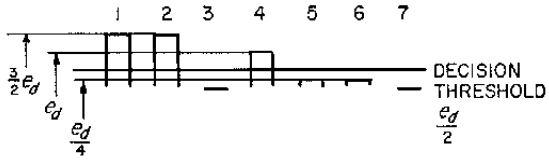


Figure 12—Coding matrix output.

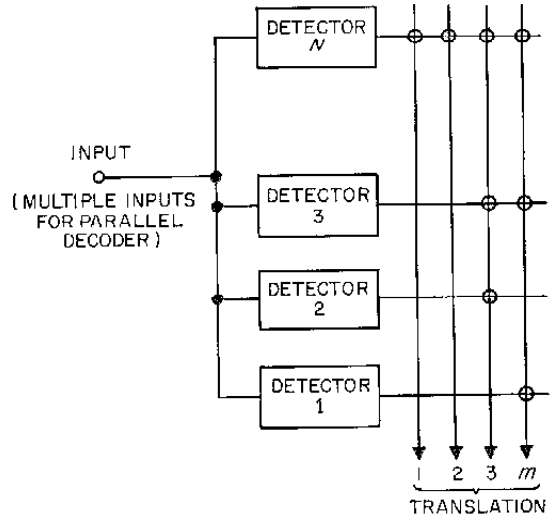


Figure 13—Code translator for arbitrary code.

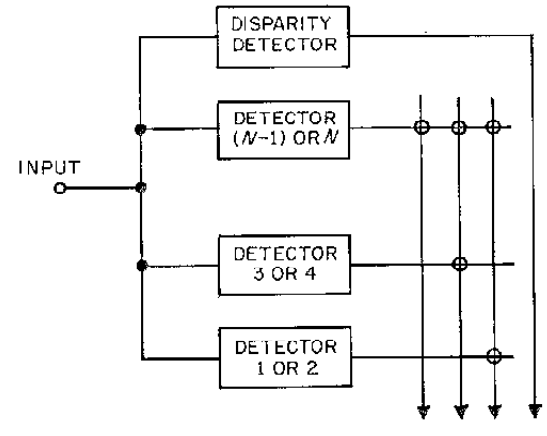


Figure 14—Code translator for unit-distance unit-disparity code.

3.5 Practical Decoder

The code translator employs an array of saturable magnetic cores, each of which is equivalent to a detector plus a line of the matrix on the scheme of Figure 14. The ideal core for this function would have no coercivity or hysteresis: its B–H characteristic would have a sharp threshold, so that an arbitrarily small magnetomotive force in either direction produced saturation flux in that direction. For simplicity we start by explaining the action in terms of ideal elements.

The array of cores comprising the code translator, shown in Figure 15, is traversed by windings carrying digit currents derived from the serial-parallel converter. Standard unit currents flow in the digit windings corresponding to marks of the input code character; no current flows in those corresponding to space. Each core is wound so as to be responsive to a different pair of characters. The method is best explained by taking a specific example. Each core is polarised by current through a bias winding in a sense that we will call reverse.

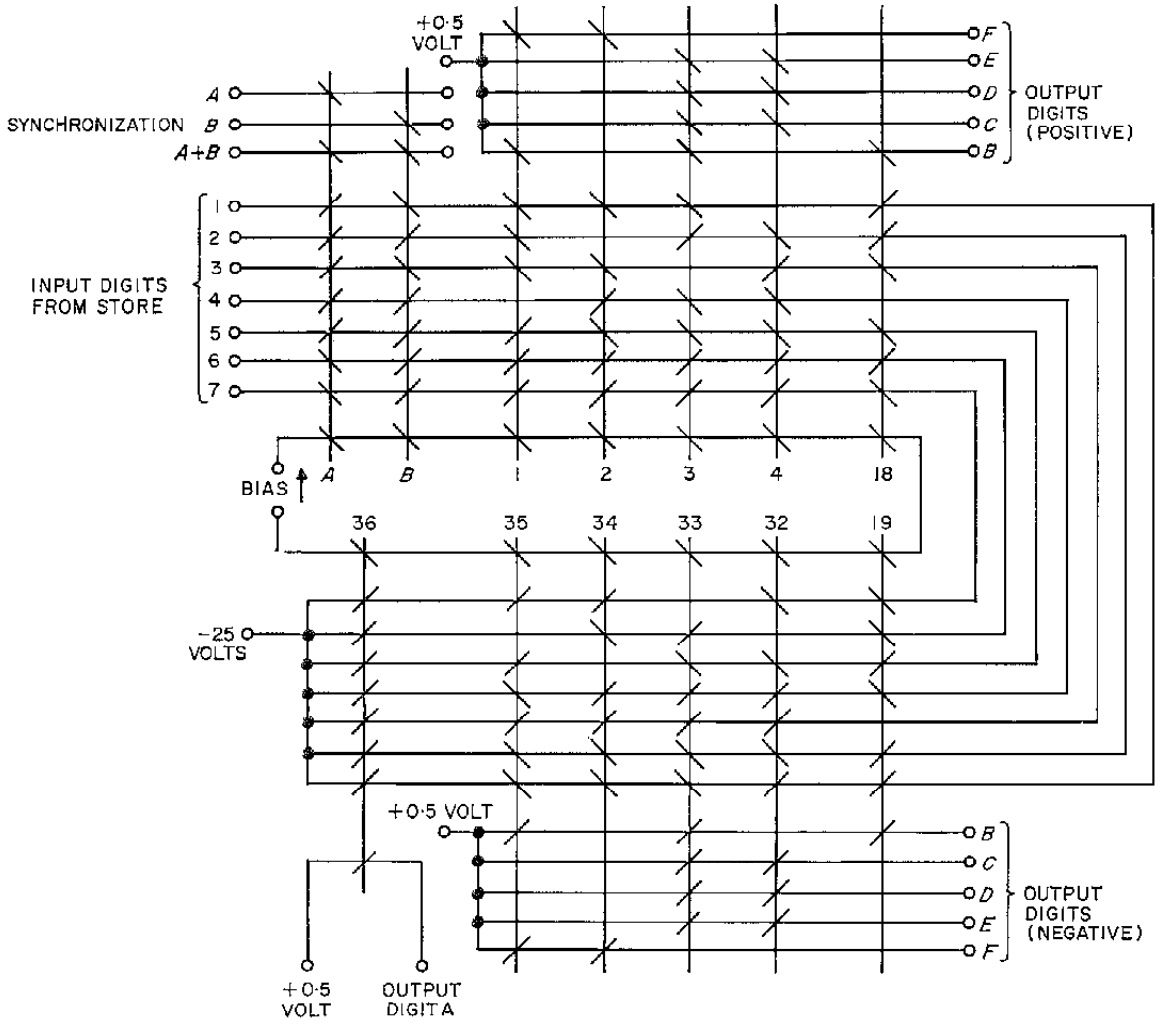


Figure 15—Code translator array.

This bias can be overcome only by unit currents in at least 3 digit windings. Core 1 has windings in the forward sense for digits 5, 6, and 7, and so is energized only if all these digits are marks. It also has windings in the reverse sense for digits 1, 2, and 3: so if a mark occurs in any of these positions it reduces the total magnetomotive force below the threshold and there is no significant change of flux. Digit 4 is not applied to this core and so may be mark or space indifferently. This core therefore responds to either of 2 characters: 0001111 or 0000111, corresponding to levels +33 and +34 respectively. Similarly, any other core responds to a pair of adjacent characters, the higher level being even numbered.

The next step is to generate the digits of a binary number corresponding to the even-numbered level. The cores are divided into 2 groups, dealing with positive and negative levels respectively. Each group has 5 output lines *B, C, D, E, F* allotted to digits with weights 2, 4, 8, 16, 32 respectively; since we are dealing with even numbers, weight 1 is not needed at this stage. The level chosen as example, namely 34, is the sum of 2 and 32; so core 1 requires to be connected to outputs *B* and *F*.

These outputs are pulsed in the negative sense whenever the core, having been switched from reverse to forward magnetization by the input digits, is reset by the bias. The negative pulses trigger digit regenerators that apply standard output pulses through appropriate weighting resistors to a common output terminal.

It remains to distinguish the pairs of adjacent levels. This is done by the supplementary core 36, which is biased and wound so that it responds to any group of 4 marks but ignores a smaller number. The output of this core, digit *A*, is given the weight −1 unit, so that, for example, level 33 is generated as the sum of +34 from the action of core 1 and −1 from the supplementary core.

Two further cores *A* and *B*, driven in the same way as the level cores, respond to the synchronizing characters 11111000 and 01011110 respectively. Individual outputs trigger the frame generator (see Figure 2) while the common output supplies the main synchronizing pulse to the search control unit.

3.6 Compression and Expansion

The number of quanta provided for each polarity, namely 35, is only of the same order as the volume range occurring in the telephone network. That is to say, if quantizing were linear and a loud talker's speech were moderately clipped, a soft talker would just excite 1 quantum of each polarity. A coder with a greater, but still practicable, number of levels would allow him to excite 3 or 4. This is so obviously inadequate that non-linear quantizing is accepted as essential. It is conveniently provided by inserting a common amplitude compressor before the coder and a complementary expander after the decoder.

The ideal compression characteristic would provide constant speech quality over a large part of the required volume range, which implies that it should be roughly logarithmic. Departure from this law, and degradation of quality, are inevitable at the lowest levels where any practical characteristic approaches linearity. The characteristic used is mainly determined by the non-linear devices available, but is near enough to the ideal to be useful; it is shown in Figure 16, with a logarithmic curve (broken line) for comparison. The ratio of the highest to the lowest threshold is raised by the compandor from 31 to about 56 decibels.

The amount of compression that may usefully be employed is limited by the departure of the quiescent level from 0. The gating pedestals and other inaccuracies in the equipment are such that a wider range of levels would not be worthwhile. Two effects of such displacements

should be distinguished. Firstly, suppose a displacement occurring at the transmitter to be conveyed through the system and reproduced at the receiver. The effect is that small signals are quantized more coarsely than they should be; the effect on quantizing noise can be computed [9]. Secondly, suppose that a displacement present at the compressor is not conveyed to the coder and so is not reproduced at the expander. The effect is that compression and expansion are no longer complementary, being centred on different points; so the speech is distorted asymmetrically. Direct coupling of the compressor, expandor, and common amplifiers was not feasible. However, the direct-current component of a complete frame may be practically annulled by reversing the sense of the gating pedestal in alternate channels.

One of the most stringent requirements imposed by the passage of multiplexed pulse-amplitude-modulated signals through common equipment is the provision of adequate low-frequency transmission to ensure satisfactory crosstalk margins. Circuits preceding the compressor need a cut-off frequency of 2 to 3 hertz, whereas in subsequent circuits compandor advantage permits the use of cut-off frequencies ranging up to 8 hertz.

The compandor is of the instantaneous type using germanium diodes as the non-linear element. The basic circuit of the compressor is shown in Figure 17. For a high degree of compression, $R1 >> R2$. The peak signal current in the diodes is chosen to be 5 milliamperes, and they then have a forward resistance that is nominally 75 ohms. To ensure an adequate match of compressor-expander characteristic over the operating temperature range, the diodes are placed in an oven. Since the equipment is designed to work at a maximum temperature of 45 degrees centigrade (40 degrees ambient + 5 degrees bay rise) the oven is set to regulate at 50 ± 1 degrees centigrade.

The output of the decoder is connected directly to the expander-amplifier. The expander is given a characteristic that is complementary to that of the compressor by using a compressor network in a feedback path. A simplified circuit is given in Figure 18, where $R1$ includes the output impedance of the amplifier viewed at the emitter, together with series resistance if necessary.

The problem of low frequency cut-off is simplified by the fact that the decoder output is a dipole pulse. Since the 2 halves, although of equal area, are not exactly the same in shape, they are not equally treated by the expander. The crosstalk advantage is, therefore, not more than about 10 decibels; nevertheless the total crosstalk of the expander-amplifier is very low.

The dipole must, of course, be converted to a single pulse for demodulation. This is done by gating out 1 half with a diode bridge.

3.7 Signalling

There is a great variety of switching plant, and a corresponding variety of signals. Although 2 signalling conditions are often enough, 3 or 4 are sometimes needed. Although a slow signalling channel is often adequate, a relatively fast one is sometimes needed.

Thus a transmission system for universal application must have a signalling channel of very much greater capacity than is needed in the less exigent situations; and in a carrier system, this is not easy to provide. In a pulse-code-modulation system, this difficulty does not arise. The addition of 1 binary digit at the usual speech sampling rate provides adequate capacity whilst occupying only a small fraction of the total bandwidth. The reason for the easy compatibility of signalling and pulse-code-modulation is that the signalling information is essentially digital, and so is well matched to the signalling channel provided by an auxiliary binary digit.

Consideration of signalling problems was forced on the designers of this system by the fact that the first installation was to be on a junction route between rotary exchanges. This type of exchange requires up to 4 signalling conditions, and the total delay of a loop path must be small

enough to permit revertive impulsing. The signalling conditions as present in the rotary exchange are encoded into 2 binary digits designated *A* and *B*, by means of a relay set at each terminal. Since the relay set takes up part of the limited time available, the permissible delay in the pulse-code-modulation equipment is only 2 milliseconds. The 4-kilohertz *A* and *B* digits, obtained by time division of digit 8 as shown in the block diagrams, can achieve this with a reasonable margin and are adequate for any known signalling system.

4. Synchronizing

It is necessary to keep the sampling, decoding, and distributing functions of the receiver in step with the incoming signal with reference to 4 distinct time scales, thus:

A. The digit period. Incoming digits must be sampled by the receive regenerator at, or near, the centre of each digit period.

B. The character period. The 8 digits of each character must be identified so that the first 7 may be decoded and the 8th routed to the signalling highway.

C. The frame period. The 24 channels of each frame must be identified and distributed.

D. The signalling-frame period. Alternate frames, although indifferent to speech samples, must be identified so that *A* and *B* signals may be distributed.

The 1536-kilohertz timing wave that defines the centre of the digit period is extracted from the incoming signal by filtering, as described in section 2. The phase relation of the timing and pulse-code-modulation waveforms issuing from the receive regenerator is determinate, so time scale *A* may be taken for granted in this context.

For the purpose of synchronizing, time scales *B* and *C* are amalgamated. Distinctive characters transmitted in channel 24 are recognized by 2 cores in the code translator (Figure 15). Their occurrence during the period allotted by the receive distributor to channel 24 confirms the receiver in its normal or LOCK condition. Thus both character and channel synchronism are necessary to maintain LOCK; failure of either initiates a SEARCH process.

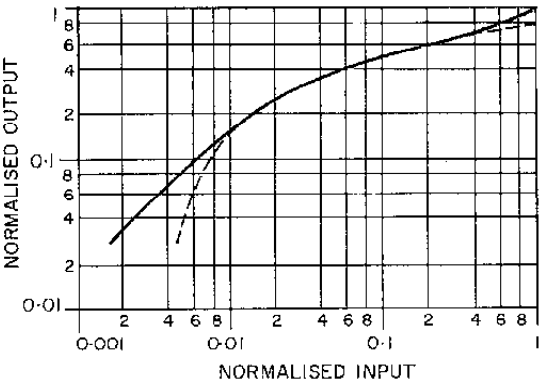


Figure 16—Compressor characteristic.

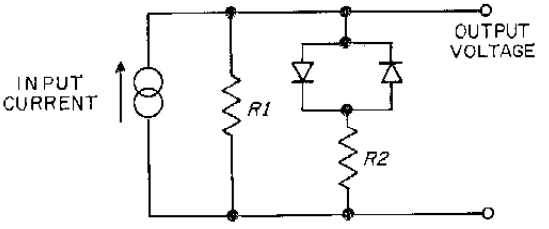


Figure 17—Compressor.

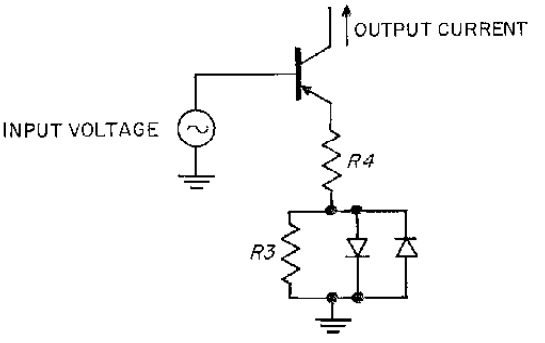


Figure 18—Expander-amplifier.

By means of a gate shown in Figure 2, the search control unit admits a group of 8 clock pulses to the digit distributor and then blanks a group of 3. This change of count enables the incoming pulse train to be scanned for synchronizing characters; every possible time position is examined during a period of 1·375 milliseconds equal to 11 frames. Meanwhile, the channel distributor is stopped. Recognition of a synchronizing character resets the control to LOCK; the distributor clock supplies are restored and the channel distributor is started at the correct time by a rephase signal.

Noise in the transmission path is just as likely to impair the synchronizing channel as any other. Searching, if it were to be initiated in response to a single imperfect synchronizing character, would be more commonly spurious than due to genuine loss of synchronism. Moreover, interruption of synchronism would disturb every channel more radically than would any error in a speech sample; so this spurious searching would make a major contribution to the system noise. For this reason, the receiver is dislodged from a sustained LOCK condition only after a succession of 3 absent synchronizing characters, although during the course of a search it responds to their individual presence or absence as rapidly as possible.

Should a search be halted fortuitously by a single simulation of a synchronizing character, it is taken up again at a corresponding point in the next frame. These properties are obtained by the use of a differential counter that registers successes and failures in identifying synchronizing characters.

These provisions [10] yield an adequately high speed of search while reducing to insignificance the hazards of falsely simulated synchronizing characters or obliterated true ones. Synchronism is restored after any momentary loss in less than 1·5 milliseconds. This enables calls in progress to be retained despite the interruption.

The 4-kilohertz framing generator that determines time scale D is directly triggered by the A and B synchronizing pulses educed respectively by the 2 synchronizing characters transmitted in alternate frames. This means that it has no protection against obliteration of synchronizing characters by noise, though there is some guard against simulation. Single errors are, however, of little importance to the signalling, because the operate or release time of even the high-speed relays is much greater than the sampling interval.

The choice of synchronizing characters is a compromise between conflicting factors. It is not possible to choose truly distinctive characters without wasting much of the informational capacity of the normal coded signals, so some risk of simulation, for example, by the end of 1 channel character plus the beginning of another, must be accepted. The probability of simulation may be reduced by some departure from the low-disparity property of the normal characters, but this obviously cannot be taken to an extreme without risk of faulty regeneration. Also, the 2 characters must not be easily confused. The final choice was 11111000 and 01011110; it will be noted that these fall within the normal disparity range but adopt an unusually close grouping of marks.

5. Conclusions

The field experiment was conducted over a junction route in Madrid on 18 lines provided by the Compañia Telefonica Nacional de España. The route joined the Norte and Delicias exchanges, on the north and south sides of the city respectively; it passed through Gran Via, the largest and busiest exchange in the city, but was not terminated there. Over the whole length of 6·2 kilometres, pairs of 0·64-millimetre diameter were available in each of 2 cables, laid in the same duct for about 4·4 kilometres but diverging for the remainder.

The pairs were part of either 900- or 600-pair cable, the remainder of which was carrying normal traffic. Repeater sites with a nominal spacing of 1 kilometre were chosen from the existing manholes; the actual section lengths ranged

from 968 to 1134 metres, with an average of 1040 metres.

Repeaters developed and manufactured by Standard Telecommunication Laboratories and Standard Electrica (Madrid) were installed by the latter on 6 pairs, 3 in each direction. The 2 directions of transmission were in separate cables. The 6 pairs chosen included those that on preliminary measurements had shown the worst crosstalk.

Terminals developed and manufactured by Standard Telecommunication Laboratories and Standard Telephones and Cables were installed at the Norte and Delicias exchanges. Each installation comprised 2 terminals, 1 fully equipped and the other having only 3 sets of channel panels. Relay sets designed and manufactured by Le Matériel Téléphonique were interposed between the rotary exchanges and the signalling circuits of the pulse-code-modulation equipment. Rotary switching plant, comprising a line finder and first selector, was installed at Norte so that calls could be originated in isolation from the main exchange. There was no special switching plant at Delicias; the incoming pulse-code-modulation junctions were connected to the main exchange in the usual way. The studies made included:

A. Measurements of transmission loss, crosstalk, and noise on the cables.

B. Checks of various parts of the equipment in isolation.

C. Objective measurements on the pulse-code-modulation system in isolation.

D. Recording of conversational speech, random sentences, and articulation material passed through the pulse-code-modulation system in isolation, for subjective evaluation.

E. Operational check with pulse-code-modulation junctions linking the exchanges.

This trial has shown the feasibility of pulse-code-modulation operation on local area cables. The more important properties observed are listed below.

A. Speech quality is good at optimum volume, tolerable over a range of perhaps 20 to 25 decibels, and intelligible over a much-wider range, perhaps 40 to 45 decibels. At optimum volume, it remains tolerable after transmission through 3 or 4 systems in tandem (simulated by recording and retransmission) and intelligible up to 9 or 10.

B. In a transmission system for universal application, moderate extension of volume range would be needed.

C. Signalling capabilities are adequate for the needs of rotary exchanges, which are the most exigent type.

D. Digit rates of 1·5 megahertz are feasible on installed paper-insulated multi-pair cables; the 1-kilometre repeater spacing could be increased.

E. Some form of pulse-code-modulated signal resistant to low-frequency cut-off is essential, and the unit-disparity code is a practicable means of achieving this.

F. Rapid recovery of synchronism enables calls to be held through momentary interruptions.

G. This equipment was used as a vehicle to try out a number of techniques, of which the more useful could form a basis for further development.

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K. W. Cattermole was born in Forest Gate, Essex, England, in 1923. He received the B.Sc. degree with first-class honours from London University in 1952.

From 1942 to 1947, he served with the Royal Electrical and Mechanical Engineers. After a year with the British Broadcasting Corporation as a transmitter engineer he joined Standard Telecommunication Laboratories in 1948. He has worked successively on the development of microwave radio links, time-division-switching, and pulse-code-modulation systems. He is currently in charge of the pulse-code-modulation laboratory in the line transmission division.

Mr. Cattermole is the author of a book on "Transistor Circuits," published in 1959.

D. R. Barber was born in Drumheller, Alberta, Canada, in 1927. He received the B.Sc. (Eng.) degree from the University of London in 1947. During the university vacations, he worked at Standard Telephones and Cables from 1944 to 1947.

From 1947 to 1950 he served with the Royal Electrical and Mechanical Engineers and subsequently returned to Canada. He joined Canadian General Electric where he was engaged in the development of microwave links.

In 1955, he joined Standard Telecommunication Laboratories, England, where he has been engaged in the development of repeater amplifiers and pulse-code-modulation inter-exchange telephony systems.

Mr. Barber is an Associate Member of the Institution of Electrical Engineers.

J. C. Price was born in Highbury, London, England, on 3 June 1921. He studied at the Sir John Cass College and in 1950 received the B.Sc. degree from the University of London.

He served in the Royal Air Force during the second world war, working on radar. From 1950 to 1960, he was a part-time science lecturer at Southgate Evening Institute. He did research work on fuses for Belling Lee from 1950 to 1953.

He joined Standard Telecommunication Laboratories in 1953 and has since worked on transistor pulse and logic circuits and on pulse-code modulation systems.

E. J. E. Smith was born in Poona, India, on 10 February 1930. He became a graduate member of the Institution of Electrical Engineers in 1956.

From 1948 to 1953, he served with the Royal Electrical and Mechanical Engineers. He then joined Standard Telephones and Cables and has been engaged in the development of pilot regulating equipment for coaxial and open-wire systems, rural area systems, and more recently in the design of local area systems. Mr. Smith is now in charge of the local area and open-wire systems development section.

SEL-Taschenbuch
(Reference Data Book for Communication Engineers)

A new reference data book, edited by H. Sarkowski, has been published in German by Standard Elektrik Lorenz. It presents brief descriptions, tables, and diagrams under the following chapter headings.

1. General Data for Communication Engineers
2. Components and Parts Used in Telecommunication
3. Electroacoustics
4. Teleprinting
5. Information Processing
6. Telephone Switching
7. Cables and Open-Wire Lines
8. Transmission
9. Radio
10. Power-Supply Equipment
11. Mathematical Section

The book is 11 by 15 centimeters (4.4 by 6 inches) and contains 312 pages. It is available from Verlag Berliner Union, Stuttgart, Urbanstrasse 12–14, at DM 11.40 per copy.

United Kingdom–Faroes–Iceland (Scotice) Submarine-Cable Telephone System

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1. Introduction

The requirements of the International Civil Aviation Organisation for improved communication facilities among its stations in the United Kingdom, Iceland, Greenland, and Newfoundland added to the growing commercial demand for improved telephone and telegraph facilities between the United Kingdom, Faroe Islands, and Iceland resulted in the Scotice [1] project. Figure 1 illustrates how the loop formed by Cantat, Scotice, and Icecan [2] with London and Cornerbrook as the eastern and western terminals respectively may be switched rapidly to provide high-quality communication between all stations of the International Civil Aviation Organization even in the presence of a fault in one of the submarine-cable links.

Scotice provides telephone and telegraph service over a band of 72 kilocycles per second in each direction. The telegraph load may occupy up to ⅓ of this band. The system uses a single 0·46-inch (11·7-millimetre) diameter polythene coaxial cable. The South section of 290 nautical miles (537 kilometres) contains 10 submerged repeaters and links Gairloch in Scotland with Torshavn in Faroe Islands. The North section of 408 miles (756 kilometres) contains 15 submerged repeaters and 1 submerged equaliser between Torshavn and Vestmannaeyjar in Iceland. The 24 traffic channels, each 3 kilocycles per second wide, are transmitted over the bands from 24 to 96 and from 120 to 192 kilocycles per second in the *A–B* and *B–A* directions respectively. In addition, 3 engineering circuits of 2-kilocycle-per-second width each are similarly provided over the bands from 18 to 24 and from 192 to 198 kilocycles per second. Supervisory facilities of the frequency-doubler type monitor the loop transmission over the submerged repeaters and occupy the bands from 96 to 100 and from 192 to 200 kilocycles per second.

The North section, with terminal *A* at Torshavn and terminal *B* at Vestmannaeyjar, was laid in November 1961 by *H.M.T.S. Alert* of the British Post Office and the South section, with terminal *A* at Torshavn and *B* at Gairloch, was laid during the following month by *C.S. John W. Mackay* of the Commercial Cable Company.

2. System Design

2.1 Line Equipment

The North and South sections must meet the requirements of the Comité Consultatif International Télégraphique et Téléphonique under such adverse temperature conditions as are encountered when the cable is installed with allowance for the addition of 2 nautical miles (3·7 kilometres) of repair cable and with ⅓ of the telephone channels loaded with 24-channel voice-frequency telegraph signals.

The mean temperature of the sea bottom is 4·2 degrees centigrade with extremes of 2·2 to 6·2 degrees for the North section. For the South section, these values are 7 and from 4·5 to 9·5 degrees centigrade respectively.

2.1.1 *Submarine Cable*

The number of telephone channels to be accommodated being relatively small, a correspondingly thin coaxial cable is used. The centre conductor consists of a single copper wire of 0·128-inch (3·3-millimetre) diameter. The diameter over the polythene insulation is 0·46 inch (11·7 millimetres) and the return conductor consists of 6 copper tapes with 1 binder tape. This construction gives a characteristic impedance of 52 ohms and an attenuation at 200 kilocycles per second of 2·36 decibels per nautical mile (0·147 neper per kilometre) at 15·5 degrees centigrade. At the

mean temperature of 4·2 degrees centigrade, the attenuation is 2·3 decibels per nautical mile (0·143 neper per kilometre) at 196 kilocycles per second. The type of armour used depends on the depth. In the deep water of both sections, type *D* cable is used, while in the South section each armour wire is separately taped. Heavier armour is used in shallower water and screened cable is installed for shore ends and for land cable. The temperature coefficient of attenuation is +0·17 per cent per degree centigrade, the pressure coefficient is +0·05 per cent per 100 fathoms, and the laying factor is −0·4 per cent.

2.1.2 *Submerged Repeaters*

The requirements for the repeaters are derived from the line characteristics and for the following conditions.

A. Changes in cable attenuation due to seasonal temperature variations at the sea bottom.

B. Assumed increased length of cable of 2 nautical miles (3·7 kilometres) due to future repairs.

C. Equalisation errors.

D. Overload point and basic noise provided by the valves employed.

An amplifier gain of 65 decibels at 196 kilocycles per second can be justified. Since the band-width of the system is relatively narrow, 40 decibels of over-all feedback can be applied to the amplifier and the intermodulation requirement as calculated by the method of Brockbank and Wass [6] is achieved. The nominal repeater levels are given in Table 1.

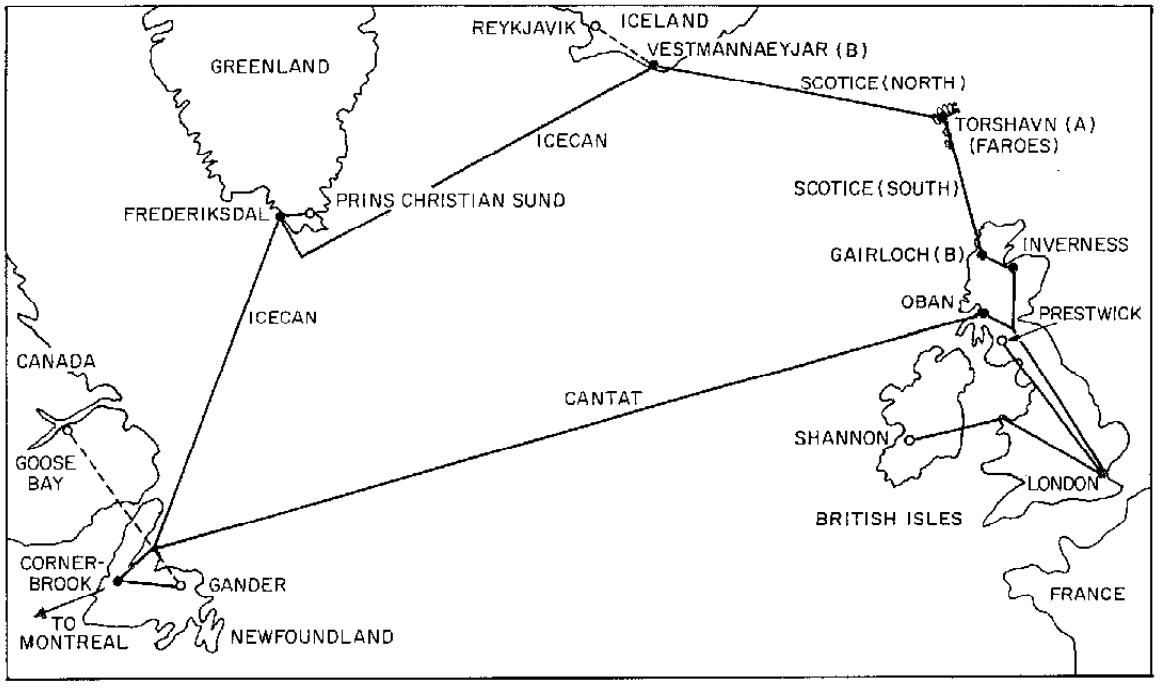


Figure 1—The Cantat, Scotice, and Icecan loop linking the air-traffic-control stations of the International Civil Aviation Organization, which are shown by circles. The broken line is a radio link from Gander to Goose Bay.

The schematic diagram of the submerged repeaters shown in Figure 2 illustrates that a common amplifier is used for both directions of transmission. The stability margin requirements of 40 decibels under normal conditions, 20 decibels under cable fault conditions in the transmission bands, and 20 decibels outside the transmission bands are achieved by the use of directional filters, frequency-selective bridge networks, and power separating filters.

The repeater amplifier providing a rising gain-frequency characteristic consists of a pair of 3-stage amplifiers connected in parallel between common input and output networks. Approximately 40 decibels of over-all voltage and current feedback is applied to both amplifiers

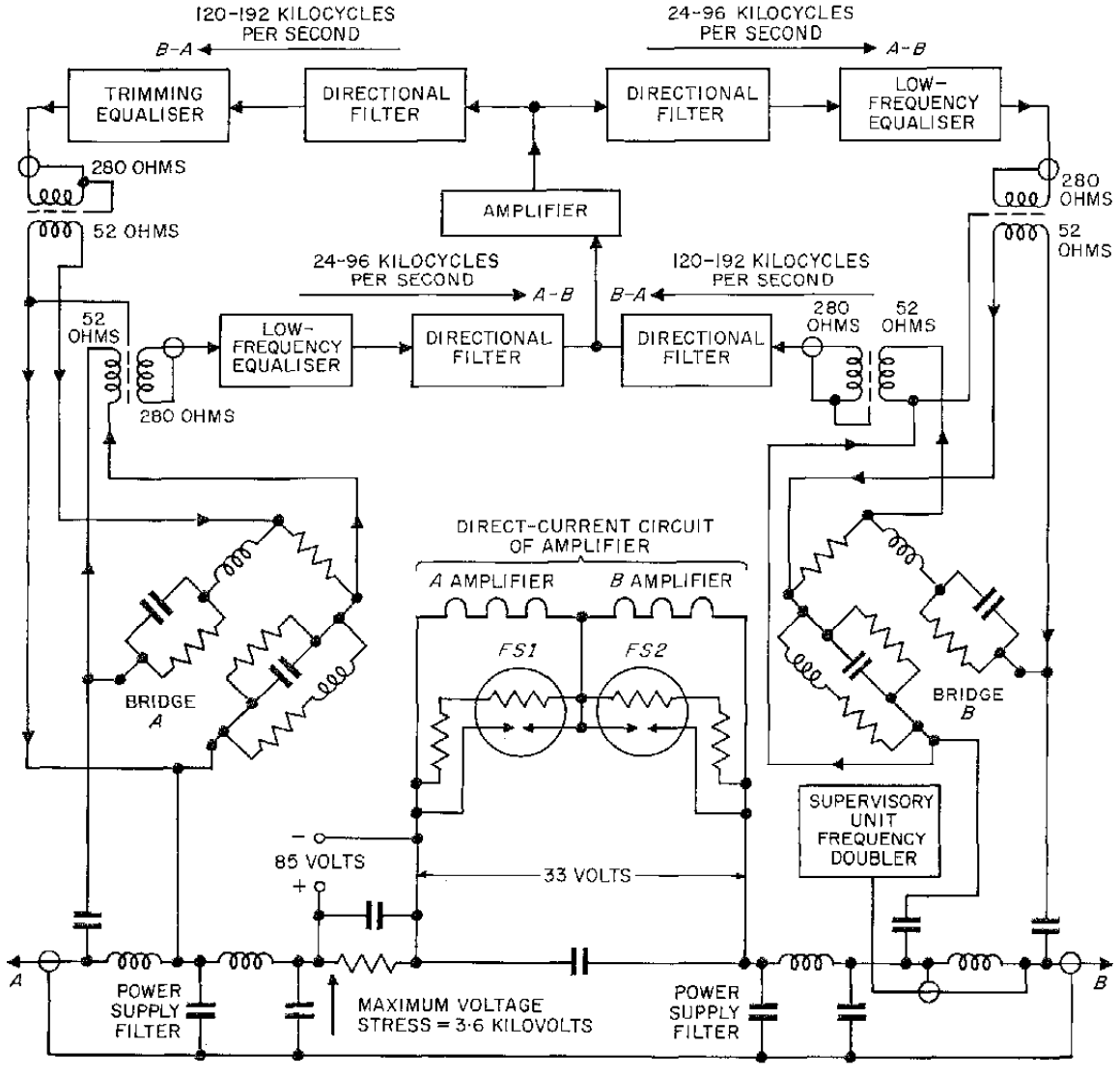


Figure 2—Diagram of submerged repeater. A direct current of 316 milliamperes flowing in the center conductor of the cable provides power to operate the amplifiers.

via a common bridged-T beta network. By this means, a fault within either amplifier changes the over-all gain by less than 0·1 decibel. The valves are special submerged-repeater types similar to the British Post Office [4] type *6P12*. The *5A/162* is used for the first 2 stages and *5A/166* for the final stage. The gain–frequency characteristic is controlled principally by the input transformer, output transformer, and beta network. Accordingly, these components are manufactured to extra-close tolerances and are hermetically sealed in dry nitrogen to insure stability.

In the event of a valve-heater failure, thermal short-circuiting fuse *FS1* or *FS2* operates to make a soldered joint across the fuse contacts, thus maintaining the power supply to the remaining amplifier.

The supervisory unit [7] contains an input crystal band-pass filter, a frequency doubler, and an output band-pass filter. Connection is made to the repeater via high-value protective resistors such that even in the event of a complete failure within the unit, the bridging loss is less than 0·15 decibel at all frequencies. In the North and South sections, each repeater has a different supervisory crystal-filtered mid-band frequency between 96 and 100 kilocycles per second. Thus, by selecting the appropriate frequency, loop transmission measurements can be made from the *A* terminal (Torshavn) to the *B* end of each repeater where the doubled frequency in the band from 192 to 200 kilocycles per second returns to the *A* terminal.

TABLE 1

NOMINAL REPEATER RELATIVE LEVELS

Frequency in Kilocycles per second	Input in Decibels	Output in Decibels
24	−39·2	−16·9
50	−43·7	−12·6
100	−51·4	− 7·6
120	−58·0	− 9·9
150	−59·6	− 5·8
196	−62·9	− 1·4

The electrical units are mounted in copper cans that, like the paper capacitor cans and connecting wire, are gold-plated to give an inert finish. The cans and the amplifier chassis are mounted on Perspex bars to withstand the maximum voltage stress of 3·6 kilovolts, then, after a period of dry-nitrogen flushing, the electric unit is sealed in a brass cylinder 7·75 inches (19·7 centimetres) in diameter. The internal unit is contained in a demountable pressure housing. The seal between the high-tensile-steel housing and the bulkheads is provided by a special arrangement of pre-pressurised *O* rings. The actual cable connections that pass through the bulkheads are tested at a pressure of 5 tons per square inch (787 kilogrammes per square centimetre) for a month before they are actually used. External to the bulkheads are the jointing chambers and cable-armour-clamping arrangements. The demountable housing is also used for the submerged equaliser and is described in more detail elsewhere [5].

2.1.3 *Submerged Equalisers*

The submerged equaliser used in the North section is described in detail elsewhere [5]. Briefly, the internal unit consists of a chassis containing power-separating and directional filters in the sealed unit. Space is provided on the chassis for mounting equaliser sections and cable simulators. The design of the equaliser may take place while the cable is being laid so as to permit compensation for changes in cable attenuation due to unpredictable laying effects. Kits of hermetically sealed capacitor units, resistor units, inductor units, and cable-simulator units are used to build up the circuit configurations required. After assembly of the necessary components on the chassis, this is sealed into its demountable housing and can be laid in the same way as a repeater. The designing, building, and testing of an equaliser, including adjusting the length of the cable and making a sea-cable joint, may take up to 16 hours. Figure 3 shows the equalisation error over the North section of Scotice.

2.2 Terminal Equipment

2.2.1 *Channelling Equipment*

High-efficiency 3-kilocycle-per-second channelling equipment is used to assemble 16 channels into a full group (60–108 kilocycles per second) and a further 8 channels into a half group (60–84 kilocycles per second).

2.2.2 *Main Transmission Path*

The full group is treated as group 5 and the half group as part of group 4 of a standard coaxial-line system. The subsequent modulation and demodulation processes are shown in the simplified block schematic diagrams of Figure 4, which shows one of the two *A* terminals at Torshavn, Faroe Islands, and in Figure 5, which shows the *B* terminal at Gairloch, Scotland, or Vestmannaeyjar, Iceland. The frequency allocation is shown in Figure 6.

2.2.3 *Speaker Equipment*

Both local and omnibus telephone and telegraph engineering channels are required. The band available in each direction is 6 kilocycles per second, some of which is, however, already permanently occupied by an edge pilot and some of which must also be available for repeater supervisory purposes. Three channels of 2-kilocycles per second each are provided, 2 of which are used for speech circuits and a portion of the third channel is available for at least 2 voice-frequency telegraph circuits. These 3 channels are translated in 2 stages to the 54-60-kilocycle-per-second band and combined at the group distribution frame with group 5 in the *A–B* direction. They are translated to the 84-90-kilocycle-per-second band and combined with group 4 in the *B–A* direction.

2.2.4 *Edge Pilots*

An edge pilot is transmitted at 23 kilocycles per second in the *A–B* direction and at 193 kilocycles per second in the *B–A* direction. Each pilot is monitored and recorded at the receiving end. A further edge pilot of 120 kilocycles per second can, however, be obtained in the *B–A* direction by injecting 60 kilocycles per second into group 5 at the *B* terminal.

2.2.5 *Carrier Supply*

This is largely conventional and based on a crystal-controlled 124-kilocycle-per-second master oscillator divided down to 4 kilocycles per second. The special frequency of 576 kilocycles per second is picked off as a harmonic of 4 kilocycles per second, and 672 is derived by modulation of 576 and 96 kilocycles per second. Since failure of either of these special-frequency supplies or of either of the two group carriers of 564 and 612 kilocycles per second would result in loss of all traffic or a considerable part of it, all carrier supply amplifiers are duplicated and change-over from standby to working is automatic in the event of failure.

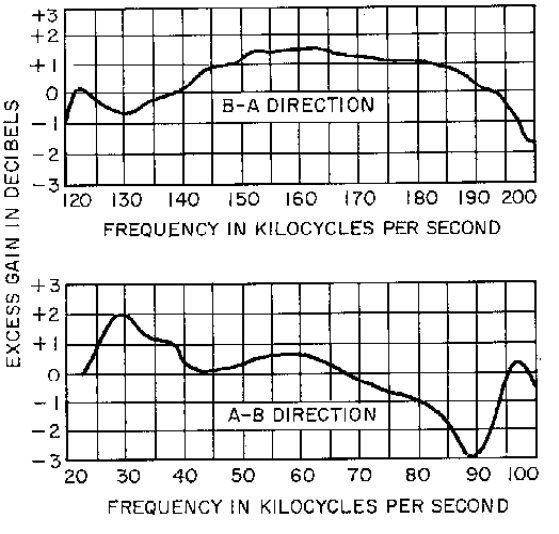


Figure 3—Equalisation error for the North section of Scotice from cable head to cable head. There are 15 repeaters and 1 equaliser in this 408-nautical mile (576-kilometre) run of cable.

2.2.6 *Frequency-Comparison Pilots*

It is desirable that the British standard of 60 kilocycles per second be accurately transmitted from Gairloch to Torshavn and Vestmannaeyjar and then be available (still as a frequency standard) for onward transmission to Reykjavik and over Icecan towards Greenland. To avoid the errors that modulation by incorrectly synchronised carrier supplies could introduce, the 60-kilocycle-per-second standard is doubled at Gairloch and transmitted to Torshavn as 120 kilocycles per second. At Torshavn, it is divided down to 60 kilocycles per second for local frequency-comparison purposes, then further divided down and transmitted as 20 kilocycles per second along to

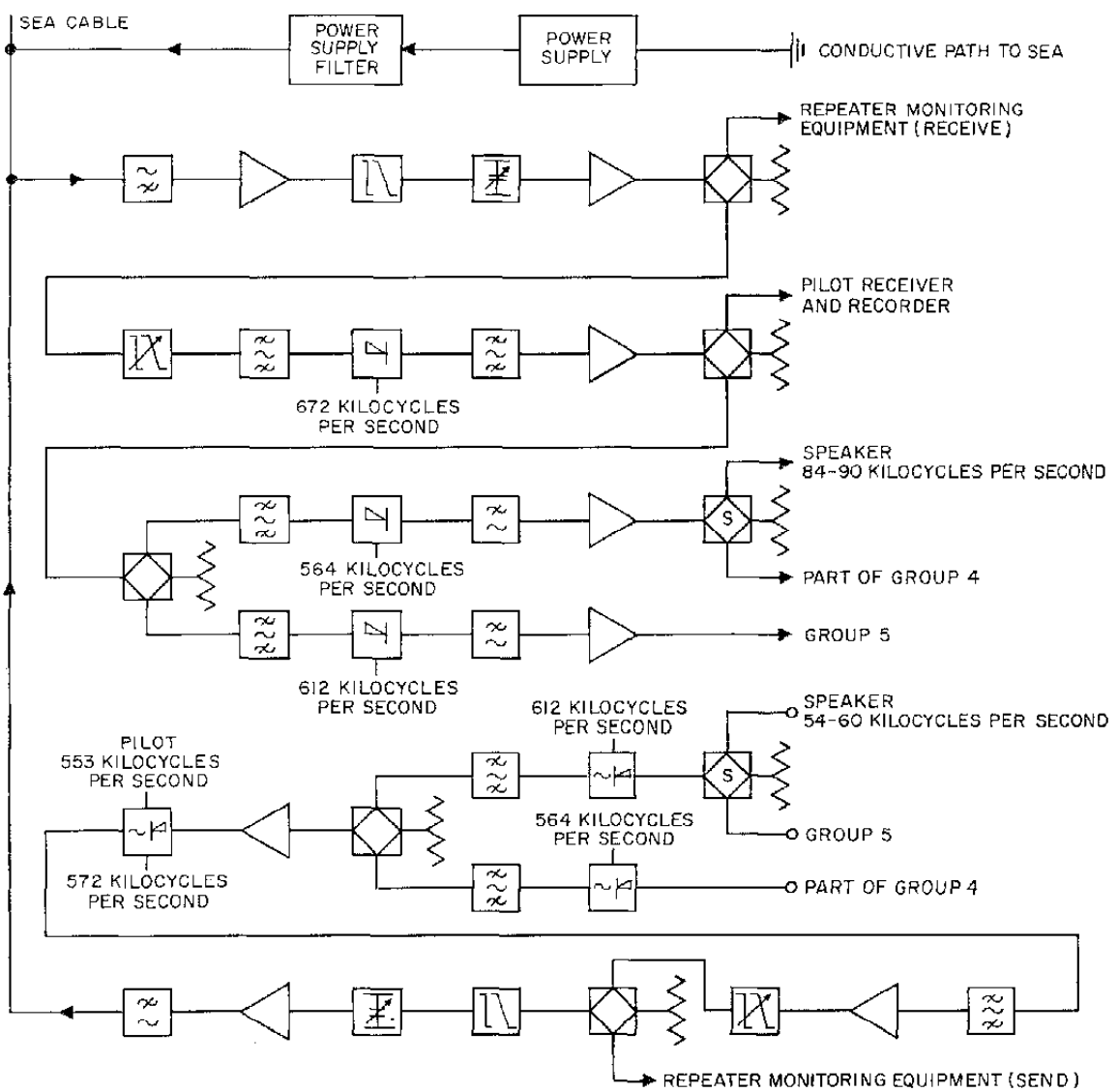


Figure 4—Simplified diagram of main transmission path at *A* terminal.

Vestmannaeyjar, where it is multiplied back to 60 kilocycles per second.

2.2.7 *Through Circuits*

Channelling equipment for all 24 channels is provided at Gairloch and Vestmannaeyjar, but at Torshavn only the half group is translated to audio frequencies, the full group being a "through" group to Gairloch and Vestmannaeyjar.

2.2.8 *Power Supply*

The power equipment is required to supply a constant direct current of 316 milliamperes that varies by not more than ±1 per cent. It is designed for double-end feeding and each terminal normally supplies half of the total system voltage. Either terminal can automatically supply the whole voltage in the event of failure at the other terminal. The equipment is connected in series with the centre conductor of the cable and return is via the sea. The total system voltages of the North and South sections are 2440 and 1670 volts respectively. Working and standby power-feeding units are provided at both Gairloch and Vestmannaeyjar. At Torshavn, in addition to the working units for each of the two terminals, there is a standby unit that can be switched into service in either the North or South section.

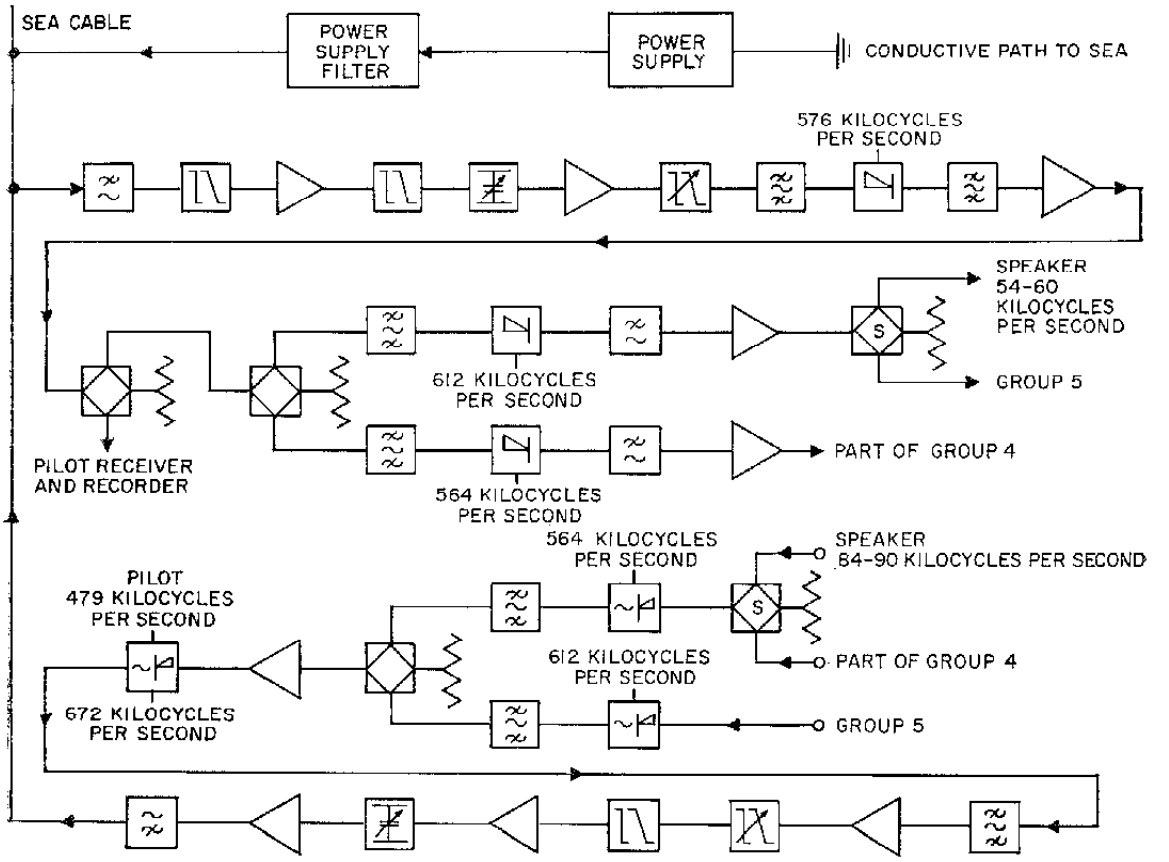


Figure 5—Simplified diagram of main transmission path at *B* terminal.

3. Manufacture of Submerged Repeaters and Equalisers

The submerged repeaters and equalisers were manufactured at North Woolwich in special fully air-conditioned shops often called "dairies." This manufacturing organisation was set up originally to produce the submerged repeaters for the Newfoundland–Nova Scotia link of the first transatlantic telephone cable [3]. Since that time, considerable experience has been gained and the facilities have been expanded and improved. To provide a thoroughly reliable repeater, considerable care must be taken with the mechanical and electrical design of its components. This care must start with the basic raw materials and continue throughout all the manufacturing processes. Most of the components are therefore manufactured within the company and those that are obtained outside are rigidly held to special specifications. The strict inspection pattern followed results in a large rejection rate for both raw materials and finished components, but this high standard of inspection in the early stages pays dividends in providing extremely uniform and stable units. The rejection of a complete repeater unit would be extremely serious in that it would indicate a weakness in the earlier inspection procedure.

A complete record of all components making up a repeater, of all processes, and of all inspection test results are filed individually for each repeater.

The submerged equalisers and their hermetically sealed components are manufactured in the same shops and to the same standard as the submerged repeaters.

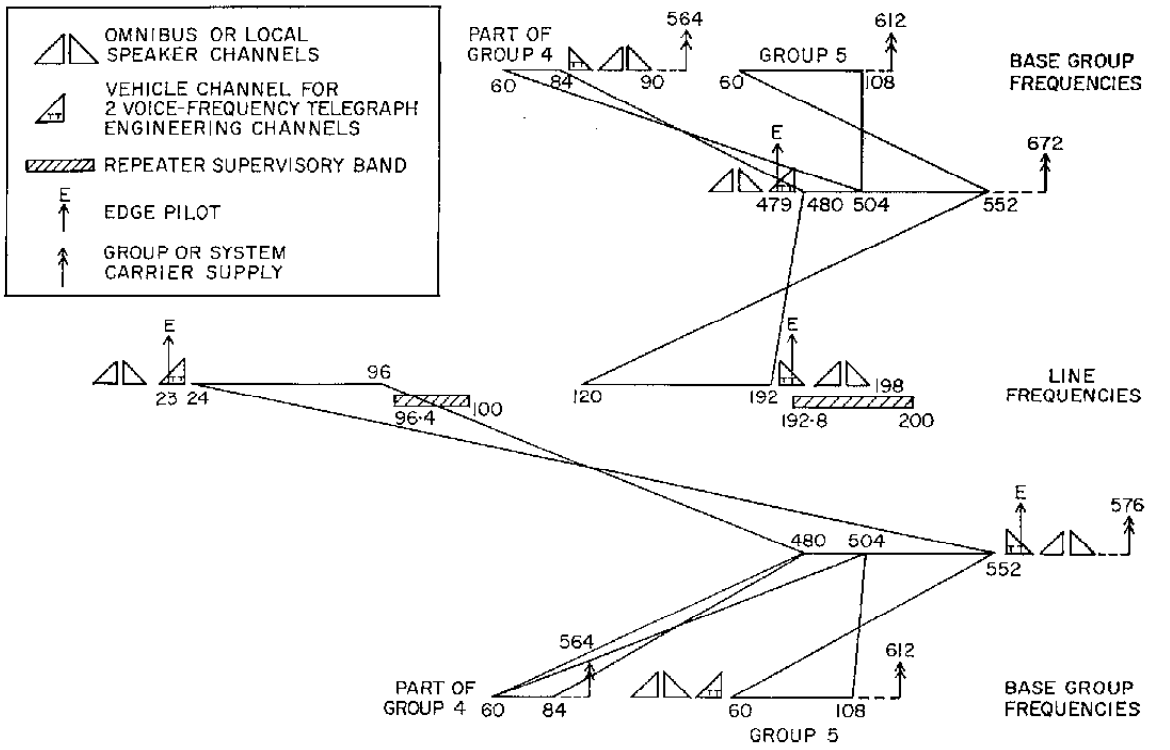


Figure 6—Frequency allocation.

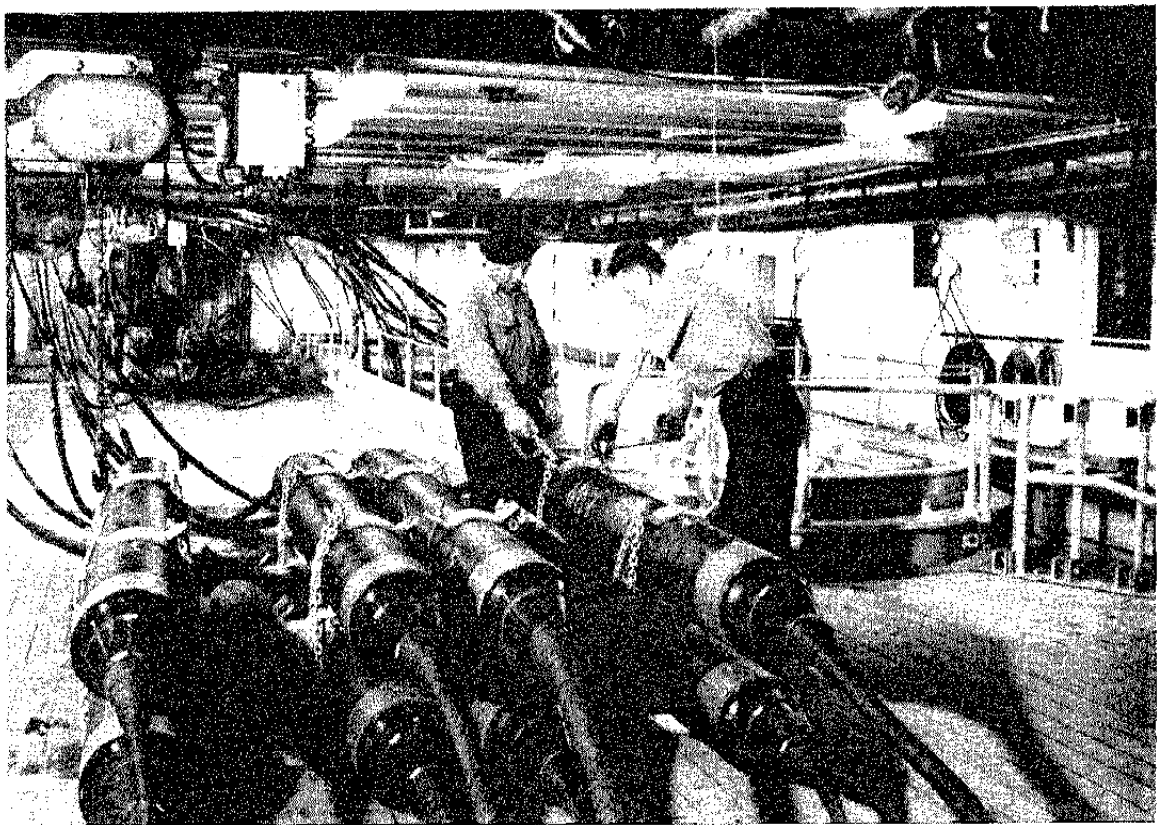


Figure 7—Centrecastle of *H.M.T.S. Alert* showing 8 repeaters in the foreground and 7 in the background for the North section of Scotice. They are connected to the cable that is stored in the cable tanks of the ship.

4. Cable Loading and Laying

The cableships were loaded direct from the submarine-cable factory at Southampton. The cable sections are approximately 27 nautical miles (50 kilometres) long and were linked by haulage hoses. While the cable loading continued at about 3 nautical miles (5·6 kilometres) per hour at Southampton, repeaters arrived at the rate of approximately three every two days from North Woolwich. Testing and jointing continued according to a coordinated schedule commencing with the last section to be laid. Both North and South sections were jointed to form 2 blocks, the first section of the second to be laid being longer than nominal to allow for length adjustment during the laying. Since the North section was longer, an equaliser was included between the 2 blocks. The following description applies to the laying of the North section.

H.M.T.S. Alert picked up the Faroes shore-end cable, which had previously been laid by a smaller cableship, made the initial splice, slipped the cable from bow to stern, and began to lay towards Iceland. Laying from the stern has the advantage of speed and the ability to work in heavy seas. The disadvantage is that there is a minimum speed, depending on weather, at which the ship can be controlled. During the weather experienced throughout



Figure 8—A repeater about to be laid.

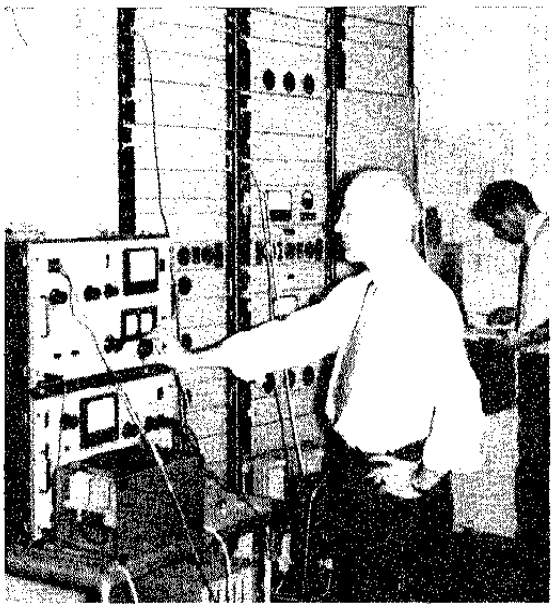


Figure 9. Testing the North section at Vestmannaeyjar.

this lay, in which the wind was usually of force 7 to 8, the minimum speed was about 2·5 to 3 knots. Thus, as the equaliser position was approached, a count-down of 16 hours was initiated, on the assumption that the ship's speed would be approaching 6 knots. The power unit aboard ship energised the repeaters continuously except during jointing operations. Thus, throughout the lay, communication with Faroes was maintained over the cable using carrier speaker equipment. As each repeater was laid, the shore station transmitted the supervisory signal to the last repeater in the block and observed the returned signal. Throughout the laying of the first block, measurements of insertion loss between Faroes and the ship as a function of frequency were made twice while laying the cable from one repeater to the next. This provided a careful check on the change in attenuation as the cable reached sea-bottom conditions. The insertion loss at each frequency was plotted against nautical miles of cable laid and by extrapolating the curves to the equaliser position a comparison with the target values for that point provided a forecast of equalisation error. The forecast equalisation requirement indicated that two bridged-T sections should be used to increase the loss at low frequencies. Since these sections had little loss at high frequencies, 1 nautical mile (1·9 kilometres) of cable simulation was added to the equalisers in each direction of transmission so that the lightly armoured deep-sea cable would not come too near the Iceland shore. The equaliser work including jointing both blocks together took less than 12 hours. The change in cable attenuation predicted for this period was very close to the actual change.

The equalisation procedure is described in detail elsewhere [5]. Throughout the remainder of the lay, insertion loss and supervisory measurements were made once per repeater section.

The over-all insertion-loss measurements made between Iceland and Faroes agreed very closely with the final measurements aboard ship, and at high frequencies with factory measurements of cable and repeaters. At low frequencies, the laying factor was slightly greater than predicted. Further over-all measurements made with equipment of the type shown in Figure 9 indicated that the design objectives were met

and the circuits between London and Reykjavik were opened to traffic on 22 January 1962.

5. References

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6. Acknowledgments

The successful conclusion of this project is the result of the work of a team too numerous to name individually. The authors express their thanks to colleagues in the Lines and Submarine Branches of the British Post Office for assistance with the cable laying and over-all testing under difficult conditions. Thanks also to their colleagues in, and to the management of, Standard Telephones and Cables Limited.

7. Appendix—Performance of Typical Submerged Repeater

Insulation resistance = 2×10^4 megohms.

Direct-current resistance at 22 degrees centigrade.

Current in Milliamperes	Direct-Current Resistance in Ohms
5	334·0
20	334·0
50	337·3
100	351·6

Voltage drop at 316 milliamperes = 120.

Insertion gain in a 52-ohm circuit.

Frequency in Kilocycles per Second	Gain in Decibels	Frequency in Kilocycles per Second	Gain in Decibels
24	22·33	120	48·18
30	24·59	130	50·17
40	27·93	140	52·07
50	31·02	150	53·95
60	33·99	160	55·71
70	36·71	170	57·41
80	39·26	180	59·03
90	41·57	190	60·60
98	43·53	196	61·47
100	43·98	200	62·02

Harmonic distortion in decibels referred to 1 milliwatt.

60-kilocycle-per-second fundamental level at *B* terminal +11

120-kilocycle-per-second-harmonic level at *A* terminal −63

180-kilocycle-per-second third-harmonic level at *A* terminal −70

Impedance, return loss in decibels against 52 ohms.

Frequency in Kilocycles per Second	*A* Terminal	*B* Terminal
20	35·5	29·0
40	24·0	27·0
70	19·0	12·0
100	12·5	8·0
120	11·5	11·0
140	14·0	15·5
170	19·5	26·0
200	21·5	22·5

Noise in decibels referred to 1 milliwatt.

A terminal between 120 and 200 kilocycles per second = −61·0.

B terminal between 20 and 100 kilocycles per second = −75·5.

Supervisory.

Fundamental Level at *B* Terminal in Decibels Referred to 1 Milliwatt	Second-Harmonic Level at *A* Terminal in Decibels Referred to 1 Milliwatt
−12	−21·6
− 2	− 8·4
+ 8	+ 2·7

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In 1946 he joined the transmission division of Standard Telephones and Cables at North Woolwich. Until 1950, he was engaged in the design of audio systems. He then transferred to submerged repeater systems, working mainly on electrical circuit design.

Mr. Archibald is an Associate Member of the Institution of Electrical Engineers.

Shipboard-Adjustable Submerged Equaliser

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1. Introduction

The continual advances in the design of submarine cables and submerged repeaters have been responsible for the rapid increase in the number of submarine telephone cables that are now in operation or are planned for the near future [1, 2].

The laying of the first transatlantic telephone cable known as *TAT-1*, in 1956 [3], was so successful in providing high-grade telephone channels between the continents that it was decided within 3 years of its opening to traffic to provide another similar system, *TAT-2*, for an additional 36 circuits between the United States and Europe.

For good-quality telephone circuits, consideration must be given to the variation of signal level along the line, including equalisation errors caused by failure of repeater gain and cable loss to compensate for each other within the design margins. The equalisation errors have to be limited to, say, 3 decibels, if the use of a "mop-up" equaliser is to be avoided. Thus, on a transoceanic submarine cable for telephony, system equalisation is of paramount importance. Unfortunately, the exact properties of the equaliser depend on the characteristics of the cable and repeaters as laid and can therefore be accomplished only during the laying of the cable, possibly in difficult weather conditions.

Early oceanic submarine cable systems using the Bell Telephone System flexible unidirectional repeaters had a number of submerged equalisers having different fixed equalisation characteristics [4]. From a stock carried on board the cableship, equalisers were selected and spliced into the cable during the laying according to the system requirements. With the relatively low line frequencies involved, this method proved to be quite successful.

However, using a single cable with bidirectional rigid repeaters and working at higher line frequencies, successful operation of a long submarine cable system will depend largely on the ability to equalise the cumulative errors along the system at those intervals where the magnitude of the error approaches the system design limit.

Because of the rather involved nature of the system errors, the equaliser should be capable of being adjusted aboard the cableship and then laid in deep water in the same manner as the repeaters.

The pressure-resisting housing developed for our submerged repeaters and first used for the Denmark–Norway cable [5], being of the so-called demountable type, offered the possibility of adjusting an equaliser and enclosing it in a pressure housing on shipboard.

The first submerged equaliser using such a housing was adjusted during the laying of the Anglo–Swedish cable [6, 7]. Several equalisers of a similar type have also been laid during the installation of the Cantat system [8].

A detailed description of the type of adjustable submerged equaliser initially developed for the recently completed Scotice and *Med-1* projects is given in this paper. The pressure housing, being the key to the problem of providing a submerged equaliser that is capable of being adjusted on board ship, is dealt with in some detail.

2. Design Objectives and Sources of Equalisation Errors

The main design objective was an equaliser capable of being adjusted aboard the cableship with relative ease and speed and that provides a loss characteristic to compensate for any possible system errors. It is therefore worthwhile to examine the sources and nature of the system equalisation errors to appreciate the design of these networks. The errors may be either predictable or unpredictable.

2.1 Predictable Errors

Predictable errors are those that are known before assembly of the system aboard the cable-ship; they are as follows:—

A. Design errors E_d are the discrepancies between the gain of the laboratory model repeater and the expected cable loss. They are normally small and known at an early stage of project engineering.

B. Cable errors E_c cover the difference between the predicted cable attenuation characteristic and the actual cable loss figures as measured on the manufactured cable lengths in the factory. Their magnitude depends on whether the cable is of a new design or of a design that is well established and within the percentage loss limits given in the cable specification.

If the cable employed for the project is of a type already used on other systems, then E_c is small. It may be significant for a cable of new design for which only estimated loss figures were available when the final repeater gain-frequency characteristic had to be accepted.

C. Repeater errors E_r involve the difference between the repeater gain-frequency characteristics of the laboratory model and of the production units. Normally E_r is small. The over-all summation of these 3 errors gives the expected equalisation error E_e.

The frequency characteristic of E_e can be predicted to a fair degree of accuracy when the values for each of the contributing factors are known. The expected error can be compensated by a fixed equaliser unit, which as a practical matter must be designed at a late stage in the production of the cable and repeaters.

2.2 Unpredictable Errors

All changes in the cable characteristic during the laying are unpredictable errors. Such changes are truly unknown as to their extent and character, but they may be of two categories:—

A. Cable-length error E_l is a change in the cable attenuation that is approximately proportional to the cable attenuation constant; this type of error can be normally corrected by provision of facilities for cable-length adjustment at the end of each ocean block.

B. Cable-shape error E_s covers those cases of change in the frequency-loss characteristic of the cable during laying. Normally, significant changes in the shape of the attenuation curve are not expected, except perhaps, at the lower end of the system frequency spectrum.

3. Submerged Equaliser Requirements

3.1 Electrical Requirements

The examination of the possible error sources indicates that the equaliser should have a fixed equaliser unit to compensate for E_e and an adjustable equaliser to compensate for the unpredicted errors and for any residue of errors from inadequate equalisation of E_e by the fixed unit.

In some cases it may be difficult to design and manufacture a fixed equaliser in time for the laying operation; the adjustable unit will then have to provide for all the resultant errors, $E = E_d + E_c + E_r + E_l + E_s$. The equaliser must also provide the facility to insert cable simulator sections as and when required. In addition, power separating filters are needed to separate the transmitted signals from the power sent through the cable to operate the repeater amplifiers.

Directional filters separate the two directions of transmission, which simplifies the design of the adjustable equaliser sections as each direction of transmission can be considered independently.

3.2 Mechanical Requirements

From the mechanical standpoint, the design should aim to meet a number of requirements.

A. The pressure housing should, of course, be capable of being opened and subsequently closed and sealed on shipboard.

B. There should be some easy means of checking the sealing of the housing.

C. The completed equaliser should have the same expected reliability as that of the submerged repeaters in the system.

D. The equaliser should be capable of being completed on any cableship and not only on one specially equipped for this purpose.

E. The number of components for the adjustable units should be as few as possible consistent with flexibility and adequate coverage of foreseeable equalisation requirements.

F. All parts and components should be robust.

G. The design should provide for simple and speedy assembly and wiring on shipboard to reduce delay during the laying of the system.

It was appreciated that some special equipment would be needed to meet requirement *A*, which would have to be portable to comply with *D*.

4. Description of Equaliser

4.1 Mechanical Construction

From the mechanical aspect, the submerged equaliser can be divided into 3 main parts; the pressure housing, the apparatus unit that carries the fixed elements and also accepts the adjustable elements, and the parts and components for making up the adjustable elements.

4.1.1 *Pressure Housing*

The pressure housing, originally designed for submerged repeaters, was developed basically over a number of years in conjunction with Messrs. Vickers Armstrong [9]. Refinements were added as a result of further development carried out in conjunction with the British Post Office. The housing is suitable for working pressures up to 4 tons per square inch (630 kilograms per square centimetre); this is equivalent to a maximum depth of water of 3400 fathoms (6218 metres).

In principle, the housing consists of a high-tensile-steel tube or casing with its ends closed by high-tensile-steel bulkheads sealed into the casing by means of *O* seals. No elaborate equipment is required on shipboard to carry out the sealing operation.

Figure 1 is a sectional view of part of one end of a pressure housing, showing the sealing of the bulkhead. The space to the left from the coil of slack cable needed during assembly to the outer face of the bulkhead is open to the sea. The bulkhead, which is a close fit in the end of the casing, has two main *O* seals carried in grooves round its periphery. The space between the seals is connected via small radial holes with 2 cylinders bored into the bulkhead from the outer face; both of these cylinders are sealed against the entry of water by close-fitting pistons equipped with *O* seals. In the lower of the two cylinders, a small hole extends through the bulkhead to permit filling the inside of the housing with dry nitrogen, and the inner end of the cylinder is also sealed by a piston and *O* seal. In addition, the small hole is also sealed off by a soft copper disk attached to the end of the piston, which is forced into intimate contact with the metal at the bottom of the cylinder by a load of 3·5 tons (3600 kilogrammes); this load is applied via the piston by tightening its clamping screw with a pre-determined torque. The purpose of the disk is to provide a diffusion seal against the possible entry of water vapour.

The enclosed space formed by the 2 main *O* seals and the sealed-off cylinders is filled with petroleum jelly, and the jelly compressed to a pressure of 4 tons per square inch (630 kilogrammes per square centimetre) at an ambient temperature of 22 degrees centigrade by means of the pressurising piston and its pressure screw in the lower of the two cylinders shown in Figure 1. Thus, in addition to the sealing afforded by the *O* seals, there is a barrier of petroleum jelly that is held at a pressure in excess of that of the surrounding water.

In very-deep water, where the pressure is high and the temperature low, possibly of the order of 5 degrees centigrade, then due to the pressure/temperature coefficient of the petroleum-jelly system, which is approximately $\frac{1}{16}$ ton per square inch per degree centigrade (9·8 kilograms per square centimetre per degree centigrade) the pressure in the system would tend to fall to a value that would possibly be less than that of the water. To maintain the

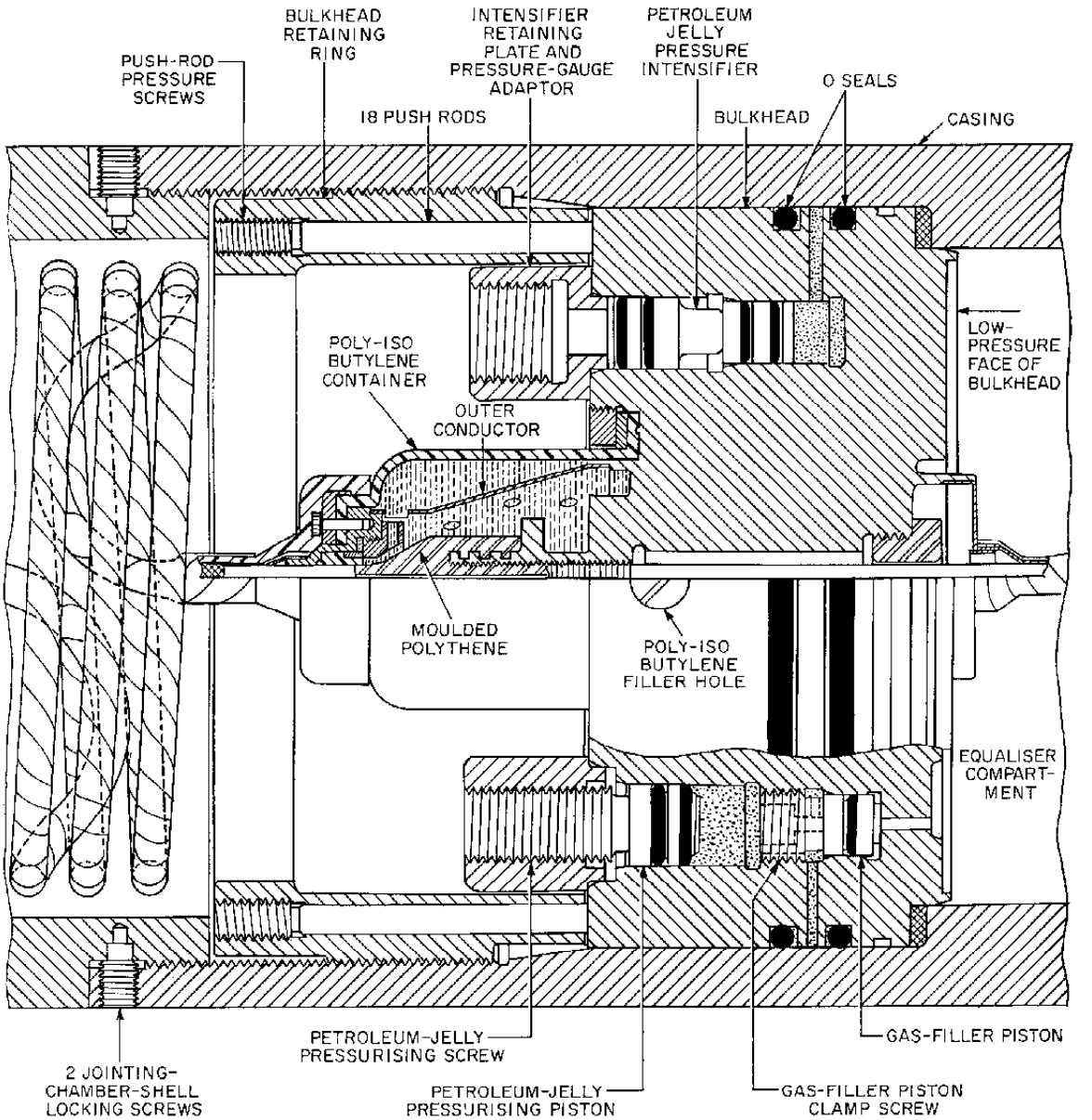


Figure 1—Section of equaliser casing containing bulkhead through which the coaxial submarine cable passes from the region of full sea pressure to the apparatus compartment at atmospheric pressure.

pressure in the petroleum-jelly system always greater than that of the water, the other cylinder in the bulkhead is provided with a pressure-intensifying device; this device will automatically raise the pressure in the system to a value that is proportionately greater than the pressure of the surrounding water. This is achieved by having the outer portion of the cylinder bored to a larger diameter than the inner portion and providing each portion with a separate piston, the inner piston being in contact with the contained petroleum jelly and the outer piston being exposed to the sea. As the ratio of the piston diameters is 1·36:1, the pressure in the system will therefore always be at least 1·36 times that of the external water pressure. The available travel of the pistons is sufficient to provide a maximum increase in pressure of the petroleum jelly of about 4·5 tons per square inch (709 kilogrammes per square centimetre).

All *O* seals are moulded of special hard-grade synthetic rubber that is particularly resistant to oil and sea water. The hardness of the seals is such that, when used with the small working clearances provided, there is no detrimental extrusion of the seals, even when they are subjected to pressures of the order of 7 tons per square inch (1102 kilogrammes per square centimetre).

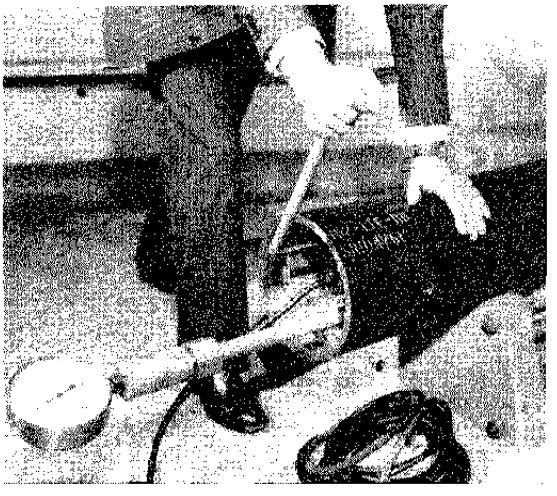


Figure 2—Measuring device showing pressure of petroleum-jelly seal. Neither pressure nor jelly is lost in connecting the device or in making a measurement so measurements may be repeated at any time without penalty.

The plate attached to the face of the bulkhead to retain the intensifier pistons is provided with a large threaded hole that is extended by a smaller plain hole through the plate to give access to the face of the outer piston. This hole, apart from giving sea water easy access to the piston, also provides for applying a pressure-measuring device to the bulkhead to determine the pressure of the petroleum jelly during the compressing operation. This measurement may be repeated at any time to check the sealing of the bulkhead. Figure 2 shows the device in operation. It should be appreciated that this pressure measuring device can be applied to, or removed from, a bulkhead without loss of pressure or petroleum jelly. This design, incorporating a pressurising system and a sensitive means of measuring the pressure, provides a ready method of checking the effectiveness of the seal. With the pressure-measuring device applied to the bulkhead during the pressurising operation, any defects in sealing will be detected by a fall in pressure. This test is quite sensitive because, as the volume of the bulkhead pressure system is very small, of the order of 2 or 3 cubic inches (33 to 49 cubic centimetres), any slight leak will show up as a drop in pressure in quite a short time. The procedure on shipboard, therefore, is to leave the measuring device in position on the bulkhead for a short time after pressurising to check sealing.

Between the seating on the inner face of the bulkhead and the bulkhead-retaining shoulder in the casing, a lead gasket is fitted as a diffusion seal against the possible entry of water vapour that may have passed the *O* seals. In fairly deep water where the pressure is of the order of, say, 1 ton per square inch (157 kilogrammes per square centimetre), it is evident that the bulkhead will be forced against the lead gasket with sufficient pressure to "flow" the lead so that it makes intimate contact with the

adjacent steel faces and provides an adequate seal. As the areas of the bulkhead and gasket are approximately 63 and 13 square inches (406 and 84 square centimetres) respectively, a water pressure of 1 ton per square inch (157 kilogrammes per square centimetre) will produce a stress on the lead of nearly 5 tons per square inch (787 kilogrammes per square centimetre). However, to cater for comparatively shallow water where the pressure would be insufficient to stress the gasket beyond the flow limit of lead, arrangements are provided to pre-stress the gasket with a load of approximately 45 tons (45 720 kilogrammes). With this load on the gasket, the stress will be approximately 3·5 tons per square inch (551 kilogrammes per square centimetre). (The flow stress of lead is approximately 2·5 tons per square inch (394 kilogrammes per square centimetre).) To achieve the initial pre-stressing of the lead, 18 equispaced push-rods, sliding in close-fitting holes in the bulkhead retaining ring, bear at their inner ends against the face of the bulkhead and are provided at their outer ends with pressure (thrust) screws. These screws are tightened with a torque that gives a thrust of 2·5 tons (2540 kilogrammes) on each rod. The compressed push-rods constitute very strong springs that provide a small amount of "follow-up" without excessive loss of pressure in the event of a subsequent slight extrusion of lead between the bulkhead and the casing. The risk of extrusion, however, is very slight, as a test on a gasket fitted in a housing having normal clearances between the bulkhead and the casing has shown that, even with the bulkhead loaded with over 300 tons (305 megagrammes), apart from an initial slight extrusion that took place during the first few hours, there was no further extrusion over the period of test which lasted for 135 days.

From the foregoing it will be seen that the sealing of the bulkhead into the casing has 3 lines of defence; namely, 2 *O* seals and the lead gasket. In addition there is a barrier of petroleum jelly held at a pressure that is always greater than that of the surrounding sea water. Thus, even in the event of the failure of an *O* seal that would completely release the petroleum-jelly pressure, there will still be an *O* seal and the lead gasket. Furthermore, it will be appreciated that the sealing operation can be carried out and checked using only a few simple tools.

The cable entry through the bulkhead is formed by a castellated stem machined integral with the bulkhead, through which passes a small-diameter polythene-insulated cable, the cable forming short tails on each side of the bulkhead. A polythene moulding is formed on the cable at the outer end which unites homogeneously with the polythene of the cable and also extends over the castellated portion of the stem. The contraction of the moulding on to the stem will provide a seal against the entry of water, but as an additional safeguard the stem is processed before moulding to ensure an effective bond between the polythene and the metal of the stem. In deep water, the contraction of the moulding on to the stem will be enhanced by the external pressure of the water. The bore of the stem is threaded to form circumferential serrations into which the softened polythene of the cable is forced during the moulding operation; this provides an effective keying that prevents extrusion of the cable through the bulkhead, even with hydraulic pressure of 5 tons per square inch (787 kilogrammes per square centimetre). In fact every production seal is subjected to a water pressure test at 5 tons per square inch for a period of a month; during the test a daily check is made of humidity in an airtight container sealed over the low-pressure inner face of the bulkhead. Whilst under pressure all seals are also electrically checked for insulation resistance and ionisation-inception voltage, and are subjected to a proof test of 40 kilovolts of direct current for 1 minute. This type of seal was developed by the British Post Office and is used on all their deep-water submerged repeaters.

The cable tail on the high-pressure side of the bulkhead is provided with a tinned-copper braid outer conductor that is connected to the

bulkhead via a perforated metal cup. Completely enclosing the moulded cable seal and the outer conductor is a moulded synthetic-rubber bell that is vacuum filled with a viscous insulating fluid (poly-iso-butylene) [10]. Thus, water is prevented from contacting the junction of dissimilar metals formed by the outer-conductor cup and the bulkhead; this latter would, in the presence of water, cause electrolytic corrosion of the metal. When subjected to a hydrostatic pressure the bell will, of course, collapse slightly until the pressure in the insulating fluid is comparable with that of the surrounding water.

The cable tail on the low-pressure side of the bulkhead is jointed to a corresponding tail from the apparatus unit, which is contained within the pressure housing. The outer-conductor braid of the apparatus unit is extended over the bulkhead tail and connected to the bulkhead via a thin metal cup; this operation is carried out during the assembly of the submerged equaliser in the factory.

When dispatched from the factory, the tail cable on the high-pressure side of the bulkhead is jointed to a co-axial test lead; during installation operations on board the cableship, this lead is removed, and the tail is jointed to a corresponding tail on the pre-terminated end of the sea cable; after jointing the outer conductors of the tails, the whole of the tail cables are protected by layers of self-adhesive polythene and polyvinyl-chloride tapes.

The method of anchoring the sea cable to the end of the housing, shown in Figure 3, follows the commonly accepted practice whereby the armour wires of the cable are formed back over a tubular coned member. This armour cone is threaded onto the cable and the wires are clamped between it and a surrounding anchor plate having a hole complementary in shape to the outside of the armour cone. The armour wires, where they are formed back round the nose of the armour cone, are individually located in radial slots in an anti-rotation plate to prevent any possible torsional movement of the sea cable relative to the pressure housing during the laying operation.

By adopting a removable collet made in two halves and fitted into the anchor plate to form the outer coned clamping member, the ends of the sea-cable sections may be terminated before loading onto the cableship. This terminating operation includes jointing-on of the small-diameter cable tail by means of a tapered moulded joint, jointing of the outer-conductor braid of the tail to the outer-conductor tapes of the sea cable, forming back of the armour wires over the armour cone, and laying-up and serving with spun yarn the turned-back armour wires. This greatly reduces the amount of work to be performed on the ship when installing the equaliser.

It is interesting to note that this pressure housing has been adopted by the British Post Office for shipboard-adjustable submerged equalisers on trans-oceanic cable systems such as Cantat and Compac.

4.1.2 *Apparatus Unit*

The apparatus unit within the pressure housing consists basically of 2 parts, a factory-sealed capsule that houses the fixed electric units of the equaliser, and an open chassis, flexibly attached to one end of the capsule, into which the adjustable units are mounted on shipboard. The complete apparatus unit, prior to assembly in the pressure housing, is shown in Figure 4. When assembled in the housing the free end of the chassis is rigidly secured to the removable bulkhead; thus on board ship, when the bulkhead is withdrawn from the casing the chassis is also withdrawn just clear of the casing to enable the adjustable units to be fitted. Polytetrafluoroethylene rubbing pads, which are a close fit in the casing, are secured to the end of the chassis and to the free end of the capsule, to allow the apparatus unit to slide freely in the casing during the withdrawal operation.

The sealed capsule houses the power-separating filters, directional filters, and a fixed equaliser, if one is needed, and is filled with dry nitrogen. Cable tails, which are subsequently jointed to the cable tails of the respective bulkheads, are led into the capsule via seals in the end plates. In this manner, all high-voltage connections are sealed into a dry atmosphere and are not exposed during shipboard operations. The connections from the directional filters to the adjustable units are made via large ceramic-insulated terminals sealed through the end of the capsule.

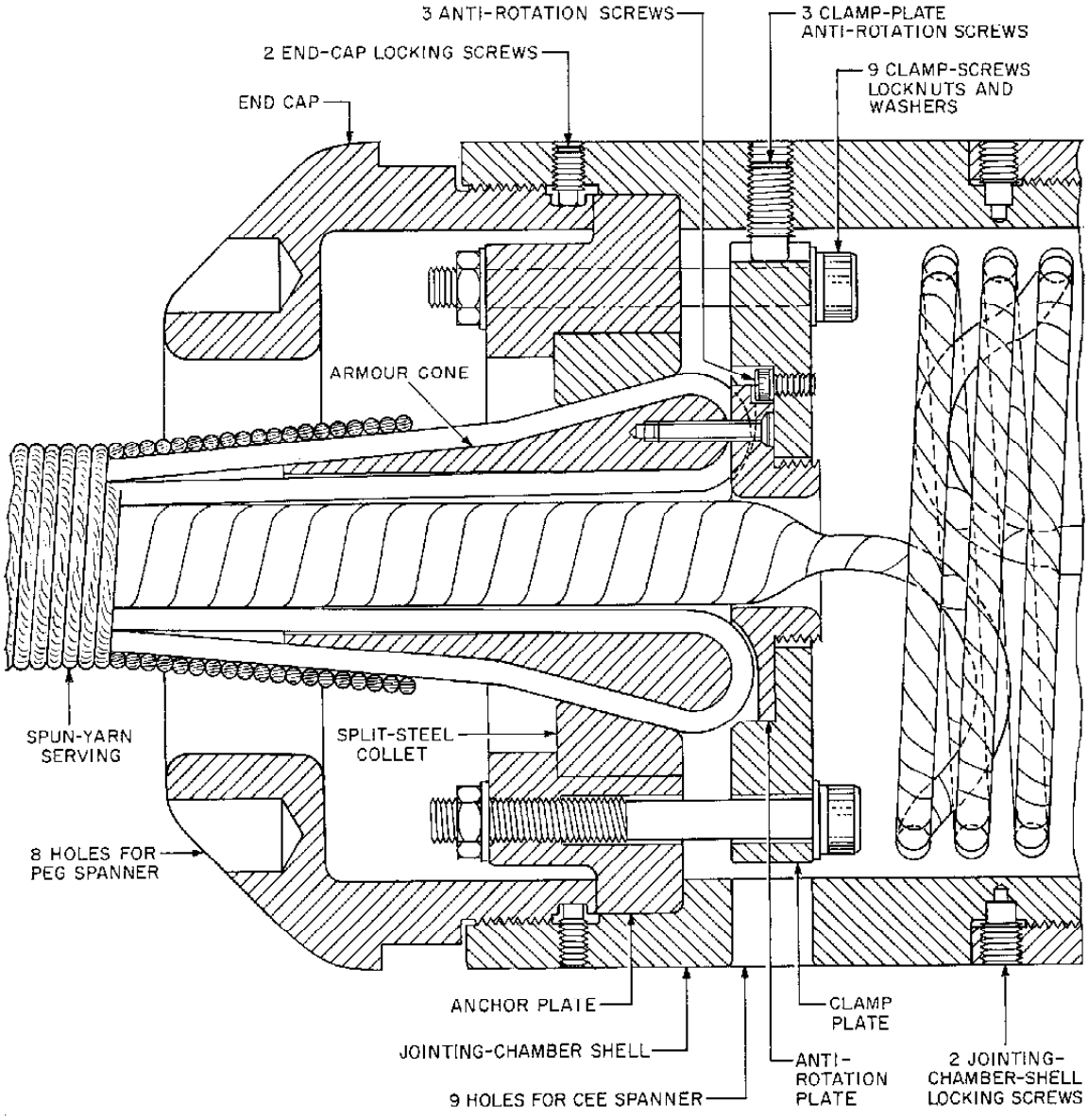


Figure 3—End of equaliser casing to which the submarine cable is anchored. All of this section is at sea pressure.

It should be noted that these terminals will be required to carry only external wiring that is of comparatively low impedance to ground.

The framework inside the sealed capsule consists of 4 longitudinal brass bars attached at their ends to the castings forming the end plates of the capsule; the framework is similar in principle to the open chassis shown in Figure 4. Within and secured to these bars, are mounting plates carrying the electric components. As in standard submerged-repeater practice, no metals are employed in the construction that are known to promote the growth of metallic whiskers. All insulation is of cast methacrylate (Perspex) and in those positions where large differences in potential may occur, long leakage paths and ample air clearances are provided that are adequate to prevent breakdown or ionisation when high line voltages have to be accommodated.

All components are made under dairy conditions of controlled humidity, temperature, and cleanliness, and are rigorously tested before acceptance. The assembly and wiring of the components are also carried out under similar dairy conditions.

After inspection of the wired assembly, the brass tube forming the shell of the capsule is fitted, and the ends are sealed to the 2 end castings. The interior of the capsule is then thoroughly dried by flushing through with dry nitrogen, introduced through small holes in the end castings, for a period of 24 hours; after which the holes in the castings are finally sealed.

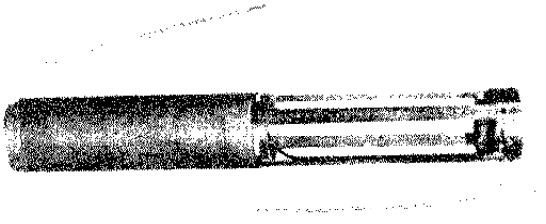


Figure 4—Apparatus unit. The closed section contains the fixed electric units assembled in the factory. The open structure will hold the adjustable electric units assembled on shipboard.

The open chassis section of the apparatus unit for mounting the adjustable units of the equaliser is formed by 4 steel mounting bars secured at their ends to gunmetal rings by screws and locating dowels; these latter are fitted to facilitate accurate replacement of a bar after it has been removed to allow the units to be mounted. The end ring of the chassis adjacent to the sealed capsule is secured to the periphery of a flexible diaphragm, which in turn is secured at its centre to the end casting of the capsule; holes in the diaphragm clear the projecting terminals and cable tail on the end of the capsule. The larger ring at the other end of the chassis carries the fixing screws for attachment to the removable bulkhead of the pressure housing and also forms a coiling-down space for the cable tail after it has been jointed to the corresponding tail of the bulkhead. Attached to the inner face of the larger ring is a pocket to accept a desiccator of silica gel which is slipped into position immediately before re-insertion of the apparatus unit into the housing after fitting the adjustable units on shipboard.

There are 6 sets of fixing holes in the mounting bars of the open chassis that will each accept either 1 adjustable equaliser section or 2 cable-simulator units. Thus, a total of 6 equaliser sections or 12 cable simulators, or a combination of both, can be accommodated. Figure 5 shows an assembly of 1 equaliser section and 1 cable simulator in each of the 2 transmission paths, mounted in the chassis and wired to the appropriate terminals of the capsule ready for re-insertion into the housing.

4.1.3 *Adjustable Units*

To meet the standard of manufacture set by the fabrication of the submerged repeaters, no component should be exposed at any time other

than in an air-conditioned dairy atmosphere. Whereas this requirement is easily met for the fixed elements by sealing them in a capsule before leaving the factory, the problem is not quite so simple for the adjustable units. Two possible lines of approach were considered initially:

A. To set up a working space on the cableship having full dairy conditions.

B. To fit all components into individual hermetically sealed cans before leaving the factory.

The objection to *A* is obvious; if the ship is not already equipped with a dairy, it would be very costly to provide the necessary air conditioning on a temporary basis, even assuming that space would be available. Although *B* also has objections, including those of cost and bulkiness, it offered the best alternative. It also offered the advantage that there would be no comparatively fragile components to handle under the far-from-ideal conditions that exist on a ship.

The final solution was to seal a number of components of the same type into 1 can to provide a range of values in the largest number of steps. It was decided to supply a number of cable simulators as factory-sealed units having different values of equivalent cable length (insertion loss).

To simplify assembly of the equaliser sections, all component units of a particular type were made identical in mechanical size, fixing, and number and position of terminals. Thus, physically, any equaliser section can be made up using only 4 basic parts; namely, resistance unit, capacitance unit, inductance unit, and mounting chassis. All component units are manufactured under full dairy conditions as outlined for the sealed capsule in Section 4.1.2 and are flushed and filled with dry nitrogen before sealing.

The resistance unit contains 10 wire-wound resistors connected to form a constant-impedance bridged-T pad having tapped series and shunt arms; all tapping points are brought out via seal terminals. Only one type of resistance unit, of the necessary impedance, is required for a particular cable system.

The capacitance unit contains 3 capacitors wired in series with their junction points and the free ends brought out via seal terminals. Thus, by suitably arranging the external wiring to the terminals, a wide range of capacitance values can be obtained from 1 unit. For a particular cable system, for example, a 120-circuit system, the required over-all range of capacitance can be covered by a total of 5 different units.

The inductance units each contain 3 inductors having carbonyl iron or ferrite cores with their leads brought out separately via seal terminals. Each unit will give a wide range of inductance values, and a total of only 4 different units are necessary to cover the required over-all inductance range for a particular system.

The mounting chassis for the component units comprises a mounting plate and a spun copper tray secured to the plate; identical drillings in both members allow the cans of components to be mounted on the underside of the plate with their terminals projecting up into the tray. The chassis will accommodate 5 component units;

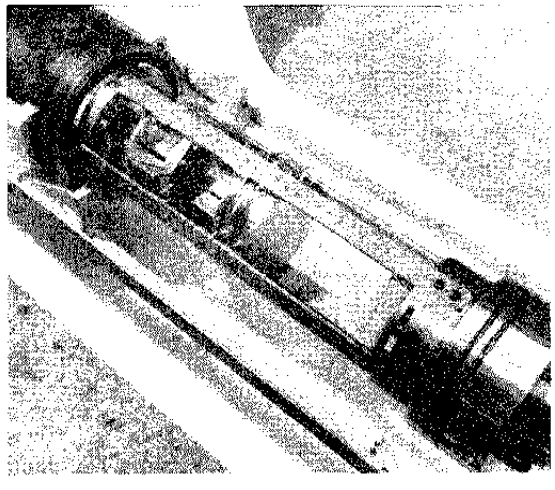


Figure 5—Adjustable units assembled and wired into chassis on shipboard before re-insertion into casing.

a resistance unit and series and shunt capacitance and inductance units. It is supplied fitted with individual blanking-off plates on the underside to cover the drillings for the series and shunt capacitance and inductance units; the plates are subsequently discarded in those positions where the electric design of the equaliser section necessitates the fitting of a component unit. Secured through the wall of the tray are 2 ceramic-insulated terminals having long leakage paths, and 2 metal terminal posts; these will form the input and output terminals of the completed equaliser section. Lugs are provided on the mounting plate for securing into the open chassis of the apparatus unit.

Figure 6 shows an assembled and wired equaliser section having resonant circuits in the series and shunt arms; thus, component units are mounted in all positions. The illustration shows clearly the simple nature of the assembly and wiring operations (to robust terminals) that are required to be performed on board ship.

It will be appreciated that the wiring to the component-unit terminals may be of high impedance to ground; to avoid deterioration of this condition, the tray containing the unit terminals and wiring is flooded with a sealing compound. The compound, which is poured at a temperature of 85 to 90 degrees centigrade, sets in approximately 20 minutes to a state that allows handling of the equaliser section. The heat-sink formed by the mounting chassis and unit cans prevents any undue rise in temperature of the components during the sealing operation.

The exposed input and output terminals of the completed equaliser section, being of low impedance, will not cause any significant deterioration in performance in the unlikely event of condensation forming on the insulators. However, as a further precaution, as has previously been stated, a silica gel desiccator is included in the open chassis of the apparatus unit to dry out the interior of the pressure housing after sealing.

The cable simulators use a mounting plate similar to that for the equaliser sections. Secured to the plate is a hermetically sealed can containing the electric components, which are mounted on a sub-plate of cast methacrylate sheet fitted into the can. For the external wiring, 2 ceramic seal terminals having long leakage paths and 2 metal earth posts project from the side of the can; as in the case of the equaliser sections, those exposed terminals are of low impedance to ground. The cable simulators are made in the factory under full submerged-repeater dairy conditions, and are flushed and filled with dry nitrogen before sealing.

The positions of the terminals on the cable simulators in relation to the lugs on the mounting plate are the same as those on the equaliser sections. This simplifies wiring between units mounted in the open chassis of the apparatus unit. Wiring bushes in the mounting plate, corresponding to similar bushes in the mounting plate of the equaliser sections, carry the interconnecting leads between units and provide intermediate supports for those leads that might otherwise be unstable due to their length. The wiring between units may be seen in Figure 5.

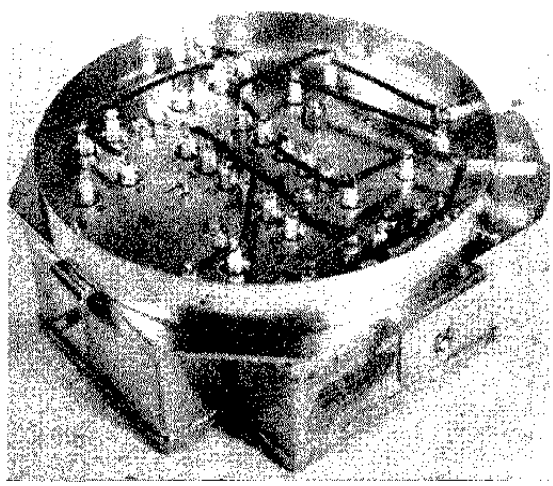


Figure 6—Assembled and wired adjustable equaliser section before the tray containing the terminals and wiring is filled with sealing compound.

4.2 Electric Circuits

At the very beginning, it was decided to provide, as far as possible, a common arrangement for the power separating filters. It was also decided that the adjustable equaliser circuits should be adequate for the 2 projects on hand, namely, the 18-circuit Scotice system with a frequency range from 20 to 200 kilocycles per second and the 120-circuit *Med-1* system with a frequency range from 60 to 1168 kilocycles per second.

A design covering both of these systems would then automatically cover a 60-circuit system with a frequency range from 20 to 552 or from 60 to 608 kilocycles per second.

4.2.1 *Power Separating Filters*

One of the basic considerations in the design of the power separating filters is the total value of capacitance required to be subjected to the line voltage stresses; it was decided to use the same volume of high-voltage capacitors in the equaliser as in the power separating filters in the submerged repeaters. On this basis, it was possible to develop the circuit shown in Figure 7, which provides for full separation of line voltage and line current from the transmission circuits. It also gives a loop-loss performance meeting the design requirement of 40 decibels minimum loss within the system frequency range.

Typical loss curves are shown in Figure 8. Curve *A* is for the 18-circuit system and curve *B* covers the case of 60- and 120-circuit systems.

To reduce the total volume of the capacitors required for the adjustable equaliser sections and directional filters, the nominal line impedance of 53 ohms is transformed to 75 ohms for 120- and 60-circuit equalisers and to 150 ohms for 18-circuit equalisers. The selected values of impedance match the impedance of standardized transmission-measuring equipment.

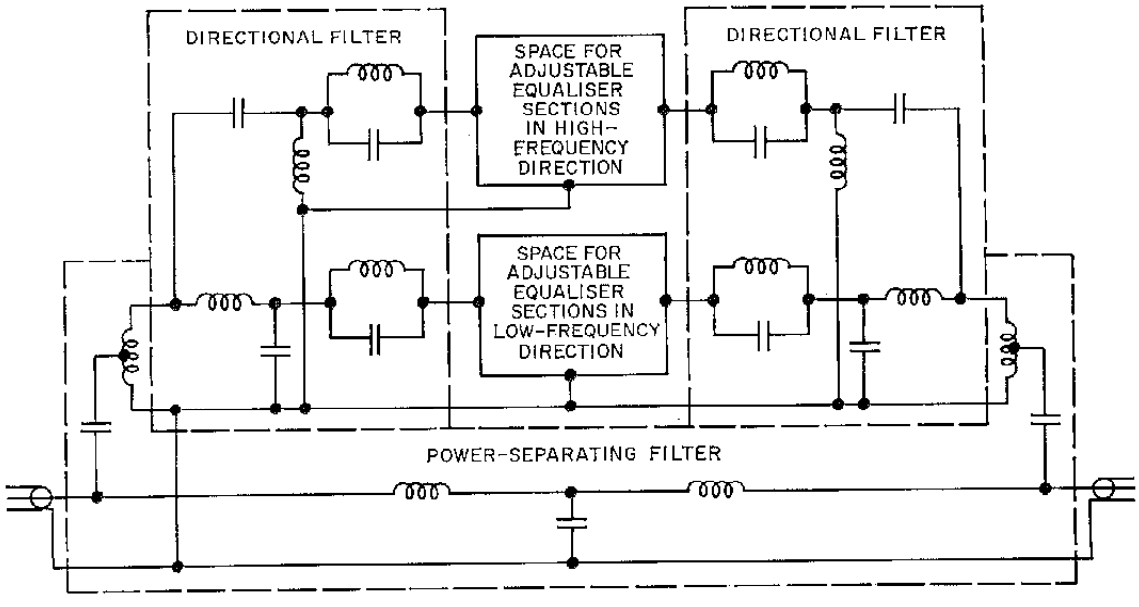


Figure 7—Submerged equaliser.

4.2.2 *Directional Filters*

Directional filters are fitted to ease the problem of designing the adjustable sections aboard the cableship. Since the equaliser is a passive device, it was decided that the minimum stop-band loss should be greater than 40 decibels. This requirement was met with only 2 half-sections in each path.

Directional filters are the only units in the equalisers that are dependent on the frequency range of the system.

4.2.3 *Adjustable Equaliser Sections*

As seen from the analysis of possible system errors outlined in Section 2, the adjustable equaliser sections must provide truly variable loss characteristics. Practical considerations, however, dictate definite limits to the possible range of available characteristics, and the following targets were set at the design stage:

A. Each adjustable equaliser section should provide any one of the 4 types of curves shown in Figure 9.

B. Each section to be of the constant-impedance type [11] with a maximum insertion loss adjustable in steps of 0·5 decibel within the range from 2 to 10 decibels.

C. The range of inductance and capacitance values must provide an insertion loss for the type-C equaliser of Figure 9 having a shape factor b from 1·2 to 4 where $b = f_r/f_b$, f_r = resonance frequency or maximum-loss point A_m for type C, and f_b = frequency where loss is half the maximum loss $A_m/2$.

In addition, the component values must provide for f_r to be any frequency between f_1 and f_2 where $f_1 = 0{\cdot}5 \times$ the lowest system frequency. For the 18-circuit system, $f_1 = 0{\cdot}5 \times 20 = 10$ kilocycles per second. $f_2 = 1{\cdot}5 \times$ the highest system frequency. For the 120-circuit system, $f_2 = 1{\cdot}5 \times 1{\cdot}2 = 1{\cdot}8$ megacycles per second.

To cover the above requirements, the range of adjustable capacitance units and adjustable inductance units were designed as follows.

4.2.3.1 *Capacitor Units*

The values follow the ratio 1:2:4, and terminals on the can allow for external series-parallel connections of the 3 capacitors.

The ratio 1:2:4 gives the widest range of values when employing all possible 16 combinations while providing values of capacitance at steps not exceeding 33 per cent. By suitably spacing the basic values of the capacitors finally selected for the 5 capacitance cans, a range from 160 picofarads to 0·407 microfarad has been obtained in steps that on the average are 10 per cent apart, with a small number of isolated cases where steps may reach 25 per cent.

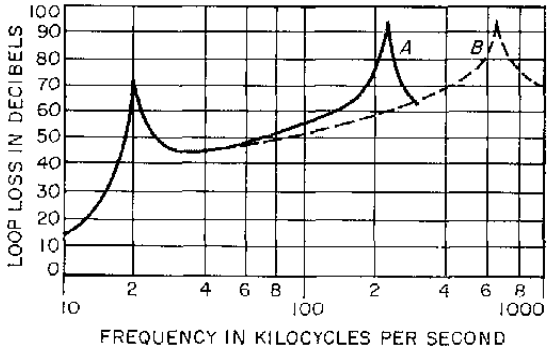


Figure 8—Loop loss of power separating filter.

4.2.3.2 *Inductance Units*

The method used to provide adjustable inductance is similar to that employed for capacitance units. Again the values are in the ratio 1:2:4, giving a total available range from 5 inductance cans of 0·77 microhenry to 6 millihenries in steps of about 10 per cent with isolated cases of a maximum 17-per-cent gap between the adjacent inductance values.

There are many intermediate values with spacings of less than 10 per cent because to cover

the required frequency range from 10 kilocycles per second to 1·8 megacycles per second and to meet a minimum Q value of 60, some coils have been designed to cover only the high frequencies, normally low inductance values, and other units are for low-frequency work and have high inductance values.

4.2.3.3 *Resistance Unit*

By suitably strapping the external terminals of the resistance unit, 23 different loss values A_m are obtained in 0·5-decibel steps between 2 and 10 decibels with additional intermediate values. Of the 2 different designs available, one is a unit of 75-ohm impedance for the 120- and 60-circuit equaliser, and the other is a 150-ohm-impedance unit for the 18-circuit equaliser.

4.2.4 *Cable Simulators*

Simple equaliser sections having loss characteristics simulating the loss of a cable length are provided with loss values at the top system frequency from 1·5 to 12 decibels in 1·5-decibel steps. A selection of cable-simulator units is

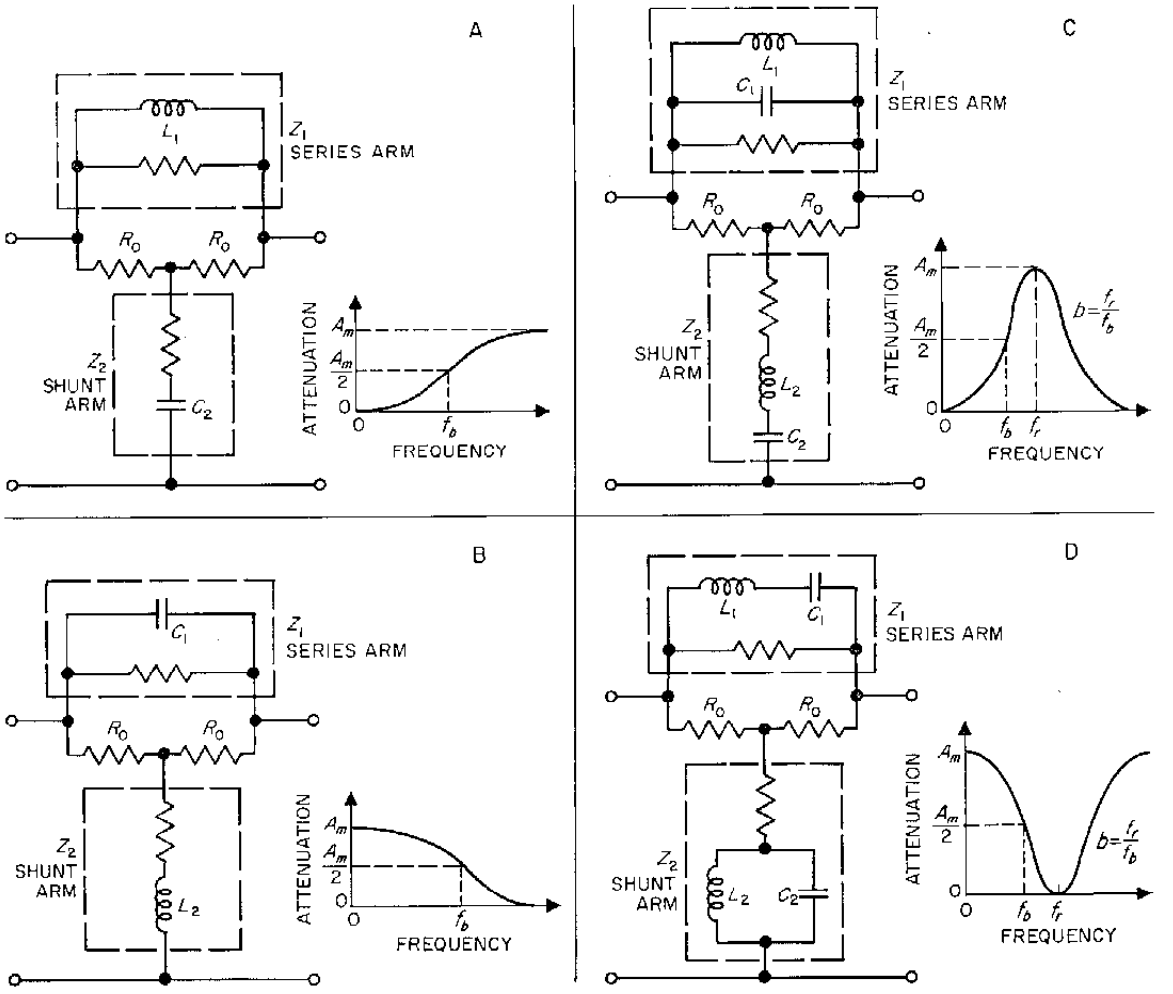


Figure 9—Circuit and attenuation-frequency characteristics of the 4 types of equaliser sections that must be provided for in the adjustable equalisers.

carried on the cableship during the laying; they are used as and when required.

5. Shipboard Assembly

One of the essential requirements during the laying of a deep-water submarine cable is to maintain a continuous paying out of the cable at normal laying speed with the correct laying tension. To avoid stopping the ship, the complete equaliser, including any cable length adjustment, must be completed before the ship reaches the position where the equaliser must be laid. Thus, from the beginning of the lay, results of the transmission measurements between the terminal station and the ship are plotted on a graph showing the system loss against the length of the laid cable. Curves are drawn for all test frequencies (say 12 frequencies in each direction of transmission). The points are normally on a straight line, so that after some time of cable laying the system loss at the end of the ocean block can be extrapolated for each of the test frequencies. The estimated system levels are then compared with the ideal system levels at this point, and the difference between the two gives the equaliser target characteristic.

5.1 Circuit Design

When the equaliser requirements are known, the designer decides on the types and number of sections to be used. The results are plotted on transparent logarithmic graph paper and are checked by superimposing the graph on curves showing the shape characteristics of equaliser sections with different values of shape factor b. A typical set of curves for a type-C section with $A_m = 3$ and 6 decibels is shown in Figure 10.

The equaliser designer then decides on the design parameters such as frequency of maximum loss, shape factor, loss value, half-loss frequency, et cetera, and estimates the basic equaliser loss. Then the cable length that is to contain the submerged equaliser is shortened to provide the effective system gain equivalent to the basic equaliser loss.

Calculated element values are checked against the available components by reference to tables. The value giving the minimum percentage difference from the design requirement is normally selected.

To provide an additional check on the calculations all available resonant frequencies have been tabulated by a computer as a function of the capacitance and inductance values, and such tables are available to the designer. The expected insertion loss of each section can be quickly calculated either by reference to a series of graphs similar to Figure 10 or by the use of tables relating design parameters to the loss characteristic of the particular type of equaliser section to be used.

After a check of expected loss characteristic the designer draws a wiring diagram from which the assembly and wiring are done. After a wiring inspection the loss characteristic is measured, and the equaliser section finally sealed as described in Section 4.1.3.

When the individual equaliser sections are complete, they are inserted into the open chassis of the apparatus unit and interconnected to give the required loss characteristic in each direction of transmission.

The total time required to design, check, wire, inspect and seal 6 equaliser sections and then fit these sections into the chassis, interconnect the units, close and seal the housing, and joint the completed equaliser into the cable, is estimated to be as long as 16 hours for a team of 5 people.

5.2 Equaliser Assembly

The description in Section 4 gives a fairly clear picture of the operations that are required on board the cableship to complete the assembly and wiring of the submerged equaliser and to seal the pressure housing; nothing has been

said however about the actual operation of withdrawal of the apparatus unit from the housing and the subsequent re-insertion. It was appreciated that sufficient head-room was not readily available on shipboard to allow this operation to be carried out in the vertical plane, which is the obvious and simple method. Therefore equipment had to be designed to perform this operation with the equaliser in the horizontal position. This created quite a problem as the bulkhead is of necessity a very close fit in the casing. Any external means for guiding the apparatus unit out of the casing would have to be aligned coaxially to the bulkhead to a very high degree of accuracy. This problem was aggravated by requiring that the equipment be portable for use on any ship.

The final solution provided withdrawal gear which has the following characteristics.

A. Alignment adjustments need not be made when erecting the equipment for operation.

B. Completely self-contained and does not require any special fitments to be provided on the ship.

C. Small enough when dismantled to pack into a travelling case approximately 60 by 18 by 14 inches (152 by 46 by 36 centimetres); this case also houses the ancillary equipment required for pressurising the petroleum-jelly system in the bulkhead of the pressure housing.

The principle of the design is to withdraw the bulkhead almost completely from its close-fitting bore in the casing whilst allowing the guiding to be done solely by the bore and not by any external guide rails. This is achieved by means of a short lead-screw attached to the face of the bulkhead and a spherical nut running in hemispherical cup washers that are free to move radially with respect to the lead-screw; the nut and washers are restrained in the carriage, which is supported on the two main members of the equipment. Rotation of the nut will withdraw the bulkhead, but the angular and radial position of the nut will be governed entirely by the position of the lead-screw; thus no radial thrust will be applied to the screw during the operation. When the bulkhead is all but fully withdrawn, the final few degrees rotation of the nut firmly clamps the lead-screw, nut, and hemispherical washers to the carriage, and so prevents any further relative movement among these members. The carriage may now be run along the guide rails formed by the main members to complete the withdrawal, the clearance

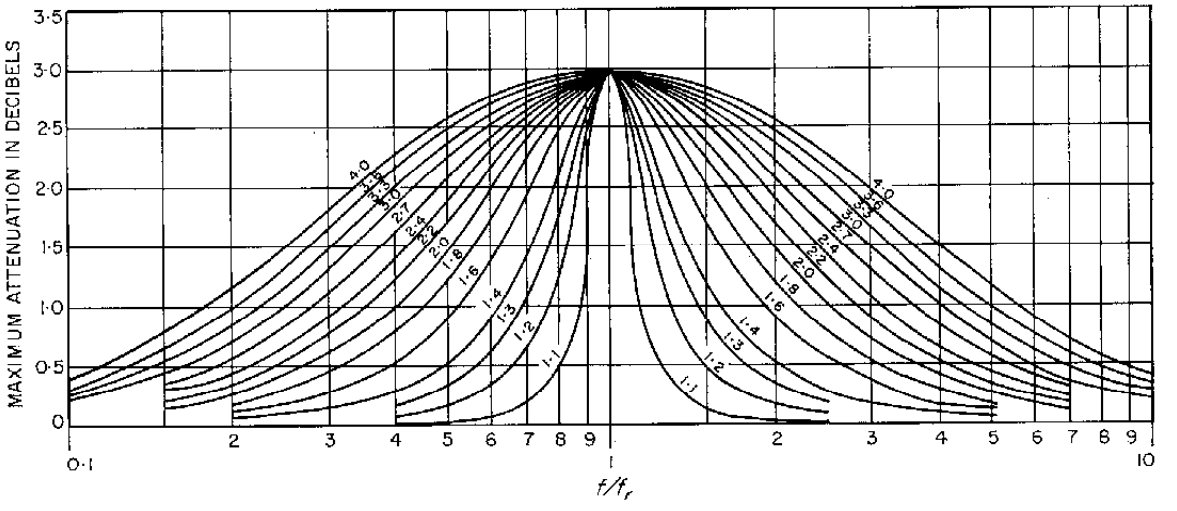


Figure 10—Normalised loss curves for a section of type-C over a range of $b = f_s/f_b$ from 1·1 to 4·0 for maximum attenuations at f_r of 3 and 6 decibels. The ordinate scale is doubled for the 6-decibel range.

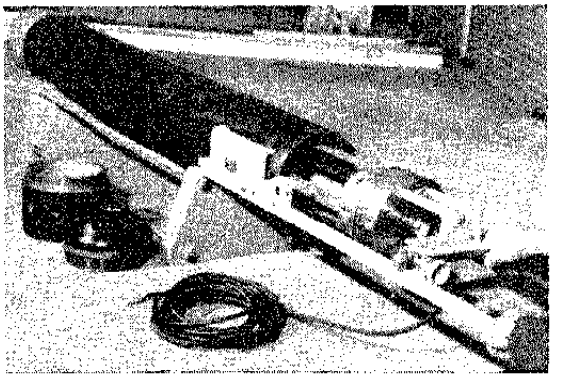


Figure 11—Withdrawing apparatus unit from the casing.

between the apparatus unit and the casing allowing for a fair degree of misalignment between the rails and the axis of the casing. The operation of completing the withdrawal by hand is illustrated in Figure 11.

For re-insertion of the apparatus unit and bulkhead into the casing, returning the carriage to its original position will just enter the bulkhead into its bore in the casing, and, after clamping the carriage to the brackets attached to the rails, rotation of the nut will drive the bulkhead fully home.

6. Conclusion

The adjustable submerged equaliser has been primarily developed for handling on board cableships, where it would be difficult and uneconomical to provide dairy conditions [12] for assembling and wiring submerged-repeater type of components. The adopted method is simple, fairly economical in the use of components, and provides truly adjustable equaliser characteristics within required limits.

Apart from the directional filters and 2 autotransformers in the power separating filter which are dependent on the system circuit capacity and, hence, on the system line frequencies, all other units are applicable to all projects within the frequency range of 20 kilocycles per second to 1200 kilocycles per second. This permits any unused inductance and capacitance units and cable simulator sections manufactured for a particular system to be available for another project.

The estimated time for building all of the possible 6 adjustable equaliser sections, together with extensive electrical and mechanical inspection and the subsequent housing operation, is 16 hours. With an average laying speed of 4 nautical miles per hour, this represents some 64 nautical miles of cable, or approximately 2½ repeater sections for 18-circuit schemes, and 5 repeater sections in the case of 120-circuit schemes. Since equalisers are normally placed every 12th or 10th repeater, there is adequate time to determine the expected system levels at the end of the ocean block to be equalised and to design and complete the assembly and wiring of the submerged equaliser before it is due to be laid.

An equaliser described above has been successfully employed during the laying of the Scotice system [13] and demonstrated its full flexibility by providing some equalisation outside the transmission band to compensate for the excess gain at frequencies around 15 kilocycles per second. The equaliser was laid in 830 fathoms. The equaliser-handling gear operated perfectly, and the housing work, including pressurising the removable bulkhead, went smoothly.

Other units of the same design have been laid in depths of 1000 to 2000 fathoms on the *Med-1* and *Med-2* projects.

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On graduation he joined the transmission laboratories of Standard Telephones and Cables. After two years on development of the short-haul *N1* carrier system, he transferred to the submarine systems development group. Since then he has been closely associated with the design and development of submerged repeaters and with the engineering of submarine telephone cable projects, including participation in the laying of several cables in the North Sea and in the Mediterranean. He is now in charge of the submerged-repeater development section. Mr. Dawidziuk is an Associate Member of the Institution of Electrical Engineers.

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From 1929 to 1935, he was with W. T. Henley's Telegraph Works, where he was employed as a draughtsman in the research laboratories.

Joining Standard Telephones and Cables in the latter part of 1935, he was responsible for the mechanical design of transmission laboratory and factory test gear throughout the war years until 1945. In the early post-war period, he was engaged on the mechanical design of waveguide components for microwave systems. Since 1949, he has been actively engaged on the mechanical design of all aspects of submerged repeaters and submerged equalisers.

Variable-Reactance Frequency Multipliers

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1. Introduction

Variable-reactance frequency multipliers using high-power transistors provide a reliable and efficient solid-state frequency source for the microwave range. Their properties and applications are extensively discussed in the literature.

Interesting results have been obtained from development work on variable-reactance or varactor frequency multipliers for the microwave range as part of a tropospheric scatter program. Some comparisons are made with nonlinear resistances. The advantages of variable reactances as well as their theoretical limitations are pointed out. Also, certain points are considered that may be of interest to the designer who often must compromise among efficiency, output power, and circuit complexity, especially in cases where frequency-multiplier chains tunable over wide frequency ranges are required. An X-band frequency multiplier is discussed briefly.

The theoretical discussion presented in the appendix is restricted to the frequency doubler, the simplest and most efficient varactor frequency multiplier.

2. General Properties of Nonlinear Devices

One property of linear time-invariant elements is that no frequency conversion can take place in them. They are generally classified as "pure lossy" when they dissipate power, or as "lossless" when they do not consume power. They are purely inductive or purely capacitive when they draw reactive power from and return it to the energy source.

These definitions do not apply to nonlinear or time-varying elements. A nonlinear resistance or a nonlinear capacitance, although free from dissipative losses and characterized by a single-valued charge-voltage equation, accepts real power and represents to the prime source a power-consuming device; this same power, of course, can not be dissipated in the lossless element, and it is released at other frequencies. Nevertheless, the nonlinear element, although lossless, appears as a power-consuming device to the prime source. Since no power is retained, the over-all power input to the device equals zero.

2.1 Nonlinear Resistance

When a nonlinear resistance is energized by a direct-current source and an alternating-current source in series, the following observations can be made regarding direct current and currents at fundamental and harmonic frequencies.

(A) At short-circuit conditions $R_{load} = 0$, the highest-order harmonic generated by the nonlinear resistance depends only on its degree of nonlinearity. A square-law characteristic, for instance, will give rise to a second harmonic, all higher-order harmonics being equal to zero.

(B) The same nonlinear resistor can give rise to any desired harmonic if external circuits are used to reapply additional voltages or currents at frequency multiples to the nonlinear device. This can be seen physically by the fact that under these conditions a highly distorted voltage (or current) is applied to the nonlinear device. These observations apply to nonlinear resistances as well as to nonlinear reactances.

The general power relations for resistive harmonic generators are governed by the following equations set forth by C. H. Page [7].

$$P_n \leq \frac{1}{n^2} P_1 \tag{1}$$

$$P_0 \geq (n^2 - 1)P_n \tag{2}$$

where P_0, P_1, and P_n are direct-current power, power at the fundamental frequency, and power at the nth harmonic, respectively; n is the order of the generated harmonic.

Page's equations state that:

(A) In any conversion process with nonlinear resistances, direct-current power is generated;

(B) Even with lossless conversion, where all fundamental power is converted and all undesirable harmonics are absent, the efficiency in generating the nth harmonic can not exceed $1/n^2$;

(C) The rest of the power at the fundamental frequency is converted into direct current.

In the case of a lossless resistive doubler, for instance, the maximum possible efficiency would be 25 percent; the rest of the input power (75 percent) would be converted into direct current. In the case of a tripler, the best efficiency would not exceed 11 percent.

2.2 Nonlinear Reactances

The nonlinear reactance as a harmonic generator differs in several significant respects from the nonlinear resistance.

(A) It does not rectify; that is, a sinusoidal voltage applied to the reactance gives rise only to different harmonic currents, with amplitudes depending on varactor characteristic and operational conditions. The biasing direct voltage created by the varactor when operated without external bias is due to the nonlinear resistance that is part of the P-N junction.

(B) The efficiency of a lossless reactance, operated as a frequency multiplier, can reach the theoretical limit of 100 percent. All power accepted at the fundamental frequency is released at one or several harmonics. This is confirmed by the Manley and Rowe equations [8], which in the case of a frequency multiplier assume the simple form:

$$P_{1,0} = \sum_{m=2}^{\infty} - P_{m,0}. \qquad (3)$$

In case the circuit configuration favors only one harmonic frequency, the full input power appears at that particular frequency, independent of the reactance characteristic, circuit, and load conditions. The load impedances for all undesired harmonics, however, must be made zero, infinite, or purely reactive.

(C) The fact that the varactor accepts real power does not mean that it has an input impedance in the common sense of the word. It has, however, a complex ratio of input voltage to input current termed conversion impedance that does not exist physically; rather, it describes the ability of the varactor to accept real power at the fundamental frequency and is largely a function of the chosen operating conditions.

The same remarks apply to the output impedance of the varactor. For these reasons, the performance of a varactor cannot be represented by an equivalent diagram. As will be seen, certain precautions can be taken to ensure maximum power output or best efficiency. However, these can hardly be described as matching procedures, in the common sense of the word.

3. *P-N* Junction

The P-N junction exhibits two important properties that lead to basically new applications: the sharp-reverse-breakdown effect and the electrically controllable capacitance.

Physically, the diode can be represented by the equivalent circuit shown in Figure 1. Here, $C_{(v)}$ is the voltage-sensitive junction capacitance; $R_{(v)}$ is the voltage-sensitive diode resistance; R_S is the junction series resistance or spreading resistance that is primarily responsible for the losses in the varactor. C_1, R_1,

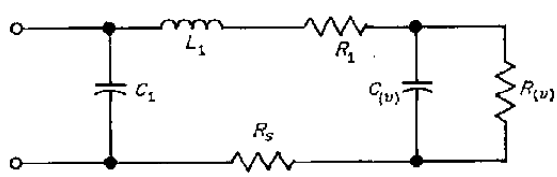
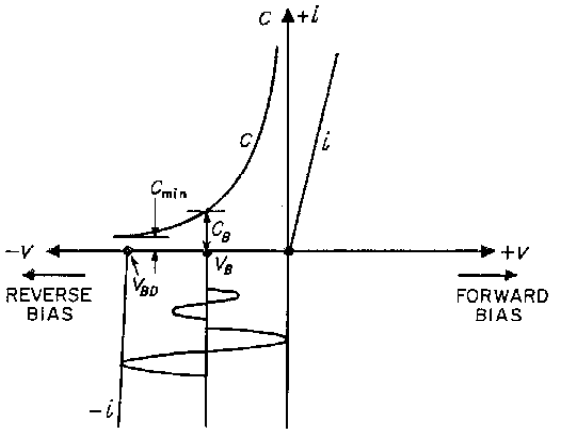


Figure 1—Equivalent diagram of a P-N junction.

Figure 2—Characteristics of a varactor.



and L_1 are case capacitance, losses, and lead inductance.

While the case components C_1, R_1, L_1 are not without influence on the proper operation for the varactor at microwave frequencies, the truly important parameters are the series resistance R_S and the capacitance-versus-voltage sensitivity of the junction capacitance $C_{(v)}$.

The cutoff frequency f_{co} is defined as the frequency at which the absolute value of the capacitive reactance of $C_{(v)}$ is equal to the series resistance R_S. In this case, the static value of $C_{(v)}$ at breakdown voltage is taken.

$$f_{co} = \frac{1}{2\pi C_{min} R_S}. \tag{4}$$

The Q factor of the junction represents the ratio of the capacitive impedance of C_{min} and the series resistance R_S.

$$Q = \frac{1}{2\pi f C_{min} R_S}. \tag{5}$$

Cutoff frequency or Q factor, together with the capacitance-voltage relationship, suffice to define a varactor for most microwave applications. The Q factor, which becomes equal to unity when the varactor is operated at cutoff frequency, is a complete measure of the efficiency of the diode element. Only high-Q diodes can ensure efficient operation at high frequencies, which means that the product fR_SC_{min} must be minimized. To ensure this is the most important problem of diode manufacture.

The static characteristics of the varactor diode are depicted in Figure 2. The capacitance-versus-voltage characteristic obeys the law $C_{(v)} = KV^{-\theta}$, θ being equal to 1/2 for the abrupt junction, and 1/3 for the graded junction. The steepness and the absolute value of the capacitance variation depends mainly on the background doping level of the P-N junction. Recent investigations [6] have shown that higher capacitance-voltage sensitivity can be achieved by using impurity profiles that have an abrupt transition from P to N at the junction, but a decreasing impurity concentration when moving away from the junction. Higher capacitance-voltage sensitivity means a greater coefficient θ; this is favorable for the application of varactors as frequency multipliers.

To operate the varactor in the purely capacitive mode, the instantaneous voltage applied to the P-N junction should remain within the limits set by V_{BD}, the breakdown voltage, and zero. For full exploration of the capacitance-voltage curve, both the bias voltage V_B and the amplitude of the alternating voltage should be chosen equal to $V_{BD}/2$ (voltage limitation). For simplicity of associated equipment, automatic bias has been adopted throughout all experimental work.

A second (thermal) limitation is set by the losses in the spreading resistance R_S, which may not exceed a safe value determined by the heat transfer properties of the varactor.

Should the losses in R_S grow beyond this point, the internal contact potential of the P-N junction will be affected. The contact potential varies from about 0.7 volt at 25 degrees to about 0.4 volt at 150 degrees centi-

grade. A change in the contact potential will result in a change of the capacitance-voltage characteristic of the varactor, which in turn will affect the tuning and cause a drop in output power and efficiency.

4. Varactor

4.1 General

The following is a study of the simplest configuration of varactor and power source. The varactor is assumed lossless; the power source has zero internal impedance and maintains a purely sinusoidal voltage across the varactor at the fundamental frequency. No resistances, filters, or reactances are present in the current path. No resonance effects take place.

The incremental capacitance is given by

$$C_{(v)} = \frac{dq}{dV} = KV^{-\theta}. \tag{6}$$

The incremental charge is

$$q = K\int V^{-\theta}dV = \frac{K}{1-\theta}V^{1-\theta} \tag{7}$$

where the integration constant is set equal to zero.

Assuming that the voltage applied to the varactor is

$$V = V_B + V\sin\omega t$$

the charge on the nonlinear capacitor will be

$$q = \frac{K}{1-\theta}(V_B + V\sin\omega t)^{1-\theta} \tag{8}$$

where V_B is the bias voltage and V is the amplitude of the sinusoidal voltage at the fundamental frequency.

The current through the varactor is given by

$$i = \frac{dq}{dt} = \frac{K\omega V\cos\omega t}{(V_B + V\sin\omega t)^{\theta}}. \tag{9}$$

The constant K can be determined from the boundary conditions. At breakdown voltage V_{BD} the varactor capacitance assumes its minimum value C_{min}, hence

$$K = C_{min}V_{BD}^{\theta}$$

and the general expressions for capacitance, charge, and current are

$$C = C_{min}V_{BD}^{\theta}(V_B + V\sin\omega t)^{-\theta} \tag{10}$$

$$q = \frac{C_{min}}{1-\theta}V_{BD}^{\theta}(V_B + V\sin\omega t)^{1-\theta} \tag{11}$$

$$i = \frac{C_{min}V_{BD}^{\theta}\omega V\cos\omega t}{(V_B + V\sin\omega t)^{\theta}}. \tag{12}$$

Setting $\theta = \frac{1}{2}$ for the abrupt P-N junction and $V = \xi V_B (V_B = V_{BD}/2)$, where ξ for operation in the purely capacitive mode approaches but never reaches unity, (10), (11), and (12) appear in the following form

$$V = V_B(1 + \xi\sin\omega t)$$

$$C = \frac{2^{\frac{1}{2}}C_{min}}{(1 + \xi\sin\omega t)^{\frac{1}{2}}} \tag{13}$$

$$q = 2^{\frac{1}{2}}C_{min}V_{BD}(1 + \xi\sin\omega t)^{\frac{1}{2}} \tag{14}$$

$$i = \frac{\xi\omega C_{min}V_{BD}}{2^{\frac{1}{2}}}\cdot\frac{\cos\omega t}{(1 + \xi\sin\omega t)^{\frac{1}{2}}} \tag{15}$$

where C, q, and i are expressed in terms of breakdown voltage and minimum capacitance, whereas the driving voltage is expressed in terms of the bias voltage.

These equations permit an insight into the mechanism of operation of the varactor.

With the driving voltage varying between approximately zero and V_{BD}, the charge on the varactor oscillates, without changing sign between a constant minimum value q_{min} (at $\omega t = 3\pi/2 + n\cdot 2\pi$) and maximum value $q_{max} \cong 2\cdot C_{min}V_{BD}$, which occurs at $\omega t = \pi/2 + n\cdot 2\pi$. The capacitance varies between C_{min} (at $\omega t = (\pi/2) + n\cdot 2\pi$), and C_{max}, which is reached at $\omega t = (3\pi/2) + n\cdot 2\pi$, and the absolute value of which depends on ξ. For $\xi = 0.95$, C_{max} is equal to $6.3C_{min}$. Thus, minimum charge value coincides with maximum capacitance value. Both occur when the voltage across the varactor is a minimum.

The current waveform closely resembles a saw tooth with zero values at $\omega t = \pi/2$ and $\omega t = 3\pi/2$. As expected, the current is symmetrical with respect to the time axis (that is, no rectifying effect takes place) and exhibits a steep transition from negative to positive values at the point $\omega t = 3\pi/2$. Further treatment of (15) reveals that both positive and negative current maxima are situated symmetrically with respect to the point $3\pi/2$. The exact phase angle at which current maximum occurs is a function of ξ and is given by

$$(\omega t)_{max} = \sin^{-1} \frac{(1 - \xi^2)^{1/2} - 1}{\xi} \qquad (16)$$

for $\xi = 0.9$, for instance, $(\omega t)_{max} = \sin^{-1}(-0.63) = -39$ degrees, and 180 degrees + 39 degrees; for $\xi = 0.95$ the phase angle is $(\omega t)_{max} = -46$ degrees and 180 + 46 degrees. Thus, with ξ approaching unity, the maximum points approach the zero transition point from both sides, which results in an increase in steepness of the saw-tooth curve.

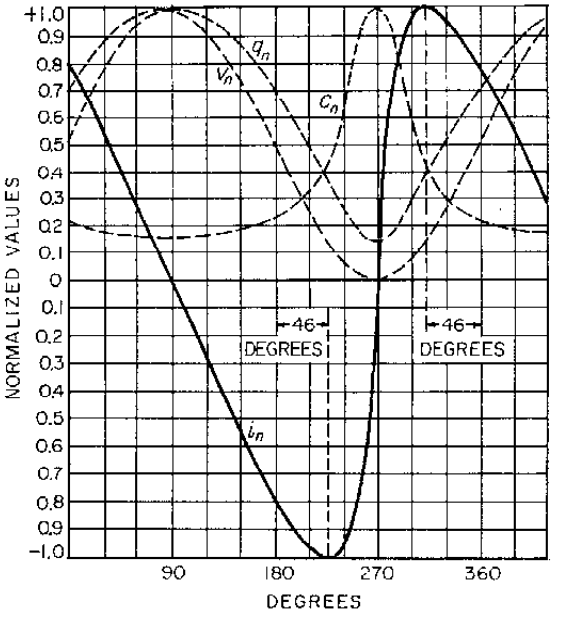


Figure 3—Normalized values of voltage, capacitance, charge, and current for $\xi = 0.95$.

Figure 3 shows the normalized values of these quantities for $\xi = 0.95$.

$$V_n = \tfrac{1}{2}(1 + \xi \sin \omega t)$$

$$C_n \cong 0.16(1 + \xi \sin \omega t)^{-1/2}$$

$$q \cong 0.71(1 + \xi \sin \omega t)^{1/2}$$

$$i \cong 1.25 \frac{\cos \omega t}{(1 + \xi \sin \omega t)^{1/2}}.$$

4.2 Circuit Analysis of a Series Doubler

This analysis is based on the following assumptions.

(A) Purely sinusoidal voltages are applied to the varactor at frequencies ω and 2ω. The harmonics appear in the charge q and the current i.

(B) Ideal conditions prevail, that is, complete harmonic separation, no power flow at any undesired harmonic, and no losses either in tuned circuits or in the varactor.

Based on the Taylor theorem, the alternating-current charge on the varactor can be expressed as a function of the applied voltage in the following manner.

$$q = \sum_{K=1}^{\infty} \alpha_K V^K \qquad (17)$$

where α_K is the Taylor coefficient of the Kth term of the Taylor expansion, and V is the applied voltage across the varactor.

In the general case, the applied voltage as well as the resulting charge can be expressed as a sum of sinusoidal voltages.

$$\sum_{\substack{n=-\infty \\ n \neq 0}}^{n=+\infty} Q_n \exp[jn\omega t] = \sum_{K=1}^{\infty} \alpha_K \left\{ \sum_{\substack{n=-\infty \\ n \neq 0}}^{n=+\infty} V_n \exp[jn\omega t] \right\}^K \qquad (18)$$

where V_n, V_{-n} and Q_n, Q_{-n} are complex conjugate vectors rotating in opposite directions at an angular velocity equal to $n\omega$.

This equation is general. Physically, it signifies that a charge component Q of frequency

$l\omega$ is built up by contributions of all those voltage components whose products yield a voltage component of frequency $l\omega$. An infinite number of such components exist, even in the case where the number of voltage components across the varactor is restricted to 2. Two restrictions reduce (18) to a form that, although less general, describes the voltage and charge relations in the series doubler with sufficient accuracy.

(A) The voltage across the varactor consists of components at the fundamental frequency and at only the second harmonic.

(B) In the case of a frequency doubler, the most significant contribution to the second-harmonic charge component is due to the square term of the Taylor series.

(C) Only current components at frequencies ω and 2ω will be selected since the external circuit configuration will assign zero-impedance paths to all higher-order currents.

The resulting equation is dealt with in Appendix 7.1. It leads to the following expression of the conversion conductance

$$G_c = \frac{\omega^2 C_B^2}{G_L} \cdot \frac{\theta^2}{4} \cdot \left(\frac{V_{in}}{V_B} \right)^2 \qquad (19)$$

where

C_B = capacitance at the bias point
θ = capacitance nonlinearity coefficient
G_L = real part of the load admittance
V_{in} = amplitude of the input voltage, and
V_B = bias voltage.

Or, the conversion resistance that appears across the input terminals of the varactor

$$R_c = \frac{1}{G_c} = \frac{1}{\omega^2 C_B^2 R_L} \cdot \frac{4}{\theta^2} \cdot \left(\frac{V_B}{V_{in}} \right)^2 \qquad (20)$$

where $R_L = 1/G_L$ is the load resistance, and $C_B = 2^{1/3} C_{min}$.

Equation (19) and the general equations for efficiency and output power derived in Appendix 7.2 allow an insight into the significant parameters that ensure efficient operation.

It can be seen from (39) that the efficiency is a maximum when both load conductance and conversion conductance are large compared to the loss conductances in secondary and primary circuits, respectively. In this case, output power at frequency 2ω will equal input power at frequency ω, and the efficiency will approach 100 percent.

While the choice of a reasonably large load conductance should not present major difficulties, the attainment of a large conversion conductance is not always possible. Examination of (19) reveals the following ways to increase G_c.

(A) By increasing the nonlinearity coefficient θ, a function of varactor physics and manufacturing.

(B) Given θ and selecting G_L from secondary-circuit considerations (mainly $G_{S2}(G_L)G_{T2}$), the only two parameters left are the ratio V_{in}/V_B, and the varactor conductance ωC_B at the bias point.

(C) V_{in}/V_B shall not exceed unity if the varactor is to be operated in the purely capacitive mode. This requirement is met by automatic bias operation.

(D) Finally, ωC_B cannot be increased beyond a certain point without affecting the loss conductance of the varactor. This can be seen from the following.

In a first approximation, the varactor spreading resistance R_S can be represented in shunt configuration as an equivalent loss conductance that must be compared to G_L and G_c in the secondary circuit and in the primary circuit, respectively. From

$$G_L > G_{S2}$$

or

$$G_L > (2\omega)^2 C_B^2 R_S \qquad (21)$$

it can be seen that it becomes increasingly difficult, with increasing frequency, to keep the load conductance larger than the equivalent loss conductance of the varactor in the secondary circuit. The use of varactors with

low capacitances at the bias point can alleviate this difficulty only to a certain point. Small varactor capacitances require careful circuit design to reduce the effect of stray capacitances that reduce the effective nonlinearity of the device and consequently degrade its performance. The same considerations, applied to the primary circuit, yield

$$G_c > G_{S1}$$
$$G_{S1} = (\omega C_B)^2 R_S$$

or

$$G_L < \frac{\theta^2}{4} \cdot \frac{1}{R_S}. \tag{22}$$

Equations (21) and (22) define a rather small margin of G_L values within which efficient varactor operation can be obtained.

$$(2\omega)^2 C_B^2 R_S < G_L < \frac{\theta^2}{4} \cdot \frac{1}{R_S}. \tag{23}$$

Assuming that the losses in the tank circuits are at least one order of magnitude lower than the losses in the varactor, (39) can be written

$$\eta = \frac{G_c}{G_c + G_{S1}} \cdot \frac{G_L}{G_L + G_{S2}}. \tag{24}$$

Examination of (24) reveals that a maximum of η can be obtained by selecting $G_L = \omega C_B \theta$. Introducing $\theta = \frac{1}{2}$ for the abrupt junction, and the cutoff frequency from (4), the efficiency can be written as

$$\eta = \left[\frac{1}{1 + 11.2(f/f_{co})} \right]^2.$$

This equation is plotted in Figure 4 with f_{co} as a parameter and f as the independent variable. It can be seen that efficiencies above 50 percent can be obtained only when the operating frequency is between one and two orders of magnitude below the cutoff frequency.

5. Tuning

The voltage-dependent capacitance of the varactor can give rise to some difficulties when the varactor is operated with automatic bias. This can be seen from the following: Figure 5 represents the radio-frequency voltage V_{RF} across the varactor (curve a) as a function of the bias voltage V_B. This is the familiar response curve of a tuned circuit, symmetrical in the vicinity of its peak point, but otherwise departing from the vertical axis of symmetry. In the same figure, the diode effect is represented by a straight line defining the relationship between radio-frequency voltage and bias voltage. The dependence of the varactor capacitance on bias voltage, C versus V_B, is also shown.

Curve a shows three points of intersection with the straight line. Of these, only points A and B are stable, whereas C is a point of unstable equilibrium. In the region where the radio-frequency response curve shows a greater radio-frequency voltage than the self-bias line, the operating point will shift toward higher bias until a new point of intersection is reached; and conversely, where the radio-frequency response curve shows values lower than the self-bias line, the operating point will shift toward lower bias values until another stable point of intersection is reached. Only at points A and B will radio-frequency voltage and varactor capacitance be in stable equilibrium. Any perturbation about these points will be encountered by a restoring effect.

A normal tuning process consists in gradually approaching the peak radio-frequency voltage across the varactor by varying one reactive component in the circuit. The peak voltage must necessarily lie on both the radio-frequency response curve and the self-bias line and will produce a bias voltage equal to one-half the breakdown voltage of the varactor. In case the tuning process is started at or below point A on curve a with a large inductance, point C is gradually approached by decreasing the inductance, and thus increasing radio-frequency and bias voltage. Increasing the bias voltage, however, reduces the varactor capacitance so that the tuning process is accelerated by a varactor capacitance variation in the right direction. An additional self-

tuning takes place that depends on the diode effect, the Q of the radio-frequency response curve, and the varactor capacitance-bias-voltage characteristic; it can be described as a regeneration process where effect and cause are interrelated and sustain each other. This effect is insignificant so long as tuning takes place far from the resonant peak. However, as tuning progresses beyond point C, it increases rapidly and culminates in a sudden jump that indicates that the stable point B, which lies beyond the maximum of the radio-frequency response curve, is reached. This point will, however, rarely correspond to the desired operating condition. The self-bias produced may be smaller or greater than the required value, and this will result in insufficient power output or overload. Consequently, subsequent adjustment of the applied radio-frequency voltage by means of a variable capacitor in series with the varactor, for instance, and retuning of the tank circuit will be needed to attain a stable operating point with the desired bias voltage V_B. This is shown on curve b.

The same line of reasoning shows that no regenerative effect will take place between points A and C, if resonance is approached by a gradual increase of a variable inductance. The resulting decrease of the capacitance of the varactor will counteract the tuning process. The system will be stabilized by what may be called an inherent negative feedback. Tuning beyond point C, however, will result in the same effect as in the first case described.

The following equations can be used to describe the tuning process.

The total incremental increase in radio-frequency voltage across the tuned circuit is

$$\Delta V_{RF} = \Delta V_{RF}' + \Delta V_{RF}'' = \frac{dV_{RF}}{dL}\Delta L + \frac{dV_{RF}}{dC}\Delta C \quad (25)$$

where L and C are tank inductance and varactor capacitance, respectively. $\Delta V_{RF}''$ is

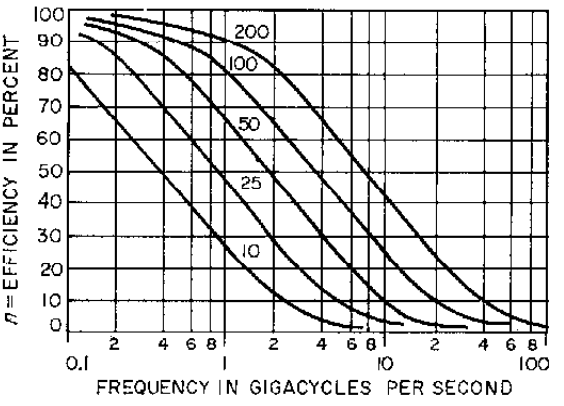


Figure 4—Efficiency of a series doubler as a function of frequency for the indicated values of cutoff frequency in gigacycles per second.

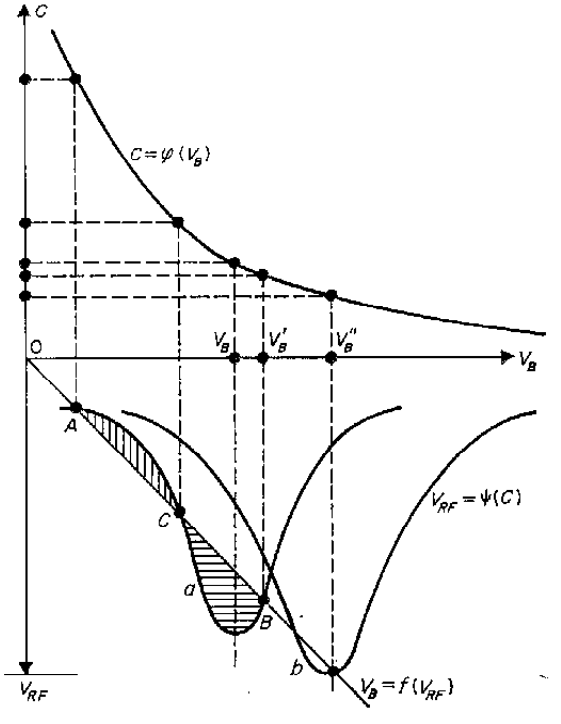


Figure 5—Tuning characteristics of a varactor.

caused by $\Delta V_{RF}'$ and can be stated as follows.

$$\Delta V_{RF}'' = \frac{dV_{RF}}{dC} \cdot \frac{dC}{dV_B} \cdot \frac{dV_B}{dV_{RF}} \Delta V_{RF}'. \quad (26)$$

Sign and absolute value of the right-hand side of this equation depend on the derivatives of three curves in Figure 5, that is,

$$V_{RF} = \psi(C)$$
$$C = \varphi(V_B)$$
$$V_B = \phi(V_{RF}).$$

A positive sign will indicate that the inductance tuning is sustained by the capacitive effect.

Substituting from

$$V_{RF} = \frac{1}{\omega C} \cdot \frac{E}{\left[R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2\right]^{1/2}} \quad (27)$$

with

$$Q = \frac{1}{\omega_r C R}$$
$$1 = \omega^2 L C_r$$
$$V_{RF} = EQ$$

(27) yields

$$V_{RF} \cong \frac{V_{RF}}{\left[1 + Q^2\left(\frac{C - C_r}{C_r}\right)^2\right]^{1/2}}$$

$$\frac{dV_{RF}}{dC} = \frac{-Q^2(C - C_r)V_{RF\ max}}{C_r^2\left[1 + \frac{Q^2}{C_r^2}(C - C_r)^2\right]^{3/2}}$$

$$\frac{dV_B}{dV_{RF}} = \xi$$

$$\frac{dC}{dV_B} = -\theta \cdot K \cdot V_B^{-1-\theta}.$$

Hence,

$$\Delta V_{RF}'' = \frac{Q^2(C - C_r) \cdot \theta \cdot K V_B^{-1-\theta} \cdot \xi}{C_r^2\left[1 + \frac{Q^2}{C_r^2}(C - C_r)^2\right]^{3/2}} \Delta V_{RF}'. \quad (28)$$

The sign of this equation is negative, since resonance is always approached from $C > C_r$. Hence, the tuning process is sustained when dV_{RF}/dL is negative also, that is, when decreasing inductance leads to resonance.

Equation (28) contains other information. The voltage increase will give rise, by virtue of the mechanism described, to an additional increase of radio-frequency voltage, $\Delta V_{RF}'''$, and this process will maintain itself and progress toward a stable point, if

$$\frac{|\Delta V_{RF}'''|}{|\Delta V_{RF}''|} > 1. \quad (29)$$

Evaluating (28), first with $\Delta V_{RF}'$ at the corresponding point of operation, then with $\Delta V_{RF}''$ at the new operating point, will reveal whether (29), and hence the condition for regeneration, is satisfied.

6. Circuit Complexity

To obtain satisfactory harmonic separation in a frequency doubler, the following tuning controls must be provided.

(A) One variable inductance in each circuit to tune out the reactance of the varactor at the fundamental frequency and at the desired harmonic.

(B) One trap in each circuit to provide isolation from the undesired harmonic voltages or currents.

(C) In case a large frequency range must be covered, a matching device in both the primary and the secondary circuits.

This amounts to six tuning controls. Omitting any one of these tuning controls will degrade the operation. In particular, omitting the traps will not allow complete current separation. Power at the fundamental frequency will be dissipated in the load resistance, and power at the generated harmonic will cause losses in the internal resistance of the power source. In addition, in the case of a series configuration, the varactor terminals will not appear short-circuited for all other harmonics. Consequently, current flow at these frequencies through the load and the internal resistance of the generator will increase the losses. Omitting the tuning inductances, on

the other hand, will not allow the two circuits to resonate properly. This will affect the input conductance, degrade the efficiency, and create abnormal matching conditions. Finally, omitting the matching controls will confine efficient operation and adequate power output to a restricted frequency range.

In microwave exciters, large multiplication coefficients are required. Assuming operations in the *X* band, for instance, and a prime frequency source of 50 to 60 megacycles per second followed by transistor stages that bring the frequency up to 150 to 200 megacycles per second, a multiplication factor of 64 is needed. This amounts to a chain of 6 doublers, which in turn sets the number of tuning controls equal to 36. The obvious way to reduce this number of controls is the use of higher-order frequency multipliers. Two quadruplers and one doubler, for instance, would yield a multiplication factor of 32, and 18 controls would be required instead of 36. The quadrupler in a simple circuit configuration, however, consisting of only two tuned circuits, would operate at lower efficiency; in many applications this is preferable to an excessive number of tuning controls. Introducing idler circuits to increase efficiency is, from this point of view, also a poor choice. In fact, a quadrupler with an idler circuit tuned to the second harmonic is equivalent to two doublers making use of the same varactor diode, and almost as complex as two separate doublers.

Figure 6 shows a frequency multiplier in which efficiency has been sacrified to reduce the number of tuning controls. Input frequency is in the range of 443.75 to 531.25 megacycles per second, and the output frequency is 7100 to 8500 megacycles per second. The multiplier makes use of two quadruplers in a series configuration; it consists of three tunable coaxial cavities and a piece of waveguide with an adjustable short. Cavity 1 receives the input signal through loop coupling; it is tunable from 443.75 to 531.25 megacycles per second by movable short (1). A varactor crystal (*MA-4345C*) is mounted at the end of the inner conductor, which reaches through an opening into the second cavity and thus establishes a capacitive link to the second cavity. The coupling is controlled by micrometer (2). Cavity 2 is tunable over the range 1775 to 2125 megacycles per second (control 3). The second and third cavities are connected by aperture coupling. The center conductor of the third cavity bears the second varactor (*MA-4343A*), which reaches (control 4) into the piece of waveguide *RG-51/U* with a movable short that is tunable from 7100 to 8500 megacycles per second, but which cannot be seen in the photograph. The third cavity is tunable by a



Figure 6—Frequency multiplier from about 500 to 8000 megacycles per second. There are 6 tuning controls, 5 of which may be seen. Controls (6) are for the band-pass filter in the output waveguide.

moving short (control 5). A two-cavity band-pass filter ensures additional attenuation of the undesired harmonics and is adjusted by controls (6).

The whole assembly has 6 tuning controls excluding the band-pass filter. Its dimensions are 4.5 inches (11.4 centimeters) by 6 inches (15.2 centimeters) by 20 inches (50.8 centimeters). The electrical performance is given in Figure 7.

7. Appendix

7.1 Series Doubler

$$\sum_{n=-1,-2}^{+1,+2} Q_n \exp[jn\omega t] = \alpha_1 \sum_{n=-1,-2}^{+1,+2} V_n \exp[jn\omega t] + \alpha_2 \{ \sum_{n=-1,-2}^{+1,-2} V_n \exp[jn\omega t] \}^2. \quad (30)$$

Separating fundamental and second-harmonic terms

$$Q_1 \exp[j\omega t] + Q_{-1} \exp[-j\omega t] = (\alpha_1 V_1 + 2\alpha_2 V_{-1} V_2) \exp[j\omega t] + (\alpha_1 V_{-1} + 2\alpha_2 V_1 V_{-2}) \exp[-j\omega t]$$

$$Q_2 \exp[j2\omega t] + Q_{-2} \exp[-j2\omega t] = (\alpha_1 V_2 + \alpha_2 V_1^2) \exp[j2\omega t] + (\alpha_1 V_{-1} + \alpha_2 V_{-1}^2) \exp[-j2\omega t].$$

Fundamental and second-harmonic current components are obtained by differentiating the Q vectors with respect to time. Letting V_1 be V_{-1}, which amounts to defining the time origin, and eliminating the time dependence in the preceeding equations, the following relationships among complex quantities are obtained.

$$\begin{aligned} I_1 &= j\omega(\alpha_1 V_1 + 2\alpha_2 V_1 V_2) \\ I_{-1} &= -j\omega(\alpha_1 V_1 + 2\alpha_2 V_1 V_{-2}) \end{aligned} \quad (31)$$

$$\begin{aligned} I_2 &= j2\omega(\alpha_1 V_2 + \alpha_2 V_1^2) \\ I_{-2} &= -j2\omega(\alpha_1 V_{-2} + \alpha_2 V_1^2). \end{aligned} \quad (32)$$

Introducing the load constraint

$$-I_2 = V_2 |Y_L| \exp[j\varphi] \quad (33)$$

where $Y_L \exp[j\varphi]$ is the load admittance, into (32) yields

$$I_2 = \frac{j2\omega\alpha_2 V_1^2 |Y_L| \exp[j\varphi]}{|Y_L| \exp[j\varphi] + j2\omega\alpha_1}. \quad (34)$$

Resonating the secondary circuit, that is

$$\mathcal{I}m\{|Y_L| \exp[j\varphi]\} = -j2\omega\alpha_1$$

annuls the imaginary part in the denominator and the secondary current appears in the following form.

$$I_2 = \frac{j2\omega\alpha_2 V_1^2 |Y_L| \exp[j\varphi]}{\mathcal{R}e\{|Y_L| \exp[j\varphi]\}}. \quad (35)$$

Substituting (33) and (35) into (31) leads to

$$I_1 = j\omega\alpha_1 V_1 + \frac{4\omega^2\alpha_2^2 V_1^3}{\mathcal{R}e\{|Y_L| \exp[j\varphi]\}}. \quad (36)$$

Then, the input or conversion conductance Y_c is

$$Y_c = \frac{I_1}{V_1} = j\omega\alpha_1 + \frac{4\omega^2\alpha_2^2 V_1^2}{\mathcal{R}e\{|Y_L| \exp[j\varphi]\}}. \quad (37)$$

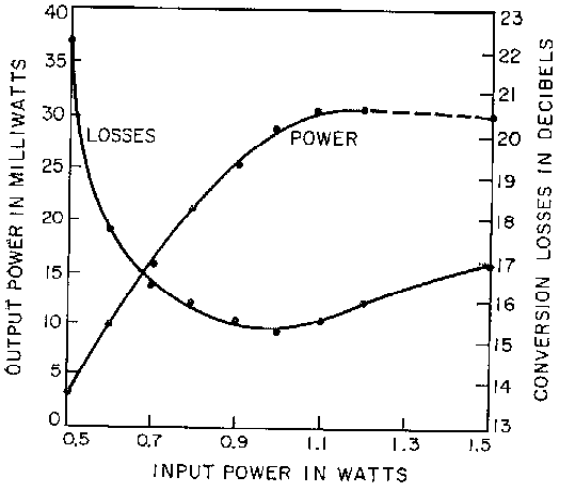


Figure 7—Output power and efficiency of a frequency multiplier operating at 7.8 gigacycles per second.

Resonating the primary circuit at the fundamental frequency will cancel the imaginary part of Y_c; introducing G_L for the real part of the load admittance, (37) can be written.

$$G_c = \frac{4\omega^2\alpha_2^2 V_1^2}{G_L}. \qquad (38)$$

One last step remains to be done: the computation of the Taylor series coefficients α_1 and α_2; α_2 will determine the numerical value of G_c, whereas α_1 determines the admittance that will have to be resonated in the primary and secondary circuits.

The Taylor expansion of the charge q as a function of the applied voltage can be written

$$q(V) = q(V_B) + \alpha_1(V - V_B) + \alpha_2(V - V_B)^2 + \cdots$$

where

$$\alpha_1 = \frac{1}{1!}\frac{dq_{(V)}}{dV}\bigg|_{V=V_B}$$

$$\alpha_2 = \frac{1}{2!}\frac{d^2q_{(V)}}{dV^2}\bigg|_{V=V_B}.$$

Making use of (7)

$$\alpha_1 = C_B$$
$$\alpha_2 = -\tfrac{1}{2}\cdot\theta\cdot C_B\cdot V_B^{-1}.$$

With this expression for α_2 and since the driving or input voltage $V_{in} = 2V_1$, (38) appears in the following final form

$$G_c = \frac{\omega^2 C_B^2}{G_L}\cdot\frac{\theta^2}{4}\cdot\left(\frac{V_{in}}{V_B}\right)^2. \qquad (19)$$

7.2 Efficiency and Power Output

An elementary circuit analysis of a varactor multiplier shows the effects of varactor losses and of the external circuit properties on the efficiency of the chosen assembly.

The following assumptions are made.

(A) The varactor is resonated at the fundamental frequency and at the desired harmonic. Power flow is restricted to only these two frequencies.

(B) The fundamental and harmonic currents are confined to strictly determined paths, and losses in the respective circuits are due to currents at only these two frequencies.

Under these conditions the multiplier circuit can be split into two separate circuits as in Figure 8. The first circuit operates at the fundamental frequency and contains the power source with its internal conductance G_1, the loss conductance of the tuned circuit G_{T1}, the loss conductance of the varactor diode at the fundamental frequency G_{S1}, and the "conversion conductance" G_c, which is the real part of the input admittance and constitutes the only useful element in the primary circuit.

The secondary circuit operates at the chosen harmonic frequency and contains the load conductance G_L, the loss conductance of the tank circuit G_{T2}, and the loss conductance of the varactor diode at the chosen harmonic G_{S2}.

7.2.1 *Primary Circuit*

Input power

$$P_1 = \frac{I_1^2}{G_{T1} + G_{S1} + G_c}.$$

Power absorbed by the conversion conductance

$$P_c = \frac{I_1^2 G_c}{(G_{T1} + G_{S1} + G_c)^2}.$$

Efficiency of the primary circuit

$$\eta_1 = \frac{P_c}{P_1} = \frac{G_c}{G_{T1} + G_{S1} + G_c}.$$

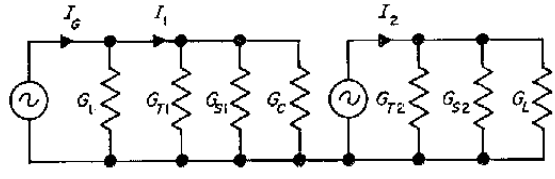


Figure 8—Equivalent circuit for a frequency multiplier.

7.2.2 Secondary Circuit

Input power

$$P_2 = \frac{I_2^2}{G_{T2} + G_{S2} + G_L}.$$

Output power

$$P_L = \frac{I_2^2 G_L}{(G_{T2} + G_{S2} + G_L)^2}.$$

Efficiency of the secondary circuit

$$\eta_2 = \frac{P_L}{P_2} = \frac{G_L}{G_{T2} + G_{S2} + G_L}.$$

Over-all efficiency

$$\eta = \eta_1 \eta_2 = \frac{G_L G_c}{(G_{T1} + G_{S1} + G_c)(G_{T2} + G_{S2} + G_L)}. \quad (39)$$

These equations show that the over-all circuit efficiency and output power at the selected harmonic depend on the loss conductances, the load conductance, and the conversion conductance. While the varactor losses are determined primarily by fabrication methods, the conversion conductance depends on operating conditions.

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9. Acknowledgments

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W. Janeff was born in Burgas, Bulgaria, in October 1920. He received a Diplom Ingenieur degree from the Technische Hochschule in Vienna, Austria, in 1942 and a Ph.D. degree in 1945 from the Technische Hochschule in Dresden, Germany.

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Institute of Radio Engineers Celebrates 50th Anniversary

Founded in 1912, the Institute of Radio Engineers published in May 1962 an unusually interesting commemorative issue of the Proceedings of the IRE celebrating its 50th anniversary.

Following a factual presentation of the history of the Institute, over 50 of its Fellows contributed to a predictive symposium on communications and electronics in 2012. Henri Busignies, Vice President and General Technical Director of International Telephone and Telegraph Corporation, wrote in two fields, "Communication and Navigation" and "Electronic Spectro-Analysis of Chemical Compounds."

The major part of the editorial contents consists of 28 sections each prepared with the assistance of the appropriate Professional Group of the Institute and covering its field. They are abundantly rich in historical and tutorial material. In the section on aerospace and navigation is a paper on "The Air Traffic Control Equipment Subsystem—Present and Future" by P. C. Sandretto of the United States Defense Group of the International Telephone and Telegraph Corporation, and in the section on communication systems is one on "Radio Receivers—Past and Present" by Christopher Buff, Vice President and Chief Engineer of American Cable & Radio Corporation.

It is interesting to note that on 1 January 1963, the Institute of Radio Engineers and the 78-year-old American Institute of Electrical Engineers will combine and become the Institute of Electrical and Electronics Engineers.

Contribution to Studies of Overflow Traffic

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1. Simple Overflow Problem

1.1 General Equations for Distribution: Adaptation of Riordan Method

Firstly consider a primary group of x trunks to which a random traffic a is offered and, secondly, a secondary group of y trunks to care for the traffic rejected by the primary group, that is, the overflow traffic. We have to calculate the probabilities $\theta(0), \theta(1), \cdots, \theta(n)$ of occupation of $0, 1, \cdots, n$ unspecified paths (and only this number) among the total number y of them, these probabilities being independent of the state of the primary group.

If we let $f(m, n)$ be the probability of m paths being busy in the lower group of x trunks and n being also busy among the y circuits of the upper group, the equations of state giving the probabilities $f(m, n)$ are

$$\left.\begin{aligned}(a+m+n)f(m,n) - (m+1)f(m+1,n) - (n+1)f(m,n+1) - af(m-1,n) &= 0\\ (a+x+n)f(x,n) - af(x,n-1) - (n+1)f(x,n+1) - af(x-1,n) &= 0\\ (x+y)f(x,y) - af(x,y-1) - af(x-1,y) &= 0.\end{aligned}\right\} \quad (1)$$

This last equation having been introduced to take into account the fact that we have to consider a limited number of overflow trunks, instead of an unlimited group as Riordan has done. [1]. The factorial moment-generating function defined by

$$M(m,t) = \sum_{k=0}^{y} M_k(m)\frac{t^k}{k!} = \sum_{n=0}^{y} f(m,n)(1+t)^n \quad (2)$$

satisfies the following equations.

$$\left.\begin{aligned}(a+m)M(m,t) + t(D/dt)M(m,t) - (m+1)M(m+1,t) - aM(m-1,t) &= 0\\ (-at+x)M(x,t) + t(D/dt)M(x,t) - aM(x-1,t) + af(x,y)t(1+t)^y &= 0.\end{aligned}\right\} \quad (3)$$

In a way similar to Riordan's, the following recurrences may be found.

$$\left.\begin{aligned}(a+m+k)M_k(m) - (m+1)M_k(m+1) - aM_k(m-1) &= 0\\ (x+k)M_k(x) - akM_{k-1}(x) - aM_k(x-1) + a\frac{k!}{y!}\binom{y}{k-1}M_y(x) &= 0.\end{aligned}\right\} \quad (4)$$

The introduction of the new generating function

$$G_k(u) = \sum_{m=0}^{x} M_k(m)u^m \quad (5)$$

leads to the same equations

$$(a+k-u)G_k(u) + (u-1)(D/du)G_k(u) = 0 \quad (6)$$

and by integration

$$G_k(u) = ce^{au}(1-u)^{-k} \quad \text{with} \quad c = M_k(0).$$

The quantities $M_k(0)$ in this case are of course different from the corresponding quantities in the case of an unlimited overflow. For their calculation, we could also use the last of (6), but in fact, as will be seen later, we shall only be interested in the values of $M_k(x)$. By replacing $M_k(x-1)$ in (4) by its value

$$M_k(x-1) = \sigma_k(x-1)M_k(0) = \frac{\sigma_k(x-1)}{\sigma_k(x)}\cdot M_k(x)$$

we obtain

$$\left[x+k-a\frac{\sigma_k(x-1)}{\sigma_k(x)}\right]M_k(x) = akM_{k-1}(x) - a\frac{k!}{y!}\binom{y}{k-1}M_y(x). \quad (7)$$

We recall that

$$\sigma_k(x) = \sum_{j=0}^{x}\binom{k-1+j}{k-1}\frac{a^{x-j}}{(x-j)!}$$

and that the quantities σ satisfy a good number of relations of which we shall use the following.

hence
$$\left.\begin{aligned}(m+k)\sigma_k(m)-a\sigma_k(m-1)&=k\sigma_{k+1}(m)\\ k\sigma_{k+1}(m)&=(m+k-a)\sigma_k(m)+a\sigma_{k-1}(m)\\ \sigma_{k+1}(m)&=\sigma_{k+1}(m-1)+\sigma_k(m).\end{aligned}\right\} \tag{8}$$

By use of the first relation for $m = x$, we arrive at the required recurrence

$$M_{k-1}(x)=\frac{1}{a}\cdot\frac{\sigma_{k+1}(x)}{\sigma_k(x)}\,M_k(x)+\frac{M_y(x)}{(y-k+1)!}. \tag{9}$$

This relation has already been given by Bech [2] and Le Gall [3] under slightly different forms.

1.2 Probabilities of Overflow Distribution

Having thus adapted Riordan's method to the particular case of a limited number of overflow trunks, we shall now calculate the distribution probabilities $\theta(n) = \sum_{m=0}^{x} f(m, n)$.

A simple result, which appears to be formulated here for the first time in studies of this question, is the following

$$\theta(n) = (a/n)f(x, n-1). \tag{10}$$

In fact, by summing m in (3), we obtain

$$aM(0,t) + t\frac{dM(0,t)}{dt} - M(1,t) = 0$$

$$(a+1)M(1,t) + t\frac{dM(1,t)}{dt} - 2M(2,t) - aM(0,t) = 0$$

$$(a+2)M(2,t) + t\frac{dM(2,t)}{dt} - 3M(3,t) - aM(1,t) = 0$$

$$\cdots$$

$$(a+x-1)M(x-1,t) + t\frac{dM(x-1,t)}{dt} - xM(x,t) - aM(x-2,t) = 0.$$

$$(-at+x)M(x,t) + t\frac{dM(x,t)}{dt} - aM(x-1,t) = af(x,y)t(1+t)^y$$

$$\cdots$$

$$-atM(x,t) + t\sum_{m=0}^{x}\frac{dM(m,t)}{dt} = af(x,y)t(1+t)^y$$

as
$$\sum\frac{dM(m,t)}{dt} = \theta(1) + 2(1+t)\theta(2) + 3(1+t)^2\theta(3) + \cdots + y(1+t)^{y-1}\theta(y).$$

We deduce $n\theta(n) = af(n, n-1)$ by equating the coefficients of the powers of t.

We have thus

$$\begin{aligned} \theta(y) &= (a/y)f(x, y-1) \\ \theta(y-1) &= (a/y-1)f(x, y-3) \\ &\cdots \\ \theta(1) &= a \cdot f(x, 0) \\ \theta(0) &= 1 - \theta(1) - \theta(2) - \cdots - \theta(y). \end{aligned}$$

To calculate the quantities

$$f(x, y-1),\ f(x, y-2),\ \text{et cetera},$$

we shall use the equations of the factorial moments.

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$$\begin{aligned} y!f(x, y) &= M_y(x) \\ (y-1)!f(x, y-1) + y!f(x, y) &= M_{y-1}(x) \\ (y-2)!f(x, y-2) + (y-1)!f(x, y-1) & \\ + \frac{y!}{2!}f(x, y) &= M_{y-2}(x), \end{aligned}$$

et cetera.

The quantities $M_k(x)$ are given by (9), which we shall write from now on as

$$M_{k-1}(x) = \frac{S_{k+1}}{a} M_k(x) - \frac{M_y(x)}{(y-k+1)!}$$

with the notation

$$S_{k+1} = \frac{\sigma_{k+1}(x)}{\sigma_k(x)}.$$

Thus we obtain

$$M_{y-1}(x) = \left(1 + \frac{S_{y+1}}{a}\right) M_y(x)$$

$$M_{y-2}(x) = \left(\frac{1}{2!} + \frac{S_y}{a} + \frac{S_y \cdot S_{y+1}}{a^2}\right) M_y(x)$$

$$M_{y-3}(x) = \left(\frac{1}{3!} + \frac{1}{2!}\frac{S_{y-1}}{a} + \frac{S_{y-1} \cdot S_y}{a^2} + \frac{S_{y-1} \cdot S_y \cdot S_{y+1}}{a^3}\right) M_y(x)$$

$$M_{y-4}(x) = \left(\frac{1}{4!} + \frac{1}{3!}\frac{S_{y-2}}{a} + \frac{1}{2!}\frac{S_{y-2} \cdot S_{y-1}}{a^2} + \frac{S_{y-2} \cdot S_{y-1} \cdot S_y}{a^3} + \frac{S_{y-2} \cdot S_{y-1} \cdot S_y \cdot S_{y+1}}{a^4}\right) M_y(x)$$

et cetera.

The equations giving the probabilities $\theta(n)$ are thus

$$\left.\begin{aligned} \theta(y) &= S_{y+1}E_{x+y}(a) \quad \text{as} \quad f(x, y) = F_{x+y}(a) \\ \theta(y-1) &= \frac{y}{a}\left[S_y \cdot S_{y+1} + a(S_y - S_{y+1})\right]E_{x+y}(a) \\ \theta(y-2) &= \frac{y(y-1)}{a^2}\left[S_{y-1} \cdot S_y \cdot S_{y+1} + aS_y(S_{y-1} - S_{y+1}) + \frac{a^2}{2!}(S_{y-1} + 2S_y + S_{y+1})\right]E_{x+y}(a) \end{aligned}\right\} \quad (11)$$

et cetera.

They may be written in the following way, which may be clearer.

$$\left.\begin{aligned} \theta(y) &= \frac{\sigma_{y+1}}{\sigma_y} \cdot E_{x+y}(a) \\ \theta(y-1) &= \frac{y}{a}\left[\frac{\sigma_{y+1}}{\sigma_{y-1}} + a\left(\frac{\sigma_y}{\sigma_{y-1}} - \frac{\sigma_{y+1}}{\sigma_y}\right)\right]E_{x+y}(a) \\ \theta(y-2) &= \frac{y(y-1)}{a^2}\left[\frac{\sigma_{y+1}}{\sigma_{y-2}} + a\left(\frac{\sigma_y}{\sigma_{y-2}} - \frac{\sigma_{y+1}}{\sigma_{y-1}}\right) + \frac{a^2}{2!}\left(\frac{\sigma_{y-1}}{\sigma_{y-2}} - 2\frac{\sigma_y}{\sigma_{y-1}} + \frac{\sigma_{y+1}}{\sigma_y}\right)\right]E_{x+y}(a) \end{aligned}\right\} \quad (12)$$

et cetera.

The above expressions become more and more complex but the most useful $\theta(y)$, $\theta(y-1)$, $\theta(y-2)$ are relatively simple especially if we can work out a simple way of obtaining the quantities S_{y+1}, S_y, S_{y-1}, et cetera.

1.3 Approximate Calculation of S

We have $S_1 = (1/E_x(a))$

and $S_2 = x + 1 - a + aE_x(a)$

and according to the recurrent relation (8)

$$kS_{k+1} = x + k - a + (a/S_k). \quad (13)$$

Thus if the number of the lines of the overflow group is not too high, it is quite easy to calculate all the values, each one in turn, starting from the first.

But when, on the contrary the number y is high enough to result in a long and tedious computation, it seems easier to calculate directly the quantities S_{y+1}, S_y, et cetera, by making use of an approximation.

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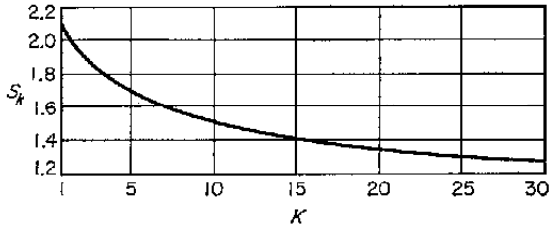


Figure 1.

If we make S_k satisfy (13), it is clear from Figure 1 that its curve quickly assumes a flat shape with an asymptote of 1, so that an approximation of S_{k+1} may be derived from

$$(k+\tfrac{1}{2})S_{k+1}^2 - (x+k+\tfrac{1}{2}-a)S_{k+1} - a = 0. \quad (14)$$

This equation has a positive root that gives a good approximate value of S_{k+1}.

1.4 Numerical Application

We shall examine a particular example using the following data.

Primary group $x = 12$ lines
Overflow group $y = 20$ lines
Offered traffic $a = 21$ erlangs.

The exact values of S_k calculated by (13) are given in Table 1 and plotted in Figure 1.

TABLE 1
Values of S_k

S_1	2.10825	S_{11}	1.48771	S_{21}	1.33205
S_2	1.96085	S_{12}	1.46506	S_{22}	1.32215
S_3	1.85482	S_{13}	1.44449	S_{23}	1.31287
S_4	1.77395	S_{14}	1.42600	S_{24}	1.30415
S_5	1.70950	S_{15}	1.40904	S_{25}	1.29593
S_6	1.65683	S_{16}	1.39359	S_{26}	1.28818
S_7	1.61247	S_{17}	1.37931	S_{27}	1.28085
S_8	1.57478	S_{18}	1.36617	S_{28}	1.27390
S_9	1.54190	S_{19}	1.35397	S_{29}	1.26731
S_{10}	1.51329	S_{20}	1.34263	S_{30}	1.26105

The exact values of the occupation probabilities are

$$\theta(20) = 1.33205 \times 0.00593 = 0.00793; \quad E_{32}(21) = 0.005953$$

$$\theta(19) = \frac{20}{21}(1.78845 + 0.22226)0.005953 = 0.01140$$

$$\theta(18) = \frac{20 \times 19}{21 \times 21}(2.42150 + 0.61804 + 0.16577)0.005953 = 0.01644$$

$$\theta(17) = \frac{20 \times 19 \times 18}{21 \times 21 \times 21}(3.3082 + 1.3028 + 0.5393 + 0.1851)0.005953 = 0.02345.$$

The approximate values of S given by (14) are

$$S_{21} = 1.3308, \quad S_{20} = 1.3413, \quad S_{19} = 1.3527, \quad S_{18} = 1.3649$$

hence

$$\theta(20) = 1.3308 \times 0.005953 = 0.0079$$

$$\theta(19) = \frac{20}{21}(1.7850 + 0.2205)0.005953 = 0.0114$$

$$\theta(18) = \frac{20 \times 19}{21 \times 21}(2.4146 + 0.6169 + 0.1984)0.005953 = 0.0166$$

$$\theta(17) = \frac{20 \times 19 \times 18}{21 \times 21 \times 21}(3.2957 + 1.2992 + 0.5621 + 0)0.005953 = 0.022.$$

As may be seen, this approximation may be considered to be satisfactory.

It should be noted that values of S_{19}, S_{18}, et cetera, drawn from the recurrent relation (13) by making use of the approximate value of S_{20} obtained from (14) are less accurate than the corresponding values directly obtained from (14).

1.5 Variance of Overflow Distribution

The mean of the distribution $\bar{m}$ equals $aE_x(a) - aE_{x+y}(a) = \alpha - \beta$ if we represent by α the overflow traffic and by β the traffic in the last instance.

The overflow distribution variance v equals $\sum_0^y n^2\theta(n) - \bar{m}^2$.

$$\begin{aligned} n^2\theta(n) &= anf(x, n-1) = a(n-1)f(x, n-1) + af(x, n-1) \\ \sum n^2\theta(n) &= a[M_1(x) - yf(x, y) + M_0(x) - f(x,y)] \\ &= a[M_1(x) + M_0(x) - (1+y)f(x, y)] \end{aligned}$$

$$\begin{aligned} M_1(x) &= E_x(a) \\ M_1(x) &= (a/S_2)[M_0(x) - f(x, y)] \end{aligned}$$

hence after reduction

$$\sum n^2\theta(n) = \frac{a(\alpha - \beta)}{x + 1 + \alpha - a} + \alpha - \beta(1 + y)$$

and

$$v = \frac{a(\alpha - \beta)}{x + 1 + \alpha - a} + \alpha - \beta(1 + y) - (\alpha - \beta)^2.$$

1.6 General Equations for Factorial Moments

By fully solving (9), it is found

$$M_y(x) = \frac{a^y\sigma_0(x)}{\sigma_{y+1}(x)} \cdot \frac{\sigma_{y+1}(x)}{\sigma_{y+1}(x) + a\sigma_y(x) + \frac{a^2}{2!}\sigma_{y-1}(x) + \cdots + \frac{a^y}{y!}\sigma_1(x)}$$

$$M_{y-1}(x) = \frac{a^{y-1}\sigma_0(x)}{\sigma_y(x)} \cdot \frac{\sigma_{y-1}(x) + a\sigma_y(x)}{\sigma_{y+1}(x) + a\sigma_y(x) + \frac{a^2}{2!}\sigma_{y-1}(x) + \cdots + \frac{a^y}{y!}\sigma_1(x)}$$

$$\cdots$$

$$M_1(x) = \frac{a\sigma_0(x)}{\sigma_2(x)} \cdot \frac{\sigma_{y+1}(x) + a\sigma_y(x) + \cdots + \frac{a^{y-1}}{(y-1)!}\sigma_2(x)}{\sigma_{y+1}(x) + \cdots + \frac{a^y}{y!}\sigma_1(x)}.$$

If we represent by $\mu_y(x)$, $\mu_{y-1}(x)$, $\cdots$, $\mu_0(x)$, the moments of the unlimited overflow distribution, $M_y(x)$ may be written

$$M_y(x) = \mu_y(x) \frac{1}{1 + \frac{\mu_y(x)}{\mu_{y-1}(x)} + \frac{1}{2!}\frac{\mu_y(x)}{\mu_{y-2}(x)} + \frac{1}{3!}\frac{\mu_y(x)}{\mu_{y-3}(x)} + \cdots + \frac{1}{y!}\frac{\mu_y(x)}{\mu_0(x)}}. \tag{15}$$

This expression appears as a generalization of the Erlang equation.

2. Composition of Overflow Traffic from 2 Primary Groups

Let us consider 2 primary groups of trunks with the following characteristics.

x trunks for the first group offered traffic (poissonian) a

x' trunks for the second group, offered traffic (poissonian) b.

We shall denote by $f(m, m', n)$ the probability for simultaneously having m and m' trunks busy in the primary group and also n trunks busy in the overflow group.

2.1 Unlimited Overflow Group. Moments of the Unlimited Overflow Distribution

The engagement of trunks in the overflow group results from the calls rejected by both primary groups. Let us write $n = \nu + \nu'$. The distributions ν and ν' are independent in the case of the unlimited overflow. The factorial moments of the resulting overflow distribution are the factorial moments of a sum of 2 independent variables. It is known that the order k moment, for the occupation situation (m, m') in the primary groups reads

$$M_k^{(m)}(m, m') = \mu_k^{(\nu)}(m) \cdot \mu_0^{(\nu')}(m') + \binom{k}{1} \mu_{k-1}^{(\nu)}(m) \cdot \mu_1^{(\nu')}(m') + \binom{k}{2} \mu_{k-2}^{(\nu)}(m) \cdot \mu_2^{(\nu')}(m') + \cdots + \mu_0^{(\nu)}(m) \cdot \mu_k^{(\nu')}(m'). \tag{16}$$

If we denote by $\mu_k^{(\nu)}(m)$ and $\mu_k^{(\nu')}(m)$ the factorial moments of order k of the distributions ν and ν', respectively.

For instance, the second moment has the following value

$$M_2(m, m') = \frac{a^2\sigma_0(x)\sigma_2(m)}{\sigma_3(x)\sigma_2(x)} \cdot \frac{\sigma_0'(m')}{\sigma_1'(x')} + 2a\frac{\sigma_0(x)\sigma_1(m)}{\sigma_2(x)\sigma_1(x)} \cdot b \cdot \frac{\sigma_0'(x)\sigma_1'(m')}{\sigma_2'(x)\sigma_1'(x')} + \frac{\sigma_0(m)}{\sigma_1(x)} \cdot b^2 \cdot \frac{\sigma_0(x')\sigma_2'(m')}{\sigma_3'(x')\sigma_2'(x')}$$

the polynomials σ and σ' having to be taken with their respective values in each group.

2.2 Limited Overflow Group

Let us now assume the overflow group to be limited to a number of trunks y. The equations of state analogous to (1) become

$$(a+b+m+m'+n)f(m,m',n)-(m+1)f(m+1,m',n)-(m'+1)f(m,m'+1,n)$$
$$-(n+1)f(m,m',n+1)-af(m-1,m',n)-bf(m,m'-1,n)=0. \quad (17)$$

If $m'=x'$

$$(a+b+m+x'+n)f(m,x',n)-(m+1)f(m+1,x',n)-f(m,x',n+1)$$
$$-af(m-1,x',n)-bf(m,x'-1,n)-bf(m,x',n-1)=0. \quad (18)$$

For $m=x$ and $m'=x'$

$$(a+b+x+x')f(x,x',n)-(n+1)f(x,x',n+1)-af(x-1,x',n)$$
$$-bf(x,x'-1,n)-(a+b)f(x,x',n-1)=0. \quad (19)$$

Finally for $n=y$, the term with $(n+1)$ disappears. If now $m=x$, the term with $(m+1)$ disappears in (18) as does the quantity a in the first term.

If lastly $m=x$, $m'=x'$, and $n=y$, (19) reduces to

$$(x+x'+y)f(x,x',y)-af(x-1,x',y)-bf(x,x'-1,y)-(a+b)f(x,x',y-1)=0. \quad (20)$$

Let us once more denote

$$M(m,m',t)=\sum_{k=0}^{y}M_k(m,m')\frac{t^k}{k!}.$$

It is found that the generating function of factorial moments, the function M, satisfies the following equations.

$$\left.\begin{aligned}
&(a+b+m+m')M(m,m',t)+t(d/dt)M(m,m',t)-(m+1)M(m+1,m',t)\\
&\quad-(m'+1)M(m,m'+1,t)-aM(m-1,m',t)-bM(m,m'-1,t)=0.\\
&(a+m+x'-bt)M(m,x',t)+t(d/dt)M(m,x',t)-(m+1)M(m+1,x',t)\\
&\quad-aM(m-1,x',t)-bM(m,x'-1,t)+bt(1+t)^yf(m,x',y)=0.\\
&(b+x+m'-at)M(x,m',t)+t(d/dt)M(x,m',t)-(m'+1)M(x,m'+1,t)\\
&\quad-aM(x-1,m',t)-bM(x,m'-1,t)+at(1+t)^yf(x,m',y)=0.\\
&[x+x'-(a+b)t]M(x,x',t)+t(d/dt)M(x,x',t)-aM(x-1,x',t)\\
&\quad-bM(x,x'-1,t)+(a+b)t(1+t)^yf(x,x',y)=0.
\end{aligned}\right\} \quad (21)$$

By making use of a process analogous to that of the first part, we obtain the recurrent relations applying to the factorial moments:

$$(a+b+m+m'+k)M_k(m,m')-(m+1)M_k(m+1,m')-(m'+1)M_k(m,m'+1)$$
$$-aM_k(m-1,m')-bM_k(m,m'-1)=0. \quad (22)$$

$$(b+x+m'+k)M_k(x,m')-(m'+1)M_k(x,m'+1)-bM_k(x,m'-1)$$
$$-aM_k(x-1,m')-akM_{k-1}(x,m')+\frac{k!}{y!}\binom{y}{k-1}aM_y(x,m')=0. \quad (23)$$

$$(a + m + x' + k)M_k(m, x') - (m + 1)M_k(m + 1, x') - aM_k(m - 1, x') - bM_k(m, x' - 1) - bkM_{k-1}(m, x') + \frac{k!}{y!}\binom{y}{k-1} bM_y(m, x') = 0. \quad (24)$$

$$(x + x' + k)M_k(x, x') - bM_k(x, x' - 1) - aM_k(x - 1, x') - (a + b)kM_{k-1}(x, x') + \frac{k!}{y!}\binom{y}{k-1}(a + b)M_y(x, x') = 0. \quad (25)$$

2.3 Loss Probabilities

The loss probability to be applied to the traffic a offered to the first group is the probability for the lines of both this primary group and the overflow group being engaged in totality. This probability is $\sum_{m'=0}^{x'} f(x, m', y)$ that we shall write in brief form as $f(x, y)$ and we have $f(x, y) = \frac{1}{y!}\sum_{m'=0}^{x'} M_y(x, m') = \frac{1}{y!} M_y(x)$.

We are thus led to sum up (23) and (25) over all possible values of m' and obtain

$$(x + k)M_k(x) - aM_k(x - 1) - akM_{k-1}(x) - bkM_{k-1}(x, x') + \frac{k!}{y!}\binom{y}{k-1} aM_y(x) + \frac{k!}{y!}\binom{y}{k-1} bM_y(x, x') = 0. \quad (26)$$

and in particular

$$(x + y)M_y(x) - aM_y(x - 1) - ayM_{y-1}(x) - b_yM_{y-1}(x, x') + ayM_y(x) + b_yM_y(x, x') = 0.$$

If one compares the latter equation to the last of (4) and notes that such quantities as $M_{k-1}(x, x')$ and $M_y(x, x')$ are comparatively small, one is led to expect that an equation analogous to (15) may give an approximate value of $M_y(x)$, provided the quantities $\mu(x)$ are given the actual values resulting from (16) of Section 2.2. The greater the number of trunks of the primary groups, the better should be the approximation, but on the other hand the result may not be so good in the case of overflow groups built up of a large number of lines.

2.4 Verification of Approximate Equation

The numerical results published hitherto are very few and concern only primary groups of rather small size. This is due to the difficulty of solving the important set of linear equations involved in any method, despite the substantial simplifications proposed by Le Gall [6] for the handling of the matrix calculations.

G. Neovius [4] has treated the case of $x = x' = 3$ for various traffic values: $a = b = 1$, 2, 3, 4, and 5 erlangs. A comparison of the exact results and the approximate values given by the proposed equation are shown in Table 2, in which E_y denotes the loss to be applied to the primary traffic (traffic offered).

TABLE 2
COMPARISON OF EXACT AND CALCULATED LOSS TO PRIMARY TRAFFIC

a	E_1 in Percent		E_2 in Percent		E_4 in Percent	
	Exact	Approximate	Exact	Approximate	Exact	Approximate
1	1.73716	1.750	0.39562	0.402	0.07678	0.0785
2	11.54760	11.684	5.62963	5.764	2.43797	2.5225
3	24.56570	24.754	16.39860	16.663	10.21233	10.465
4	35.93526	36.095	27.70477	27.968	20.53455	20.842
5	44.91999	45.039	37.33561	37.547	30.30786	30.435

2.5 Case of Several Primary Groups

The generalization of (16) for the case of more than 2 primary groups is self-evident in the same way that M_2, for instance, corresponds to a symbolic power $(a + b)^2$ in the case of 2 primary groups, the same quantity corresponds to the symbolic power $(a + b + c + d)^2$ in the case of 4 primary groups.

G. Bretschneider [7] has published some results concerning 4 primary groups of $x = x' = x'' = x''' = 4$ lines for traffic values $a = b = c = d = 0.75$, 2.25, 3, 4.50, and 10 erlangs, and an overflow group of $y = 6$ lines. Comparative results for $a = 2.25$ and 3 erlangs are also given in Table 3.

TABLE 3
LOSS PROBABILITY

a	Exact	Approximate
2.25	0.15425×10^{-2}	0.1675×10^{-2}
3	0.15539×10^{-1}	0.1676×10^{-1}

The approximate equations, satisfactory as they be, fail however to give a better accuracy than the approximate method of R. I. Wilkinson. Nevertheless, they may be used to investigate the important problem of determining the loss that is to be assigned to each element of traffic among those that compose the total resulting overflow.

2.6 Remarks

In the same way as Poisson's equation gives a lower, though valuable approximation for the Erlang equation, we may also consider the probability $f(x, y)$ of having x busy lines in the concerned primary group and simultaneously y calls in the unlimited overflow group. The latter probability is given by

$$f(x, y) = \frac{1}{y!}\left[M_y(x) - M_{y+1}(x) + \frac{1}{2!}M_{y+2}(x) - \frac{1}{3!}M_{y+3}(x)\cdots\right]$$

in which for instance

$$M_y(x) = \sum_{m'=0}^{x'} M_y(x, m') = \frac{a^y\sigma_0(x)}{\sigma_{y+1}(x)} + \binom{y}{1}\frac{a^{y-1}\sigma_0(x)}{\sigma_y(x)}\cdot\frac{b\sigma_0'(x)}{\sigma_1'(x')} + \binom{y}{2}\frac{a^{y-2}\sigma_0(x)}{\sigma_{y-1}(x)}\cdot\frac{b^2\sigma_0'(x')}{\sigma_2'(x')} + \cdots.$$

The series giving $f(x, y)$ converges slowly in most cases and requires a long calculation. It gives a result that is, by far, less accurate than the result of the method of Section 2.4 even in the case of overflow groups of very-low loss probabilities.

2.7 Remarks

By summing up (22) $\cdots$ (25) over all possible values of m and m', it is found for the M_k

factorial moment of distribution of the overflow group limited to y lines

$$M_k = a\left[M_{k-1}(x) - \frac{k!}{y!}\binom{y}{k-1} M_y(x)\right] + b\left[M_{k-1}(x') - \frac{k!}{y!}\binom{y}{k-1} M_y(x')\right].$$

This result is nothing more than a generalization of the results of Section 1.2, which permits us to write

$$y\theta(y) = af(x, y-1) + bf(x', y-1)$$
$$(y-1)\theta(y-1) = af(x, y-2) + bf(x', y-2).$$

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Interregister Multifrequency-Code Signalling for Telephone Switching in Europe

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Important developments in the technique of signalling used in telephone switching have taken place during the past few years in Europe and will have a far-reaching effect on the basic principles underlying switching methods as well as on the operating facilities that will be provided for both national networks and for international connections in Europe. Since agreement on these methods has now been reached among many of the leading telephone administrations in Europe, a description of these developments, their basic philosophy, and their implications is thought to be timely.

1. History

The need for fast and reliable signalling methods for telephone switching was recognized independently around 1956 or 1957 by at least three organizations working on new or improved switching systems in Europe.

At this time, the Netherlands administration considered the use of multifrequency-code signalling for the transfer of numerical information on their toll network as an essential requirement for reducing connection time. In Belgium a fast signalling method was needed for international subscriber's toll dialling to signal the identity of the calling subscriber, during the establishment of the connection, via national toll lines to recording equipment that was to be provided for automatic message accounting at the centrally located international toll office. More than one manufacturer was engaged in the development of crossbar switching equipment in which a fast signalling system operating via 2-wire interoffice junctions would provide means for simplifying the common control equipment. For example, at the Bell Telephone Manufacturing Company in Antwerp, Belgium, plans were conceived for using a uniform code of multifrequency signals for transmitting all numerical and other information needed in establishing telephone connections and for signalling between different stages of selection in the same central office or in different offices, both on a national and international scale.

The issuance of technical recommendations on signalling methods for international toll lines being in the province of the Comité Consultatif International Télégraphique et Téléphonique, with which international body many administrations had collaborated to design two recommended signalling systems for use on international toll lines in Europe and the Mediterranean basin, it was recognized that any system proposed for international use would have to be submitted to this body for discussion.

Informal discussions with several European administrations showed, however, a widespread desire to consider improved methods of signalling with emphasis on their use for national systems. Late in 1957 the Netherlands administration took the initiative to call a conference on these questions; all administrations and the leading manufacturers of telephone equipment in Western Europe were invited. This conference took place in March 1958 and was attended by representatives of 8 administrations, 3 large manufacturing organizations, and furthermore by observers from France, the United Kingdom, and in the person of its director by the Comité Consultatif International Télégraphique et Téléphonique. Contributions were made by the administrations of the Netherlands, Denmark, the United Kingdom, and by each of the manufacturers.

A second meeting, again on the initiative of the Netherlands administration, took place in September 1959 at Brussels between the administrations of the common-market countries, and a third, at the invitation of the European Conference of the Administrations of Post and Telecommunications in October 1960 at Paris, in which 22 countries participated.

The voluminous documents that had been prepared at Brussels for further study could not,

however, be fully discussed at Paris, at which delegates were present from many countries that had not previously participated in these studies, and no decisions were reached.

Since then, discussions among various administrations and manufacturers* have led to a crystallization of requirements and to a large degree of agreement on the signalling code to be used. At a meeting at Munich in March 1962, which was attended by delegates of the administrations of Belgium, Denmark, Italy, The Netherlands, Norway, Sweden, Switzerland, and Western Germany, at the invitation of the latter, general agreement was reached on a code of signals and the consequent methods of operation. A complete specification on this system will be prepared for discussion at a (it is hoped) final meeting in the autumn of 1962, to which other administrations will also be invited.

2. General Principles

The general principle on which the plan is based is that the signals used in the establishment and supervision of telephone connections will be divided into two categories.

2.1 Line Signals

Line signals make use of signal transmitters and receivers associated with each individual line or speech channel. They comprise all signals that must be exchanged between the line terminations when no common-control or register equipment is connected at either or both of these terminations. The signals shown in Table 1 form part of this category; their definitions and functions are explained in Section 5.2.

As all of these signals, if provided, make use of the line signalling equipment they may be transmitted in various manners as determined by the characteristic properties of this signalling equipment and have the same signal composition as employed for corresponding signals on existing line equipment on which all signals are line signals. In other words, existing junction and toll line equipment may be used unchanged, whether these lines are provided with various types of in-band voice-frequency signalling or out-of-band signalling, whether the signalling system is of the continuous or pulse (spurt) type, whether the toll lines are used for one-way or both-way operation, whether signal discrimination is obtained by signal length or frequency, or whether the lines or junctions are of the 2-wire or 4-wire type.

TABLE 1
Line Signals

Forward Signals	Backward Signals
Seizure	Acknowledgment of Seizure*
Clear Forward	Answer or Reanswer
Forward Transfer*	Clear Back
Toll Offering*	Release Guard*
Reringing*	Blocking*
	Acknowledgment of Optional Forward Signals*

* Optional

The only requirement the line signalling equipment should meet to be used with the new system is that the signal receivers must not respond to frequencies between 500 and 2000 hertz which requirement is fulfilled by modern equipment meeting recommendations of the Comité Consultatif International Télégraphique et Téléphonique.

It should be observed that, in comparison with methods employing line signals exclusively, fewer line signals are needed with the new methods, so that it may be possible to simplify the relay sets associated with the line terminations for new equipment. Further, the new methods, employing a combination of line-and-register signals, do not in any way influence the introduction of changes in existing line signalling codes if such were found desirable to increase reliability and/or transmission speed of line signals.

* H. Pausch, A. Pfau, and F. Pfleiderer, "Multifrequency Code Dialing Method for Worldwide Telephone Traffic," *NTZ Communications Journal*, volume 1, number 1, page 13; 1962.

As line signals are independent of the particular subject of register signals to be dealt with here, they will not be considered further in the following section.

2.2 Register Signals

Register signals make use of signal transmitters and receivers associated with the common control equipments or registers that are temporarily engaged during the establishment of a connection. They comprise all signals exchanged between these common circuits. Although these signals are transmitted via the speech channels by which these common circuits are interconnected during certain periods in the establishment of the connections, no mutual interference can take place between the register and line signals as they employ different frequencies.

The following signals may form part of this category.

Forward Signals

Numerical signals

A. Constituting number of called line
B. Constituting number of calling line
C. Indicating call to operator
D. Indicating call to routine test equipment

End of sending

Affirmation that echo suppressor is needed

Rejection of request for information

Category of caller

A. Subscriber without preference
B. Subscriber with preference
C. Operator
D. Test call
E. Data transmission

Backward Signals

Proceed-to-send with

A. Next digit
B. Digit transmitted last but 1
C. Digit transmitted last but 2
D. Digit transmitted last but 3
E. First digit of international country code
F. First digit of national number

Request for transmission of

A. Local number of calling line
B. National number of calling line
C. Code of originating country
D. Category of caller
E. Need for echo suppressor
F. Subsequent figures with dial pulsing

End-of-selection

A. Without further information on condition of called line
B. Preparative to transmitting information on condition of called line

Condition of called line

A. Free with call charging
B. Free without call charging
C. Free with holding of malicious callers
D. Busy
E. Out-of-order
F. Changed number
G. Not used

Congestion

A. In toll switching equipment
B. In local switching equipment

Dead level

It should be understood that the above list of register signals shows only the possibilities incorporated in the system. All common equipments do not have to be arranged for transmitting or receiving all of these signals. A number of them are for optional use, such as the signals relating to transmission of the number of the calling line, which are to be provided

only if the local switching equipment is arranged for identifying the calling line.

The signals relating to the category of the caller and to the condition of the called line need be provided similarly only if the local switching equipment (and possibly the national toll dialling system) are capable of providing corresponding electric signals.

Other signals shown will be required only by the common equipments located at a relatively small number of particular types of offices, for example a number of signals are used only on international connections and will be employed only by registers controlling the establishment of international connections. Some other signals will be required only for national toll connections and will be provided only at the national toll registers. None of these two classes of signals will be required at the registers of local offices.

This will be discussed further hereafter.

2.3 Signals for Interregister Signalling

It follows from the above that these so-called register signals are used for the transmission of digits and associated orders for establishing connections, as well as for transmitting information on the condition of certain equipments engaged in the connection. It is, therefore, necessary to transmit information in both a forward and backward direction relative to the direction in which the connection is set up. For the signals in each direction, a self-checking multifrequency code is used, in which each signal is composed of 2 frequencies out of a group of 6.

In this manner, 15 signal combinations are available for each direction. Arrangements may be made so that in both directions—by the transmission of appropriate "shift" signals—some or all of these signals may be given new meanings during different stages of the establishment of the connection, so that the total number of signal meanings that may be transmitted is larger than 15 in each direction.

As it is desirable for signalling systems to be effective on both 2-wire and 4-wire circuits, separate groups of 6 frequencies are employed for each direction, so that a total of 12 frequencies is provided. In each group the 6 frequencies are spaced at intervals of 120 hertz and the space between the 2 groups has been fixed at 240 hertz with a highest frequency of 1980 hertz; the lowest frequency that may be used thereby becomes 540 hertz. The following frequency allocation has been accepted.

Signals in forward direction use 1380 through 1980 hertz.

Signals in backward direction use 540 through 1140 hertz.

A complete list of register signals is given in Table 2. Each signal comprises 2 of the frequencies indicated by the 2 crosses in the corresponding frequency columns. The code is of the additive type. In the table, a broken line separates the 2 frequencies used for each signal. The frequency indicated to the left of this line corresponds to an index figure for each frequency, shown at the top, and that shown at the right of the line corresponds to a weight figure shown at the bottom of the table. The code number or numerical value of each signal may be found by adding the values of the index and weight figures indicated.

2.4 Compelled Operation

A system has been adopted in which one signal in the forward direction and one signal in the backward direction are always used in combination, so that the length of each of these signals is determined by the other. This has been called the compelled method of operation for this reason. The exchange of signals always commences with a forward signal that has no definite length but continues until a backward signal is received that is transmitted on receipt and in acknowledgment of the forward signal. In its turn, the backward acknowledgment signal continues until the termination of the

forward signal is recognized. The next forward signal may be transmitted from the moment the termination of the backward signal is recognized at the originating end. The principle of this operation has been indicated diagrammatically in Figure 1.

The fact that signals in the two directions of operation are available permits the adoption of this compelled type of operation, which has the following advantages.

A. The length of each signal and the intervals between signals need not be specified accurately, as is the case with systems using pulses and intervals of predetermined duration without acknowledgment of each signal. This avoids the use of apparatus capable of providing and accepting these pulses within prescribed limits, as well as maintenance to ensure continued operation within these limits.

B. The system is insensitive to receiver distortion and will work as long as the receiver responds to the received signal. This renders the performance specification of the receiver less stringent and permits operation over a wider range of signal levels than if the receiver would perform only within narrow limits of distortion.

C. The arrangement is less sensitive to disturbances in the transmission media than noncompelled systems. Neither short openings nor the presence of other frequencies is able to disturb a signal, since its transmission is continued until it has been properly received and acknowledged. Once a signal has been recognized, it will be recorded so that no disturbance may be caused by subsequent irregularities in the transmission equipment during the time the signal continues. It should be noted that the codes are self-checking, for only when 2 frequencies are received will a signal be accepted. Signals of 1 frequency or of more than 2 frequencies will be ignored.

TABLE 2

DENOMINATION AND COMPOSITION OF REGISTER SIGNALS

Code	Numerical Value of Code		Frequency in Cycles per Second (Hertz)					
		Forward	1380	1500	1620	1740	1860	1980
		Backward	1140	1020	900	780	660	540
		Index	f0	f1	f2	f3	f4	f5
15	4 + 11		Rank Frequency				x	x
14	3 + 11					x		x
13	2 + 11				x			x
12	1 + 11			x				x
11	0 + 11		x					x
10	3 + 7					x	x	
9	2 + 7				x		x	
8	1 + 7			x			x	
7	0 + 7		x				x	
6	2 + 4				x	x		
5	1 + 4			x		x		
4	0 + 4		x			x		
3	1 + 2			x	x			
2	0 + 2		x		x			
1	0 + 1		x	x			Base Frequency	
		Weight	0	1	2	4	7	11

2.5 Link-by-Link and End-to-End Signalling

With the type of signalling described above, use may be made of either of 2 different principles for the exchange of signals between registers, which are known as "link-by-link" and "end-to-end" signalling, respectively. *Link-by-link signalling* implies that the register at each, except the last, office involved in a connection must receive all figures of the called number, including those needed for controlling the selecting operations at that office and other figures that may be needed only at offices engaged later in the connection. Either all of these figures or the latter part mentioned must be repeated (re-transmitted) from each to the next engaged register during the process of establishing the connection.

With *end-to-end signalling*, the register at a transit office accepts only that few of the initial digits of the called number that indicate the direction to be selected at that transit office. After the necessary number of figures have been received at the register of the transit office and the call has been extended in the wanted direction, this register will be released immediately. The numerical information required at the next office will be transmitted from a "leading" register at a preceding office, which will remain attached until the connection is completed. This leading register may be either at the office nearest to the calling subscriber or be located for example at a toll center acting as a pivotal point for a number of local areas.

2.6 Leading Registers

By a leading register is meant a register that is located relatively near the calling subscriber and that remains connected to control the connection process until the called line has been selected. This register will receive the complete numerical information identifying the called subscriber and by means of significant backward signals it will be ordered to transmit any part of this information to any other office that may be involved in the establishment of the connection.

For certain connections, the leading register will be the local register at the office to which the calling subscriber is connected. This is notably the case for calls that remain within the local area or for calls to other areas in which no office of higher order will be involved.

In other cases, the leading register may be located at the toll office serving one or a number of local areas. In such a case, the originating local register in the office to which the calling

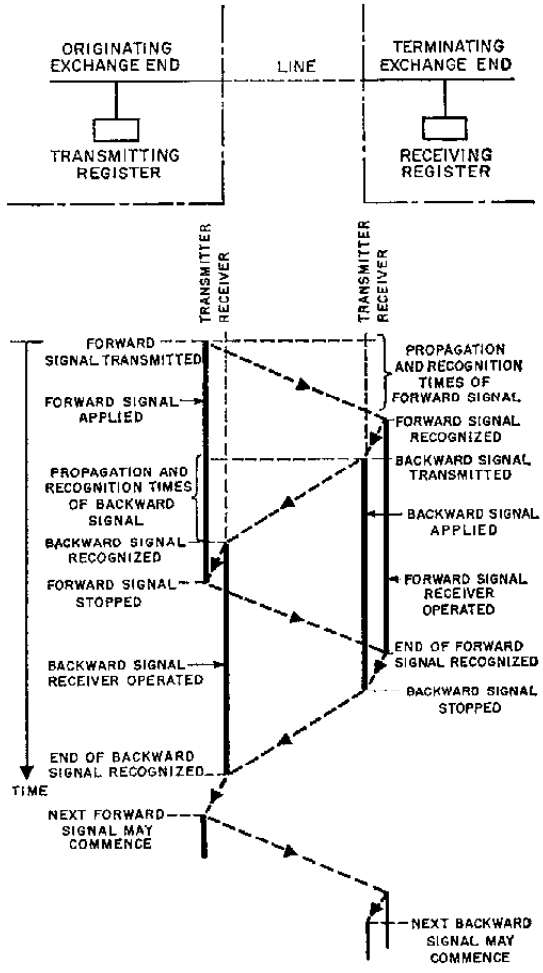


Figure 1—Principles of compelled signalling.

subscriber is connected will transmit all numerical information to the leading register in the toll office, so that all operations that are required for establishing a connection through a toll network may be controlled by the latter register. It is therefore unnecessary that the registers at the local offices be loaded with all the complications required for the establishment of toll connections.

For international connections, the leading register may be located at other offices that are of a higher order than the toll office serving the calling subscriber. Such so-called international registers may be located at regional centers or at one or more international toll offices. Again, the purpose of concentrating these registers at a few offices of higher order is to simplify the registers of lower order, so that the complications for international connections will be concentrated at a relatively small number of registers.

In this case also, all numerical information will be transferred from the originating local register to the leading register at the regional or international toll office. It will be evident that for this transfer of numerical information it may be necessary to transmit the signals through one or more transit offices and that end-to-end signalling may take place from the register at the originating local office to the leading register. For the establishment of the connection beyond the office of the leading register, the latter register takes control and the signalling may again be an end-to-end operation starting from the leading register.

It is not mandatory to employ end-to-end register signalling in the strict sense of the word between the first and last offices engaged in a long-distance connection. Such connections may be made in 2 or more sections with end-to-end signalling taking place within each section. Thus, each section may contain one or more transit offices the registers of which receive only a part of the numerical information and are released immediately after the connection has been extended to the next office.

2.7 Advantages of End-to-End Signalling

The advantages of end-to-end signalling when applied in the manner described in Section 2.6 are as follows.

2.7.1 The toll transit registers will be few in number because of their short holding time and of a simple type as they need receive only a restricted number of digits and need not retransmit any digits. If the transit traffic is small, the functions of the toll transit registers may be combined with those of the terminating registers in an economical manner.

2.7.2 The total number of operations for the transfer of numerical information in a connection with end-to-end signalling involving one or more toll transit exchanges, will be less than in a system with link-by-link signalling. For example, assuming a connection including two toll transit offices for which the complete called number comprises 10 digits, of which only 3 are required for completing the connection at a toll transit office, the total number of digits that must be transmitted with link-by-link signalling is 10 on the first two links and 7 on the last link, or 27 in all; with end-to-end signalling, this number is 3 on the first 2 links and 7 on the last link, or 13 in total. This will tend to increase the speed of operation and to reduce the error liability.

2.7.3 With end-to-end signalling, the leading register always remains connected until the connection has been completed. On receipt of a signal indicating that the call cannot be completed because of the state of the called line, the leading register will immediately release that part of the connection beyond its own office. Another possibility is to use the signal indicating the free condition of the called line for the connection of ringing tone to the caller from the originating office, or from the office at which the leading register is located, when the called office provides a type of ringing tone to which the caller is not accustomed, for example, on international connections.

2.7.4 Similarly as for 2.7.3, the leading register is available for receiving signals indicating congestion or any need for alternative routing. It may from the absence or presence of the latter signal determine the route of the call, which is particularly important in international connections if revenues are to be shared by different administrations. It also permits the leading register to receive reorder signals so that a different route may be taken from the office at which the leading register is located, in case of congestion at a further transit office through which the call was first routed.

2.7.5 The waiting time between the moment the last digit has been dialled and the moment the called line has been reached may be reduced by starting selection at the leading register before all numerical information has been received. With link-by-link signalling, this would not be possible without holding all transit registers engaged in the connection until all numerical information had been received and retransmitted, which would very considerably increase their holding time. With end-to-end signalling, it is sufficient to wait only until enough digits have been received to permit the transit registers to complete their operation and to release immediately thereafter.

2.7.6 End-to-end signalling facilitates operation under the following conditions.

2.7.6.1 In case of nonuniform numbering of subscribers within local areas. Subscribers numbers with either 5 or 6 digits are sometimes employed in one area or even in one office.

2.7.6.2 For the interworking of local areas having a different number of digits for their subscribers numbers. Various areas in one national network may have local subscribers numbers ranging from 3 to 7 digits. Also, the total number of digits to be dialled for a national toll call may vary for different areas.

2.7.6.3 If provision has to be made for dialling into private automatic branch exchanges by supplementary digits, resulting in a total number of digits larger than that for regular subscribers numbers.

All these cases present the problem of how to determine when all digits have been received at the originating and transit registers and when these registers have completed their functions, so that they may disconnect. This holds true particularly in the case of automatic service, because subscribers have not the possibility afforded to operators to transmit an end-of-sending signal. The automatic equipment must determine itself when the numerical information is complete.

In the cases of irregular numbering mentioned above, it becomes very complicated, if not impossible, to do this at the originating (leading) or transit registers, because in principle these registers must control the operation to many areas, each of which may have a different numbering pattern. It is, however, relatively easy, and in many cases quite simple, to determine the moment the full selection information has been received at the terminating area, because either the switching equipment serving this area may already provide a signal for this purpose from the final selection stage to the terminating register, or number discriminators may be associated with these registers that are able, in a rather inexpensive and rapid way, to determine the number of digits required for completing the connection within the terminating area alone.

It may now be seen that end-to-end signalling presents an easy and straightforward solution to the problem mentioned by a combination of the following circumstances.

Firstly, with end-to-end signalling, transit registers need not receive and retransmit the complete numerical information and may release immediately after having received a few initial digits and after having completed the required selection at that transit office. There is, therefore, no problem of indicating the receipt of all

numerical information or of the completion of the selection, as far as the transit registers are concerned.

Secondly, end-to-end signalling is based on the condition that a leading register remains connected until the connection has been completed and that backward signals may be transmitted to this register at any time until this moment.

Thirdly, the fact that the receipt of the last numerical information or the completion of selection may conveniently be determined at the terminating office, leads to the concept of using backward signals from the terminating to the leading register to signal this fact. The leading registers will not need to determine themselves that no more numerical information is to be expected.

Fourthly, in case the leading register is preceded by another part of the connection controlled by the originating register, as explained in Section 2.6, the leading register may retransmit the backward signal it receives to the originating register.

It should be observed that this backward signal may contain further information, such as on the state of the called line, as already explained under 2.7.3.

2.8 Signals Employed

The meanings of the accepted signals are shown in Tables 3 and 4 for the forward and backward register signals, respectively. The following may serve as an explanation of these tables.

2.8.1 *Signals in Forward Direction*

Table 3 shows that each of the 15 signals available in the forward direction may have two different meanings as given in groups *I* and *II*, the normal meaning being that for group *I*. Group *II* will become effective on receipt of a particular backward signal, as will be explained later.

2.8.1.1 *Group I*

The first 10 signal numbers in group *I* are allotted to the 10 numerical digits, and will in general be used for the transmission of numerical information such as that in the called number. Signal *11* is used only for semiautomatic international operation where it may be necessary for an outgoing international operator to select an operator at the distant country, for example, for handling delayed or person-to-person calls. This signal may be followed by other numerical signals indicating the particular operator's position wanted. Such additional numerical indications may be omitted if any free position is acceptable. In either case, the numerical information is always followed by the transmission of signal *15*, which indicates that no further digits are to be accepted. Signal *15* will be transmitted by the operator who will depress an end-of-sending key for this purpose. It will be evident that subscribers who are permitted to establish automatic connections via the same international lines that are accessible to operators cannot reach the operator's positions for delayed or person-to-person calls because subscribers cannot make use of code signals *11* and *15*. The operation as here described corresponds to that covered by signals *11* and *12* in the recommendations of the Comité Consultatif International Télégraphique et Téléphonique for semiautomatic international operation.

Signal *12* is transmitted in the forward direction from a register when it receives a backward signal from a distant office indicating a request that cannot be complied with. This may be the case, for example, when the distant office sends a backward signal asking the originating or leading register to transmit the identity or the category of the calling subscriber and the automatic equipment preceding this originating or leading register is not able to furnish this information. Other cases in which forward signal *12* may be used will be discussed later.

TABLE 3

LIST OF FORWARD REGISTER SIGNALS

Signal Number	Group I	Group II, Category of Caller (Responding to A-3 and A-5)	
1	digit 1	Subscriber	For national service (2)
2	digit 2	Subscriber with preference	
3	digit 3	Maintenance equipment	
4	digit 4	Reserve	
5	digit 5	Operator's position	
6	digit 6	Data transmission	
7	digit 7	Subscriber	For international service
8	digit 8	Reserve	
9	digit 9	Reserve	
10	digit 0	Operator's position	
11	Access to operator's positions (1)	Spare for optional use	For national service (3)
12	Demand rejected (responding to A-9, A-10, or A-13)	Spare for optional use	
13	Access to maintenance equipment (1)	Spare for optional use	
14	Insert echo suppressor	Spare for optional use	
15	End of sending	Spare for optional use	

(1) These signals may be followed by other numerical signals and by signal *I-15* or by *I-15* alone.
(2) The international outgoing registers must convert signals *II-1* to *II-4*, *II-6* into *II-7*, and *II-5* into *II-10*.
(3) These signals will be converted by international outgoing registers into signal *II-7*.

TABLE 4

LIST OF BACKWARD REGISTER SIGNALS

Signal Number	Group *A*	Group *B*
1	Send next digit	Prepare for malicious call tracing
2	Send next digit but one (4)	Changed number
3	Switch over to *B*-signals (1)	Called line busy
4	Congestion (4)	Congestion
5	Transmit category of caller (1)	Dead line or dead level
6	Switch through speech path	Called line free, with call charging
7	Send last digit but two	Called line free, no call charging
8	Send last digit but three	Called line out-of-order
9	Spare for optional national use	Spare for optional national use (3)
10	Spare for optional national use	Spare for optional national use (3)
11	Send first digit of international country code	Reserve
12	Send first digit of national number	Reserve
13	Identification: send first digit of international country code (2)	Reserve
14	Inform whether or not echo suppressor is needed	Reserve
15	Reserve	Reserve

(1) Must be responded to by a forward signal of group *II*.
(2) Repeat signal *A-13* = send next digit of calling line's number.
(3) On international connections, signals *B-9* and *B-10* must initiate the connection of an information tone to the caller, similarly to signal *B-5*.
(4) On lines from outlying offices preceding leading registers and on which 4 frequencies for *A* signals without *B* signals are employed, signal *A-4* may initiate the connection of information tone and signal *A-2* may be used to initiate the transmission of a predetermined digit.

Signal *13* is a numerical signal that can be transmitted only from test equipment; it initiates the selection of other test equipment at the distant end so that the transmission conditions of the connection may be tested. If necessary, this signal may be followed by further numerical signals to indicate which of several different types of test equipment should be selected.

The only forward signal of group *I* not discussed so far is *14*, which has been provided in case it is necessary to call for the insertion of echo suppressors. The exact manner in which this signal and the corresponding backward signal will be used has not yet been determined.

2.8.1.2 *Group II*

All the forward signals of group *II* are concerned with the particular category of a caller. The first 6 signals in this group have been reserved for national use and identify 5 different categories of callers with one signal in reserve. These forward signals will not be transmitted on international connections. If a request for information on the category of a caller is required on an international connection, one of signals *7* through *10* will be transmitted. The leading register controlling international connections is able to distinguish the fact that an international connection is concerned and will, therefore, send the appropriate signal. This leading register must obtain the necessary information on the category of the caller from the register at the originating office and this latter register will be able to send only one of the signals *1* to *6*. On receiving this information from the originating register, the leading register will convert signals *1, 2, 3, 4,* and *6* into the international signal *7*, and will convert signal *5* into international signal *10*. Signals *11* to *15* have not been assigned but are reserved for only national service and may be employed by each administration for such purposes as may be required in its country.

A few words may be said here regarding the need for providing these category-of-caller signals. A distinction between calls originating from *subscribers* and from *operators* may be needed for various reasons. In certain national networks an operator, by sending offering signals, may through-connect to a busy called line to inform the busy subscriber that a national or international toll call is in preparation for him. In such cases, it is, therefore, necessary that a connection to a busy subscriber's line be not broken down by the leading register if it receives a backward signal indicating the busy condition. In order that subscribers or other unauthorized callers will not be able to effect the offering operations, it is useful to receive an indication at the terminating office that offering should be permitted. Accordingly, when a called line is found busy, a backward signal will be transmitted initiating the transmission of the forward signal indicating the category of the caller and only if this indicates an operator's call will the terminating equipment respond to offering signals. It will not respond when such offering signals have been imitated by subscribers. Another reason for distinguishing between operators and subscribers is that in cases of overload or congestion, arrangements may be provided so that preference is given to operator's calls, and a congestion signal is always given to subscribers even if free outlets become momentarily available.

Arrangements are provided also to signal *two different classes of the calling subscribers*, of which only those having preference will be able, in cases of emergency, to obtain certain connections.

A further category-of-caller signal has been provided to indicate calls originating from *subscriber stations at which data-transmission equipment is connected*. Such a signal may be needed in the future to apply the proper tariffs for data-transmission connections.

On international connections, a distinction is provided only between operator's and subscriber's calls.

2.8.1.3 *Optional Extension of Forward Signals*

Apart from the signals indicated in Table 3, administrations are free to provide in their national network additional forward signals not in this list. Thus the forward signals of group I may receive additional meanings on receipt of a particular backward signal. For example, in certain countries it may be necessary to transmit the identity of the calling subscriber to recording equipment for automatic message accounting concentrated at certain toll offices. Toll calls on which information for automatic message accounting is to be provided, will cause the transmission of a backward signal that changes the meaning of the forward signals in such a manner, for example, that signals *1* to *10* no longer signify the digits of the called subscriber's number but those of the calling subscriber's number. In case such a request for identification of the calling subscriber's number is required at originating offices where this request cannot be complied with (for example, when no automatic identification equipment has been provided to determine the identity of the calling subscriber), this request may be answered with signal *12*, which in this case will indicate that no identification is possible, as a result of which such a toll call may be refused.

2.8.2 *Signals in Backward Direction*

From the above description of the forward signals, it will be obvious that a minimum of 10 signals is always needed; they require the use of combinations of a minimum of 5 frequencies in the forward direction. In most cases, as soon as one supplementary signal has to be provided, the use of the 6th frequency will be required.

Contrary to this condition, the number of signals in the backward direction does not always require the use of as many as 5 or 6 frequencies. In many cases, the necessary number of signal combinations may be provided by only 4 frequencies, such as when the number of backward signals does not exceed 6 combinations. Table 4 takes into account the desire to reduce the number of frequencies employed for backward signals wherever possible. The signals have been grouped so that either 4, 5, or 6 frequencies may be used according to the number of combinations required, which will be 6, 10, and 15, respectively, in these 3 cases. The first thing to be observed about the table for backward signals is, therefore, that it has been divided by broken lines into these 3 subgroups of signals.

Furthermore, in a manner similar to that for forward signals, Table 4 has two columns of different meanings for each of the signals from *1* to *15*. Signals indicating only one of these columns, or both, may be employed. This implies that several different possibilities exist, the simplest being to use only the 6 signals in the first subgroup for the meanings given in *A*. This simplest case may be extended in either of two ways, namely, by extending into subgroups *2* and *3* in the same vertical column or by adding signals to indicate a transfer to the meanings shown in the second column, group *B*. This same principle may be applied to the case of the 10 *A* signals shown in the first column. Supplementary signals in this case may be obtained either by adding the last 5 signals in group *A* or by making use of signals *1* to *10* in group *B*. It will, therefore, be seen that a large number of combinations of signals is possible that are all compatible with one another, so that a restricted number of signals may be used in part of the offices and a more-extended number in other offices.

The number of signals that may be employed at any one office is determined by the characteristic operating features of the local switching equipment. For the simplest possible type of local switching equipment, which does not provide any electric criterion on the condition of the called line, it is sufficient to provide only signals *1* to *6* of group *A*, which even then do not all need to be used. Local switching equipment that is arranged to provide a restricted number of different electric signals to indicate the condition of the called line, may also employ

the first group of 6 signals by adding the second meanings shown for group *B* of these signals. Additional features may be provided in each of these two cases by making use of 10 signals, as will be explained hereafter.

Signals *11* through *15* appear only in group *A* and have been reserved for use on international connections exclusively, so that normally none of the registers in local switching equipments will have to provide for transmitting or receiving these signals.

The purpose of each of the signals shown in Table 4 will now be explained by describing how some typical connections are established. It will be evident that the backward signals are very important because they exercise a necessary control on the operations to be performed.

3. Typical Cases of Operation

3.1 Interoffice Connection in Local Multioffice Area

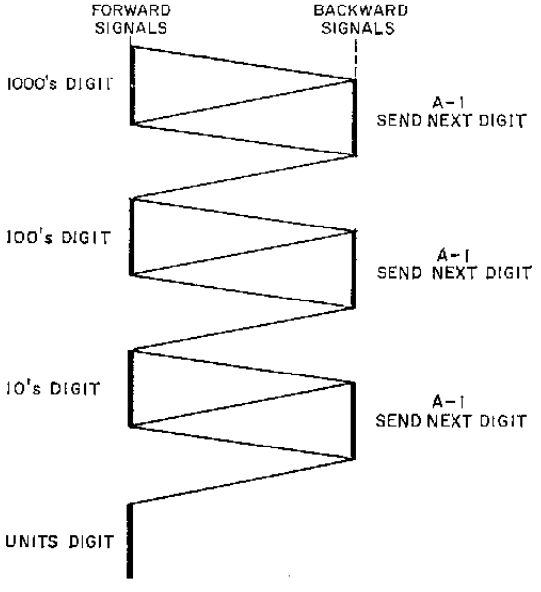


Figure 2—Transmission of 4 digits from one local exchange to another.

3.1.1 *Direct Connection between Two Local Offices*

The subscribers' numbers used in a local area have an initial part identifying the local office to which the subscriber is connected and a second part identifying each particular subscriber of this office. With offices having the usual 10 000-line capacity, this latter part consists of 4 digits. The initial part, which is the office code, may have from 1 to 3 digits depending on the size of the area. In the simplest case, when a direct junction interconnects the two offices concerned in the connection, the selection at the originating office is controlled by the office code; only the last 4 digits of the subscriber's number need be transferred to the wanted office. The operation in this case is shown in Figure 2.

As soon as the junction to the wanted office has been seized, a signal is transmitted thereon which initiates the connection to the incoming end of this junction of an incoming register at the wanted office. Concurrently, the register at the originating office will connect to the junction the combination of frequencies indicating the value of the thousands figure, and this signal will be extended to the incoming register as soon as it is connected. On recognizing this first digit, the incoming register transmits to the originating register backward signal *1* of

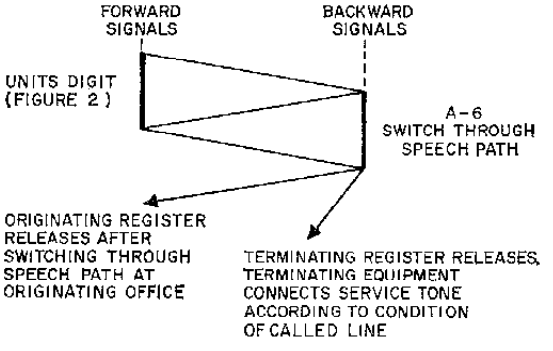


Figure 3—End of selection without signalling condition of called line.

group *A* (signal *A-1*), which has the dual purposes of informing the originating register that the first digit has been properly received, which causes it to stop transmission of the first digit, and also of signalling to the originating register that the next operation required is the transmission of the following digit. The termination of the forward signal in its turn, by being recognized at the incoming register, causes the backward signal to stop. Recognition of this at the originating register permits it to commence transmitting the next digit. It will be obvious that this transmission will take place only when this digit has already been dialled by the calling subscriber. In this manner, all 4 digits of the number wanted at the distant exchange will be transmitted by mutual control of forward and backward signals, and each time the next digit must be transmitted, the forward signal will be acknowledged by the transmission of the backward signal *A-1*.

This situation changes at the moment the units digit has been transmitted to and is recognized at the incoming register, and different operations may now take place according to the possibilities provided at the terminating office. The simplest case is when the terminating office is *not capable of giving any indication of the condition of the called line*. As shown in Figure 3, recognition by the incoming register of the units digit will immediately be acknowledged by signal *A-6*, which causes the originating register to release immediately and the speech path to be closed at the originating office. Also, the register at the terminating office will disconnect and the incoming junction will be connected through to the selector equipment so that the service tone indicating the condition of the called line (ringing, busy, or any other tone) may be transmitted to the calling subscriber from the terminating office.

A somewhat-more-complicated condition exists when the terminating equipment is capable of *providing electric criteria on the free and busy conditions of the called line*. With reference to Figure 4, the units digit will be acknowledged by the transmission of the backward signal *A-3*, instructing both originating and incoming registers to use the meanings given in group *II* of Table 3 and group *B* of Table 4. When the signal *A-3* has stopped, the originating register proceeds immediately with the transmission of one of the signals shown in group *II* of Table 3, depending on the origin of the call. In this case, where a call is assumed to be originated by a subscriber, either signal *II-1* or *II-2* will be transmitted. Assuming that it will be possible within a very short interval after the units figure has been received to select the called line and determine its condition, the forward signal

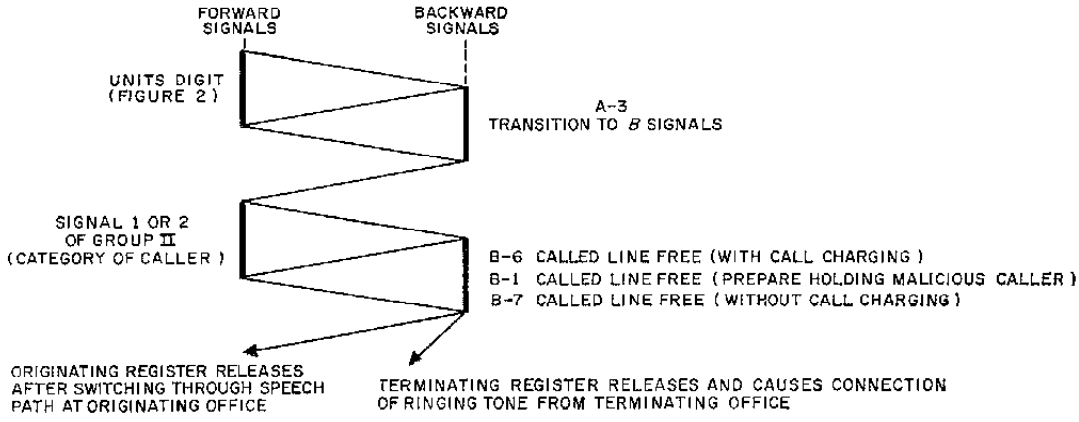


Figure 4—End of selection when called line is free.

II-1 or *II-2* may within a very short time be acknowledged by one of the signals in group *B* of Table 4, indicating the condition in which the called line has been found. If the called line is free and answering should cause the call to be charged, signal *B-6* must be transmitted. Both originating and terminating registers will disconnect on termination of the *B-6* signal, and the switching equipment will connect the speech path through at both offices so the calling subscriber may hear the ringing tone from the terminating office. Signals *B-1* and *B-7* have this same effect but initiate supplementary operations at the originating office, signal *B-1* causing the equipment at this office to prepare for holding the connection under the control of the called subscriber to permit the tracing of a malicious caller, and signal *B-7* operates a relay to prevent the call from being charged when the answer signal is received.

Signal *B-3* will be transmitted when the line of the called subscriber is busy, as indicated in Figure 5. In this case, the switching equipment may be arranged so that on termination of the signal the originating register causes busy tone to be connected to the calling subscriber and releases the junction to the distant office, as a consequence of which all equipment engaged at the distant office will be set free. Signals *B-2, B-5, B-8, B-9,* and *B-10* are available for other conditions of the called line that do not permit the connection to be completed to the called subscriber. All of these signals may cause the register at the originating office to set free the connection to the distant office and connect to the calling subscriber either a special tone (so-called information tone) or a recorded message to inform him why the connection cannot be completed.

If the connection cannot be completed because of congestion in the switching equipment at the terminating office, the incoming register will transmit the signal *B-4,* congestion, to acknowledge one of the forward signals of group *II* that is transmitted in response to signal *A-3* as shown in Figure 6. This causes the originating register to release the junction to the distant office and to connect an appropriate tone signal to the caller.

If the terminating selector equipment is of a type in which *a few seconds may elapse before the called line may be reached* after the receipt of the units figure, it is not desirable to ac-

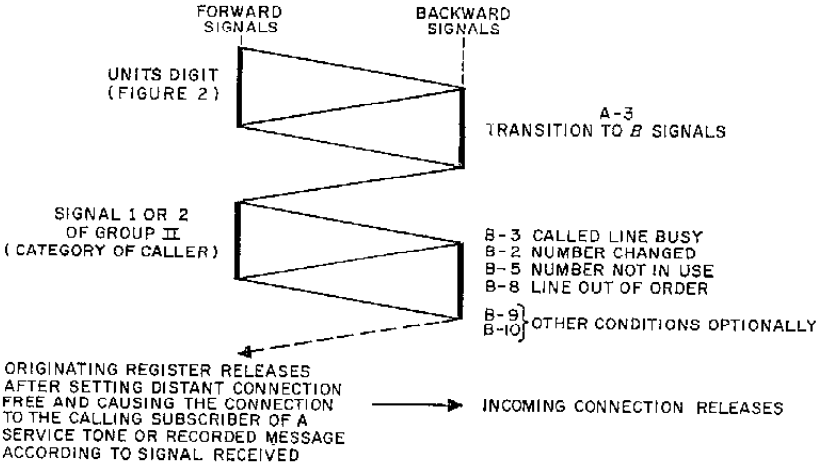


Figure 5—End of selection when called line is in a condition other than free.

knowledge the units figure by signal *A-3* as described above, because this would be immediately responded to by the transmission of the signal *II-1* or *II-2*, which would have to remain connected until it can be acknowledged by a *B* signal after the completion of selection. Although this is not objectionable in a local connection as now considered, it becomes objectionable when a connection is extended via a toll line employing carrier equipment in which the amount of signalling energy should be made the minimum possible. For this reason, the units digit will not be acknowledged by either signal *A-3* or *A-6* but by signal *A-1*, which invites the originating register to transmit the next digit. See Figure 7. Since, however, no next digit is available, no forward signal will be transmitted during the period the terminating selector equipment needs to complete the selection. After selection and the condition of the called line has been determined, further operations are then initiated by the incoming equipment, which will send a pulse of predetermined duration. If this pulse is of signal *A-3*, the operation might be continued in the same manner as described above for the cases in which the backward signal *A-3* was transmitted immediately on receipt of the units digit. In this case, however, it would be necessary, when the called line is found free, to transmit two cycles of signals before this free condition can be signalled, namely, firstly by the cycle commencing with backward signal *A-3* and secondly by the cycle consisting of a forward signal of group *II* and the backward signal *B-6*. This may be objectionable, because for a quick reply, the signalling operations

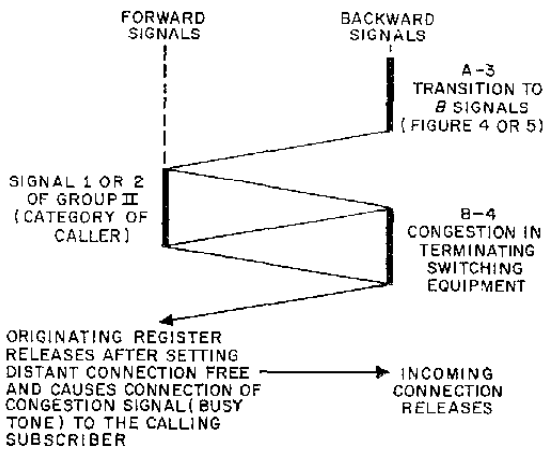


Figure 6—Congestion in terminating exchange switching equipment.

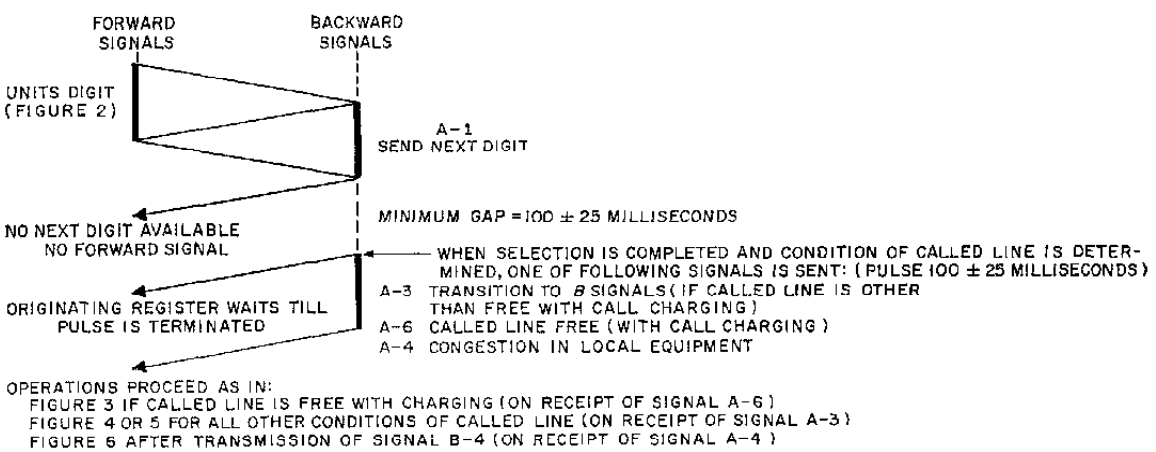


Figure 7—Suspension of signalling after units digit if selection is delayed.

would not be completed before the called subscriber has challenged the caller. Therefore, when it can be determined that the line is free, the free condition in this case will not be signalled by signal *B-6*, but by a pulse of signal *A-6*, which is then transmitted instead of the pulse of signal *A-3* referred to above. In this way, it is ensured that the speech path may be closed with the minimum of time after the free condition of the called subscriber has been determined. In the case of congestion in the terminating switching equipment, a pulse of signal *A-4* will be transmitted instead of *A-3*, signal *A-4* having the same significance as signal *B-4*.

3.1.1.1 *Interoffice Junctions Providing Access to 2 or More Units of 10 000 Lines*

It is obvious that when two or more office units of 10 000 lines each are located in the same building and are served by common groups of junctions, the originating register, on seizing a junction to the wanted office, will first have to transmit the ten-thousands digit to the incoming register. All other operations as described in the former case remain the same.

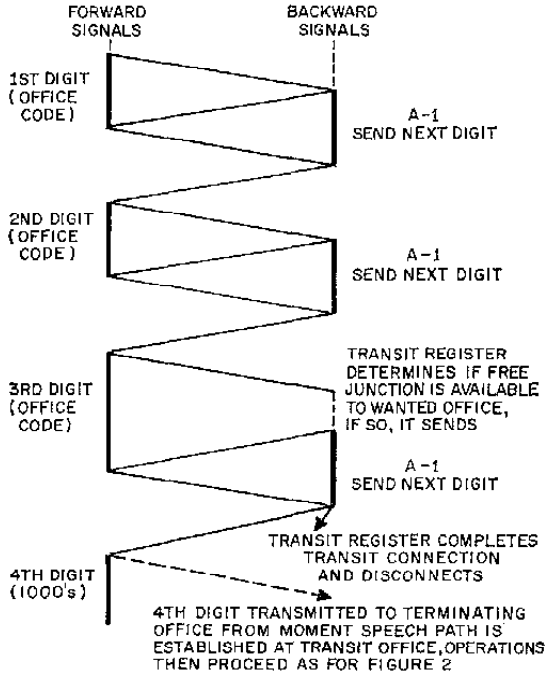


Figure 8—Operation at transit office if junction to the wanted office is available.

3.1.2 *Transit Connection in Local Area*

When two offices in the same local area are interconnected via a transit office, the operations are as follows.

The register at the originating office will determine from the office-code part of the called subscriber's number that the connection must be routed through a transit office and will cause the selection of a free junction to that office. It will then proceed by transmitting the first digit of the called subscriber's number, as shown in Figure 8. The register at the transit office will accept this first digit and by successive transmissions of backward signal *A-1* it will consecutively initiate the transmission of those digits from the originating register that are required at the transit office to determine the wanted office. The transit register will, before acknowledging the last of these digits, determine if a free junction to the wanted office is

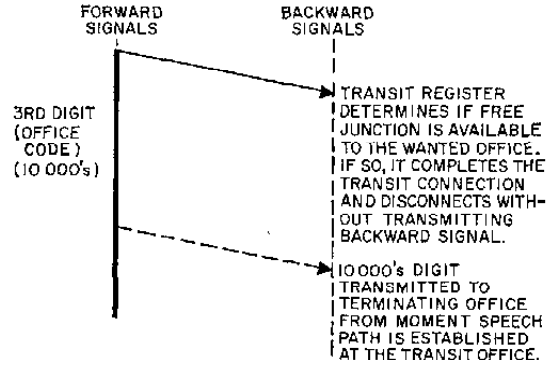


Figure 9—Modified operation of Figure 8 if the last digit of the office code is required as the first digit at the terminating office.

available. If so, it will acknowledge this digit with signal *A-1*, which will cause the originating register to prepare for transmitting the thousands figure. This will be extended to the wanted office as soon as the selection at the transit office has been completed. The register at the transit office will disconnect after it has connected the speech path through at the transit office and on having recognized the termination of the transmission of the last digit of the office code, so that this cannot be switched through to the terminating office. Figure 8 shows the case of an office code consisting of 3 digits.

In case the terminating office is connected via a group of junctions providing access to more than 10 000 lines, it will be necessary to transmit the ten-thousands figure to this office. If this ten-thousands figure had already been received at the transit office because it was required at this office to determine the identity of the wanted office, the receipt of the ten-thousands figure at the transit office is not acknowledged by any backward signal, so that the transmission of this figure from the originating office continues and it will be extended to the terminating office as soon as the connection is switched through at the transit office. It is obvious that in this case the transit register will switch through without awaiting the termination of the ten-thousands digit, as in Figure 9.

3.1.3 *Alternative Routing from Originating Office*

Assuming the originating and wanted offices are interconnected by a group of direct junctions but that these junctions are fully occupied, the connection may be routed alternatively via a transit office.

The originating register will first attempt to select a free junction in the direct route. If successful, it will transmit the first digit of that part of the subscriber's number that is required at the terminating office (see Figure 2). If no free direct junctions are available, the register will attempt to select a free junction to the transit office. When successful, it will transmit the first digit of the called subscriber's number to this office as explained above for operation through a transit office (see Figure 8). Also all further operations will be as there described.

3.1.4 *Alternative Routing from First Transit Office through Second Transit Office*

The operations in this case will commence in the same way as described for the case of a call through a single transit office (Figure 8) up to the point where the register at a first transit office has received the last digit of the office code but before this has been acknowledged by a backward signal. The transit register will first attempt to select a direct junction to the wanted office and, if this is not available, it will next attempt to select an alternative route through a second transit office. When extending the connection to another transit office, the register at the first transit office has to ensure that the register at the second transit office receives the complete office code to determine the identity of the wanted office. To do this, the register at the first transit office will acknowledge the last digit of the office code by one of the signals, *A-2*, *A-7*, or *A-8*, depending on whether the office code characterizing the wanted office is of 2, 3, or 4 digits as shown in Figure 10. The transmission of one of these backward signals will cause the originating register to respond with the transmission of the first digit dialled by the calling subscriber and this is now transmitted via the first transit office to the register at the next transit office, which acts in exactly the same way as described for the case of a single transit office (see Figure 8).

3.1.5 *Congestion at Transit Office—Rerouting*

In case a connection cannot be completed at a transit office because of the unavailability of junctions both directly to the wanted office and to any other transit office through which the connection might alternatively be routed, the last digit of the office code will be acknowledged by signal *A-4*.

Receipt of this signal will cause the originating register to break down the connection to the transit office. Next, it will either cause the connection of an appropriate tone to the caller to inform him of the congestion condition or attempt to set up the connection from the originating office via a different route if such possibility exists. This is shown in Figure 11.

3.1.6 *Interworking between Offices with Different Operating Facilities*

The case may be assumed now that in a multioffice area some offices of older design are not able to provide or respond to electric signals indicating the condition of the called line and that others of more modern design incorporate this possibility. At the first-mentioned offices, the incoming registers associated with the incoming junctions from the latter type of offices will always acknowledge the receipt of the units figure with signal *A-6* as in Figure 3 independent of the called line's condition. This signal will at modern offices cause the disconnection of the originating register and the through connection of the speech path, so that the caller will hear the service tone (ringing, busy, et cetera) transmitted from the switching equipment at the terminating office.

On the other hand, the incoming registers at offices capable of providing an electric signal on the condition of the called line will normally transmit a signal *A-3* (transition to *B* signals and demand to transmit category of caller) to acknowledge the receipt of the units digit (Figures 4 and 5). When this signal is received at the originating register of an office that cannot respond appropriately to the various conditions signalled by the *B* signals and is also unable to provide a signal indicating the category of the caller, this register will in response to signal *A-3* transmit the forward signal *I-12*, which is a rejection of the demand for information.

As shown in Figure 12, this signal will be acknowledged by transmitting from the terminating register the signal *A-6*, which will act in the normal way to establish the speech path. The equipment at the terminating office will therefore not be released, even if the condition of the called line (busy, et cetera) prevents completion of the connection to the called subscriber, and the terminating equipment must be arranged to supply a service tone to signal this condition to the caller.

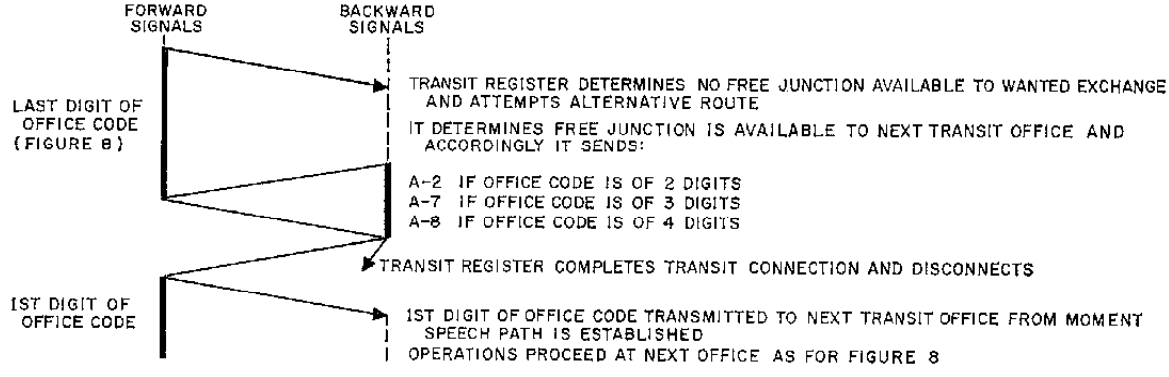


Figure 10—Alternative routing at transit office.

3.2 Operation in a National Toll Network

The methods of operation employed in establishing national toll calls are very similar to those described in conjunction with the operation between offices in a local network. The principal difference resides in the fact that the selection process in the toll network will be governed by the digits constituting the national toll code, instead of by the digits constituting the office code as described for the operation in local multioffice areas. In other words, the possibility is provided by this system to use identical methods of operation and identical signals for the transmission of numerical information on any type of connection within a national network.

3.2.1 *National Networks with Closed and Open Numbering of Network Sections*

A few words should first be said on the different methods used in various national networks to divide the total area into sections from a point of view of numbering subscribers. These various methods of numbering will influence to a certain extent the signals required for the establishment of connections between different local areas. Because the terminology used on this subject is not identical in various countries, a definition of the terminology used here has been given in the appendix, Section 5.2.

All national networks are characterized by different orders or ranks of switching centers used to establish connections among local areas. The lowest rank is that serving one local area or a few small localities considered as one area from the point of view of numbering. These switching centers will be called "toll end offices."

Furthermore, all national networks are characterized by the existence of a higher rank of switching center, which may be considered as the pivotal point of a network group, through which all traffic (with some exceptions) of this network group will be routed to and from other

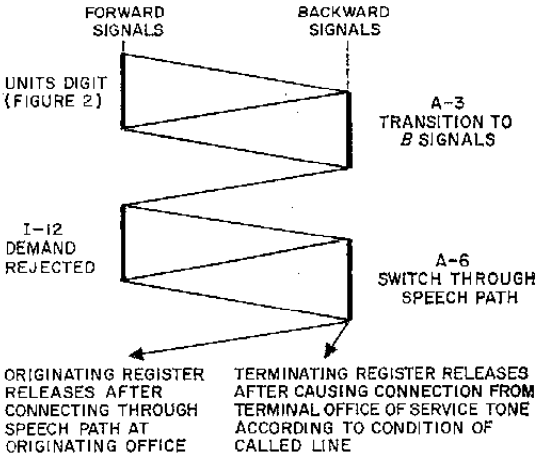


Figure 11—Congestion at transit office.

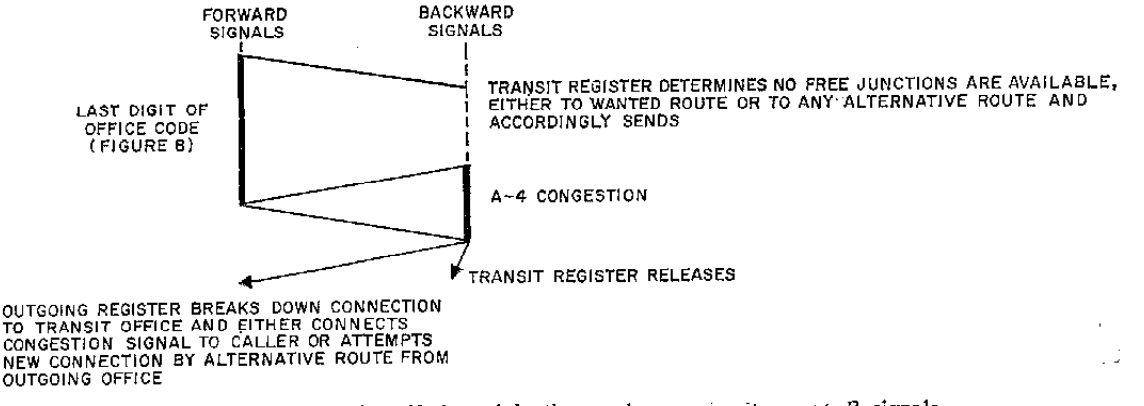


Figure 12—Procedure if the originating exchange cannot accept B signals.

similar network groups. In medium- and small-size countries, these pivotal switching centers constitute the highest rank of switching centers and many or all of them are interconnected by direct toll lines. The switching centers referred to here will be called "primary toll offices." (Very large countries will be divided into regions in which one of the primary toll offices acts as a regional center, all regional centers being directly interconnected, but the situation does not differ essentially from that in small- or medium-size countries.)

The network group served by a primary toll office is designated in some countries as a district or zone. In the following, such a network group will be called a "primary toll area." Its network in principle is of the star type with one or two intermediate ranks of switching centers between the primary toll office and the distant toll end offices located within the primary toll area.

In descending order, these switching centers will be called "secondary toll office" and "tertiary toll office" and the areas served by them "secondary toll area" and "tertiary toll area." The network within each secondary toll area and tertiary toll area (if existing) is again in principle of the star type, the toll end offices constituting the outer periphery of the integrated star-shaped network.

Evidently, offices of lower rank will be incorporated in each office of higher rank to serve the lower rank of area directly associated with each of these offices of higher rank.

From the point of view of numbering, there is a difference between some countries and others as concerns the area over which subscribers may obtain connection with one another by the use of only the local calling number, that is, the number used for connections within the local area.

A primary toll area is said to have open numbering when the use of local numbers is restricted to calls within each local area. In such a case, for establishing a connection between any two different local areas, even if they are within the same secondary or tertiary toll area, the local calling number of the wanted subscriber must be preceded by dialling the national "toll code" of the wanted toll end office, and each toll end office will have a distinct toll code.

A consequence of the use of open numbering is that the number of digits of the local calling number is determined independently for each local area according to the numbering capacity required for this area alone.

A primary toll area is said to have closed numbering when the use of local numbers is extended to calls for all subscribers within this area. This means that all subscribers within the primary toll area share a common block of numbers that will be divided among all local offices within the area according to the number of subscribers they have to serve or will be expected to serve in the future. Usually, all subscribers' numbers within this "closed numbering area" have an equal number of digits determined by the total numbering capacity required for the whole of the area, although the possibility of using nonuniform numbers is not excluded. In the case now considered, it will be evident that local calling numbers must be dialled for all calls completed within each primary toll area and that this number should be preceded by a national toll code only for calls between subscribers located in different primary toll areas. It further follows that only one toll code is needed for each primary toll area and these toll codes are usually assigned to the primary toll offices to identify the primary toll area served by them.

In the following sections (3.2.2 through 3.2.4), it will be shown that this difference in numbering influences the signals employed for making connections between different local areas.

In some countries, a numbering plan is employed that may be considered to be somewhat intermediary between the two cases described above, in that connection may be obtained between adjacent primary toll areas by using

the local number preceded by only the last or the last two digits of the toll code instead of by the whole toll code. This is not further considered, as it may be regarded as a special case of the closed numbering plan referred to above.

3.2.2 *Connections within Primary Toll Area*

3.2.2.1 *Closed Numbering*

With the use of closed numbering for a primary toll area, the whole of this area may be considered from the point of view of switching and signalling as one large multioffice local network. As all connections employ only the local calling number, the operations proceed exactly as described in Sections 3.1.1 through 3.1.6 for a local multioffice area, and will normally take place under the control of leading registers located at the toll end office associated with the originating local area.

In some cases, however, it may be considered propitious to provide leading registers at the primary toll office or at a secondary or tertiary toll office on which the originating office depends for outgoing traffic. For connections switched through one of these offices of higher rank, the leading register will take control of that part of the connection at and beyond the office at which it is located.

This method of operation may be found particularly desirable to concentrate at the toll office, at which these leading registers are located, the equipment required for determining the tariff and for applying call charging on calls leaving these toll end offices.

It should be noted that in connection with primary toll areas with closed numbering as now considered, a particular terminology is in use and is defined in the terms used here under a separate heading in the appendix, Section 5.2.

3.2.2.2 *Open Numbering*

For any call leaving the local area, the calling subscriber must commence by dialling the national toll code of the wanted end office, comprising the national access code and a significant number identifying the wanted toll end office.

The national *access code* is used to obtain access from the originating local office to the *toll end office serving the originating local area.* In large local areas, the toll office may be in a separate building and reached from the originating local office either directly or via a local transit office. Any selection operation will be controlled by the toll access digit(s), using the principles of operation described for calls within a local area. In local areas served by a single local office, this may be located in the same building as, and be integrated with, the toll end office.

The significant number of the toll code is used to establish the connection to the wanted toll end office, which connection may be routed either via a direct toll line or via one or more of the primary, secondary, or tertiary toll offices (within the primary toll area, in the case now considered).

When a direct line is available to the wanted toll end office, the leading register controlling the establishment of the toll part of the connection (this will usually be the outgoing toll register located at the originating toll end office) recognizes this fact and will, having selected a line to the wanted end office, immediately transmit the first digit of the wanted subscriber's local number after which the operation will proceed exactly as described for a local call in Section 3.1.1.

When, however, the wanted toll end office is integrated with a toll office of higher rank, the selection of a direct line may not be followed by the transmission of the wanted local subscriber's number, as this line may have to serve also for connections to be switched in transit through this office.

In such a case, it is necessary to transmit to this office at least that part of the toll code that will permit its incoming registers to determine

whether a terminating or a transit connection is involved. This number of digits varies with the numbering plan and with the rank of the toll office concerned.

The numbering plan will normally identify toll offices of higher rank by fewer digits of the toll code than those of lower rank; the number of digits usually increases by one for each rank of lower order.

In each case, the incoming register at the toll office to which connection is obtained will, by the successive transmission of signals *A-1*, call for as many significant digits of the toll code as is required to determine how the connection should be routed.

If the call terminates at the toll office concerned, the complete toll code must be received. After this, the transmission of the first digit of the local number is initiated by signal *A-1* and the connection proceeds as for a local interoffice connection (Section 3.1). Either the incoming toll register may perform the same functions as a local transit register in a local connection or it may delegate these functions to separate registers that are provided for handling terminating connections and are associated with outgoing junctions from the toll office to the local office equipment.

For a connection within the primary toll area that must be switched through one or more primary, secondary, or tertiary toll offices in this area, the operations at each of these offices preceding the primary toll office and including the latter will commence with the transmission of the first significant figure of the toll code followed by as many more requested digits as are necessary to determine the direction to be selected, which may be the terminating local area, another toll office of lower rank dependent on the office considered, another office of equal rank to which direct lines are provided, or an office of higher rank on which the office considered is dependent.

If the connection is directed to a toll office of higher rank, it is necessary that the leading register transmit to this office the information starting with the first of the significant digits of the toll code. This is brought about by acknowledging the last digit received at each office by signals *A-2*, *A-7*, or *A-8*, depending on whether 2, 3, or 4 digits were received.

In case the connection is directed to a toll office of equal or lower rank, it is usually unnecessary to transmit to this office all significant digits of the toll code; the digits identifying the area associated with the office of the next higher rank on which it is dependent may be omitted.

This again is brought about by the use of the appropriate proceed-to-send signals *A-1*, *A-2*, *A-7*, or *A-8* for the acknowledgment of the last digit received.

3.2.3 *Connections between Different Primary Toll Areas*

3.2.3.1 *Closed Numbering*

For calls leaving the primary toll area of the calling subscriber, the latter must commence by dialling the national toll code of the wanted primary toll area (or primary toll office). This code comprises one or two access digits and a significant number identifying the wanted primary toll office.

The *national access code* is used to obtain access from the originating local office to the *primary toll office serving the originating local area*. This part of the called number controls the connection through the toll end office serving the local area and if necessary through a secondary and a tertiary toll office to the primary toll office. At all of these offices, receipt of the toll access code alone will cause the selection of the shortest possible route to the primary toll office. Furthermore, this same numerical information will also be transmitted first to the incoming register at the primary toll office, where it serves to select the outgoing toll equipment, which comprises toll registers acting as leading registers for handling the connection on the toll network beyond the primary toll office.

It will be obvious that each of the offices involved in establishing the connection up to this point, by omitting the transmission of an acknowledgment signal if the access code is only one digit and by using signal *A-2* for acknowledgment of the second access digit if two access digits are employed, will cause the transmission of the access digit(s) in the manner described. The incoming register at the primary toll office is the last to receive this information and will release after it has caused the attachment of a leading toll register.

This procedure is employed so that the same switching and line equipment available for interconnecting the toll end offices with the primary toll office may serve both of the categories of connections dealt with here and in Section 3.2.2.1, which use different numerical information for controlling the connection.

The leading toll register at the primary toll office now initiates transmission from the originating register of all numerical information commencing with the first of the significant digits of the toll code. This first digit is transmitted because the incoming register at the primary toll office acknowledged the (last) digit it received with signal *A-1*.

The establishment of the connections via the toll network that interconnects the primary toll centers is controlled by the leading toll register and the significant part of the toll code, which will be transmitted to any other primary toll office involved in the connection, to select the shortest possible route to the wanted primary toll office. This part of the operation may be compared with that taking place in a large local multioffice area. If the primary toll offices are interconnected by direct lines, the operation is similar to that described in Section 3.1.1 for a direct connection between two local offices. If one or more of the primary toll offices may act as toll transit offices for primary toll offices not directly interconnected, operation is similar to that described in Section 3.1.2 for a transit connection in a local area. Also alternative routing or rerouting may take place in the toll network as described for similar circumstances in Sections 3.1.3 through 3.1.5 for a local area. However, in all of these cases, the selection will be controlled by the significant part of the toll code, instead of by the office code as described in connection with local calls.

3.2.3.2 *Open Numbering*

The description given in Section 3.2.2.2 will apply to the establishment of connections between different primary toll offices, taking into consideration that all of these offices are of the highest rank. This implies that at all of these offices only enough significant digits of the toll code will be received (commencing with the first of these digits) as are needed to determine whether the connection should terminate in the associated primary toll area or whether it must be switched in transit to another primary toll office (considering also that, in the same manner as described in Section 3.2.3.1, one or more of the primary toll offices may act as toll transit offices for primary toll offices not directly interconnected). It will also be evident that in this case, in the same manner as described in Section 3.2.3.1, the operation on the part of the toll network interconnecting the primary toll offices may be compared with that of a large local multioffice area, so that all operating possibilities may be provided as described in Sections 3.1.1 through 3.1.5 in connection with a local area.

As exactly the same types of signals are employed for the switching and line equipments within the primary toll areas as on the toll network interconnecting the primary toll offices, there is no reason, as in the case described in Section 3.2.2.1, to provide separate leading toll registers. The incoming registers may assume the function of leading registers for the type of connection now considered, without appreciable additional complication of these registers.

3.2.4 *Connections Terminating in Primary Toll Area*

3.2.4.1. *Closed Numbering*

The establishment of connections entering from other primary toll areas into a primary toll office will be completed in the terminating primary toll area under the control of incoming toll registers at the terminating primary toll office, which will perform the same functions as leading registers at this office charged with the completion of connections within the primary toll area. In the same manner as these leading registers, these incoming registers will receive the complete local number and control all subsequent operations within the terminating primary toll area.

So that the incoming register at the primary toll office will receive only the local subscriber's number and not the toll code, receipt of the last digit of the toll code at the preceding office will be recognized by signal *A-1*, which may be performed by the register at this preceding office based on its selection of a direct line to the terminating primary toll area.

In case the primary toll office acts as a toll transit office for connections between primary toll offices not directly interconnected or for the purpose of alternative routing between primary toll offices, it may be economical to combine the functions of the incoming toll registers with those of the toll transit registers, if the amount of toll transit traffic is insufficient to warrant provision of separate registers.

The use of a combined register for terminating and toll transit connections implies, however, that discriminative information be transmitted to this register to indicate whether the connection is terminating or in transit. Because this information may not be derived from the significant part of the toll code or from the local number, an obvious solution is to precede the toll code, in case of a toll transit connection, by the toll access code.

3.2.4.2 *Open Numbering*

For connections entering a primary toll office from other primary toll offices, the description given in Section 3.2.2.2 applies, taking into consideration that any office through which the connection must be routed subsequently will be of lower rank. Consequently, the part of the significant number of the toll code identifying the primary toll area need not be transmitted to subsequent offices.

3.2.5 *Control of Selection at Terminating Toll Area*

The control of the selection at the terminating toll area may be arranged in different ways depending on the facilities afforded by the switching equipment existing there.

If the equipment is capable of accepting multifrequency register signals, the operations will proceed directly under the control of the leading toll register at the originating toll office. An incoming register at the terminating office will then successively initiate the transmission from the leading register of the digits constituting the local office code. In all further respects, this incoming register also will act as a local transit register as shown in Figures 8 through 11, in which operations the leading toll register will perform the functions of an originating register.

If the equipment at the terminating area does not accept multifrequency signals, convertor registers will be needed to convert the multifrequency signals to the type of selective information required by the terminating switching equipment. These convertor registers may be located either at the terminating office to serve the whole of the area for all incoming calls or at individual offices if only some of the offices in the terminating area require different selective information. In the latter case, these convertor registers may handle toll calls in addition to incoming calls from other offices employing multifrequency signalling within the terminating area.

In each of these cases, the convertor register will cause the leading register to transmit all digits required to complete the connection by the successive transmission of acknowledgment signals *A-1*, until the units digit has been received. The units digit may be acknowledged by an appropriate backward signal from the convertor register, depending on the possibilities of the terminating switching equipment to signal the condition of the called line to this register, as explained in conjunction with Figures 4 through 7.

Another possibility consists in the use of the backward signal *A-9* or *A-10* from the last transit register in the connection that may accept multifrequency signals, for the purpose of informing the leading register that this has to transmit all remaining numerical information in a different form, for example, by decimal pulses. This will be possible only when all toll lines and junctions involved in the connection between the office at which the leading register is located and the terminating office are equipped for the transmission of decimal pulses.

3.3 International Connections

Disregarding certain differences that will be explained later, *the operating conditions with the new signalling methods as applied to international connections in many respects are comparable to those prevailing in a national toll network* employing a closed numbering system for each toll area.

Similarly to the assignment in a national network of so-called national toll codes to each primary toll area, which codes must be dialled ahead of the local subscriber's number only for connections between different primary toll areas, in the international network so-called international toll codes are assigned to each country and have to precede the national number only in case of international connections.

In a national network, each primary toll area is served by a primary toll office. All primary toll offices are interconnected by the national toll line network either with direct lines or via one or more of these primary toll offices serving as national toll transit offices. The selection at each of the toll offices involved in a connection through this network is determined by the significant part of the national toll code. In the international network, each country in principle is served by one principal "international toll office" via which access may be obtained to all subscribers within the country. These international toll offices are mutually interconnected by the international toll line network, either with direct lines or via one or more of them serving as international toll transit offices. The selection at each of the international toll offices involved in a connection is determined by the significant part of the international toll code dialled.

In a national toll network, the establishment of toll connections is under the control of leading national toll registers located at the originating toll office. Similarly, for international connections, leading international toll registers will control the establishment of the connection on the international network and will be preferably located at the originating international toll office.

Also comparable to a national system in which a larger number of register signals is provided for the exchange of information between toll registers than is employed by the local registers handling the connections within each toll area, the number of signals employed for international connections is larger than that used within national systems, as may be seen by referring to Tables 3 and 4.

Another point of resemblance is that in a national toll network distinction between local and toll calls is provided by dialling a national toll access code for the latter calls (that is, a figure or combination of figures not employed in the local numbering system) in a position similar to that for the local numbers; for international calls it is necessary to dial an international access code, after which the significant part of the international toll code may be dialled

to signify the distant country to which the connection should be directed.

In some other respects, the comparison between national and international operating methods does not hold. A few cases will be mentioned.

3.3.1 *More Than One International Toll Office*

Some countries may have more than one international toll office to permit more economical routing to and from different adjacent countries than would be provided by a single international toll office. However, the same international toll code must be dialled for calls to such countries, independently of the incoming international toll office through which they will enter the country. A particular case influencing the signalling may arise if calls from other countries may enter through any of several of these incoming international toll offices, depending on the location of the called subscriber within the country of destination. Then, the registers at the outgoing international toll office of the neighboring country (or at the last international toll transit office as the case may be) have to determine from the significant part of the called subscriber's national toll code to which of the incoming international toll offices at the country of destination the call should be routed. Obviously, these registers receiving the last digit of the international toll code indicating the country of destination, must, before effecting a selection, call for a sufficient number of digits of the national toll code, which follows the international toll code, to determine the most-economical routing. After one of the incoming international toll offices at the country of destination has then been selected, it will normally be necessary to transmit to this office the complete national toll code. The backward signal *A-13* (send first digit of national number) will then be used for acknowledging the last digit received by the register of this incoming international toll office.

If the incoming international toll office serving the country of destination may also be used as an international toll transit office, it may be necessary to transmit to this office the international country code so it may determine that the call is terminating in the country concerned. The backward signal *A-11* (send first digit of international country code) is then employed by the register at the neighboring country for acknowledging the last digit received, so that the desired effect is produced.

3.3.2 *Special Requirements Applying Only to International Operation*

There are a number of special requirements applying only to international operation, such as recording at each originating country the charges and routing of calls for use in the division of revenues on international calls between different administrations. This does not affect the signalling methods, however. In certain cases, congestion signals transmitted from international transit toll offices to the originating international toll registers may signal the fact that an alternative route has been taken at a transit office. This permits a record of the routing to be kept at the originating end.

3.3.3 *Echo Suppressors*

Another question affecting signalling is the need *for the insertion of echo suppressors* in certain long-distance connections. Signals are given for this purpose in Tables 3 and 4.

Although no definite method of operation for the insertion of echo suppressors has been decided on, it should be noted that during the process of compelled end-to-end register signalling echo suppressors may not be used because they would prevent the simultaneous transmission of signals in both directions. Arrangements will, therefore, have to be provided to insert echo suppressors in the speech path if the routing information indicates need for them but only after the exchange of register signals is completed.

Furthermore, it may also be found necessary to disable echo suppressors after a connection has

been completely established to permit data transmission with some form of error correction.

The most suitable location for echo suppressors would be at the outgoing and/or incoming international toll offices, at which the speech path is through connected only after the international registers are disconnected.

3.3.4 *Transborder Connection*

A particular type of international connection is the *transborder connection.* This is a relatively short connection between towns on opposite sides of an international border that is not routed through either of the principal international toll offices of the two countries concerned. To establish such connections, it is necessary that the register at the border town in the originating country examines the significant part of the international toll code and at least a part of the called subscriber's national toll code to determine whether or not the connection may be established via the transborder line. The fact that this is *not* possible and the call should be routed through the originating country's principal international toll office can be established only after a variable number of digits have been examined, namely, the international toll code alone or this code plus a variable number of digits of the national toll code that will provide the necessary distinction.

This may consequently require the repetition of a variable number of digits from the originating register, which must then be transmitted to the principal outgoing international toll register. This will be effected by the transmission of an appropriate backward signal from the register at the border town, when it has determined that no transborder connection is permitted.

3.4 Identification of Calling Line's Number

A number of optional signals have been included in the Tables of Signals that administrations are free to use in their own network for specific national purposes. Two of these signals are backward signals *A-9* and *A-10,* of which some use has already been described (see Section 3.2.5), if it is desired to change over from multifrequency-code signalling to dial pulsing at a certain point during the establishment of the connection.

Another purpose for these signals is to identify the calling subscriber's number as may be needed for automatic message recording and accounting on toll connections.

These signals permit the concentration of automatic message recording equipment at important toll offices and for the transmission to that toll office of the calling number via the toll lines used for establishing the toll connection from the originating office. The use of multifrequency-code signalling permits the calling number to be transmitted very rapidly while the switching operations take place at the toll office.

For this purpose, after the register at the toll office has received the last digit of the called number needed to effect its selection, it will acknowledge this digit by backward signal *A-9* or *A-10,* which requests transmission of the first digit of the calling line's local number or national number, respectively.

Two distinct signals are needed because in certain cases where the automatic message recording equipment is located at the primary toll office serving a primary toll area with closed numbering, it is sufficient to obtain the local number alone as all subscribers will have the same national toll code. In other cases, the recording equipment may serve several closed-numbering areas, for example, when this equipment serves for recording international calls exclusively and is located at the international toll office. It will then be necessary to obtain the complete national number of the called subscriber, less the access code.

Receipt of either *A-9* or *A-10* by the originating register will cause the calling number to be identified and will connect a forward signal

indicating the value of the first digit of this number. The receipt of this and all following digits except the last of the calling number will be acknowledged by the same backward signal *A-9* or *A-10*, which thereby assumes the meaning of transmit-next digit of calling number.

The last digit is acknowledged (if this is known to be the last digit at the receiving register) by backward signal *A-1*. This restores both the transmitting and receiving registers to the condition in which selection may be resumed and which prevailed before signal *A-9* or *A-10* was sent first, except that the originating register will be initiated to transmit the next digit of the called number. See Figure 13.

If it is not known at the receiving register how many digits are in the calling number, as in the case of nonuniform numbering, the transmitting register, after having received the signal *A-9* or *A-10* acknowledging the receipt of the last digit of the calling number, will send the forward signal *A-12* (demand rejected). This will in its turn be acknowledged by backward signal *A-1*, after which both registers are again in the condition in which selection may be resumed with the transmission of the next digit of the called number. See Figure 14.

On international calls it is possible by using backward signal *A-13*, to obtain in a similar manner from the originating country the significant number of the international toll code signifying this country, and, if desired, part or all of the national number of the calling subscriber. The maintenance staff may thus be informed of the origin of the connection.

4. Conclusion

Agreement has been reached among many of the leading telephone administrations in Europe on a flexible and adaptable system of signalling in the control of telephone switching equipment for national and international networks. It uses the same principles of operation throughout these networks and provides high operating speed and inherent reliability.

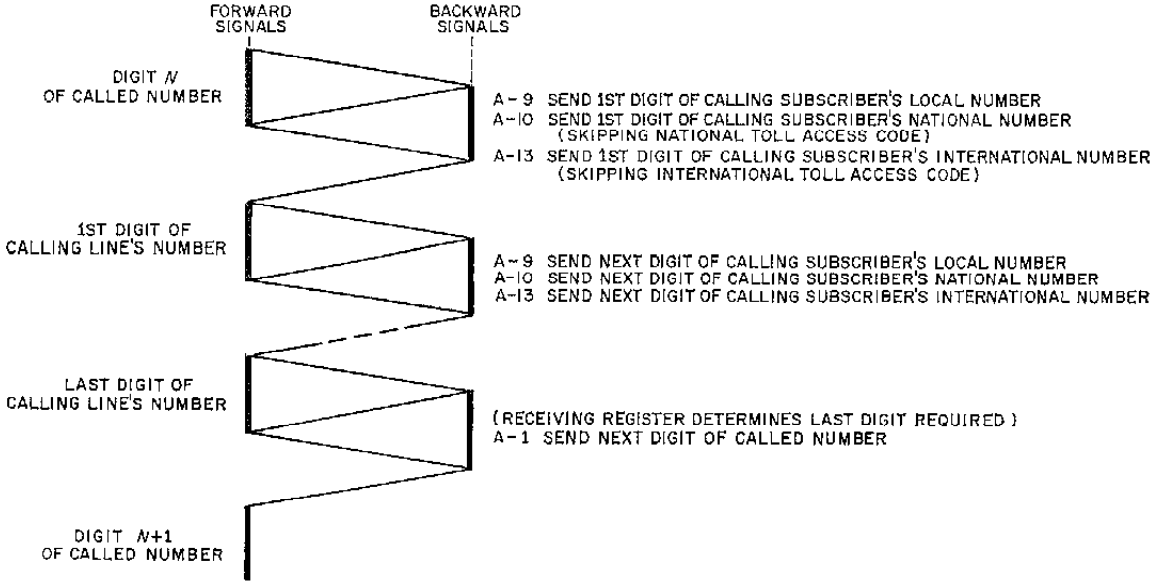


Figure 13—Identification of calling number if the last digit can be determined by the receiving register.

5. Appendix

5.1 Definitions and Functions of Line Signals

The definitions and functions of the line signals mentioned in Section 2.1 may be described similarly to those listed under the corresponding headings in "Recommendation Q 60" contained in the Red Book of the Comité Consultatif International Télégraphique et Téléphonique, volume 6, pages 93 through 98.

As some of the signals mentioned in Section 2.1 do not appear in the Red Book and others have a slightly different function, a complete list follows.

In these definitions the terms "outgoing termination" and "incoming termination" refer to that line termination, and the terms "forward signal" and "backward signal" refer to that direction of transmitting a signal to which these terms apply with respect to the direction in which the connection is established. This means that the terminations of both-way circuits may be either outgoing or incoming, depending respectively on whether they are located at the office from which the line is seized for a particular connection or at the office to which this connection is directed via the line.

Seizure (Forward Signal): This signal is transmitted on a line circuit after it has been engaged for a connection at its outgoing termination in order to switch the incoming termination into the "nonidle" condition and to initiate the connection to the incoming termination of an equipment capable of controlling the switching operations at the incoming office. It

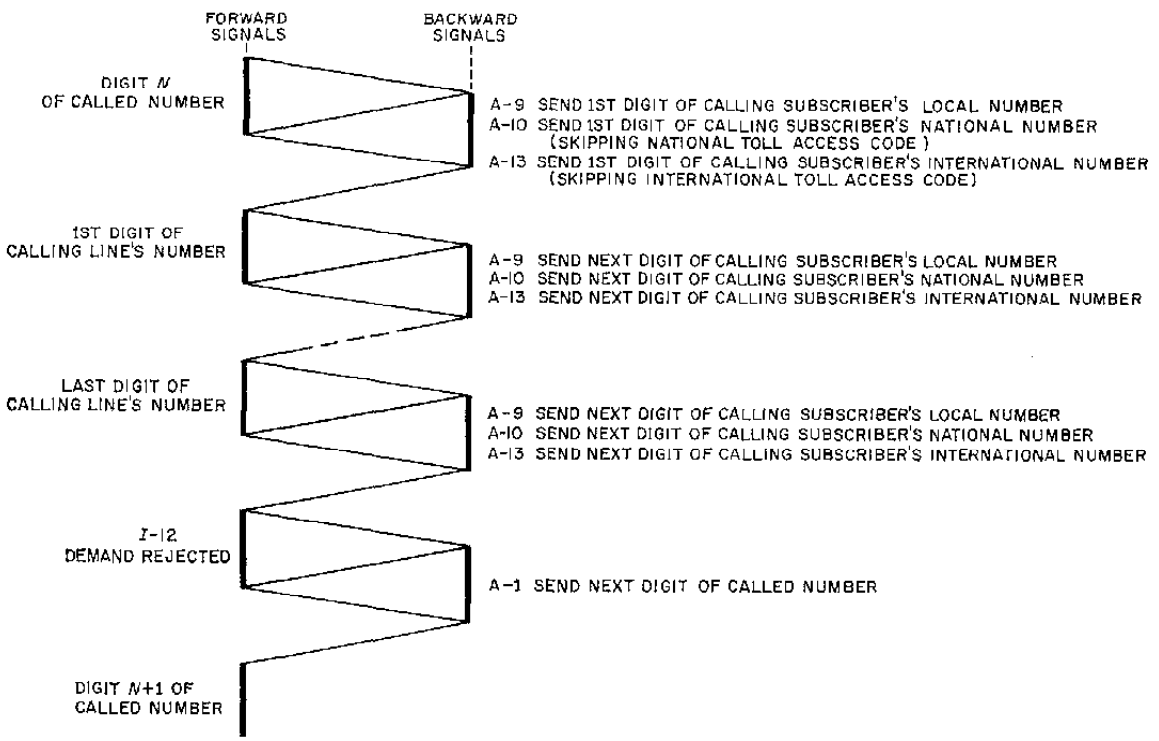


Figure 14—Identification of calling number if the last digit cannot be determined by the receiving register but can be by the transmitting register.

should be noted that no distinction is provided between terminating and transit connections by using two corresponding seizure signals, as it is considered preferable to provide this distinction by numerical signals.

Clear Forward (Forward Signal): This signal is transmitted from the outgoing to the incoming termination of a line circuit at the end of a connection, when (*A*) in the case of semiautomatic operation, the operator controlling the connection withdraws her plug from the line jack or accomplishes an equivalent operation, or (*B*) in the case of automatic operation, the calling subscriber replaces his telephone on the switch hook or accomplishes an equivalent operation as in the case of a subscriber's installation with supplementary telephone sets.

In automatic operation, this signal is automatically sent on receipt of certain backward signals indicating that the connection cannot be completed. This may be due to many reasons, such as: congestion in the switching or line equipment; called line being busy, out-of-order, et cetera; need for forced release because of the failure of the calling subscriber to clear within a predetermined interval subsequent to the receipt of the "clear back" signal; no reply, with calling subscriber failing to clear within a predetermined interval subsequent to the receipt of the signal indicating "end-of-selection" or "called subscriber free"; and abnormal operation at a transit center, when a "release guard" signal is received without a preceding "clear forward" signal.

At the termination of the "clear forward" signal, all switching equipment engaged in the connection must be liberated at both terminations of a line circuit, that is, at the outgoing, incoming, or transit office at which the line circuit terminates. (See, however, "Release Guard", below.)

Forward Transfer (Optional Forward Signal): This signal is transmitted in semiautomatic connections on international lines when the controlling operator at the outgoing intercontinental toll office wishes the assistance o an operator at the incoming international tol office. It will be transmitted from an outgoing line termination at the first-mentioned office tc an incoming line termination at the last-mentioned office, without taking effect at transit offices through which the connection is switched and will serve normally to call for an assistance operator on a connection switched automatically through the incoming international toll office If the connection at this office was switched manually by an operator, the signal will serve to recall this operator (operators with code *11* or *12*).

Toll Offering (Optional Forward Signal): This signal is transmitted in semiautomatic toll connections within a national network arranged for this type of service. It is used when the controlling toll operator, on finding a wanted number busy, and without breaking down that connection, wishes to communicate with this subscriber to inform him that a toll connection is in preparation for him. The toll offering condition may be undone by the transmission of the "end of toll offering" signal.

Reringing (Optional Forward Signal): This signal is transmitted in semiautomatic toll connections within a national network arranged for this type of service when the controlling toll operator wishes to rering a subscriber who had cleared after having been in communication with the operator previously. Normally, ringing is started by this signal, which is transmitted when the operator throws the ringing key, and is continued until, by restoring the ringing key, the ringing condition is undone. This may be produced by using a pulse signal for starting the ringing and another pulse signal "end-of-ringing" to stop the ringing, or alternatively, by using for the reringing signal a train of pulses that continues as long as the ringing key is thrown.

In particular cases, for example, when the toll switching circuits serving a local network are so arranged that the automatic ringing of the called subscriber on the selection of the called line is suppressed for certain types of operator's connections, these operators may initiate either manual or machine ringing by the use of the reringing signal.

Acknowledgment of Seizure (Optional Backward Signal): This signal is transmitted from the incoming to the outgoing termination of a line circuit immediately on receipt and recognition at the incoming termination of a seizure signal. It will be transmitted without awaiting the result of the switching operations (connection of a free register) that are initiated by the seizure signal and will merely serve to inform the outgoing termination that it may cease the transmission of the seizure signal. This signal is normally employed in conjunction with the seizure signal to provide for "compelled operation" by mutual control of these two signals.

Answer or Reanswer (Backward Signal): This signal is transmitted from the incoming to the outgoing termination of a line circuit to signal that the called subscriber has replied and, in general, to indicate the off-hook condition of the called line.

On semiautomatic connections, this signal will permit the controlling operator to supervise the answer of the called subscriber by visual means.

On automatic connections, this signal will initiate:

A. Commencement of charging for the connection by any appropriate means provided for this purpose, or of measuring the chargeable period.

B. Discontinuation of measuring the time interval before answering, which, if exceeded, would cause the connection to release forcibly (if provided).

C. Completion of the forward speech channel in certain cases.

Clear Back (Backward Signal): This signal is transmitted from the incoming to the outgoing termination of a line circuit to signal that the called subscriber has replaced his receiver on the switch hook, and, in general, to indicate the on-hook condition of the called line. In semiautomatic operation, this signal will permit the controlling operator to supervise the clearing of the called subscriber by visual means.

In automatic operation this signal will initiate:

A. Start of measuring a time interval, after which charging of the calling subscriber will be interrupted (or the measurement of the chargeable period will be discontinued) and the connection will be forcibly released if the calling subscriber had not cleared previously.

B. Interruption of the forward speech channel in certain cases.

Release Guard (Optional Backward Signal): This signal is transmitted from the incoming termination of a line circuit, either at an incoming or a transit office, to its outgoing termination, either at a transit or an outgoing office, immediately on the receipt and recognition at the incoming termination of a clear forward signal. It will be continued until the moment the incoming termination, in response to the clear forward signal, has been caused to switch into the idle condition, excepting from this condition the fact that the backward release guard signal is transmitted. (Which means that also the clear forward signal must have terminated.) The discontinuation of the release guard signal at this moment will cause both outgoing and incoming terminations to become simultaneously available for a next connection.

The presence of this signal will permit the clear forward signal to continue indefinitely, until its recognition at the incoming termination has been signalled to the outgoing termination by the release guard signal, so that compelled operation is thereby provided by the mutual control of these two signals.

Blocking (Optional Backward Signal): This signal is transmitted from an incoming to an outgoing termination of a line when it is desired at the incoming office to render the circuit inaccessible at its outgoing termination. The receipt of this signal at the outgoing termination in its idle condition causes the signalling equipment to mark the circuit "busy" to operators or automatic selectors normally having access to it. The blocking condition is different from a genuine busy condition in that it will produce an alarm at the outgoing termination after a predetermined delay, so that if the blocking condition is continued after this delay, the maintenance staff is warned to investigate this abnormal case and take appropriate measures. Normally, the blocking signal should be employed for a short period only, for example during routine testing of the incoming terminating equipment.

Acknowledgment of Optional Forward Signals (Optional Backward Signals): In conjunction with each of the forward signals mentioned in items 5.1.3 through 5.1.5, a backward acknowledgment signal may be optionally provided so that each pair of forward and backward signals thus constituted will provide for compelled signalling by mutual control.

5.2 Definitions of Terminology

Office: General designation of a switching center.

Local Office: Office to which the subscribers in one locality (or its equivalent) or part of a locality are connected.

Originating Office: Office to which the subscriber is connected who originates the connection considered.

Terminating Office: Office to which the called subscriber is connected for the connection considered.

Local Area: Area served by the office(s) of one locality.

Toll Office: Office employed for the establishment of connections between different local areas.

Primary Toll Office: Toll office of highest rank in a national toll network. Many or all of these offices are directly interconnected by toll lines, constituting the principal national long-distance toll line network.

Transit Toll Office: Primary toll office used for switching transit connections between other primary toll offices not directly interconnected or for alternative routing of connections between other primary toll offices interconnected by direct lines.

Primary Toll Area: Area for which a primary toll office is the pivotal switching center for connections to and from other similar areas, served by other primary toll offices.

Secondary Toll Office: Toll office of second rank within a primary toll area and that is directly dependent on a primary toll office for connections to and from other primary toll areas.

Secondary Toll Area: Area for which a secondary toll office is the pivotal switching center for connections to and from other offices of equal and higher rank.

Tertiary Toll Office: Toll office of third rank within a secondary toll area and that is directly dependent on a secondary toll office for connections to and from other toll offices of higher rank.

Tertiary Toll Area: Area for which a tertiary toll office is the pivotal switching center for connections to and from other areas of equal or higher rank.

Toll End Office: Toll office serving one local area. Toll end offices constitute the toll offices of lowest rank in a national toll network and are dependent on a toll office of the next higher rank (primary, secondary, or tertiary toll office depending on how many other ranks of toll offices exist), for connections to other toll offices of higher rank.

Originating Area: Area in which the connection considered is originated.

Terminating Area: Area in which the connection considered is terminated.

Originating Toll Office: Toll office that incorporates the leading registers controlling the establishment of the toll connection considered.

Terminating Toll Office: Toll office that incorporates the registers controlling the establishment of an incoming toll connection within the terminating area.

Single-Office Area: Locality (or its equivalent) served by one local office.

Multioffice Area: Locality served by more than one local office.

Local Number: Subscriber's number used for establishing connections between subscribers connected to the same office. This same number may be used by subscribers connected to offices other than that to which the called subscriber is connected, when these offices are located in the same closed numbering area as the terminating office.

Closed Numbering Area: Area comprising all offices the subscribers of which may mutually call one another by using the local number only.

Closed Numbering System: System of allotting subscriber's numbers in a national toll network according to which each closed-numbering area comprises several local areas. In most cases, this area may be identified with a primary toll area.

Open Numbering System: System of allotting subscriber's numbers in a national toll network according to which each closed-numbering area comprises one local area only.

National Number: Subscriber's number to be used for a connection within a national network when originated outside of the closed-numbering area in which the designated subscriber is connected. The national number is the local number designating a subscriber preceded by the national toll code, designating the closed-numbering area in which this subscriber is connected.

National Toll Code: Part of a national number preceding the local number and comprising the national toll access code and a significant number identifying the closed-numbering area in which the designated subscriber is connected.

National Access Code: Code of one or more digits by which national numbers commence and by which these numbers are distinguished from local and international numbers.

International Number: Subscriber's number to be used for a connection when originated outside of the national network in which the designated subscriber is connected. The international number is the national number less the national access code, preceded by the international toll code.

International Toll Code: Part of the international number preceding the national number less the national access code and comprising the international access code and a significant number identifying the country in which the designated subscriber is connected.

International Access Code: Code of one or more digits by which international numbers commence and by which these numbers are distinguished from local and national numbers.

International Toll Office: Office at which toll lines terminate interconnecting different national toll networks and through which the international traffic handled via these toll lines is switched to and from the whole or part of the country in which the office is located.

International Transit Toll Office: International toll office used for switching transit connections between other international toll offices not directly interconnected, or for alternative routing of connections between other international toll offices interconnected by direct lines.

Local Register: Register serving principally for the establishment of connections within a local area and provided at local offices.

Toll Register: Register provided at toll offices.

Outgoing Toll Register: Register at the toll end office serving the calling subscriber for a connection considered.

Incoming Toll Register: Register controlling the completion of connections within the terminating toll area.

Leading Register: Register located relatively near to the calling subscriber and charged with supervising the establishment of toll connections. This register may be located at the originating toll end office or at a toll office subsequently engaged. It will receive the full numerical information required for completing the connection at and beyond the toll office at which it is located and supply any part of this information to any subsequent office engaged, on receipt of appropriate backward signals. It will remain connected until it receives a backward signal informing it on the final result of the selective operations.

Leading National Toll Register: Leading register supervising the establishment of national toll connections. This may be located at the primary toll office serving the calling subscriber or at any preceding toll office.

Leading International Toll Register: Leading register supervising the establishment of international toll connections. This may be located at the international toll office serving the calling subscriber or at any preceding toll office.

Toll Line: Any type of transmission channel providing both-way speech possibilities between any two toll offices.

Terminology in use particularly in connection with primary toll areas with closed numbering (applied sometimes also to such areas with open numbering)

Rural Area: Primary toll area, but usually not including the local area directly associated with the primary toll office.

City Area: Local area associated directly with primary toll office.

Rural Main Office: Part of primary toll office required for switching connections betweer local areas within the primary toll area.

Toll Office: Part of primary toll office used for switching connections between any local area within the primary toll area on the one hand and other primary toll areas on the other hand

Rural Office: Office equipment serving any local area within the primary toll area, excepting that directly associated with a primary toll office, and usually including both the local and toll offices associated with that local area.

(Rural) Sector Center Office: Office equipment associated with a local area and comprising a secondary toll office.

(Rural) Subcenter Office: Office equipment associated with a local area and comprising a tertiary toll office.

(Rural) End Office: Office equipment associated with a local area and comprising a toll end office.

(Rural) Sector: Secondary toll area.

(Rural) Subsector: Tertiary toll area.

Rural Junction: Toll line within primary toll area.

Toll Line: Toll line interconnecting primary toll offices.

Tie Line: Toll line interconnecting two toll offices in nonhierarchical manner.

Martinus den Hertog was born in Utrecht, The Netherlands, in 1901.

He became a student engineer for Bell Telephone Manufacturing Company in 1921, working until 1925 as an equipment engineer. He then joined the circuit engineering group and served as its head from 1933 to 1953.

From 1953 to 1959, Mr. den Hertog was chief engineer and manager of the switching systems group. He became commercial manager of the company in 1959 and two years later was appointed assistant to the technical director of ITT Europe, where he has charge of the coordination of the development of switching systems.

United States Patents Issued to International Telephone and Telegraph System; November 1961–January 1962

Between 1 November 1961 and 31 January 1962, the United States Patent Office issued 36 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

R. T. Adams, B. M. Mindes, and Z. G. Lyon, ITT Federal Laboratories, Post-Detection Diversity Combining System, 3 013 151.

L. D. Apt, ITT Kellogg, Dial Controlled Intercommunication Systems, 3 014 097.

A. H. W. Beck, Standard Telephones and Cables (London), Thermionic Cathodes, 3 013 171.

T. E. Beling, ITT Laboratories, System for Providing Short High-Amplitude Square Voltage Pulse, 3 015 777.

J. I. Bellamy, ITT Kellogg, Decade-Counting Device, 3 007 638.

T. G. Brown, Jr., ITT Laboratories, Decoder for Line Printer, 3 013 119.

P. Cheilik, ITT Federal Laboratories, Magnetic Read and Write System, 3 013 255.

B. Dal Bianco and M. Scata, Fabbrica Apparecchiature per Comunicazoni Elettriche (Milan), Electromagnetic Switch Apparatus, 3 014-102.

C. L. Day, Farnsworth Electronics Company, Method of Making Bowl-Shaped Fine-Mesh Screens for Electron Discharge Tubes, 3 017-687.

E. M. Deloraine and H. G. Busignies, International Telephone & Radio Manufacturing Corp., Radar Counter-Measure Radio Repeater, 3 015 096.

M. J. DiToro, ITT Federal Laboratories, Aural Detection Apparatus Comprising an Acoustical Delay Line Having External Feed-back Circuit, 3 013 212.

A. G. Field, Standard Telephones and Cables (London), Thermionic Valves, 3 013 176.

L. G. Fischer, ITT Federal Laboratories, Radio Range Receiver System, 3 013 264.

J. C. Gibson, ITT Kellogg, Primary-Secondary Link-Spread Crossbar Selector System, 3 017-465.

F. S. Gutleber, ITT Federal Laboratories, Diversity Receiving System Having Separate Phase-Angle Indicators, 3 013 150.

L. B. Haigh and T. W. Tuttle, ITT Federal Laboratories, Telecommunication Link, 3 013-111.

H. Herbig and E. Schmidt, Federal Telecommunication Laboratories, Transistor Oscillators, 3 015 696.

R. C. P. Hinton, B. Dzula, and A. Fettweis, ITT Federal Laboratories, Recording System, 3 013 253.

R. L. M. LeQueau, Compagnie Générale de Constructions Téléphoniques and Le Matériel Téléphonique (Paris), Telephone Systems, 3 008 010.

L. Lewin and Y. Klinger, Standard Telecommunication Laboratories (London), Maser-Type Oscillator, 3 013 217.

F. P. Mason and R. G. Stemp, Creed & Company (Croydon), Facsimile Apparatus, 3 015 690.

B. Minakovic, Standard Telephones and Cables (London), Traveling-Wave Tubes, 3 013 177.

H. Reiner, Mix & Genest (Stuttgart), Transistor Double-Integrating Circuit, 3 013 160.

R. Scheidig, Standard Elektrik Lorenz (Stuttgart), Reed-Contact Thermo Relay, 3 008 019.

R. Schultz and J. Sullivan, Standard Telephones and Cables (London), Trigger Circuit, 3 015 736.

J. Schockaert, Bell Telephone Manufacturing Company (Antwerp), Carrier Supports for Sorting Machine, 3 015 379.

J. R. Simkovich, ITT Federal Laboratories, Radial Resonant Cavities, 3 013 230.

A. T. Starr, Standard Telecommunication Laboratories (London), Semi-Conductor Devices, 3 013 192.

R. Thyen, International Standard Electric Corporation, Circuit Arrangement for Encoding Devices, 3 015 805.

M. Toohig and C. L. Day, Farnsworth Electronics Company, Method of Making Charge-Storage Electrodes for Charge-Storage Tubes, 3 015 586.

A. W. Wallens, Creed & Company (Croydon), Teleprinter Communication Systems, 3 013 112.

J. Weber and C. Heck, Süddeutsche Apparate Fabrik (Nuremberg), Method of Making Ferrites, 3 015 538.

E. P. G. Wright, A. D. Odell, and C. Parker, Standard Telecommunication Laboratories (London), Electronic Equipment Practice 3 015 755.

E. P. G. Wright, Standard Telecommunication Laboratories (London), Data Processing Systems, 3 013 120.

E. P. G. Wright, Standard Telecommunication Laboratories (London), Data Processing Equipment, 3 013 251.

E. Zaratkiewicz and E. Smith, ITT Laboratories, Semi-Conductor Rectifiers and Method of Manufacture, 3 015 591.

Electronic Equipment Practice

3 015 755

E. P. G. Wright, A. D. Odell, and C. Parker.

A mounting rack and functional sub-unit assemblies in which all of the sub-units are identical. These units are shown as basic on-off switching units in the form of sub-panels. These sub-panels are provided with jack terminals for plugging into a prewired rack to provide complete functional equipment.

Method of Making Ferrites

3 015 538

J. Weber and C. Heck

Manganese-zinc mixed ferrites annealed in an atmosphere containing a small amount of oxygen and about ten percent sulphur dioxide. This process achieves a substantial improvement in magnetic properties of the ferrites, particularly a large increase in permeability.

Post-Detection Diversity Combining System

3 013 151

R. T. Adams, B. M. Mindes, and Z. C. Lyon

A diversity receiver post-detection combiner circuit for phase- or frequency-modulated waves provided with devices having a limiting action preceding the individual discriminators, to maintain equal gain in all the channels of the diversity receiving system.

Thermionic Cathodes

3 013 171

A. H. W. Beck

A thermionic dispenser cathode in the form of a relatively large block of sintered refractory metal powder mixed with emissive material, having a very small block of sintered emissive material integral with the larger block and serving to provide the emissive surface.

Electromagnetic Switch Apparatus

3 014 102

B. Dal Bianco and M. Scata

A magnetic-reed-type switch in which the switch envelope is mounted between two plates magnetically energized to normally operate the switch. A magnetic shield is positioned around the switch envelope and is movable out of shielding relation to control operation of the switch.

Principal ITT System Products

Telecommunication Equipment and Systems

Automatic telephone and telegraph central office switching systems
Private telephone and telegraph exchanges—PABX and PAX, electromechanical and electronic
Carrier systems: telephone, telegraph, power-line
Long-distance dialing and signaling equipment
Automatic message accounting and ticketing equipment
Switchboards: manual, central office, toll
Telephones: desk, wall, pay-station
Automatic answering and recording equipment
Microwave radio systems: line-of-sight, over-the-horizon
Radio multiplex equipment
Coaxial cable systems
Submarine cable systems, including repeaters
Data-transmission systems
Teleprinters and facsimile equipment

Military/Space Equipment and Systems

Aircraft weapon systems
Missile fuzing, launching, guidance, tracking, recording, and control systems
Missile-range control and instrumentation
Electronic countermeasures
Electronic navigation
Power systems: ground-support, aircraft, spacecraft, missile
Radar
Simulators: missile, aircraft, radar
Ground and environmental test equipment
Programmers, automatic
Infrared detection and guidance equipment
Global and space communication, control, and data systems
Nuclear instrumentation
Antisubmarine warfare systems
System management: worldwide, local

Industrial/Commercial Equipment and Systems

Distance-measuring and bearing systems: Tacan, DMET, Vortac, Loran
Instrument Landing Systems (ILS)
Air-traffic control systems
Direction finders: aircraft and marine
Ground and airborne communication
Data-link systems
Inverters: static, high-power
Power-supply systems
Altimeters
Flight systems
Railway and power control and signaling systems
Information-processing and document-handling systems
Analog-digital converters
Mail-handling systems
Pneumatic tube systems
Broadcast transmitters: AM, FM, TV
Studio equipment
Point-to-point radio communication
Marine radio
Mobile communication: air, ground, marine, portable
Closed-circuit television: industrial, aircraft, and nuclear radiation
Slow-scan television
Instruments: test, measuring
Oscilloscopes: large-screen, bar-graph
Vibration test equipment
Magnetic amplifiers and systems
Alarm and signaling systems
Telemetering
Intercommunication, paging, and public-address systems

Consumer Products

Television and radio receivers
High-fidelity phonographs and equipment
Tape recorders
Microphones and loudspeakers
Refrigerators, freezers
Air conditioners
Hearing aids
Incandescent lamps
Home intercommunication equipment
Electrical housewares

Cable and Wire Products

Multiconductor telephone cable
Telephone wire: bridle, distribution, drop
Switchboard and terminating cable
Telephone cords
Submarine cable
Coaxial cable, air and solid dielectric
Waveguides
Aircraft cable
Power cable
Domestic cord sets
Fuses and wiring devices
Wire, general-purpose

Components and Materials

Power rectifiers: selenium, silicon
Parametric amplifiers
Transistors
Diodes: tunnel, zener, parametric
Semiconductor materials: selenium, germanium, silicon
Capacitors: wet, dry, ceramic
Ferrites
Tubes: power, transmitting, traveling-wave, rectifier, receiving, thyratron
Picture tubes
Relays and switches: telephone, industrial
Magnetic counters
Resistors
Varistors
Fluorescent starters
Transformers
Quartz crystals
Crystal filters
Printed circuits
Hermetic seals
Magnetic cores

Application of Pulse-Code Modulation to an Integrated Telephone Network

Experimental Pulse-Code-Modulation Transmission for Local-Area Telephony

United Kingdom–Faroes–Iceland (Scotice) Submarine-Cable Telephone System

Shipboard-Adjustable Submerged Equaliser

Variable-Reactance Frequency Multipliers

Contribution to Studies of Overflow Traffic

Interregister Multifrequency-Code Signalling for Telephone Switching in Europe

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EDITOR, H. P. Westman **Subscription: $2.00 per year; 50¢ per copy**

This Issue in Brief

Probable Evolution of Telephony—Despite rather large short-term variations, the long-term development in telephony shows strong trends that can be described mathematically for planning purposes.

If the number of telephone lines per 100 population is called the line density and its development over the years is plotted individually for many countries, it is found that each country follows the same general law but may lead or lag other countries in time as to its present stage of development. The composite curves of many countries may be described as an exponential function or as a hyperbolic tangent. It shows an annual increase of almost 5 percent. Saturation is not yet evident.

The number of subscriber sets per line varies slowly and should reach a peak when business use, with large numbers of sets per line connected through private branch exchanges, reaches saturation. As private homes then account for further increases, the sets per line will be reduced. This may again be reversed as homes are equipped with more than one set.

The annual number of conversations per capita differs greatly among countries, even with the same line densities. It appears to increase with line density and a rough estimate can be made of it.

1000B Pentaconta Crossbar Switching System—The original Pentaconta design grouped subscriber lines in 500 units served by 2 markers. The improved *1000B* type uses 1000-subscriber groups served by 2 markers that can handle 6000 calls per hour.

Two crossbar switching stages are employed. Of the 52 outlets in each primary-stage selector, 40 are used for direct junctions to the secondary stage and 12 are reserved for alternative routing of calls within the primary units. This means of providing mutual aid or reciprocal overflow is called entraide and avoids internal blocking.

For maintenance, any subscriber line can be called from the test desk by its directory number. All switching operations are timed and if the allotted interval is exceeded a fault record of the identity and position of all engaged switching elements is automatically punched on a card. Every subscriber line is checked for insulation and abnormal voltages immediately after the dialled pulses are recorded in the register. Registers, markers, and junctions are equipped for monitoring traffic loads. The usual alarms are provided for blown fuses and abnormal currents.

The *1000B* improvements in switching, equipment, and wrapped wiring over the original *500* system are significant with regard to reliability, speed, flexibility, and interconnection with other systems including national, continental, and intercontinental networks. There are now 850 000 lines of Pentaconta in service or in course of completion.

Pentaconta Line Concentrators—A concentrator is a telephone switching means to serve a relatively large group of subscribers located near each other over a small number of rather long common trunks connected to the central office. Although concentrators are not new, the Pentaconta system with its simple circuits and robust equipment not requiring constant maintenance, has generated much interest in this method of extending the capacity of existing cable. It is particularly adaptable to residential subscribers whose average utilization is of the order of 10 minutes per day.

Three designs of Pentaconta concentrators are described. Two serve 52 subscribers over 12 and 8 common 2-wire trunks to the exchange and the other serves 24 subscribers over 6 similar trunks. The remote unit is near the subscribers and simply extends the engaged subscribers' set to the exchange over the trunk pair without effecting its electrical properties. All classes of service may thus be provided. The number of common trunks are calculated so as not to loose as many calls as are lost in the terminal selection stage at the exchange. Only 800 milliseconds are lost in obtaining dial tone and

ringing over a direct connection to the central office.

As subscribers' lines represent about 60 percent of the cost of a local telephone network, concentrators offer a significant economic possibility. For a distance of 1.55 miles (2.5 kilometers) between the exchange and a group of 56 subscribers, the cost of 2 overhead 28-pair distribution cables for normal direct connections to the exchange and of one similar 14-pair cable connected to a 52-line 8-trunk concentrator, with 4 subscribers directly connected, will be equivalent.

New networks are usually designed for future expansion and have spare pairs. Concentrators would then be uneconomical for they would simply increase the number of spares. However, they may be useful in new plant serving areas distant from the exchange and having low population density. Their outstanding value is in providing for expansion of overloaded systems without laying new cable. A newer design will permit combining the exchange concentrator equipment with existing terminal selection apparatus to ease its accommodation in already crowded rooms.

Telegraph Transit Exchange Using Crossbar Pentaconta Design—The telegraph transit exchange for the Orly airport near Paris employs semiautomatic continuous-type operation using crossbar selectors and relays designed for the Pentaconta telephone switching system. Daily, it handles an average of 19 000 messages having either single or multiple addresses.

Messages of standard format are reproduced at the incoming position on a typing reperforator, the tape from which is directly threaded to a transmitter distributor. The operator notes the routing indication and selects the required outgoing junction via a crossbar multiselector. When the connection is made and the receiver motor at the destination office is running, the message is sent from the transmitter. If the outgoing channel is to a radio station, the message is first stored in a reperforator at the radio teleprinter position to permit any delay required in completing the radio circuit.

If the incoming crossbar selector is busy, the call goes to an overflow selector. If the outgoing channel is busy, the message is accepted at an overflow position which can store one extra message. If this is occupied, the message goes to the multiple-address position. Correction and transmission of messages received with faulty addresses is done at a supervisory position.

Assistant Type Telephone Set—A new series of subscribers sets called the Assistant type have been developed. In compact form with miniature components mounted on printed-circuit boards suitable for dip soldering, they are available in four different circuits to meet the needs of various administrations and in both desk and wall models.

Several types of transmitters and receivers may be assembled in the handset to produce desired operating characteristics. The handset is short to give a high signal-to-noise ratio based on presented statistics on human ear-to-mouth measurements. Amplitude-versus-frequency characteristics are given for several transmitters, receivers, and combinations.

Intercommunication Telephone Sets—There are several types of intercommunication systems, some are connected to the public telephone network while others are private. As the terminal equipment must provide for switching not normally required of a subscribers' set, the intercommunication sets are correspondingly larger. This emphasizes the need for small components and for combining functions to reduce size.

One system permits up to 5 extensions to be connected to one line to the public network. A green lamp operates on the extension connected to the exchange and red lamps on all other extensions warn that the party-line is in use. The exchange line may be held while another extension is called and the call then transferred to the second extension. Up to 10 extensions may also be served over 2 local exchange lines.

The secretarial system bridges the executive's and secretary's phones so all calls are handled by the secretary and transferred to the executive when this is desirable. The secretary may also complete outgoing calls and transfer them to the executive when the called party has been reached.

While private automatic exchanges normally provide intercommunication among all extensions, special sets permit additional services such as loudspeaker handsfree operation, direct selection, staff location, transfer, camp-on busy, priority call, and group hunting. Transistor amplifiers provide for loudspeaker operation or a handset can be used for privacy or under noisy conditions.

Another system provides for a master station and up to 10 extensions, any one of which can be called by or can call the master station. In an alternative arrangement, any of the 11 stations can call any other station and may also set up conference calls.

Matrix Multiplication in Search for Alternate Routes—A high degree of adaptability is a prime requirement of a military communication system. One of the most important characteristics of such a system is the multialternate routing scheme by means of which outage of a link or of a communication center due to war damage, or virtual outage due to a local traffic congestion, will not seriously impair the functioning of the system. The pattern of the multialternate routing must be adjusted as quickly as possible whenever the damage or the congestion becomes more widespread.

The routing pattern must be stable; the possibility of the phenomenon known as the "ring-around-the-rosie," which is a snowballing loop connection that ties up the system, has to be eliminated.

A procedure for determination of an optimum alternate routing pattern by a computer has been developed. The procedure incorporates a step by which routes that can lead to ring-around-the-rosie are eliminated as allowed routes.

The computer program can be used in the central control facility of a communication net. This facility would receive status information on the whole network, compute the optimum routing for the whole network, and send the solution to all the nodes, where the new route translators are simultaneously implemented for the whole network.

The procedure has been programed for the IBM 704 computer.

Theoretical and Practical Aspects of Telephone Traffic—The mathematical description of telephone traffic employs three distinct classes of distribution functions; the holding times of connections, the arrival times of calls, and the number of calls simultaneously in progress. An equation is derived relating these three classes.

The agreement between actual and theoretical traffic is considered. The distribution functions describing the arrival of calls agree quite well, but the simple negative exponential law for holding time does not and cannot correspond with actual conditions.

Methods of observing and evaluating actual traffic data are examined to make sure that they yield quantities which, when used in the calculations, give reliable results.

The concepts of busy hour and of quality of service are discussed as are the arithmetical rules to be applied when traffic flows merge or divide.

The strangling effect of inadequate switching is considered. Some methods are presented for estimating the traffic offered to a group of switches from observations of the traffic carried by the group.

Slow-Wave Structures for Millimetre-Wavelength Backward-Wave Oscillators—Progress in millimetre-wavelength beam tubes is closely linked with the design of the slow-wave struc-

ture that propagates the electromagnetic wave. These structures must have phase and group velocities that will minimize the effects of losses and must also avoid random reflections without incurring highly critical mechanical tolerances. Both attenuation and the interaction impedance are inversely proportional to the group velocity, which is therefore a critical design factor. It may be chosen so as to minimize the starting current or to maximize the efficiency.

The use of multiple internally combined slow-wave structures corresponds to operating several tubes in parallel. Parallel operation could, however, impose insuperable difficulties. Theoretically, where M is the number of slow-wave structures in a multiple array, the gain in power output should be about $2 \cdot 5M$. A 4-ladder array should produce an order of magnitude higher power than a single-ladder design.

Ladder-shaped structures are readily adapted to multiple assemblies and, importantly, dissipate heat readily through short elements that are connected to relatively massive heat sinks. The simple symmetrical ladder does not propagate a wave as its rungs, acting as resonators, are equally and oppositely coupled by the electric and magnetic fields. Several methods are shown of perturbing one field and so providing a pass band. A design having a wide pass band and means for matching the slow-wave multiple ladder structure to a waveguide is discussed.

Lewin Receives Papers Awards

Leonard Lewin of Standard Telecommunication Laboratories at Harlow, Essex, England, has received two awards for his paper "On the Resolution of a Class of Waveguide Discontinuity Problems by the Use of Singular Integral Equations" that appeared in the IRE Transactions of the Professional Group on Microwave Theory and Techniques for July 1961.

The Professional Group awarded him its annual Microwave Prize of $100 and the parent organization, the Institute of Radio Engineers, gave him its annual W. R. G. Baker prize of $1000 for the best paper published in all of the transactions of its Professional Groups.

Mr. Lewin joined Standard Telecommunication Laboratories in 1946 and is now assistant manager of the Telecommunication Transmission Systems Division in charge of developing microwave systems. His book on the *Advanced Theory of Waveguides* was published in 1951.

Recent Achievements

Data-Transmission Error-Rate Analyzer—An error-rate analyzer employing transistors has been produced for development, factory testing, alignment, and maintenance of data-transmission systems. It is shown in Figure 1. Measurements may be made on actual traffic. Each analyzer contains all the equipment needed for the measurements, which are made on a synchronous basis.

Element error rates and various kinds of time distortion can be measured. The probability of time distortion exceeding a selected degree may be evaluated. Mutilation disturbances may be studied.

For point-to-point measurements, 24 crystal-controlled modulation rates are available between 45 and 3000 bauds and, for closed-loop measurements, the modulation rate is continuously adjustable from 25 to 3000 bauds.

A variety of modem characteristics can be tested with a white-noise generator. The error rate may be obtained for different signal-to-noise ratios. Phase and attenuation distortion of a modem and line may be expressed as impairment in decibels. The relative merits of different midchannel frequencies may be determined. With the use of an error-distribution analyzer, detailed studies can be made of error occurrence.

Standard Radio & Telefon
Sweden

Laser Displayed to French Society of Physics—Four laser equipments were displayed and demonstrated at the Paris exhibition of the French Society of Physics in October 1962.

Of the two gas lasers shown, one was a simplified design primarily for instruction use and the other was a high-stability unit for research. In the latter, the gas tube and mirrors are readily adapted to the study of a wide range of wavelengths. Two low-power pulsed ruby lasers were also exhibited. See Figure 2.

More recently, at the Third Symposium of Quantum Electronics, there was presented a high-stability gas laser emitting continuously a strong red-orange light beam at a wavelength of 0.63 micron.

Laboratoire Central de Télécommunications
France

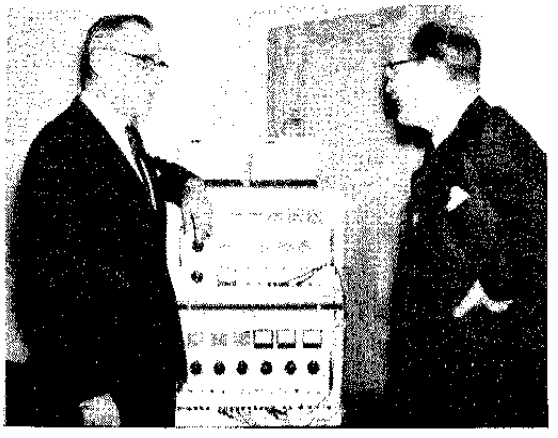


Figure 1—Dr. H. Sterky, General Director of the Swedish Board of Telecommunications, left, and Managing Director F. Hammar of Standard Radio & Telefon discuss the error-distribution and error-rate analyzers shown in the picture.

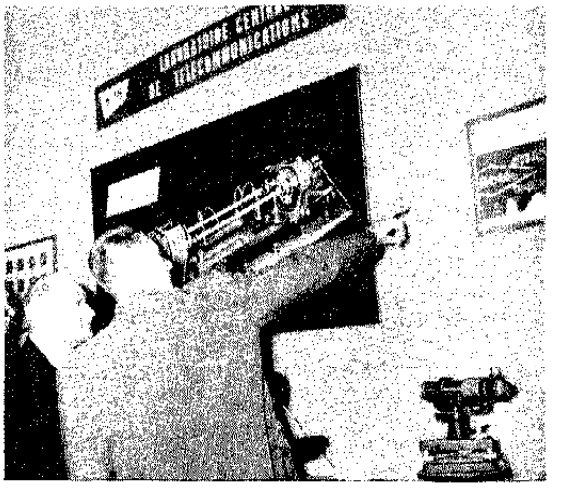


Figure 2—Mr. G. Palewsky, Minister in charge of Scientific and Technical Research, listens to a description of how a ruby laser beam may machine small holes in hard razor-blade steel given by Mr. Cagnard, head of the physics department of Laboratoire Central de Télécommunications.

Record-Communication Automatic Switching System—Fast switching, great flexibility, and inherent security are achieved in a record communication system by using a digital computer having a stored program to control the receiving, analyzing, processing, storing, re-routing, and transmitting of messages and data. It is capable of handling various transmission speeds, codes, and message formats to and from hundreds of addresses.

Installed in the United States embassy in Paris, the initial ITT 7300 ADX automatic exchange is the center for the European communications of the State Department. An on-line real-time fully automatic system, information may be received simultaneously from high-speed data transmitters and standard 60-word-per-minute teleprinters. Lower-priority traffic will be stored automatically to make way for urgent messages.

ITT Information Sytems Division
United States

New Zealand-Fiji Cable Inaugurated—The light-weight deep-sea telephone cable linking Auckland, New Zealand, with Suva, Fiji Islands, over a stretch of 1150 nautical miles (2130 kilometers) was placed in service on 3 December 1962. It is the second link in Compac, the trans-Pacific submarine telephone cable system.

The new cable provides for 80 simultaneous 2-way telephone conversations. It is equipped with 37 submerged repeaters and 6 submerged equalisers.

Standard Telephones and Cables
United Kingdom

Gallium-Arsenide Laser—New techniques have produced gallium-arsenide of both very-high purity and crystal perfection, and point-contact and diffused-junction diodes made of it have emitted noncoherent light. This suggested the development of a laser, which was successfully operated in January 1963.

Laser action occurs above a threshold current density of about 2000 amperes per square centimeter, which is considerably lower than previously reported values. The infra-red emission is polarized and spatially coherent. Above the threshold, the bandwidth is less than 5 angstrom units, while below the threshold it is about 150 angstrom units.

In Figure 4, the emission pattern observed on an image converter tube shows interference fringing caused by reflections from the base plate and, therefore, parallel to the junction plane. At right angles to this plane is the bright band of laser emission with line narrowing. It is polarized along the line of the junction.



Figure 3—Installation work at the Suva terminal of the New Zealand-Fiji submarine telephone cable installation.

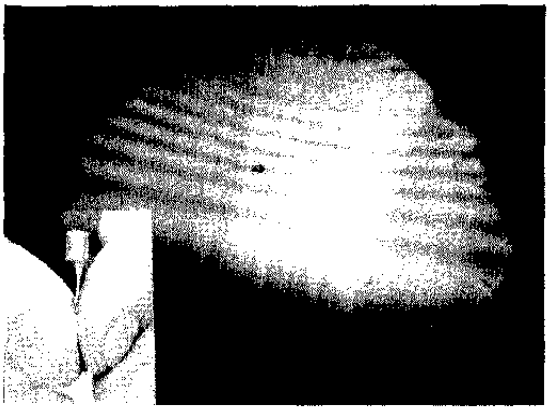


Figure 4—Gallium-arsenide laser capsule and emission pattern.

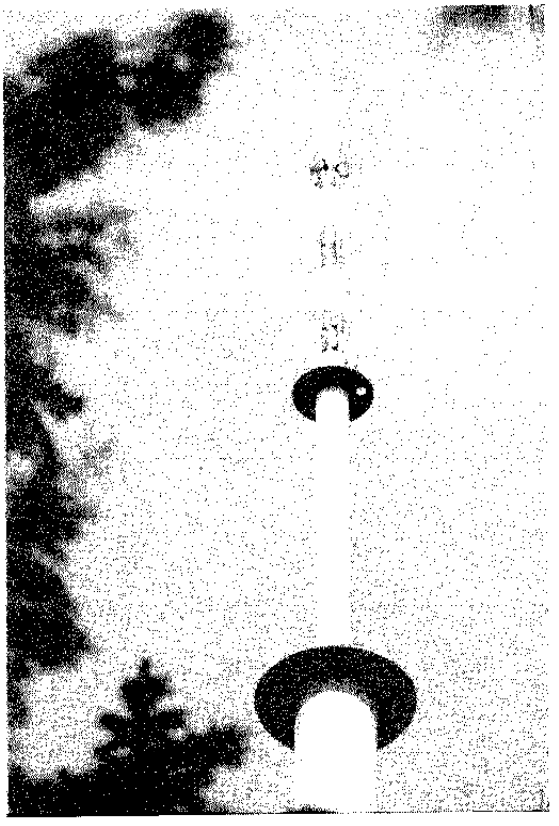


Figure 5—Antennas for television transmission on bands III and V.

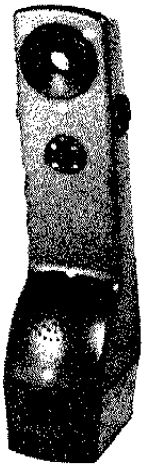


Figure 6—Lightweight field telephone set.

An interference pattern, probably resulting from light from different points of the junction, indicates spatial coherence.

Standard Telecommunication Laboratories
United Kingdom

Radio Guidance for Blind Landing—Under completely fogbound conditions that grounded all other air traffic, a Ministry of Aviation aircraft recently made the first fully automatic blind landing at London airport. The perfect landing was controlled by information supplied to the automatic pilot from a radio altimeter and from the STAN.7 localizer and STAN.8 glide-slope equipments that are now standard at Britain's main airports.

Standard Telephones and Cables
United Kingdom

Transmitting Antenna for Television Bands III and V—A 172-meter (563-foot) circular reinforced-concrete tower on the Donnersberg in Palatina supports an antenna array for television band V. It consists of a 20-meter (65.6-foot) steel hexagonal mast, on each face of which antenna arrays are arranged in 9 tiers, as may be seen in Figure 5. The center section receives more power than the others and this produces a smooth vertical radiation pattern for angles from 0 to −10 degrees. The vertical gain is approximately 16 decibels over a half-wave dipole.

Atop the band-V antenna, which is used for the second program and will be used later for a third program, is a 13-meter (42.6-foot) cylindrical slot antenna for band III. Four stacked cylinders each having four vertical slots provide a vertical gain of about 11 decibels.

The plain cylindrical surfaces of these antennas minimize the accumulation of ice and offer low resistance to wind, which can be very strong at their elevation of about 900 meters (3000 feet) above sea level.

Standard Elektrik Lorenz
Germany

Lightweight Field Telephone—Designed primarily for military field service and suitable also for civilian field activities, the watertight telephone set shown in Figure 6 weighs only 750 grams (26.5 ounces).

It is a complete telephone system that may be connected to the public telephone network or to a radio system. In addition to the telephone set, there is a switching unit, a call converter for connection to a local or to a common-battery telephone system, and a remote control unit for use with a radio transmitter and receiver. It was developed in cooperation with Standard Elektrik Lorenz.

Bell Telephone Manufacturing Company
Belgium

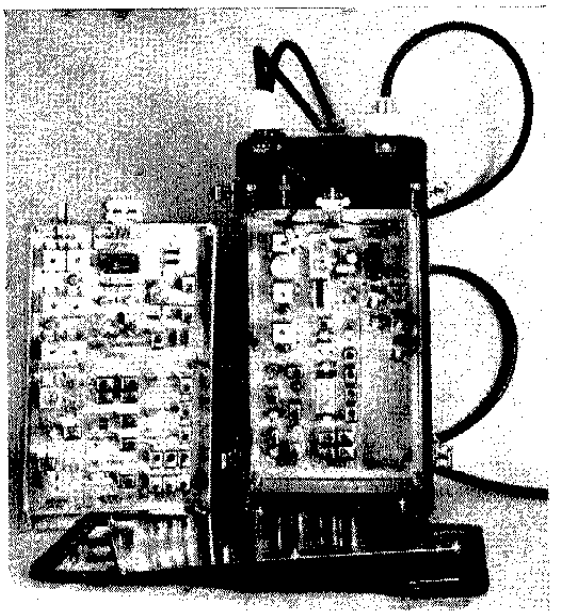


Figure 7—The receiver (left) and transmitter for the radio terminal to connect a subscriber to the nearest telephone exchange.

Radio Link for Rural Telephony—For use in sparsely populated regions, a radio link to connect a subscriber over distances up to 30 kilometers (18.6 miles) to a telephone central office has been developed. The 500-milliwatt transmitter uses frequency modulation in the 148–174-megahertz band. Signalling is done by keying the carrier.

The use of transistors permits operation from dry cells with reasonable replacement demands. The power required in the stand-by condition has been minimized. A subscriber terminal is shown in Figure 7.

Standard Eléctrica
Spain

Pentaconta Exchanges in Chile and Philippines—Three Pentaconta telephone exchanges were cut over in October 1962 at Antofagasta, La Cisterna, and San Bernardo in Chile. They have 3800, 3000, and 1500 lines, respectively. Mr. Jaime Silva, Home Undersecretary of State of the Chile government, officiated in placing the exchanges in service.

A 1500-line Pentaconta exchange was officially opened in Zamboanga on Mindanao island by the wife of President Macapagal of the Republic of the Philippines.

These new exchanges bring to 17 for Chile and to 4 for the Philippines the installations of Pentaconta in these countries since 1955.

Compagnie Générale de Constructions Téléphoniques
France

Multiple-Angle Diversity Radio Transmission—An experimental radio system uses a 28-foot (8.5-meter) paraboloidal reflector to radiate 7 separate beams, each from its own transmitter. Two corresponding receiving antennas, each with 7 collector horns, parametric amplifiers, and recorders compile statistics on the 14 received signals for analysis by data-processing equipment. This experiment, under contract

with the United States Air Force Rome Air Development Center, permits a single paraboloidal antenna to transmit or receive over several diverse paths.

ITT Federal Laboratories
United States of America

Telephone Traffic Measuring Set—Measurement of the occupation of junction groups, registers, selection stages, and other facilities in telephone toll offices and rural main exchanges may be made automatically and recorded by a teleprinter with a new test set.

The program circuit controlling the sequence of operations and the connecting of the measuring circuits to the equipment under observation use relays whereas the measuring, summing, translation, and printing are done electronically. Finders are used as storage elements and also as the time base for the teleprinter.

In Figure 8, the access selectors within the exchange are marked by a scanning circuit that is at the top of the bay. The electronic scanner is mounted below and explores the terminals under test at the appropriate times and determines the number of occupancies. The internal timing circuit is controlled by the exchange master clock.

An electronic translator and converter transform the stored information into teleprinter code for recording. Operation is from the 48-volt exchange battery, augmented by built-in power supplies.

Standard Telephone & Radio
Switzerland

Error Correction of Data Transmission at 170 Characters per Second—Tests over the public switched telephone network of Sweden confirm the high accuracy of an error-correction system used in transmitting data between tape, cards, or two computers. As no error has passed uncorrected in the tests, it is probable that only a component failure will disable the equipment.

Error detection is performed on a character-by-character basis and employs an analog-type disturbance detector in combination with a simple code control. Errors are detected at the receiver and correction is achieved by retransmission. The modems use frequency-shift modulation and are suitable for speeds of 600 to 1200 bauds. The return channel over which repetition is requested operates at only 75 bauds.

An installation with tape input and output operating at 133 characters per second has been delivered to the Swedish Board of Telecommunications. A receive terminal is shown in Figure 9.

Standard Radio & Telefon
Sweden

Three-Dimensional Electrocardiograph—By applying three pairs of electrodes to the human body to sample the electric activity of the heart in three orthogonal planes, a single pattern can be produced on a cathode-ray tube in which variation in brightness is the third "dimension."

The single presentation shows the instantaneous relations among the three directional activities, which would not be clearly evident if the three views were displayed separately. Now under evaluation in a research hospital, it is anticipated that this will become another diagnostic tool in the war against heart disease.

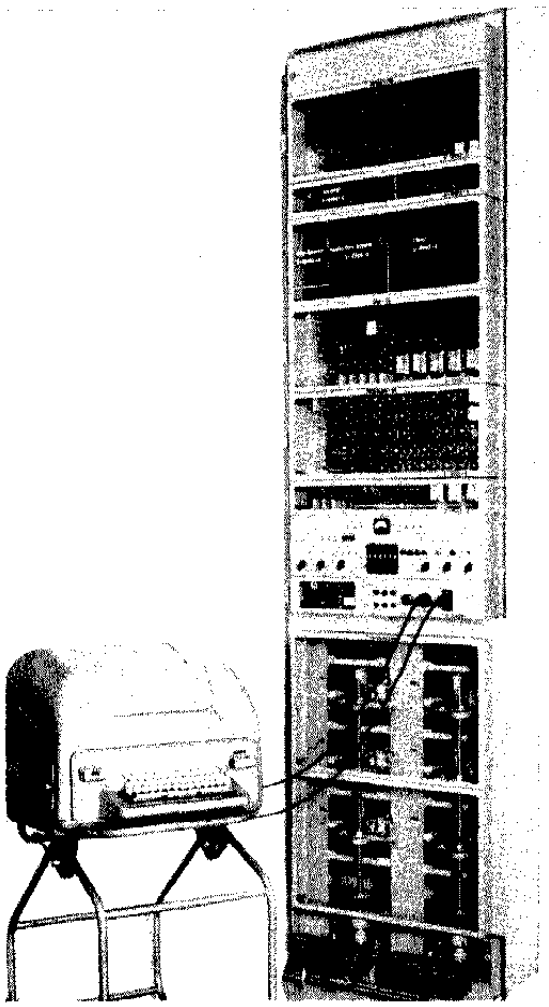
ITT Federal Laboratories
United States of America

Data Transmission Demonstration—Throughout the last quarter of 1962, demonstrations were made of a 1000-baud data transmission system that connected by telephone line a data processing center and a large number of users.

Four agents equipped with special keysets were connected, via a station capable of accommodating 100 agents, to a message control cubicle in which a simulator substituted for a central processor. Each agent sent queries to the simulator and received information in return within 1 or 2 seconds.

Another arrangement connected 12 agents with the simulator over 50-baud telegraph circuits

Figure 8—Telephone traffic measuring equipment.



through a parallel-to-serial converter and a subscriber concentrator.

To ensure accurate transmission of all messages, automatic repetitions were made each time a parity fault was detected.

Le Matériel Téléphonique
France

7E Telex Exchanges in Italy—During 1962, automatic telegraph exchanges were installed in Cagliari, Catania, Catanzaro, Florence, Genoa, Naples, Palermo, and Rome, for the Italian Ministry of Post and Telecommunication. The Rome exchange, in addition to handling national calls, also controls international and intercontinental traffic. Bell Telephone Manufacturing Company of Belgium collaborated in this work.

Fabbrica Apparecchiature per Comunicazioni Elettriche Standard
Italy

Infrared Tracking—A high-resolution infrared tracking system has been placed in operation at the Cape Canaveral missile range. It is particularly useful during the launching period of a missile when ground clutter masks the significant radar signals.

Mounted on the antenna pedestal of an AN/FPS-16 instrumentation radar, as shown in Figure 10, the angle-error signals from the infrared tracker keep the pedestal on target until the ground clutter disappears and reliable radar tracking can be obtained. Switching between infrared and radar may be either manual or automatic.

The bright Florida skies place great importance on a high degree of background suppression, while close proximity to large missiles at

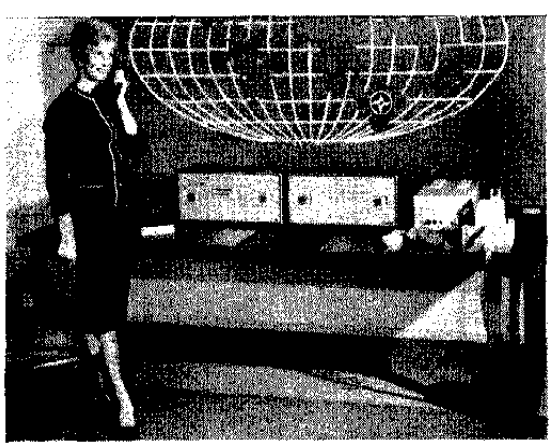


Figure 9—Receive terminal of data communication system with error correction equipment.

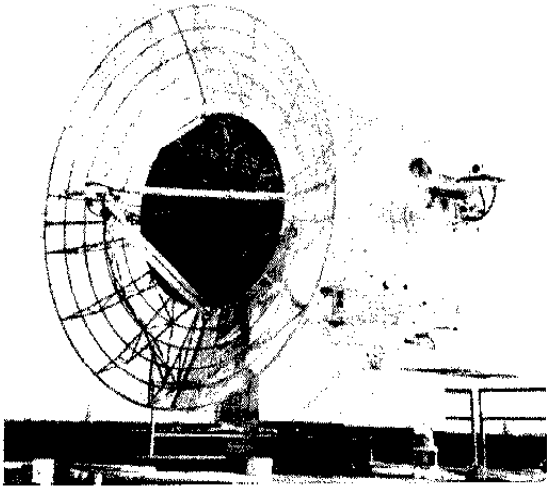


Figure 10—The infrared tracker is in the box mounted at the upper right of the AN/FPS-16 radar.

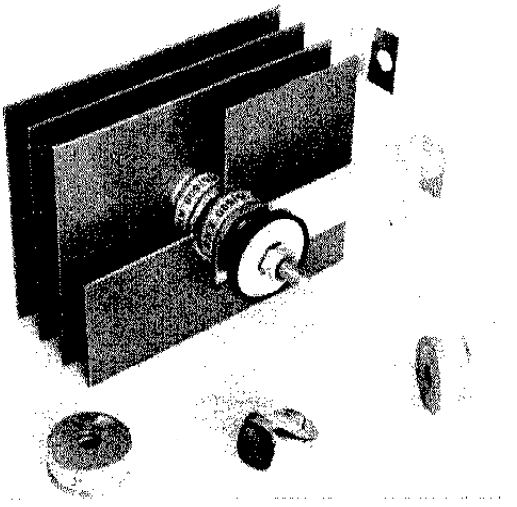


Figure 11—Silicon ring-type rectifier cells and arrangement for mounting in a stack of fins similar to selenium-rectifier practice.

launch and the necessity of tracking small missiles at great range necessitate a dynamic range of sensitivity of 100 000 to 1. A single detector cell provides both radiometric and tracking information. It is protected from damage from direct sunlight by an automatically operated shutter.

ITT Federal Laboratories
United States of America

Silicon Rectifier Stacks—Compact, easily mounted, high-capacity silicon rectifiers are now available in the stack-type of assembly well established over the years for selenium rectifiers.

A diffusion-type silicon rectifier is mounted in a ceramic ring sandwiched between large contact plates that readily transfer heat to suitably dimensioned fins. Any required number of these Silring rectifiers may be assembled with cooling fins and connecting strips on an insulated bolt, which also provides for mounting the stack. The design is evident from Figure 11.

Ratings between 2.5 and 280 amperes are available and may be doubled for forced-air cooling. At an ambient temperature of 50 degrees centigrade, a 2.5-ampere unit without fins will handle 5 amperes with 50-by-50-millimeter (2-by-2-inch) fins, 10 amperes with 75-by-100-millimeter (3-by-4-inch) fins, and 15 amperes with 125-by-125-millimeter (5-by-5-inch) fins.

Standard Elektrik Lorenz
Germany

Ground Beacon for Air Navigation Distance Measuring—A ground beacon for civilian use as part of the Vordac air-navigation system has been developed basically on the designs by ITT Federal Laboratories and Standard Telephones and Cables (London) of military tacan beacons. It meets the specifications of the International Civil Aviation Organization.

The *FSD.1* beacon shown in Figure 12 includes duplicate transponders and monitors, antenna system, remote control unit for both beacon and monitor operable over 36 kilometers (22.4 miles), and other accessory equipment.

If the monitor indicates a fault in transmission, the standby monitor will be placed in operation and only if it confirms the fault will the standby beacon be put in service.

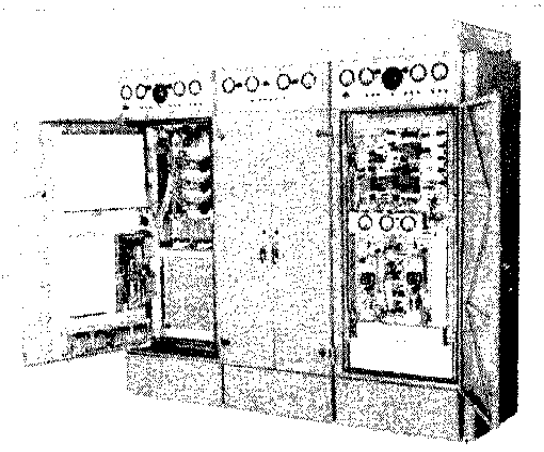


Figure 12—Ground beacon for distance-measuring element of Vordac air-navigation system.

The antenna system produces an omnidirectional horizontal pattern and in the vertical plane a gain of at least 9 decibels referred to a half-wave dipole.

Emphasis has been placed on reliability and accuracy. Direct access from the front is provided to all vacuum tubes and tuning circuits. Long-life klystrons also have high linearity for stability. All components are of established reliability as is essential for equipment for unattended operation.

Fabbrica Apparecchiature per Comunicazioni Elettriche Standard
Italy

Multicolor Processor and Projector for Computer Readout—This processor and projector is capable of working with any computer that provides for multicolor or black-and-white readout. The three primary colors, their complementary colors, and white on black may be produced by the color-addition principle.

The demonstration model projects the processed image on a 12-inch (305-millimeter) screen but this can be increased by suitable optics. The exposure and processing time, which precedes display, is only 8 seconds. The accuracy to which an image can be positioned depends only on the precision of the symbol generator and is very high. Among the uses of such a device are military command, air-traffic control, and the presentation of complex data.

ITT Federal Laboratories
United States of America

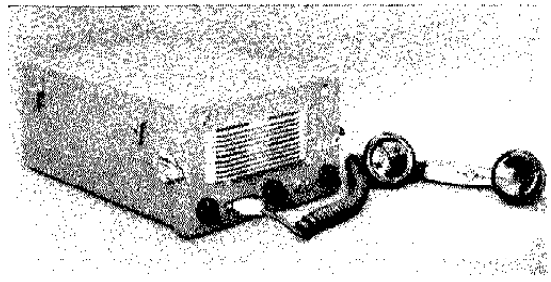


Figure 13—Very-high-frequency transceivers used by the French security services.

Mobile Transceiver—Since 1959, about 3000 transceivers of the type shown in Figure 13 have been manufactured by Le Matériel Téléphonique to equip most of the vehicles of the French police, fire, and ambulance services and of important administration officials.

Making extensive use of transistors, the equipment operates in the 70-to-90-megahertz band. The transmitter delivers about 20 watts to the antenna.

Le Matériel Téléphonique
France

Small Radio Transceivers—In the survival unit of astronaut Walter M. Shirra, Jr. in his MA-8 flight was a 15-ounce (425-gram) radio transceiver that fortunately he did not need. Had the astronaut landed outside the planned area, as happened in an earlier test, he could have been in voice communication over a distance of about 20 miles (37 kilometers) with aircraft flying at 5000 to 10 000 feet (1524 to 3048 meters). The crystal-controlled transceiver is operated from two mercury cells. Over a hundred homing receivers are installed in the search aircraft.

For intercommunication within a team such as in a space craft, for missile refueling, fire fighting, and warehouse operating, where high noise levels or other hazardous conditions exist, a 9-ounce (255-gram) compact radio transmitter and receiver provide duplex hands-free operation through the use of voice switching of the transmitter. The equipments may be operated with each other and a base station may also be employed for supervisory control.

ITT Kellogg Communications Systems
United States of America

Air-Raid-Warning Receiver—The air-raid-warning receiver shown in Figure 14 covers the band from 35.85 to 36.55 megahertz on which early warnings and reports will be broadcast for use by emergency services and places such as theaters in which there are large assemblies of people.

The double-heterodyne transistor receiver employs a 12-circuit filter in the second intermediate-frequency section to provide high selectivity. A switch permits selection among 15

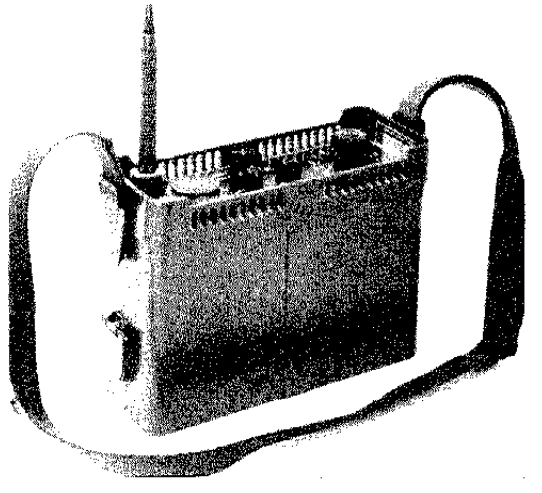


Figure 14—Portable air-raid-warning receiver operating in the very-high-frequency band.



Figure 15—Magnetically shielded loudspeakers that avoid picture distortion in portable television receivers where the loudspeaker is mounted near the picture tube.

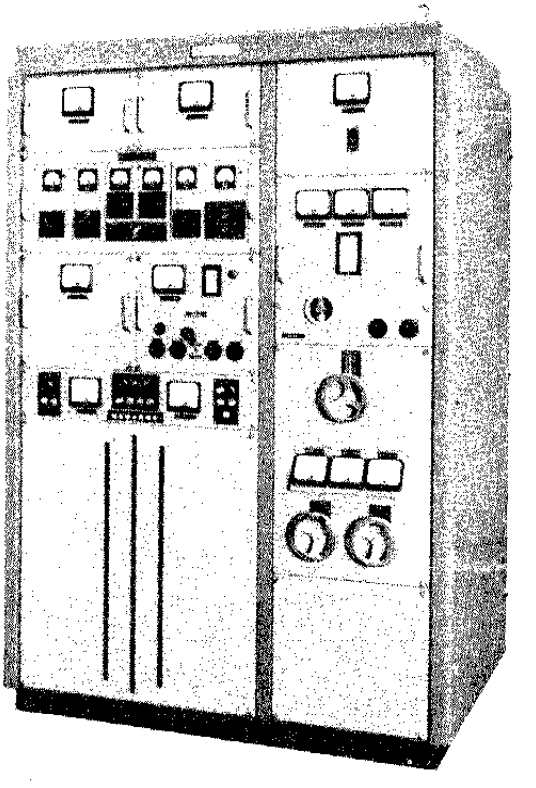


Figure 16—High-frequency 4-kilowatt marine radio transmitter.

channels and crystal control of the oscillator ensures adequate frequency stability.

A replaceable battery operates the set for 50 hours and it may be connected to the alternating-current mains for continuous operation and battery charging. Alternatively, it will work from any 6-to-24-volt vehicle battery.

The sturdy spray-waterproof case is 187 by 245 by 95 millimeters (7.4 by 8.6 by 3.7 inches) and has a belt for over-the-shoulder or arm carrying.

Standard Elektrik Lorenz
Germany

Magnetically Shielded Loudspeaker—Portable television receivers often place the loudspeaker very close to the picture tube and nonlinearity of scanning may result from the magnetic field of the loudspeaker.

A new series of oval-diaphragm loudspeakers, shown in Figure 15, having specially shielded 7000- and 8000-gauss magnets have been produced for this service. The basket dimensions of the 3 sizes are 130 by 180 millimeters (5.1 by 7.1 inches), 150 by 210 millimeters (5.9 by 8.3 inches), and 130 by 260 millimeters (5.1 by 10.2 inches).

Standard Elektrik Lorenz
Germany

Multifrequency Code Signalling Between Pentaconta Exchanges—The first installation for multifrequency code signalling between Pentaconta liaison telephone exchanges in France became operative between the Albi and Mazamet exchanges in the Toulouse area. Further installations are being made in this and in the Lyons areas.

Each signalling code group consists of the simultaneous transmission of 2- out-of-5 frequencies between 700 and 1700 hertz. The called number is sent digit by digit at the request of the distant exchange, which also signals its receipt and termination of that transmission. The distant exchange also uses this code to signal the condition, free or busy, of the called line.

Compagnie Générale Constructions Téléphoniques
France

Single-Sideband 4-Kilowatt Radio Transmitter—The new ocean liners: *Marconi, Galilei, Michelangelo,* and *Raffaello* will be fitted with recently developed 4-kilowatt radio transmitters for operation between 2.5 and 22 megahertz.

The F.912 radio transmitter is shown in Figure 16 and is designed primarily for high-quality radiotelephony. In addition to double-sideband operation, either single-sideband or independent-sideband transmission may be used and each sideband may provide for 2 commercial-quality telephone channels and 3 teleprinter channels. Amplitude and frequency-shift telegraph keying is also available.

The units include exciters, modulation equipment, and power supplies. They feature rapid frequency change and simple tuning. The telephone terminal equipment and radio receivers will also be supplied.

Fabbrica Apparecchiature per
Comunicazioni Elettriche Standard
Italy

Probable Evolution of Telephony *

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1. Introduction

In planning the future development of a telephone network, it is necessary to estimate how the present number of telephone lines, sets, or conversations will expand during the next 5 or 10 years. Although many personal estimates are founded on experience, systematic and generalized studies are rather scarce. Conclusions reached in this way, even though sketchy or incomplete, should provide useful guidance.

A consideration of statistical data on telephone development in a number of countries may permit the derivation of some laws or rules for predicting future numbers of telephone lines, sets, or conversations. As telephone developments are easily influenced by numerous unpredictable factors, all conclusions reached in this study must be regarded only as probable for the time being and far from being certain.

* This paper is a revision of one presented at the Third International Teletraffic Congress in Paris on 15 September 1961.

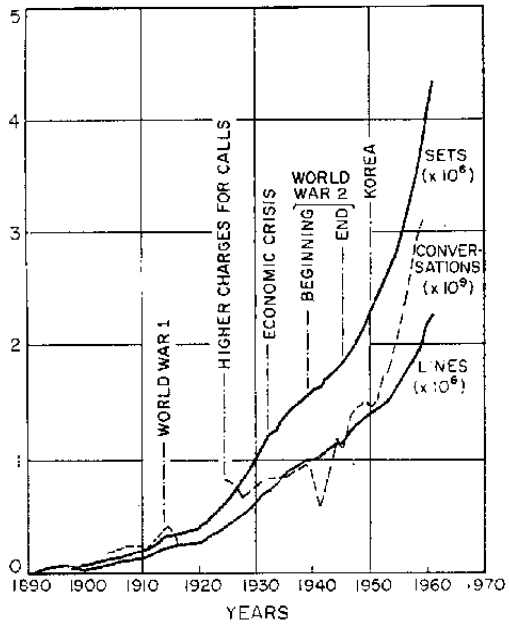


Figure 1—Telephone development in France.

2. Disturbances in Evolution

The use and number of telephones in any country increases regularly with time but with occasional irregularities due to economic crises, wars, or other events. Figure 1 shows such an evolution for France since 1890.

The influence of such disturbing elements on the normal trend of the growth curve will be briefly investigated with the help of some characteristic examples.

2.1 Economic Crisis of the 1930's

Not every country was upset to the same extent by the economic crisis of the 1930's. From the telephone viewpoint, it was most heavily felt in such countries as the United States of America, Canada, and Germany, while others suffered only mildly, if at all.

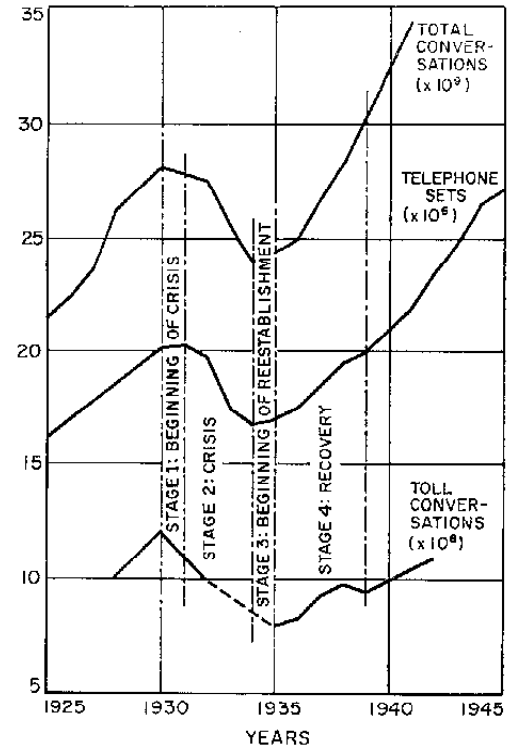


Figure 2—United States telephone sets and conversations during the economic depression of the 1930's.

Starting late in October 1929, this crisis developed in several phases, not all of which occurred in every country or at the same time in various countries.

An early effect was a slowdown in growth of or even a reduction in the total number of telephone conversations, which was most marked for toll calls. The number of subscriber sets remained constant or increased less rapidly. These changes stemmed from lessened commercial activity with its reduced need for long-distance business calls, while local business and social conversations were not affected to the same degree. As everybody hoped that the depression would be short lived, telephones were retained although there was little inclination to install new ones.

Figure 2 illustrates this history for the United States. Similar diagrams can be drawn for other countries among which Germany and Canada show corresponding sharp regressions, while others are less markedly affected.

The second stage showed a still sharper decline in communications, which now also included local traffic, as well as a decrease in the number of telephone sets as businesses were then forced to reduce expenses or even to discontinue operations entirely.

When the lowest point (1933 in the United States) was passed and business recovered slowly, the number of conversations increased again while the number of sets tended to stabilize as there was still some apprehension for the future.

At last after the general restoration of commerce, all the curves resumed a regular increase and, after some 5 to 8 years, the situation existing before the crisis was again reached. As the Second World War broke out shortly afterwards, it is difficult to prove that the re-establishment of telephone communication had a tendency to accelerate and thus to compensate for the losses during the crisis, but every indication points in that direction.

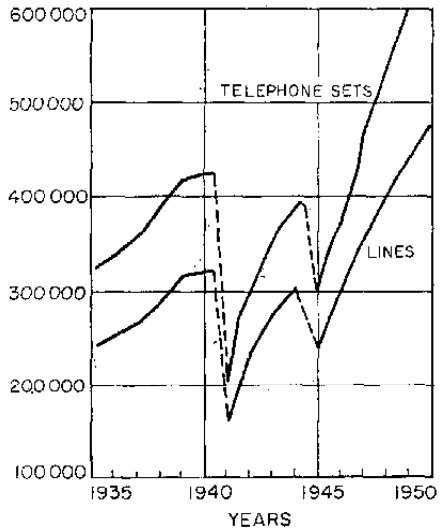


Figure 3—Effects of Second World War on telephony in Belgium.

2.2 Second World War

A considerable amount of information is missing for the period of the Second World War so that an over-all interpretation seems impossible. Nevertheless, there are no apparent signs that telephone development in countries that did not suffer from direct war activities and devastations was influenced by the hostilities.

Even in the countries of Western Europe in which the actual fighting occurred, the number of sets kept increasing although often at a slower pace because of shortages of raw materials. Sharp drops can be noted, however, where the actual fighting occurred with the concomitant destruction of equipment. After the battle front moved, there was a quick recovery that, however, was also incomplete as all equipment could not be repaired immediately. An example of this can be seen in Figure 3, which gives the number of lines and sets in Belgium from 1935 to 1950.

The number of conversations, especially for long distances, seems still more closely related to the events as can be seen in Figure 4, which shows their history in Belgium during the war and the immediate postwar period. Similar sharp drops in the conversations at the beginning and the end of the war can be seen for France in Figure 1, while the number of lines and sets grew steadily.

After the war, all countries experienced a very rapid improvement as repairs and extensions of equipment in a few years compensated for the losses. Normal expansion was then resumed.

2.3 Other Disturbances

Next to these general crises, the particular national tribulations of the different countries are also reflected in their telephone communications.

Some events extend their influence over a considerable area. Thus, the Korean war in 1950 caused a marked drop in conversations as can be seen in Figure 1.

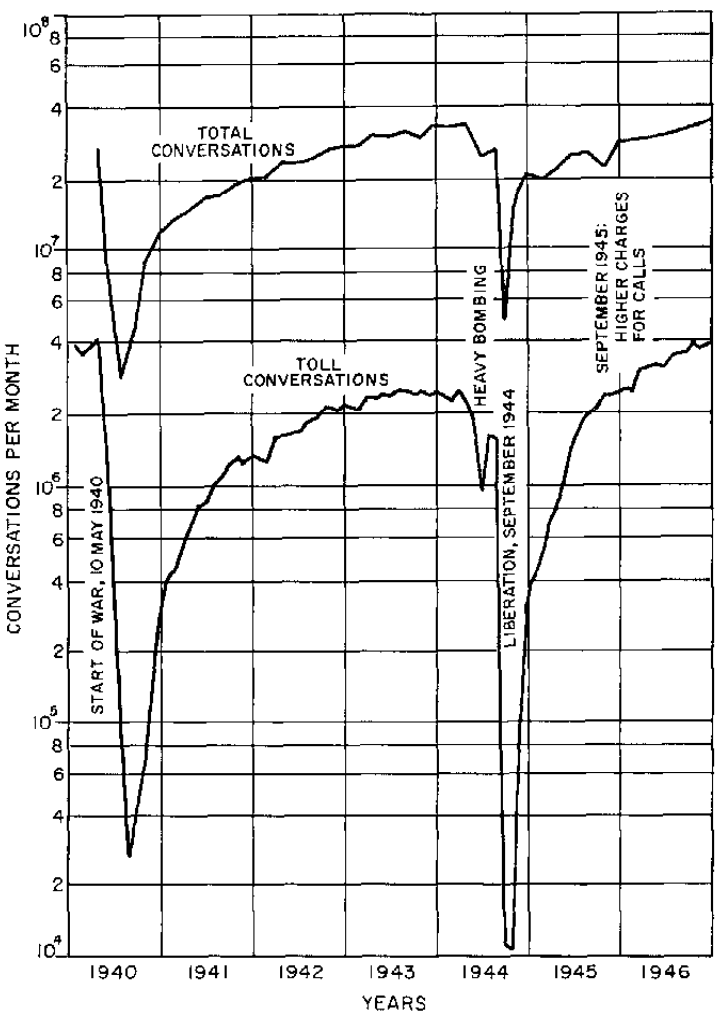


Figure 4—Conversations in Belgium in the war period.

Increased charges for calls reduces traffic, at least for a certain time. Figure 4 shows the result in Belgium of somewhat higher charges imposed in September 1945, and Figure 1 the effects in France in 1924 of imposing charges for conversations that had previously been partially free.

In general, social, economic, and political events have a tendency to influence toll traffic first; local traffic is affected somewhat later if the crisis continues. The number of operative telephone lines and sets decreases only after a long and violent economic crisis or through direct destruction.

In every case, each drop is compensated for after a few years and we can disregard those temporary crises in a general study.

3. Number of Telephone Lines

The term *telephone line* can be misleading when encountered in statistics. Often the term subscriber's line is used instead but this does not necessarily imply that every city line of a private branch exchange is taken into account. Another question is whether in party-line operation each party is counted as an individual line. As information on the number of lines is readily found in the yearly statistical survey of the International Telecommunications Union, its definition of line is adopted.

The number of telephone lines in a country can be related to its population. The line density is the number of lines L divided by the number of inhabitants N, or in percent

$$d = 100\ (L/N). \qquad (1)$$

This line density can be very different from one country to another as it depends on economic, technical, and social levels. It increases steadily when temporary crises, which are compensated for in the following years, are ignored.

As quite accurate forecasts can be made of population increases, if a general law for the growth in line density could be found, it would be possible to predict the probable number of lines in a country at a future date.

3.1 General Law of Line Density

It is remarkable how the line-density curves of some countries coincide with each other. The density in Switzerland from 1940 to 1961, for instance, is substantially the same as that in Sweden from 1931 to 1952. Thus, the Swiss line density seems to lag that of Sweden by some 9 years. Figure 5 shows this superposition.

Application of the same procedure to many countries produces a composite development curve from which it is evident that within a small deviation, each country follows the same general law in the evolution of its line density. Its present or future state of development can be expressed as a function of time.

$$d = f\ (T + c) \qquad (2)$$

where T is time in years and c is a constant indicating the time lead or lag with respect to other countries.

The trend of this general law of evolution can be seen when the density curves of several countries are superposed in such a way as to cover a minimal area. The result is shown in Figure 6, where a number of density curves from 1930 to 1961 are superposed. The zero of the time scale coincides with the smallest density used in the superposition operation. The figure shows the line density for each indicated country as of 1 January 1961.

The effectiveness of this procedure indicates clearly that growth of line density follows some general rule. Certain features should be noted.

(A) The relative deviation from the average is more pronounced in the early years. The fact that many individual density curves occupy this region tends to produce a higher probability of deviation. The war period for many countries is also situated in this part of the diagram, resulting in some erratic behavior due to partial destruction of the telephone plant.

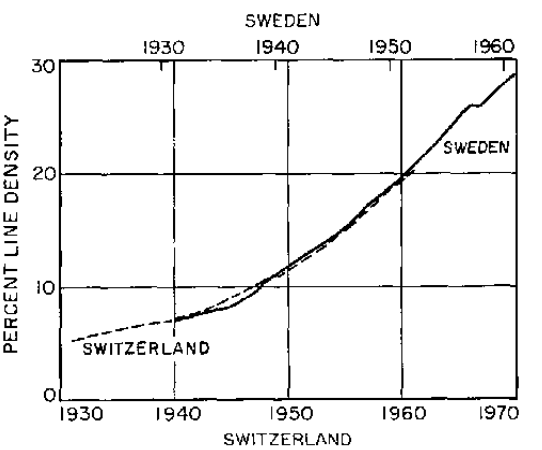


Figure 5—The line-density development in Switzerland is similar to, but 9 years later than, in Sweden.

As the trend changes substantially in the reconstruction period after the war, these short-time departures from the general trend do not invalidate the over-all superposition procedure. Also, the line density being a percentage of population, it will change most rapidly with increases in lines in countries having only a small number of telephones. In such cases, greater deviations from the average must be considered to be normal.

(B) There are countries that do not conform to this pattern. Denmark and Argentina seem to be developing at half speed, and France is also slowed down but to a lesser degree. When investments in the telecommunications system do not entirely cover the necessary expansion, there will be a serious handicap in development. This seems to be the case for France, where budget reasons delay normal development; there are, consequently, thousands of unsatisfied applications for telephones (some 115 000 at the beginning of 1959). If for social or economic reasons, a large part of a population is not interested in having telephones, the over-all picture of that country will show a slower development even if the remaining part of the population follows the general trend. This might explain the evolution in Argentina and similar countries. The reason for the slower evolution in Denmark is not evident, but the fact that only 56 percent of the network is automatic has perhaps some influence.

Figure 6—Composite line-density curves for various countries as a function of time in years t and a constant c of relative lag or lead with respect to all others. The development of each country is indicated as of 1 January 1961.

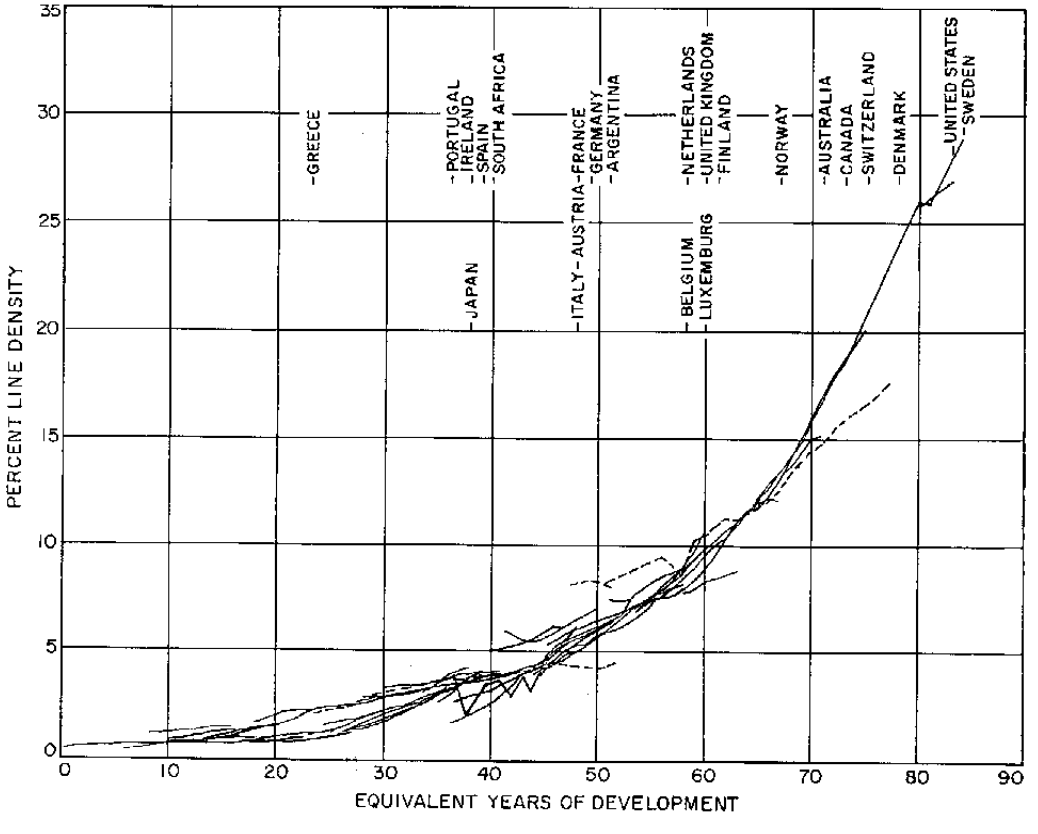


Figure 7 shows the average of the composite curves of Figure 6. It represents an experimentally derived general law of the line-density evolution. Two limit curves only 1 line per 100 inhabitants distant from the average, enclose quite satisfactorily all the component curves.

Expression (2) can be formulated as

$$d = f\,(t + c). \qquad (3)$$

Here $(t + c)$ is time as shown in Figures 6 and 7 and defined for 1961 by

$$t = T - 1961.$$

Figures 6 and 7 are based on $T = 1961$ and include $t = 0$, and the values of c in years for the different countries, can be read directly on the abscissa scales of Figures 6 and 7.

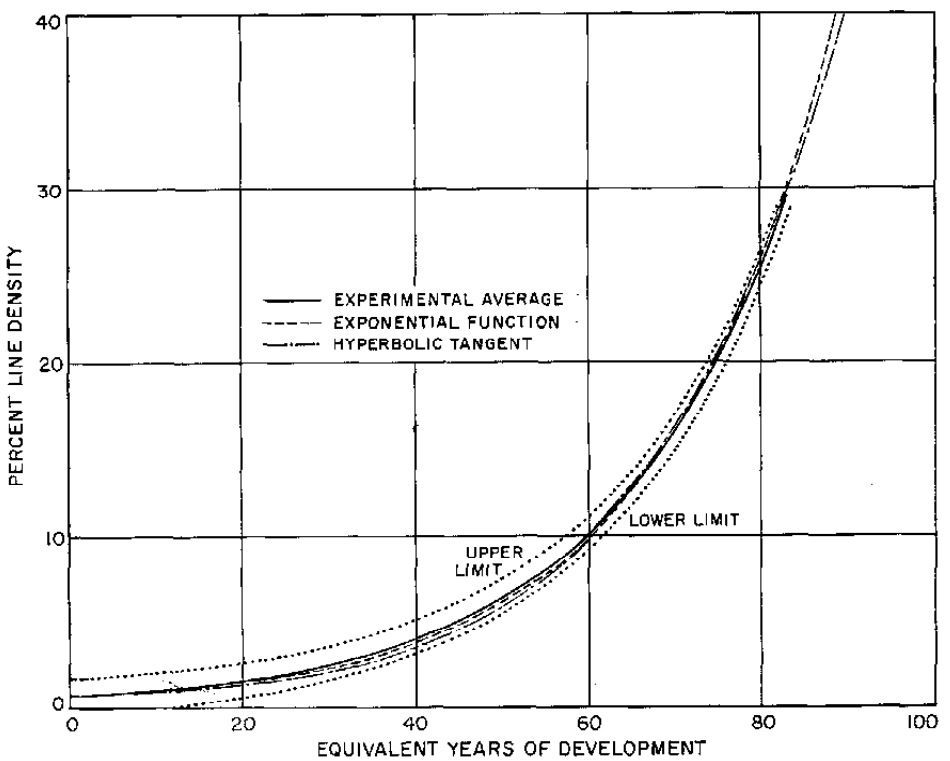
The higher the value of c, the higher is the line density; the difference between c for two countries gives the time lag between the present states of evolution for these countries. This time difference is, however, not a static figure but a dynamic characteristic of the actual development for 1961 when the general law was experimentally established.

3.2 Mathematical Forms of the Law

The average curve of the experimental law of evolution can now be used for forecasting the line density of a country in the near future.

However, when establishing a projection, account must be taken of the immediate past. This should be done as close as possible to the

Figure 7—The solid line is the average of the composite curves of Figure 6. The upper and lower limits, which would enclose all the smoothed individual curves, are only ±1 line per 100 inhabitants from it.



date of making the forecast; when this precaution is neglected, serious errors can be introduced. Even if the line-density curve of the country involved is up to date, it should again be superposed on the average curve to obtain the best possible coincidence to confirm the c value. This emphasizes again the fact that the time difference is not to be considered as a static and invariable figure.

Forecasts of future line densities will, however, be much easier to project with a mathematical form of the evolution law.

3.2.1 *Exponential Function*

The regularity of the average evolution curve suggests a constant increase of line density with time and the similarity of the average curve to an exponential function.

$$d = a^{[(t+c)+b]}. \tag{5}$$

In fact, a number of telephone administrations use such an exponential for their planning and assume a constant rate of change of line density. Figure 8 shows yearly density growth as a function of time for half a dozen countries. As can be seen, each country has a pronounced characteristic growth rate. Although the curves oscillate violently, there is no visible trend toward a steady increase or decrease in any of them. On the contrary, all the oscillations take place around the same common average of almost 5 percent per year. These irregularities stem from various temporary influences such as local crises, budget allocations, postwar reconstruction, and so on.

A good approximation, as plotted in Figure 7, of the experimental average is given by the following exponential expression.

$$d = 1.05^{[(t+c)-13.4]} \tag{6}$$

or, with $t = T - 1961$

$$d = 1.05^{(T-1974.4+c)}. \tag{7}$$

This exponential function can be used for calculating future line densities of a country if $(t + c)$ is smaller than about 80, this means during some 20 years from now on for a country with a present c value of about 60. Expressions (6) and (7) suppose a yearly growth of 5 percent, a value that could be foreseen by the study of Figure 8.

Although this exponential function now gives acceptable forecasts, the fact that such a function tends to infinity indicates a future failure of the law. Caution is therefore necessary in making a line-density forecast beyond the presently known part of this experimentally determined law.

3.2.2 *Hyperbolic Tangent*

As in the long run the exponential law will give highly improbable, if not quite impossible, values, another approach should be tried.

Consider a network serving an isolated area of C elementary communities (families, commercial, and industrial concerns) that can have telephone connections [1]. At a given moment, L of these communities have their telephones; $p = L/C$ is then the saturation quotient.

The probability W that a contact requiring telephone communication is established between a subscriber, group L, and a nonsubscriber, group C-L, is given by

$$W = 2\frac{L}{C}\cdot\frac{C-L}{C} = 2p(1-p). \tag{8}$$

The frequency f of the contacts between a given elementary community and another community can, as a first approximation, be considered as constant. On the other hand, if n is the average number of contacts between a nonsubscriber and the subscribers, sufficient to induce the former to request a telephone line, the saturation quotient increases during a time dt with

$$dp = W(f/n)dt = 2(f/n)p(1-p)dt \tag{9}$$

or, with $f/n = \omega$

$$dp = 2\omega p(1-p)dt. \tag{10}$$

This is the differential equation of p, the proportion of elementary communities having a telephone line in the considered network. Re-

solving this equation, we get the evolution in time.

$$\int \frac{dp}{p(1-p)} = 2\omega t \qquad (11)$$

$$p = \tfrac{1}{2}\,[1 + \text{tgh}\,(\omega t - k)]. \qquad (12)$$

This equation, giving the variation of p in time, is a hyperbolic tangent translated in the coordinate system. This curve, known also as a logistic curve [2], is much used to represent the trend of various growth phenomena in the demographic and biological sciences.

As $p = L/C$, and with the help of (1), it is now possible to formulate a second mathematical expression of the general law for line density.

$$d = \frac{100C}{N}\,\tfrac{1}{2}\,[1 + \text{tgh}\,(\omega t - k)]. \qquad (13)$$

A good approximation of the experimental average shown in Figure 7 is given by the following logistic curve

$$d = 115.81\left(1 + \text{tgh}\,\frac{T - 2080.6 + c}{38.3}\right). \qquad (14)$$

This hyperbolic tangent coincides nearly completely with the known part of the experimental average curve as does the exponential function given by (7). It similarly shares the necessity of being used with great caution when extrapolating beyond the known part of the experimental curve.

Indeed, complete saturation, $p = 1$, will be reached when each elementary community has its own line. This gives for (14) a density of nearly 232, or that number of lines per 100 inhabitants. It seems unlikely that the line density will ever reach such a high level even after an infinitely long time and assuming such

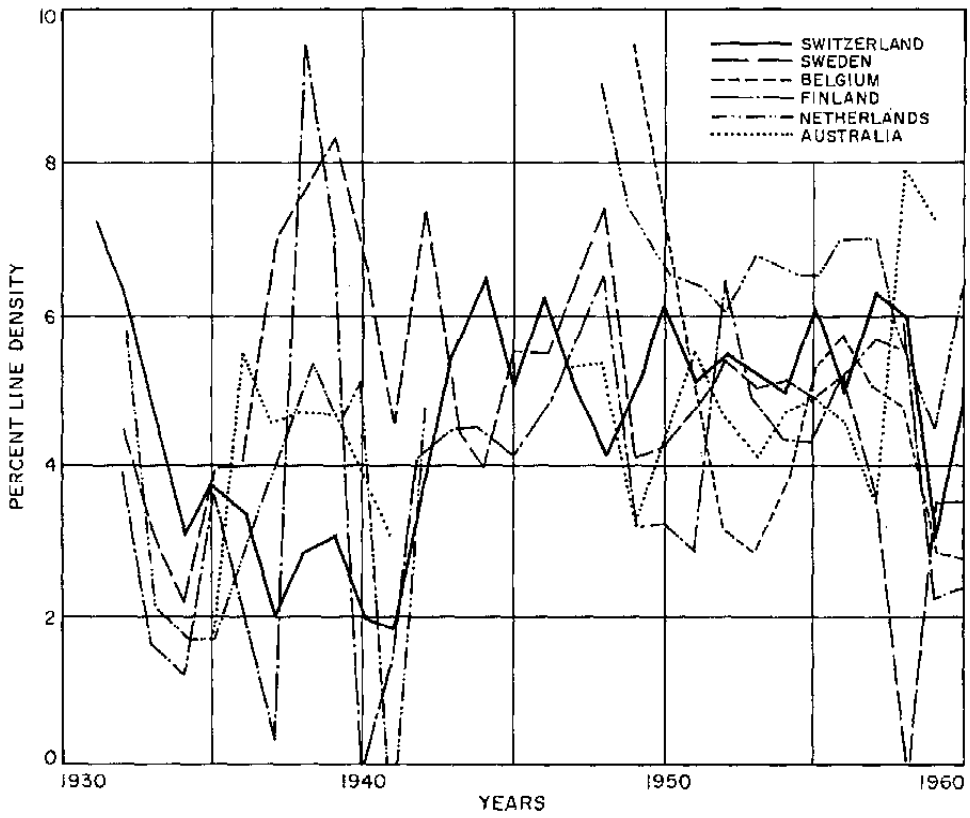


Figure 8—Yearly density growth of these countries oscillates around 5 percent.

developments as a second line in many homes, telephones in many cars, in trains and planes, more pay stations, emergency sets along the road, et cetera.

The actual development of the line density will probably follow more or less the logistic curve till a certain density is reached and will then gradually tend to a saturation density lower than the maximum given by the logistic curve. Both this departure from the common curve and the corresponding saturation may be different for each country, depending on local factors. In this way, the general law of density evolution would be described with the help of two or more mathematical functions, one of them being the logistic curve. The other factors in the evolution will be expressed otherwise but the data necessary for computation seem to be incomplete or missing now.

A factor that can be interesting in this respect is the coefficient $\omega = f/n$ in (9). The frequency f of the contacts between subscribers and nonsubscribers, and the threshold value n of the contacts sufficient to stimulate a request for a telephone connection were both considered as being constant. This is perhaps a simplification of the real situation. The frequency of contacts will probably increase and the threshold value will decrease as the public subscribes more and more readily to the telephone. Both factors are influenced by the social level but it is not certain that they are functions of time only, if at all.

3.3 Maximum Line Density

The line density in a country will reach its maximum when each elementary community has its own telephone connection, at least for the assumption that no communities will be granted more than one line while others have no telephones. When the ultimate number of communities among a population can be evaluated, an acceptable estimate of the saturating line density becomes possible.

The elementary communities can be divided in two main groups: residential and professional.

3.3.1 *Residential Communities (Households)*

To the residential group belong families, persons living alone, and also all sorts of communities of people living together or sharing the same quarters as in hotels. The number of families per 100 persons is not constant in time. In Belgium, there were 24.66 families per 100 inhabitants in 1910; 26.79 in 1920; 29.34 in 1930; and 33.33 in 1947. Today there may be 35 families per 100 inhabitants.

This family density varies also from one country to another and was for instance: 31.37 in France (1954), 27.83 in Switzerland (1950), 24.86 in Italy (1951), and 28.4 in the United States (1958).

Not all families will ever have their own telephones: some related families, for example, parents and married children, may live in the same house and share a line. Also, a sufficient economic status must be reached before a family will install a telephone in preference to buying anything else. But there will also be households with a second and perhaps even a third line.

3.3.2 *Professional Communities*

Commercial and industrial enterprises and offices form the professional group. Not all of them will have a telephone; many small shops will use the family line for business purposes. In contrast, the bigger firms will have several connections between the public exchanges and their private branch exchanges. It is therefore nearly impossible to estimate the line density for this group. Other factors are that time seems to make the telephone more of a necessity in all countries, and that a fully industrialized country will have more lines in this group than an agricultural country.

Also in this group are pay stations, sets in public offices, and emergency sets along the roads. Finally, there must be included sets in

cars, trains, planes, and ships, connected by radio with the general network. For the moment their number is still small but it will certainly grow.

It will be extremely difficult to estimate a saturation value for the line density. There are, however, no indications of an imminent saturation and a line density of at least 40 seems attainable in countries having high industrial and social development as is the case in Western Europe.

4. Subscriber Sets per Telephone Line

The first people to install telephones were certainly businessmen, who considered the new invention as a complement of the already widespread telegraph network. As telephony was a novelty, there was only one set in a building for the exclusive use of a few selected employees. Seldom was a second set found. Residence telephones were rare for there was almost nobody to talk with. So in those early days the relation of sets to lines (S/L) exceeded unity only by a small fraction.

As the practical value of telephones became apparent, more and more lines were installed for business purposes, but still with only one or at best a small number of sets per line. In the meantime, the residence telephone made its entry into the higher social levels.

Later on more and more homes were equipped, but, because business and industry installed larger and larger private branch exchanges, the relation of sets to lines increased steadily. Most countries are today in this stage of telephone expansion.

After a certain time, each firm reaches saturation. Service to houses and apartments becomes commonplace but tends to retain its one set per line status. This causes a decrease in the S/L relation; at this moment the United States, New Zealand, Switzerland, and maybe a few other countries seem to have reached this stage.

In the future, many homes will have more than one set per line and an increased S/L may be expected.

As a result of these conditions, the S/L relation develops as indicated in Figure 9.

Figure 10 shows the sets-per-line evolution with time for a few countries for which sufficient information was available to give a large part of the development curve, or which are to some extent especially characteristic. Sweden, for instance, a country where the telephone has always been a widespread social object, shows a remarkable stability of the S/L relation. The curves for Switzerland, New Zealand, and the United States indicate that these countries have passed the maximum; unfortunately some data for this last country are missing and the new increase is difficult to interpret. A maximum need not always be conclusive as is emphasized

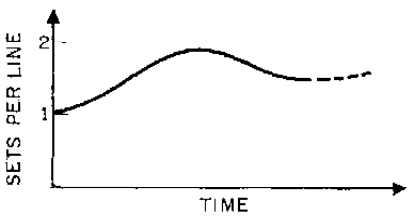


Figure 9—General development of sets per line.

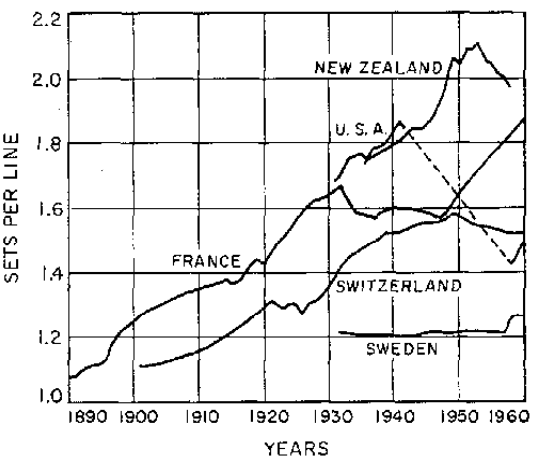


Figure 10—Development in sets per line for several countries

for France, where a new increase can be seen after a first maximum and subsequent decrease.

How the relation between sets and lines will be at a given moment depends on the industrial and commercial conditions of the country and also whether the population considers the residence telephone to be a social necessity. The first factor tends to increase the S/L relation, while the second has an opposite effect.

The density of subscriber sets s can now be defined as a function of the line density.

$$s = (S/L)\ d. \qquad (15)$$

As stated before, the line density can be forecast with some accuracy but it is difficult to predict the evolution of S/L. Because the variation of S/L over a relatively short period of, say, about 10 years, is normally not larger than some 10 percent, it is possible to obtain a fairly good approximation of the future density of sets in a given country by multiplying the projected line density with its present S/L relation. If out of the normal trend of this relation, a future value can be estimated, still better results will be obtained.

5. Telephone Conversations

With the help of the above-stated law of evolution, future line densities can be forecast within acceptable limits. As explained, the density of sets is more difficult to foresee because the relation of sets to lines depends on many local factors that can easily change in time. Despite this, set density can be projected albeit with a larger deviation possibility.

The number of conversations depends strongly on and responds immediately to the social, economic, and political situations of the moment. Examples of this close dependency have already been cited.

The number of conversations is also directly related to the character of the people in the different countries. No single rule would be applicable everywhere. For instance, in Ireland there are 51 conversations per capita per year while in Japan, which has the same line density, each inhabitant makes an average of 164 calls per year. These differences exist even between neighboring countries; Belgians now make some 79 calls per year while the Dutch have 114 conversations for the same line density.

However, if large enough deviation limits are accepted, there seems to be a roughly linear relation between the total number of conversations per capita and per year g and the line density in a given country.

$$g = a\,d + b. \qquad (16)$$

For the post-war values, this expression gives an approximation that is generally within 15 percent.

With a rare exception, most countries have an increasing number of conversations per capita per year as the line density grows, but it is quite possible that this trend will be reversed as the line density increases further.

6. Conclusion

No attempt is made to present the basic reasons that control the evolution of telephony. A rigorous mathematical treatment of this evolution may even be impossible as it depends on a large number of economic, social, and psychological influences. The purpose of the paper is to emphasize that the individual development curves of most countries are not as incoherent as they appear to be at first glance, but that they have as a general characteristic a nearly identical growth rate.

This common characteristic is formulated in a general law of line-density evolution and, mathematically expressed, even though incomplete and approximate, can make planning for the future expansion of telephony less obscure and unsurveyable.

7. Acknowledgment

The author expresses his thanks to Mr. J. Mallet of Laboratoire Central de Télécommuni-

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1000B Pentaconta Crossbar Switching System

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1. History of Pentaconta

The Pentaconta crossbar dial telephone switching system was developed in 1953. Its basic element is a crossbar switch having 14 horizontal bars, one of which is used as a change-over bar, to provide for switching 52 outlets. This switch and the relays mainly used for other switching functions were described in this magazine.[1]

The first design was called the Pentaconta *500* because the subscribers' lines and their associated equipment were divided into 500-subscriber units each served by 2 markers. The first exchange of this type was cut over at the end of 1954 at Cento, near Ferrara, Italy. Others were installed during the next four years in France, Chile, Brazil, and Argentina, the largest being at Bergamo, Italy, with 17 000 subscribers' lines.

From experience with this equipment, it was decided to increase the capacity of the line unit to 1000 subscribers, while retaining the same principles of operation.

Actually, 2 markers can handle 6000 calls per hour with a marker holding time per call ranging between 450 and 500 milliseconds. Other detailed improvements were effected in circuit operation, the equipment layout was revised, and it was decided to use wrapped wiring connections for both internal wiring and outside cabling of the equipment units. This second design is referred to as the *1000A* type. It was first installed in Albi, France, in 1959. A large number of these exchanges are operating all over the world.

Experience with the *1000A* system, the requirements for the establishment of international connections, and the decision of the French Administration to adopt Pentaconta for the Paris network, led to the development of the *1000B* type, which is suitable for public exchanges of 1000 to 50 000 lines.

Some of the most-outstanding features of the *1000B* system are as follows.

(**A**) Full accessibility to large groups in the group selection units (1040 terminals) with negligible internal blocking so that outgoing junctions may be calculated on the basis of ideal groups.

(**B**) Second attempt in case of alternative routing or faulty operation.

(**C**) The group markers can bring about overflow operations by providing for paths of first and second choice.

(**D**) The group markers will effect alternative routing, provided the original program for such selections is not modified by the translator.

(**E**) The translators will carry out two alternative routings in case of a change of program. It must be pointed out that the holding time of a translator is about 300 milliseconds. A set of 2 translators can handle 14 000 calls an hour. Each selection coupler has access to 2 translators. The number of translators will be based on the traffic of the exchange.

(**F**) It is possible to incorporate in the translators a nucleus of subscriber translation, enabling the connection of the outgoing lines of very large private branch exchanges directly to the outlets of the group selection units. This will reduce the traffic to the line selection units.

(**G**) Any type of signalling code may be provided for through auxiliary control circuits that are available to the registers.

(**H**) Provision for automatic message recording.

2. Switching Functions

Figure 1 shows a block diagram of a Pentaconta exchange of the *1000B* type. The outgoing and incoming junctions use only relay groups, whereas the line and group selection units use

[1] F. Gohorel, "Pentaconta Dial Telephone Switching System," *Electrical Communication*, volume 31, pages 75–106; June 1954.

2-stage crossbar switches under control of markers for conditional-selection operation.

Figure 2 shows the arrangement of a line or group selection unit. The two crossbar switching stages are called the primary stage and secondary or terminal stage; each stage consists of several sections made up of one or more multiswitches.

By way of example, the standard group selection unit for a large exchange may include 7 primary sections, each consisting of 2 multiswitches having 22 verticals to provide $2 \times 7 \times 22 = 308$ inlets, and 20 secondary sections, each of 1 multiswitch with 14 verticals. Each secondary section gives access to 52 outlets so that an incoming junction connected to an inlet of the primary stage has access to $52 \times 20 = 1040$ outlets. The 2-stage link arrangement is therefore the equivalent of a selector with 1040 terminals.

The 52 outlets of the primary sections are divided into two classes: 40 outlets (20×2) are assigned to the direct junctions between primary and secondary sections, and 12 outlets are reserved for alternative routing of calls within a selection unit, which will be referred to as an "entraide." *

* The French term "entraide" has the meaning of "mutual aid" and "reciprocal overflow," which differs widely from the usual overflow process.

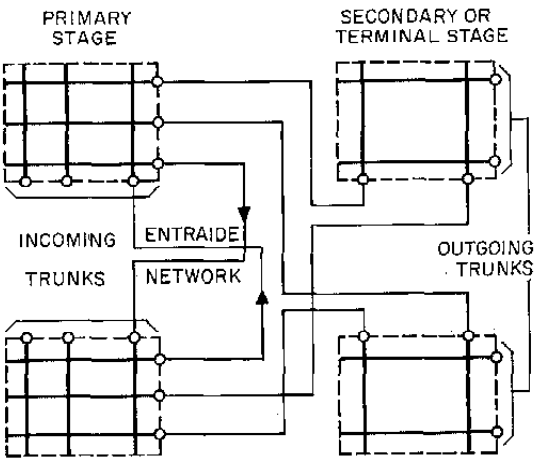


Figure 2—Arrangement of links in line and group units.

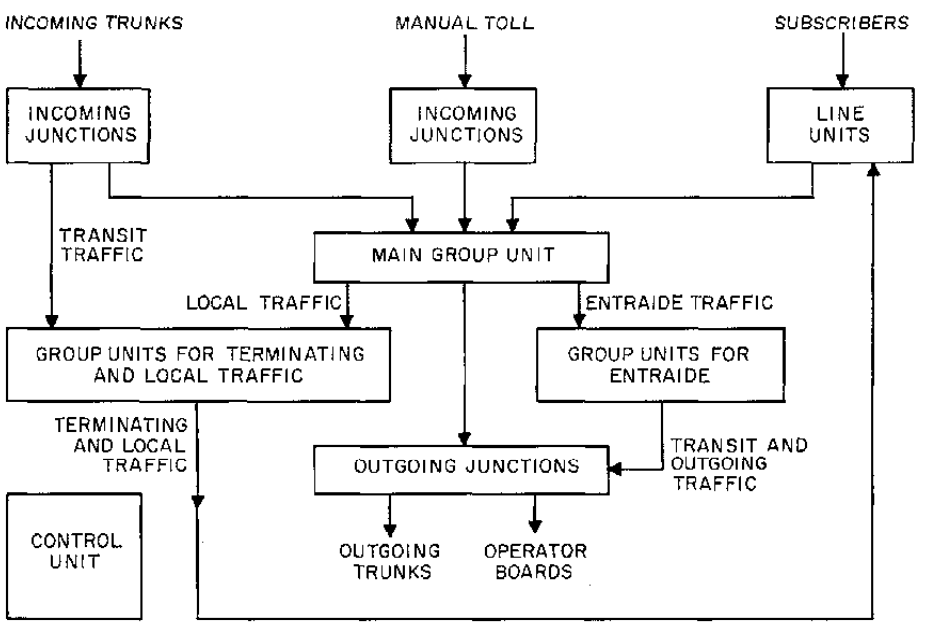


Figure 1—Functions of a Pentaconta exchange.

In conditional selection, the marking of a group of outgoing junctions causes a certain number of secondary sections that have free paths of access to the primary section in which the call originates to be selected as being available for handling the call. A lack of free paths between the originating primary section and those secondary sections having access to free outlets to the desired outgoing route (even if free paths of access to other secondary sections are available) causes an internal blocking of the link system. See Figure 3.

Assuming that b is the average load of a junction between the primary and secondary stages, c the average load per line of the outgoing group under consideration (the route), f the number of links between each primary section and each secondary section, q the number of lines constituting the route, and m the number of secondary sections, the probability of congestion E is given by

$$E = [b^f + c^q (1 - b^f)]^m$$
$$= [c^q + b^f (1 - c^q)]^m$$

if the independence of the 2 stages is admitted. This equation applies to a link system without concentration or expansion, that is, a link system in which the primary sections contain an equal number of links and inlets, which is the most-frequent case in the Pentaconta group selection unit.

Calculations show that a 2-stage link system is not fully "permeable"; a 3-stage link system must be used to reduce internal blocking to an acceptable value. The function of the entraide arrangement is to provide a third stage only for those times when no free access path is available directly between the primary and secondary sections. The total congestion will obviously depend on the permeability of the entraide network. See Figure 4.

It must be pointed out that a 2-stage arrangement of multiswitches will provide accessibility without an unreasonable amount of internal blocking only if q the number of lines of the routes is considerably higher than 1 and these lines are distributed equally over the secondary sections. This is the case for a group of outgoing junctions or for a number of subscriber's lines constituting a private-branch-exchange group. On the other hand, a single subscriber's line has only one position in one terminal section; the entraide network then provides a 2-stage conditional selection and the necessary accessibility.

Figure 5 shows the arrangement of a group selection unit of 7 primary sections having 40 inlets each, 4 entraide selectors per primary section, and 20 secondary sections with 14 verticals each, providing a total of 280 inlets and 1040 outlets. The largest line selector designed so far to cater for the maximum traffic per subscriber's line comprises 8 primary sections (containing so-called call finders for

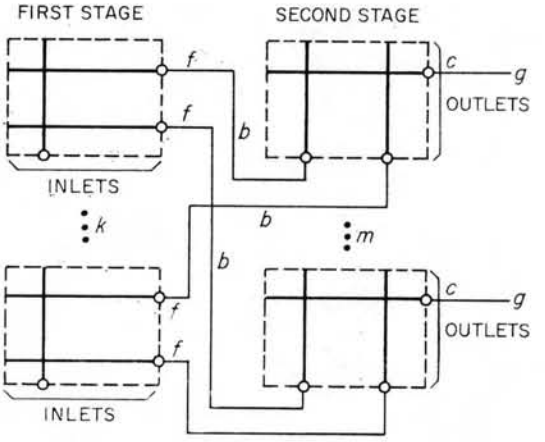


Figure 3—Link system having 2 stages.

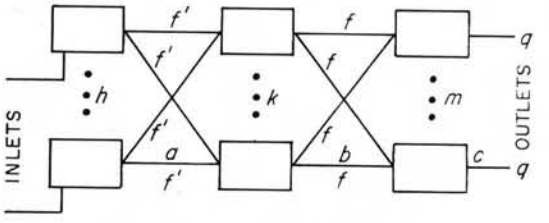


Figure 4—Link system having 3 stages.

handling originating calls and so-called "fifties" selectors for handling terminating calls) and 20 terminal sections equipped with a maximum of 16 verticals, each terminal section handling the originating and terminating traffic for 52 subscribers' lines.

Figure 6 shows the symbol of a line or group selection unit with entraide.

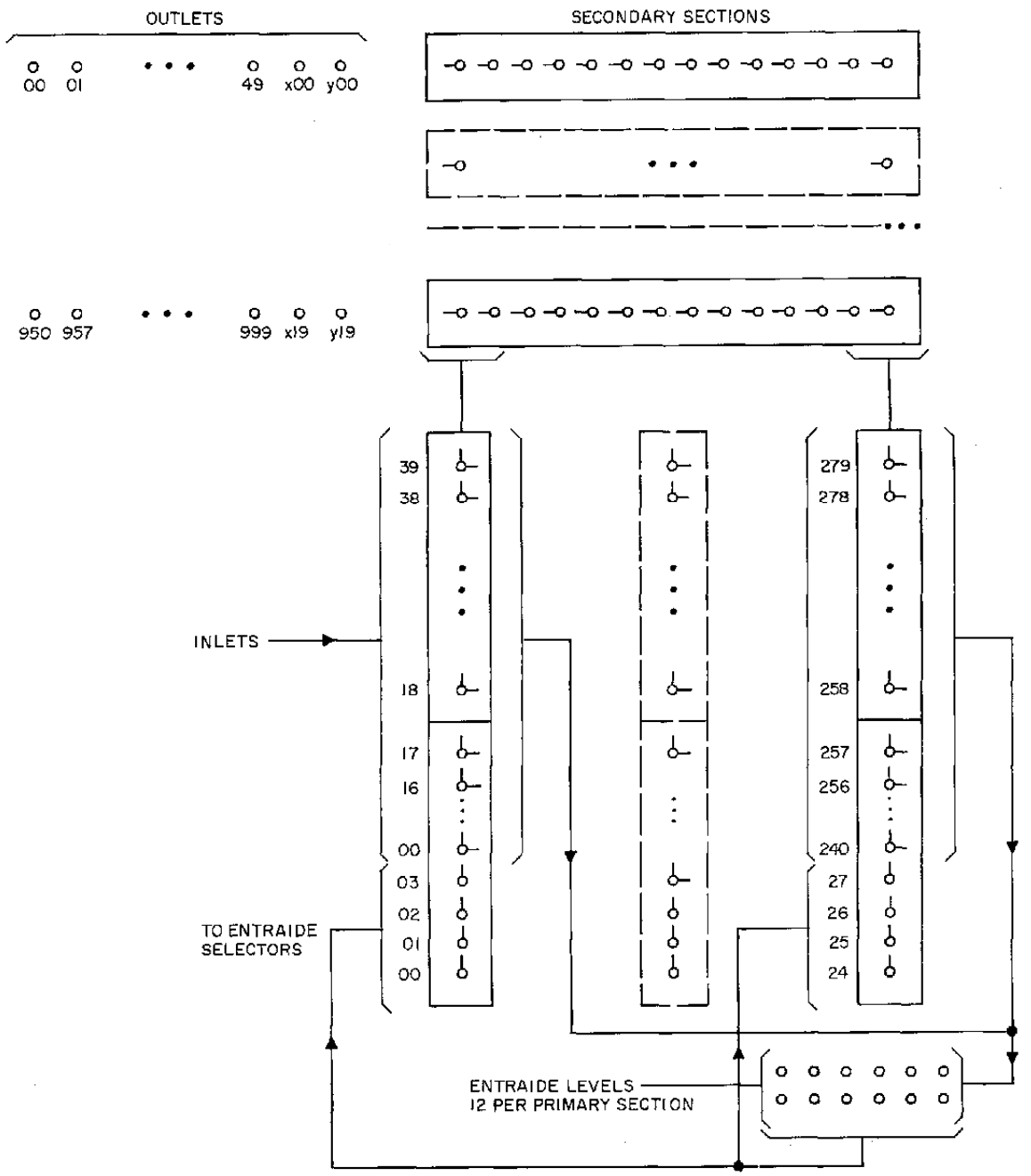


Figure 5—Typical arrangement of a group selection unit.

3. Local Calls

Figure 7 is a block diagram indicating the devices and selecting units engaged in the completion of a local call.

3.1 Line-Finder Operation

When subscriber *A* lifts his handset, the common relays controlling the line-finder operation ascertain which primary sections have access to free registers via free call finders. A register is chosen by the register finder as is also a single primary section having access to the terminal section to which the subscriber is connected.

One of the 2 markers is seized by the primary section designated to handle the call to mark the subscriber and primary section outlets. Simultaneously, the register is connected to a line-finder coupler, 2 of which may be provided for 12 registers. The marker records the class of the calling subscriber. For this purpose, all subscriber lines as well as incoming and outgoing junctions are characterized by a class-of-line indication, which serves to signal their class of service, type of signalling (loop break, revertive impulses, multifrequency code), supervision method, et cetera, to the marker controlling the connection. Once this is done, the marker is connected to a free path to the information bypass as well as to the coupler by means of a simple process of identification. When the register has received the calling line's class via the bypass, it proceeds to operate the vertical bar magnets. The subscriber is thereby connected to the register, after which the marker, line-finder coupler, and information bypass are released. Information is transmitted via the bypass by a self-checking 2-out-of-5 code.

If several subscribers in the same line unit call simultaneously, each of the 2 markers will serve one of these lines provided they belong to different fifties groups.

It must be noted that when establishing a connection from a calling line to a free register, a conditional selection is made via a 3-stage link system (group of registers, call finders, and terminal selectors) without use of the available entraide. Only in the very few occasions when all primary sections having access paths to free registers cannot reach directly the terminal section in which the calling subscriber is located, does the entraide mechanism come into action. The conditional selection is then made by a 4-stage link system, which practically affords full accessibility to all the groups of devices involved.

It is worth mentioning that the line-finder operation provides a delay function so that calls encountering occasional congestion will be served after a short wait instead of being rejected.

3.2 Group Selection

The calling subscriber dials after receiving dial tone from the register. When the register has received enough digits to undertake the first group selection (exchange selection), it engages a selection coupler, which simultaneously connects itself to both a translator and a group marker, the latter via a primary section of the group unit. The group marker as well as the coupler are then connected to the information bypass. The translated route is transmitted from the register to the marker via the coupler and the information bypass. Once informed of the route, the marker releases the bypass and also forces the coupler and translator to do so. The group marker marks the outlets belonging to the desired route (100 markings are possible) and chooses among them a particular one that can be reached by the primary inlet used for the call, preferably through a direct link or alternatively through the entraide stage. The marker receives the class-of-line indication from the chosen outgoing line and controls the connection of the coupler to the information bypass. The register next receives from the marker the class of the outgoing junction and possibly further indications concerning the selection (case of overflow for instance). It then controls the operation of the vertical bars. On connection of the outgoing junction through the main group unit, the marker is released as well as the multichannel bypass and the selection coupler.

A second group selection will be made in the terminating group unit following the same process when the register has received the required number of digits. It should be noted that one of the two group units may not be provided if this is not justified by the size of the exchange.

3.3 Line Selection

The subscribers' lines are divided into numerical groups for 1000 numbers, each served by a line selection unit. As the selection in a 1000-line group takes place in one operation, this can start only after the last digit has been received.

The selection process is the same as for a group selection; however, the selection coupler does not use the translator; the last 3 digits of the subscriber's number are transmitted untranslated to the line marker. In the Pentaconta system, a form of translation of these 3 digits takes place inside every line unit by a marker distributing frame enabling the assignment of any outlet within the line selection unit to any number of the thousands group. The individual lines of a private-automatic-branch-exchange group that share a common directory number may be distributed over different terminal sections. No additional equipment needs be provided for private-branch-exchange lines; ordinary single subscribers are treated as private-branch-exchange subscribers having only one line. Complete similarity exists between the marking of a route in group selection and the marking of a group of private-branch-exchange lines, and also for an ordinary subscriber. The information on the condition of the called line (subscriber free, with or without charge on answer, subscriber busy, absentee subscriber, et cetera) is transmitted during the line selection to the register, which may if necessary carry out a rerouting operation to direct the connection to special groups of junctions connected to the outlets of the group selection units, particularly in the case of absentees and unobtainable numbers.

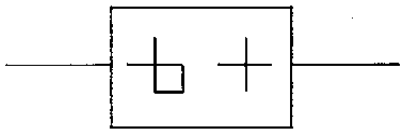


Figure 6—Symbol of a selection unit with entraide. The symbol at the right is for a crossbar switch and that at the left indicates the entraide arrangement.

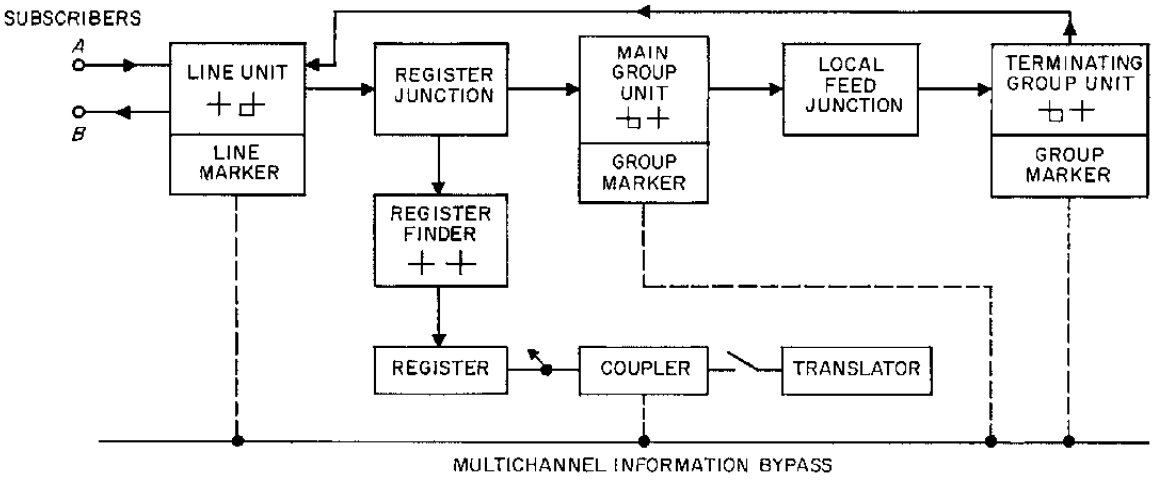


Figure 7—Switching network for local calls.

3.4 Establishing Speech Connection

After it has connected the called line B, the register informs the local feed junction of the type of operation it will have to provide (charge, supervision adaptation) by means of a direct-current compelled code, after which it releases. Subscribers A and B are now interconnected through the local feed junction, which controls the further operations required, such as ringing, metering, and release of the switching train at the end of the conversation.

4. Call to Another Exchange

Figure 8 is a block diagram showing the route and the switching and controlling units used for an outgoing call.

4.1 Line-Finder Operation

The line-finder operations, including seizure of a register, are described in Section 3.1.

4.2 Selection of Outgoing Junction

When the register has received sufficient information to determine the route, it calls the translator. When all elements required for choosing an outgoing junction have been collected, the group selection is effected as in Section 3.2. The class of the outgoing junction and the information supplied by the translator to the register, inform the register of the operations it will have to perform.

4.3 Distant Selections

Once connected to the outgoing junction, the register also seeks through a finder an auxiliary control circuit that provides a decimal sender, multifrequency sender, or receiver of revertive impulses to meet the operational requirements of the selection at the terminating exchange. The auxiliary control circuit controls the distant selections, receiving the called numbers from the register in which they were stored. When the numerical information must be translated for selecting purposes, the auxiliary control circuit obtains the translated information via the multichannel bypass from the translator and its associated selection coupler.

For interconnection with a step-by-step exchange, the register and the decimal sender to which it is temporarily connected are immediately released at the moment digit sending from the subscriber keeps pace with the retransmission of digits by the sender.

Thereafter the subscriber dials directly through the outgoing junction, which repeats the impulses towards the distant exchange.

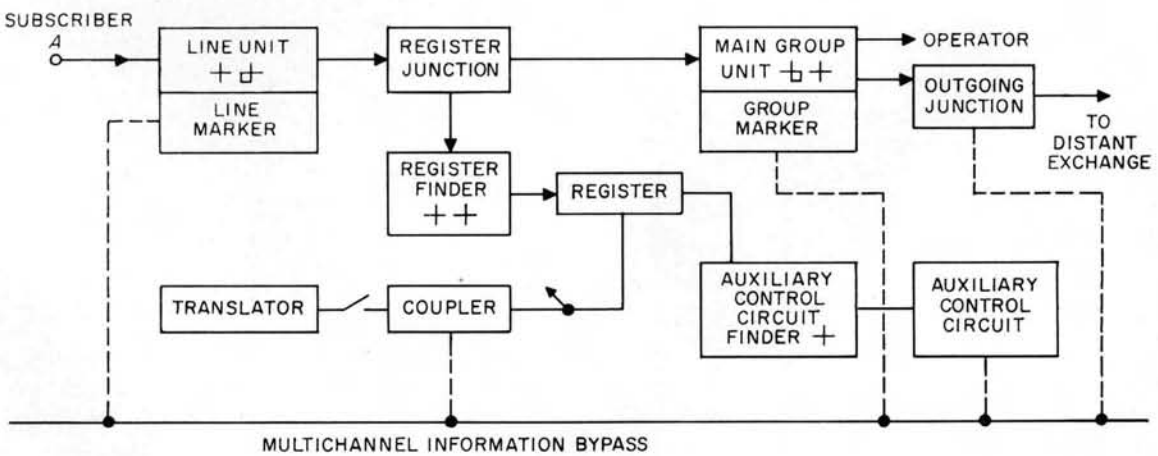


Figure 8—Switching network for outgoing calls.

4.4 Tariff Indication

The charges for connections over any outgoing junction may differ depending on the final destination of each call. If the outgoing junction must control the call charge, the tariff to be applied must be supplied when the register and its auxiliary control circuit have completed the distant selection. At this time, the auxiliary control circuit is released but the register recalls the translator and informs the junction to prepare for recording the tariff from the translator via the multichannel bypass. The register is then released normally; the outgoing junction ensures the functions of current supply, supervision, charge, and release of the engaged selectors.

5. Call Originating from a Distant Exchange

Figure 9 is a block diagram showing the route and the equipment involved in the completion of an incoming call.

5.1 Call from Exchange with Registers

A register is connected to the incoming junction by the incoming junction finder and the register finder, a 2-stage chain of multiswitches. It receives the class of this junction and calls for an auxiliary control circuit adapted to the particular type of signals to be received. The register, after informing the calling register that it is ready, receives the digits required for the selection.

The group and line selections are controlled by the register in the same manner as for a local call. If a large number of incoming junctions employ the same signalling system, as in a large urban network, it may be more economical to provide incoming registers incorporating the functions of the auxiliary control circuits so that the latter may be omitted.

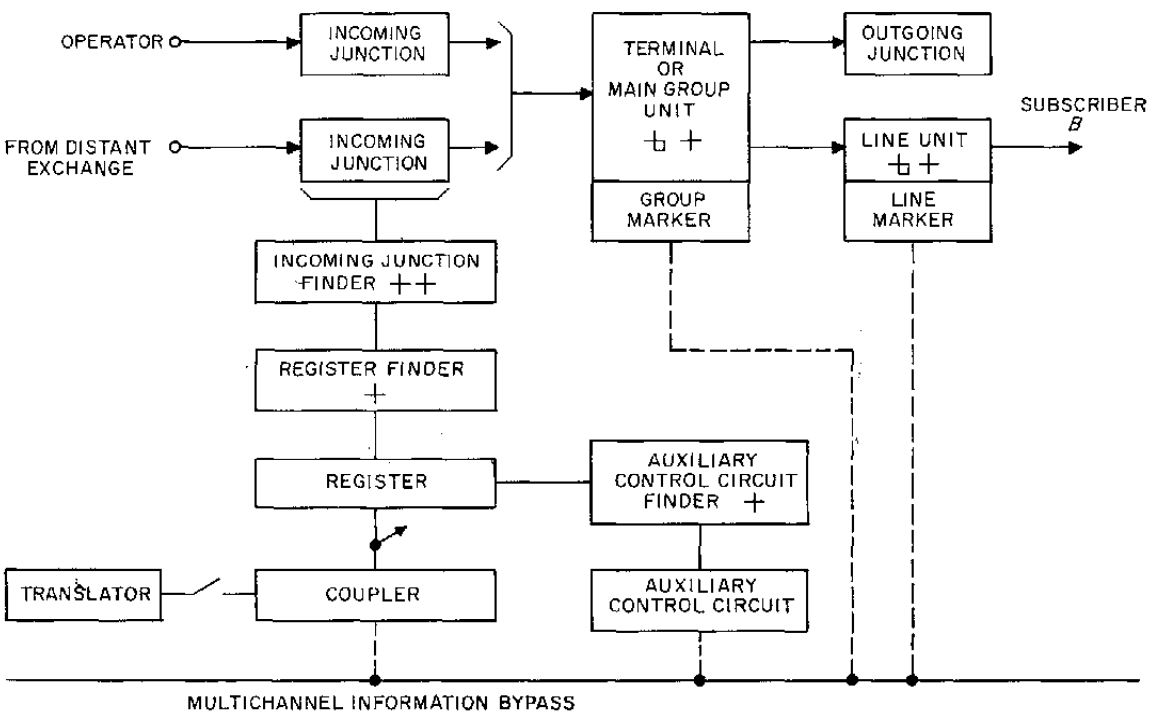


Figure 9—Switching network for incoming calls.

5.2 Call from a Step-By-Step Exchange

The time between the seizure of a junction and the moment the first digit must be accepted may be very short. Consequently, only one stage of finders must be used to minimize the time required for connecting a free register. Furthermore, no time will be available for the register to seek an auxiliary control circuit, so that this must be dispensed with and the register must be arranged to receive the dial impulses immediately. In some cases, it is necessary to record the impulses in the junctions themselves and transfer the digits subsequently to the register in a coded form.

6. Transit Call

This type of call is handled as an outgoing call. See Section 4. An outgoing junction is connected to the incoming junction. If a conversion of the selection code or a regeneration of the dial pulses has to be performed, the register will engage an auxiliary control circuit as in Section 4.2. It should be noted that the register may be connected simultaneously to two auxiliary control circuits, one for adaptation to the incoming signalling and the other for controlling the distant selections.

7. Call to an Operator

Referring to Figure 8, it will be seen that the junctions to operators' positions, in a manual toll exchange or for special services, are connected to the outlets of the group selection units. The selection process is the same as for an outgoing call except that no subsequent selections are to be made.

8. Call Originating from an Operator

As shown by Figure 9, a junction from an operator is treated as an incoming junction. When the operator's position is provided for key-sending, the register will call an appropriate auxiliary control circuit.

9. Multichannel Information Bypass

The multichannel information bypass used for exchanging information among the registers and the various common devices in the exchange consists of a number of conductor pairs. Such units as markers, auxiliary control circuits, and outgoing junctions have access to only one pair, which they can use indiscriminately. They are said to be active units for they seize the bypass themselves, through concentration devices, if necessary. The register couplers are said to be passive units but have access to all paths in the exchange. The routes of the bypass are identified as families that are at the disposal of the various stages of selection. Within each family, identification of any pair is made either by direct-current polarities (plus or minus 48 volts) or by 2-out-of-5 voice frequencies. The holding time of the bypass per call is about 100 milliseconds and 24 000 calls can be handled per hour per pair.

10. Easy Maintenance

10.1 Test Chain

Any subscriber line can be called from the test desk by its directory number. For the 40 lines in each line selection unit in excess of 1000 numbers, an access code is used (digit 11).

10.2 Fault Recording

Selections at each stage have to be completed within a given time. Thus, when each selection begins, the pilot unit starts a time-limiting device. If the selection takes too long, the pilot unit automatically connects a localizer that in its turn calls a fault recorder. All equipment units engaged for this selection will have their numbers recorded, that is, the numbers of the calling subscriber, selection units, register, and outgoing or incoming line. Furthermore, for each of these, the positions of its main relays will be recorded. The fault register will punch this information on a card, together with the time of occurrence. Every call handled by the

exchange is thus subjected to constant supervision.

10.3 False Calls

Each subscriber line equipment includes a line lockout arrangement. After a register is seized and dial pulses are received, the line is switched to an analyser that measures the loop and insulation resistances and detects any accidental earth potential. This analysis is recorded on a punched card in large exchanges or is shown on a luminous number display in small exchanges.

10.4 Traffic Observation

Registers, markers, auxiliaries, and junctions are equipped for traffic observations such as the number of seizures and the traffic carried.

10.5 Alarm Supervision

The alarm produced by blown fuses, special currents, et cetera, cause alarm lamps to light on the bay concerned, on the row of bays, and on the centralized supervisory desk; they also operate various alarm bells.

11. Exchange Equipment

The components are mounted on bays accessible from both sides. Assembled on the installation site, the bays stand side by side to form rows of varying lengths to fit the dimensions of the room. Aisles between rows give access at the front of the bays to the selection units and the cable connecting strips and, at the rear, to the internal factory-made wiring. The aisles between the front sides are used more frequently than those at the rear and are therefore the wider of the two.

The multiswitches and relays are housed in metal frames that are either 1 meter or 1.29 meters (39.4 or 50.8 inches) wide, 390 millimeters (15.4 inches) high, and 200 millimeters (7.9 inches) deep.

The wide frame can accommodate a multiswitch equipped with 22 verticals and in addition has 2 vertical holders each mounting 22 relays of the flat type used for positioning the selectors. The narrow frame takes a multiswitch having 17 verticals and a single holder for 22 flat relays. The number of switches in these standard frames can be reduced in favor of additional relay holders as some frames, such as markers, registers, and junctions, do not have multiswitches.

A frame only 235 millimeters (9.3 inches) high is often used at the upper part of a bay to house either 10 flat relays or the connecting blocks for the internal distributing frames of the selection units.

The blocks for connecting the outside cabling are at either side of the frame and as the terminals of the multiswitch operating magnets are at the front, full operation of the multiswitches and testing of the frames may be done without recourse to the rear side. Designation strips at right angles to the connection blocks identify the distribution of terminals among the blocks.

Figures 10 and 11 are front and rear views of a frame of call finders and fifties selectors. Figures 12 and 13 are similar views of a line marker and its distributing frame for a bay 1.45 meters (57.1 inches) wide.

Solderless wrapping is used for all connections inside the frames and between the various apparatus and distributing frames.

Metal covers protect against dust and shock. The frames are supported on uprights of folded sheet iron which house the wiring forms interconnecting the various frames of the bay and the switchboard cables coming from the intermediate distributing frames and other bays. Figure 14 shows this arrangement. The uprights are closed at both rear and front by cover plates that can easily be removed.

The two bay widths of 1.16 meters (45.7 inches) and 1.45 meters (57.1 inches), having a ratio of 4 to 5, correspond to the two standard widths of selector and relay frames. The bay height of 3.47 meters (137 inches) permits

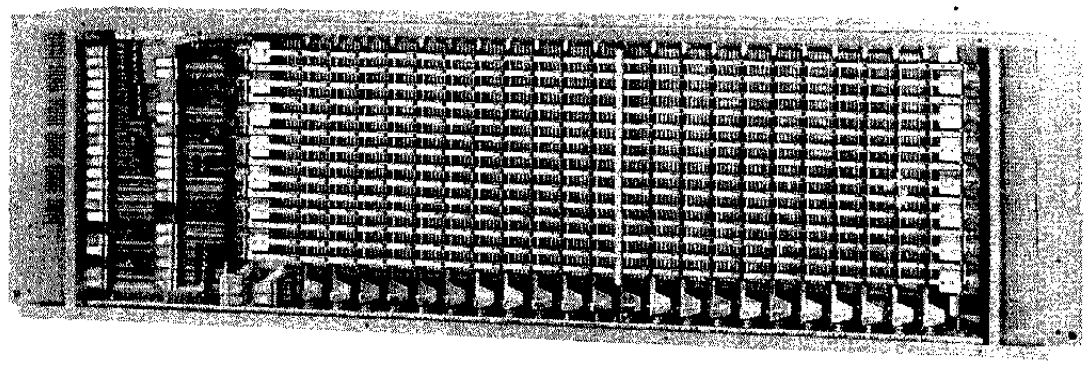


Figure 10—Front of frame of call finders and fifties selectors.

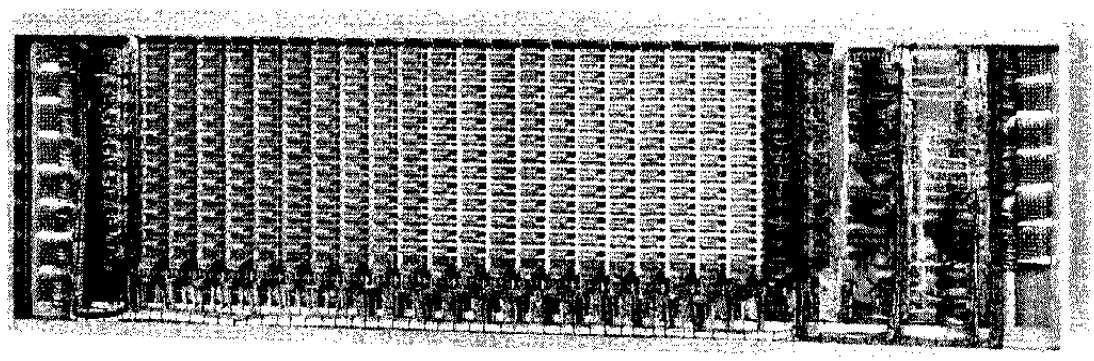


Figure 11—Rear of frame of call finders and fifties selectors.

mounting several frames arranged from bottom to top as follows.

(A) Power supply, protecting fuses, and fuse-alarm relays for the 48-volt system and for special currents.

(B) Three selector or relay frames.

(C) Supervisory panel for alarm and supervision lamps, listening-in and busying jacks, as well as a number of buttons used for maintenance purposes.

(D) Four selector or relay frames.

(E) If required, a 235-millimeter (9.2-inch) frame may be placed at the top of the bay.

Figure 15 represents a bay of couplers and registers.

If the ceiling of the switch room is too low for standard bays, the bay height may be reduced by mounting fewer selector or relay frames, which are 390 millimeters (15.4 inches) high.

Figure 16 shows the equipment layout of the frames of a 1000-line selection unit capable of handling a traffic of 0.09 erlang per line.

The line selection units are connected by cables to the main distributing frame, the subscriber meters, and the intermediate distributing frame of the register junctions. Jumpers run in this distributing frame permit the ratio of call

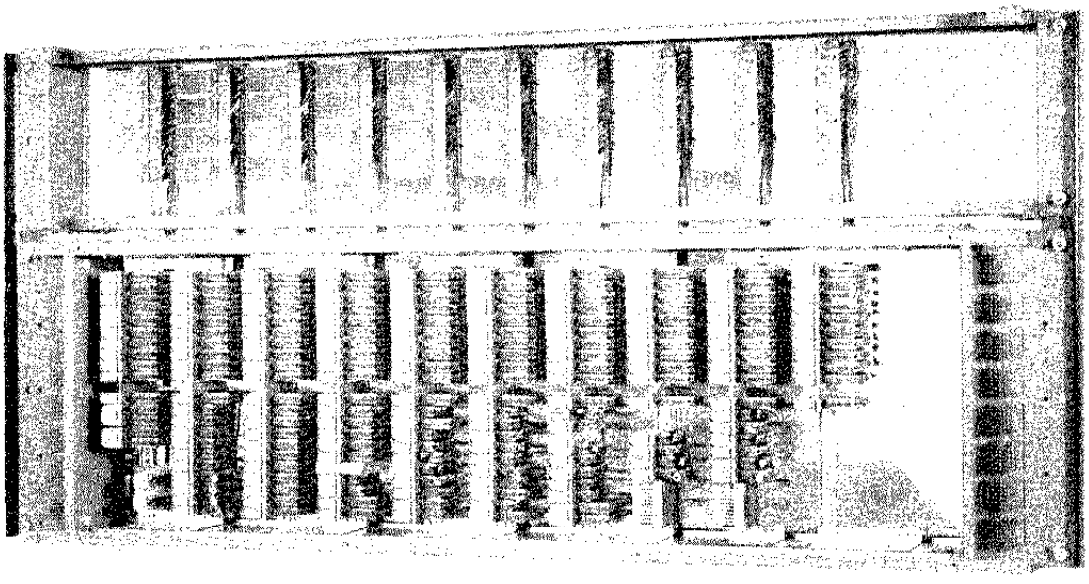


Figure 12—Front of frame of line marker and its associate distributing frame.

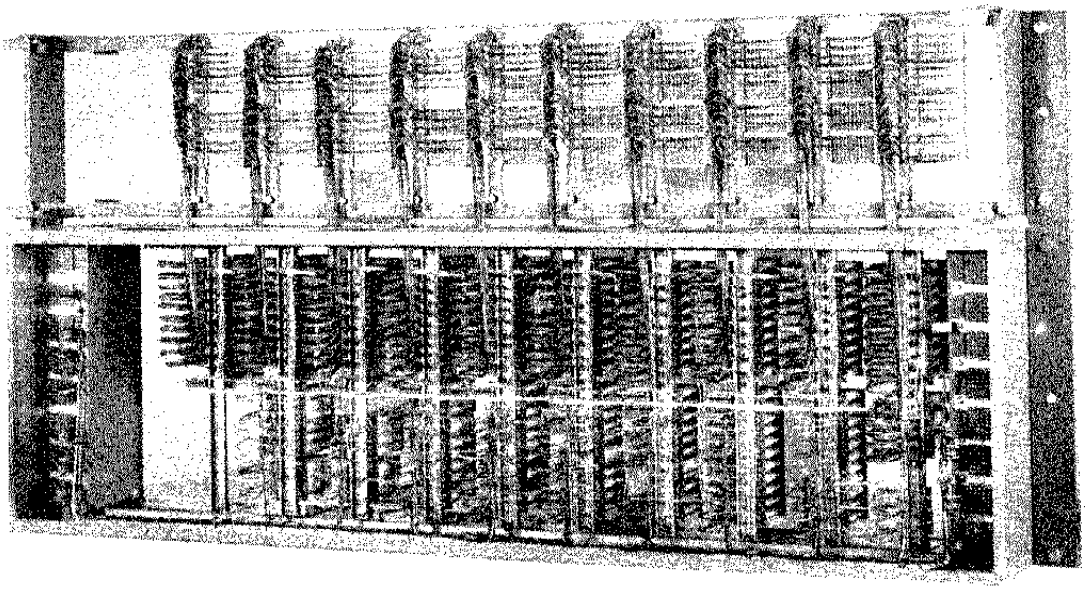


Figure 13—Rear of line marker and its distributing frame.

finders to fifties selectors to be adapted to the outgoing and incoming traffic rates. The assignment of the line relays to the subscribers' lines is made on the markers' distributing frames, which permits partial separation of directory numbers from equipment location for traffic balancing. As the position of each line generally corresponds to its directory number, in most cases this position can be marked by merely interconnecting two adjoining connector tags of the distributing frame. Such interconnections are made at the factory by bridges that can easily be removed when required during service. It should be noted that bridges are also used in the class distributing frames and for the connection of a subscriber's meter to each line.

For a change in traffic, the number of terminal selectors per frame, as well as the number of frames of call finders and fifties selectors, is suitably modified.

For traffics above 0.1 erlang per line, an additional frame that can mount 2×11 supplementary selectors is supplied for every two sections of call finders and fifties selectors, so

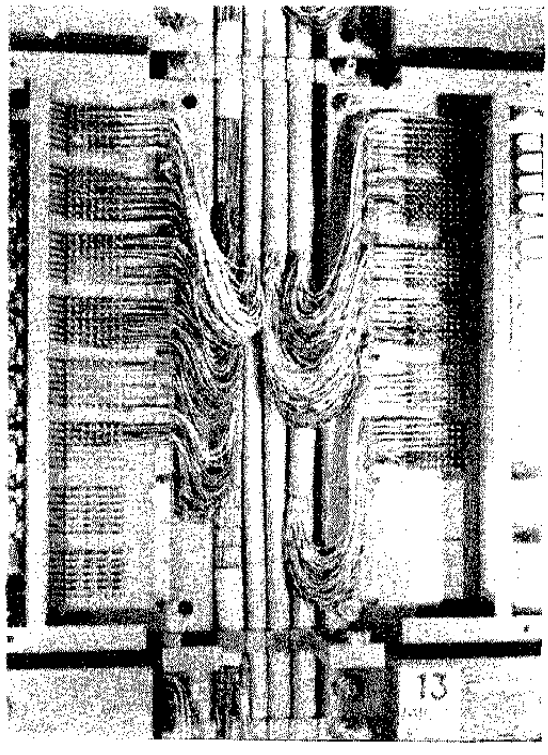


Figure 14—Cable run, connections, and wiring forms.

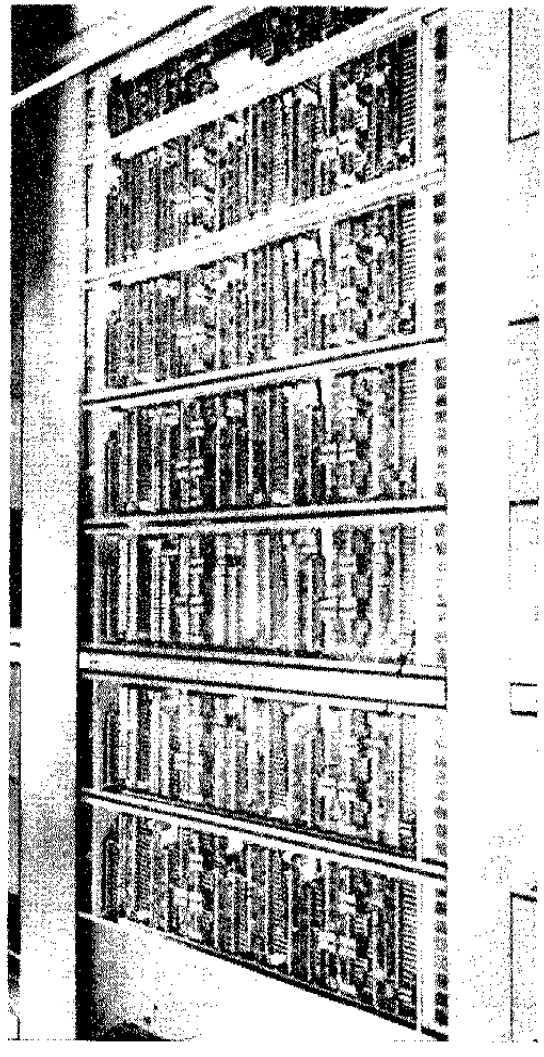


Figure 15—Bay of registers and couplers.

that their total number may be increased to 33 per section.

Figure 17 represents a group selection unit equipped with 1040 outlets. The marker relays associated with a marking distributing frame enable each outlet of the secondary selectors to be assigned to any group of outgoing lines.

The distribution of outlets of the group selection

GNF	GNF		IDF
MARKER 1	MARKER 2	12	PF SECTION 0
00	06	13	PF SECTION 1
01	07	14	PF SECTION 2
02	08	15	PF SECTION 3
S	S	S	S
03	09	16	PF SECTION 4
04	10	17	PF SECTION 5
05	11	18	19
POWER	POWER	POWER	POWER

Figure 16—Line selection unit. *GNF* is group number frame, the frames numbered 00–19 are terminal selectors. *S* is supervisory jacks and lamps, *IDF* is an internal distributing frame, and *PF* indicates call finders, fifties selectors, and entraide selectors.

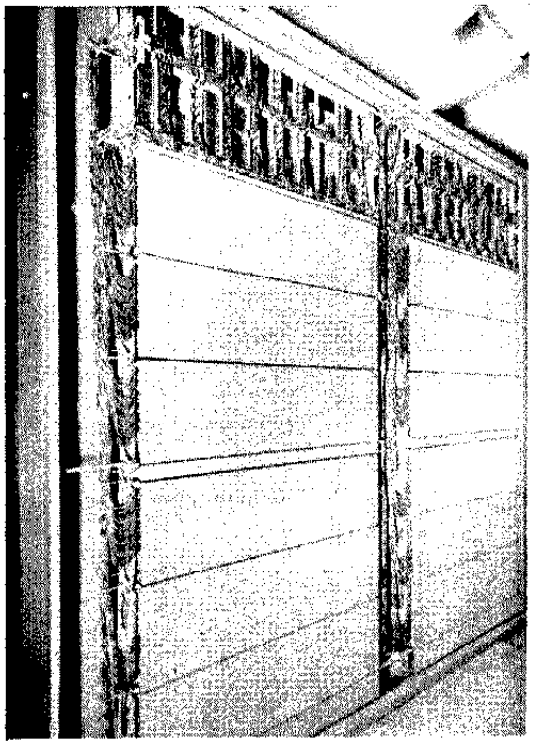


Figure 18—Bays of junctions and register finders.

Figure 17—Group selection unit. The frames numbered 00–19 are secondary selectors, *S* is supervisory jacks and lamps, *IDF* is internal distributing frame, *BP* is information bypass, *PS* is primary selectors, and *ES* is entraide selectors. The supplementary units are at the right.

	IDF	IDF	IDF	IDF	
MARKERS 1 AND 2	MARKER RELAYS 1	BP	MARKER RELAYS 2	PS SECTION 0	PS, ES
00	05	10	15	PS SECTION 1	PS, ES
01	06	11	16	PS SECTION 2	PS, ES
02	07	12	17	PS SECTION 3	PS, ES
S	S	S	S	S	
03	08	13	18	PS SECTION 4	PS, ES
04	09	14	19	PS SECTION 5	PS, ES
				PS SECTION 6	PS, ES
POWER	POWER	POWER	POWER	POWER	POWER

units between the fifties selectors of the line selection units and the junctions to other exchanges and to special services is provided for on one side of a distributing frame, the number of bays of which vary with the size of the exchange.

The distribution between the call finders and the register junctions, on the one hand, and the primary selectors, on the other hand, is effected by jumpers run inside the uprights between the upper distributing frame and the frames of junctions and register finders mounted on each bay. Figure 18 shows this arrangement.

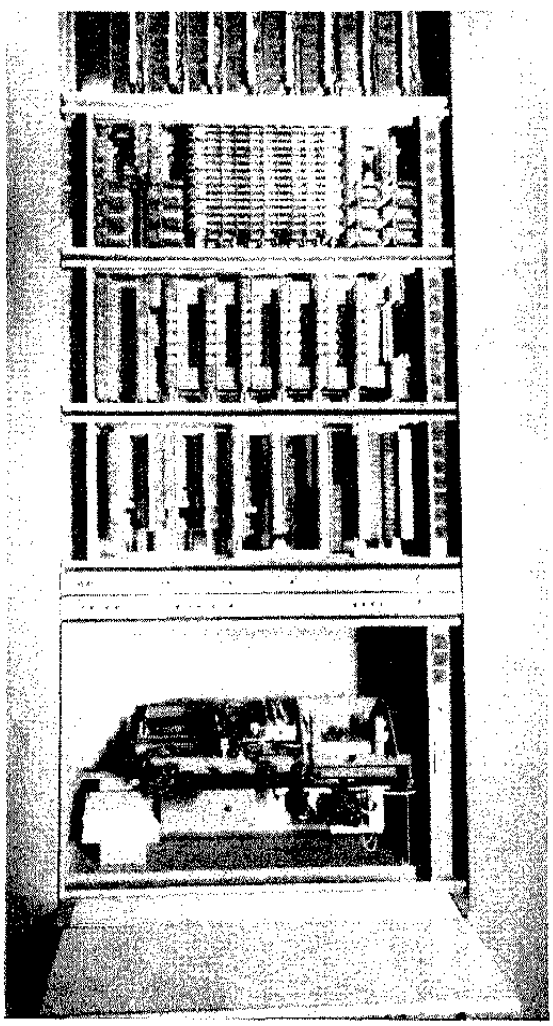
In medium-size offices (4000 to 5000 lines) use is made of group selection units equipped with only one selector stage, entraide selectors being provided in addition. The maximum capacity of the unit then is 6 primary sections with 44 switches giving access to a maximum of 240 outgoing junctions. Such units are housed in two wide-type bays 3.47 meters (137 inches) high.

The frames for information bypass are equipped with multirelay units, each having 20 spring pile-ups so as to simplify internal wiring. These relays are similar to the selectors as regards their vertical multiple arrangement (built-in) and horizontal wiring (layers of bare wires) and to the standard relays as regards their operating magnetic circuit.

The fault-recording bay shown in Figure 19 provides room for a special frame to house the card punch.

A typical floor plan of 10 000-line exchange operated as a single office with a traffic of 0.09 erlang per line is given in Figure 20.

Figure 19—Fault recordings bay with card punch and control relays.



12. Conclusions

These improvements in switching, equipment, and wiring of the Pentaconta system have augmented its original intrinsic qualities of reliability, speed, flexibility, and easy interconnection with other systems including national, continental, and intercontinental dial networks. In addition, space requirements and the time required for its installation have both been reduced.

The 850 000 lines in service or in course of completion in both private and public installations on five continents testify to the success of the Pentaconta system.

J. P. Basset was born in Lyons, France, on 16 October 1925. He is a certificated engineer from the Institut Electrotechnique de Grenoble. In 1954, he joined the Compagnie Générale de Constructions Téléphoniques to develop circuits for the Pentaconta system. He participated in a series of training lectures on Pentaconta for engineers of the International System.

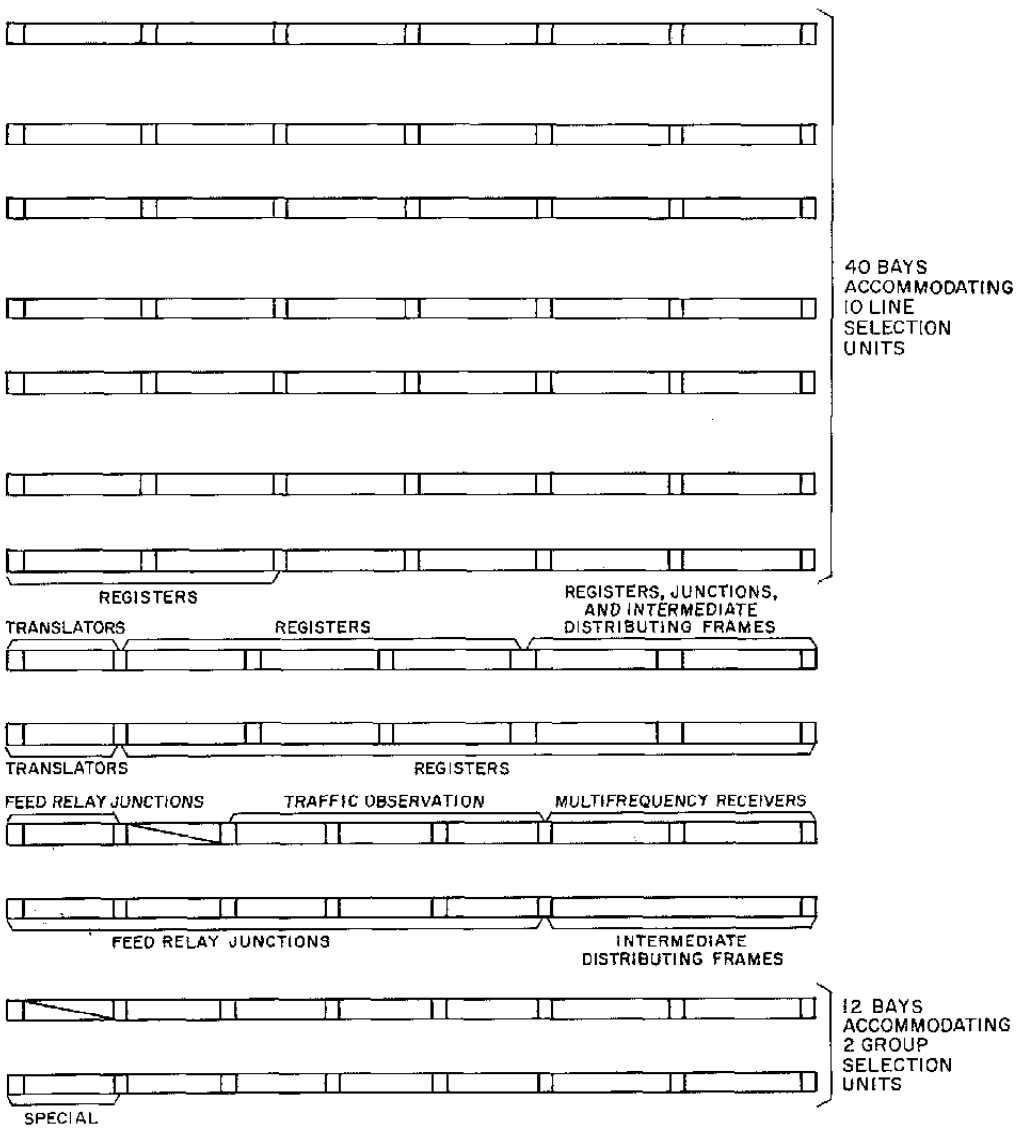


Figure 20—Typical floor plan of a 10 000-line office for a single-office area with a traffic rate of 0.09 erlang per line. The bays are 210 millimeters (8.3 inches) deep. The wider aisles between bay fronts are 900 millimeters (35.4 inches) wide and those between bay rears are 630 millimeters (24.8 inches) wide. The equipment rows are 8.8 meters (29 feet) long and the entire installation can be accommodated in a room 14 by 11 by 3.84 meters (46 by 36.1 by 12.6 feet).

Since 1961, Mr. Basset has been developing circuits for the Pentaconta *1000B*, including its adaptation to the Paris network and its international connections, to 4-wire transit exchanges, and to rural dial switching systems.

Pierre Camus was born on 18 September 1913 at Renwez, France. He received an engineering degree from the Ecoles Nationales d'Arts et Métiers, where he was awarded the silver medal in 1933. In 1934, he received the engineer diploma of l'Ecole Supérieure d'Electricité, Paris.

He served in the military from 1934 to 1936 and from 1939 to 1945.

From 1936 to 1939 and after 1945, he was with Compagnie Générale de Constructions Téléphoniques. He worked on R6 and Pentaconta switching systems. He also was responsible for training courses on Pentaconta for engineers in the International System. He is now chief of the department for Pentaconta development.

An Introduction to Pert

Pert is an acronym for Program Evaluation and Review Technique. It was developed for the management of programs involving many stages of different activities that may take various amounts of time and that are interrelated in the sense that some must be completed before others may be started.

Based on estimates of the minimum, maximum, and probable times that would be taken by each activity stage in a program, the statistical mean and variance of each activity interval are calculated. The earliest possible time at which each stage of activity can be completed is found and the critical path, that series of interrelated activities running from start to finish taking the greatest accumulated time, is determined. By working backward from the termination, the latest time at which each stage must be completed may be found and stages having excess time may be borrowed from to speed up the critical stages. The system is amenable to either manual or computer operation depending on the complexity and size of the program.

This programed learning course presents information, poses a problem based on it, provides several possible answers, and refers the reader to a different page for each answer. If the selected answer is incorrect, the referenced page will state why it is wrong and will redirect the reader to the problem to attempt a new answer or, if the answer is correct, the next item will be introduced.

This programmed learning course is presented on 140 pages measuring 7¾ by 10½ inches (20 by 27 centimeters) that are held between plastic covers by a plastic comb passing through perforations along one edge of each sheet.

It is available from Federal Electric Corporation, 621 Industrial Avenue, Paramus, New Jersey, at $5.95 per copy.

Pentaconta Line Concentrators

A. J. HENQUET
Le Matériel Téléphonique; Paris, France

A concentrator is a switching device used in telephone networks to connect a number of subscribers located in the same area to the parent exchange over a relatively small number of common trunks accessible to all of them when originating or receiving a call.

In the original conception, concentrators were exclusively associated with the subscriber line networks. They are intended to reduce the number of pairs of wires required to connect a group of subscribers' lines to the exchange. They are not part of the exchange switching

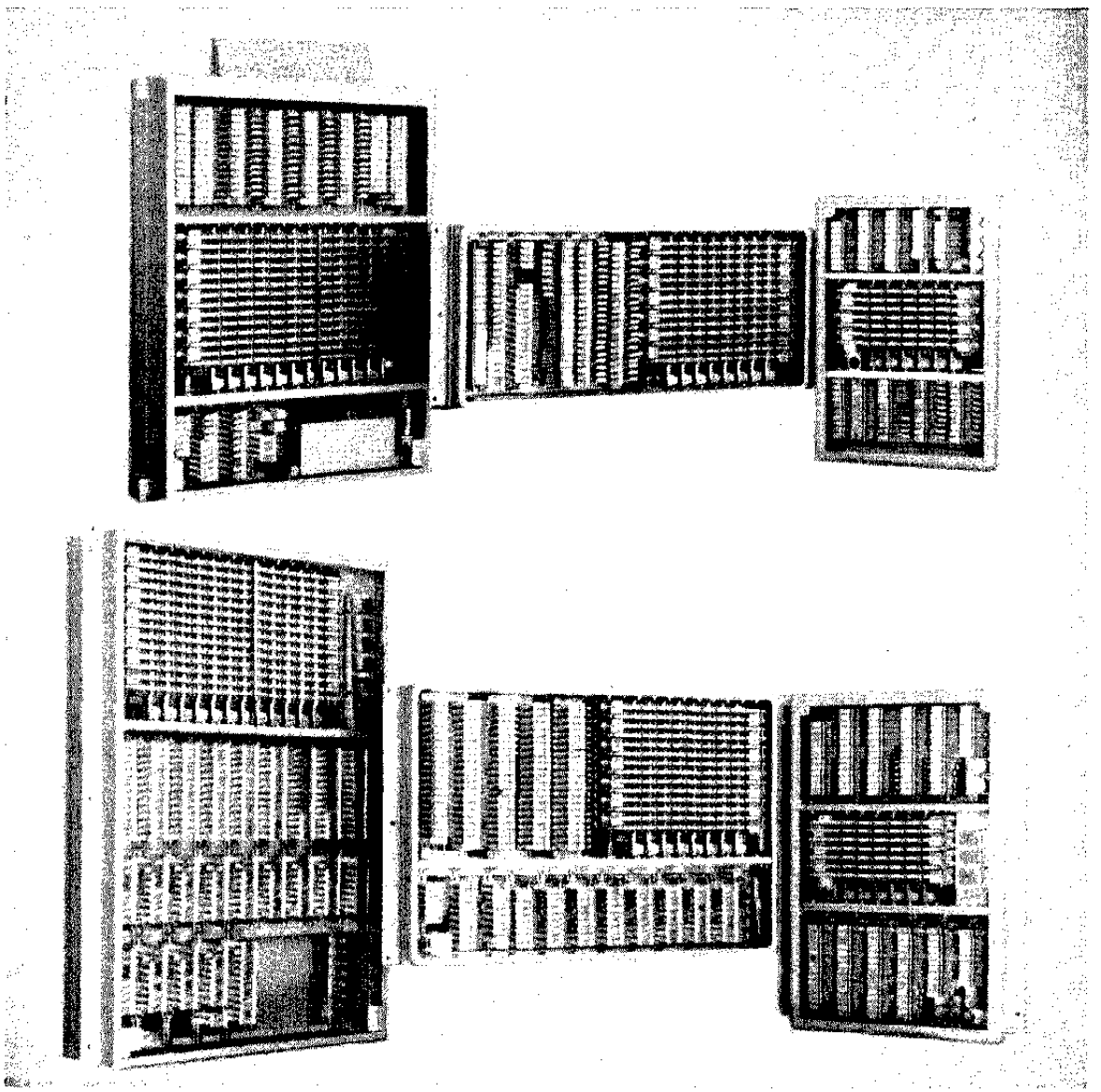


Figure 1—Above are the subscribers units and below the exchange units for, from left to right, the 52 × 12, 52 × 8, and 24 × 6 concentrators.

system through which a particular subscriber's line is selected for service.

A concentrator consists of two switching units, one at each end of a group of common trunks. One unit is in the central office and the other is at the geographical center of the group of subscribers. Each switching unit concentrates and redistributes all incoming and outgoing calls among the group of subscribers. Relays, step-by-step, or crossbar selectors may be used.

Concentrators are no novelty, but the old installations using relays or step-by-step selectors either were too bulky or required maintenance that was uneconomical because the switching unit was distant from the exchange. The development of crossbar multiselectors has renewed interest in concentrators.

1. Pentaconta Concentrators

Pentaconta switches are especially suitable for concentrators [1]. The 14-selection-bar multiselector, using one of its 14 selection bars as a multiplying bar, can serve 52 outlets. Two typical designs have been standardized to concentrate the traffic from 52 subscribers over 8 or 12 speech trunks.

A third version uses the 7-selection-bar Pentaconta multiselector with one bar for multiplying to serve 24 subscribers over 6 speech trunks.

Figure 1 shows the 3 types of Pentaconta concentrators. Their characteristics and sizes are given in Table 1.

Figure 2 represents the junction diagram of a Pentaconta crossbar concentrator connecting 52 subscriber lines to a rotary exchange. The same individual subscriber line equipment is used either for the concentrator or when connected directly to the rotary equipment. It must be pointed out that the third or sleeve wire of the subscriber line circuit must also be connected to the main distributing frame. The line equipment is connected to the concentrator by a jumper at the main distributing frame. The 3 wires of each subscriber line circuit are connected to one of the horizontal levels of the crossbar multiselector of the concentrator connecting unit in the exchange. At the remote unit of the concentrator, the corresponding level of the crossbar multiselector is connected to the 2-wire line to the subscriber's set.

A 2-wire trunk connects, 2 by 2, the corresponding verticals of the two crossbar multiselectors. A control circuit at each end determines the setting of its multiselector and the two control circuits are interconnected by 2-wire control junctions. There are 2 of these control junctions, except in the most recent designs, where only one is required. The orders enabling the control circuits to set the multiselectors are transmitted through these junctions.

TABLE 1
CHARACTERISTICS OF CONCENTRATORS

	52 × 12	52 × 8	24 × 6
Subscribers	52	52	24
Trunks, Traffic	12	8	6
Trunks, Control	2	2	1
Traffic in Erlangs	5.4	2.6	1.6
Subscriber Units			
Inches	39 × 27 × 9	38 × 16 × 7	25 × 17 × 8
Centimeters	97 × 67 × 22	96 × 39 × 18	62 × 43 × 20
Pounds	243	154	84
Kilograms	110	70	38
Exchange Units			
Inches	42 × 30 × 9	38 × 24 × 10	30 × 19 × 8
Centimeters	107 × 77 × 22	96 × 59 × 24	75 × 49 × 20
Pounds	254	198	128
Kilograms	115	90	58

2. General Characteristics

The subscriber's set that is served by a concentrator is connected to the exchange by the same kind of a 2-wire line used for ordinary subscribers.

No supplementary resistance or impedance is introduced either in series or shunt of each wire and ground for connection to the concentrator. Thus the connection established by a concentrator has the identical electrical properties of a normal line connected directly to the exchange and any variations of its electric state may be utilized by the switching system in the exchange including recognition of the different classes of service.

All classes of service may be provided through a concentrator because, up to the main distributing frame, the exchange sees only the normal line equipment. Thus, subscribers grouped on one concentrator can be assigned any directory number. This permits assigning a subscriber to a concentrator without changing his number. Message registers, if used, are connected in normal fashion in the parent exchange.

The number of common speech trunks linking the distant part of a concentrator to its exchange is determined by allowing for a probable number of lost calls equal to or less than that for the terminal selection stages; the reduced

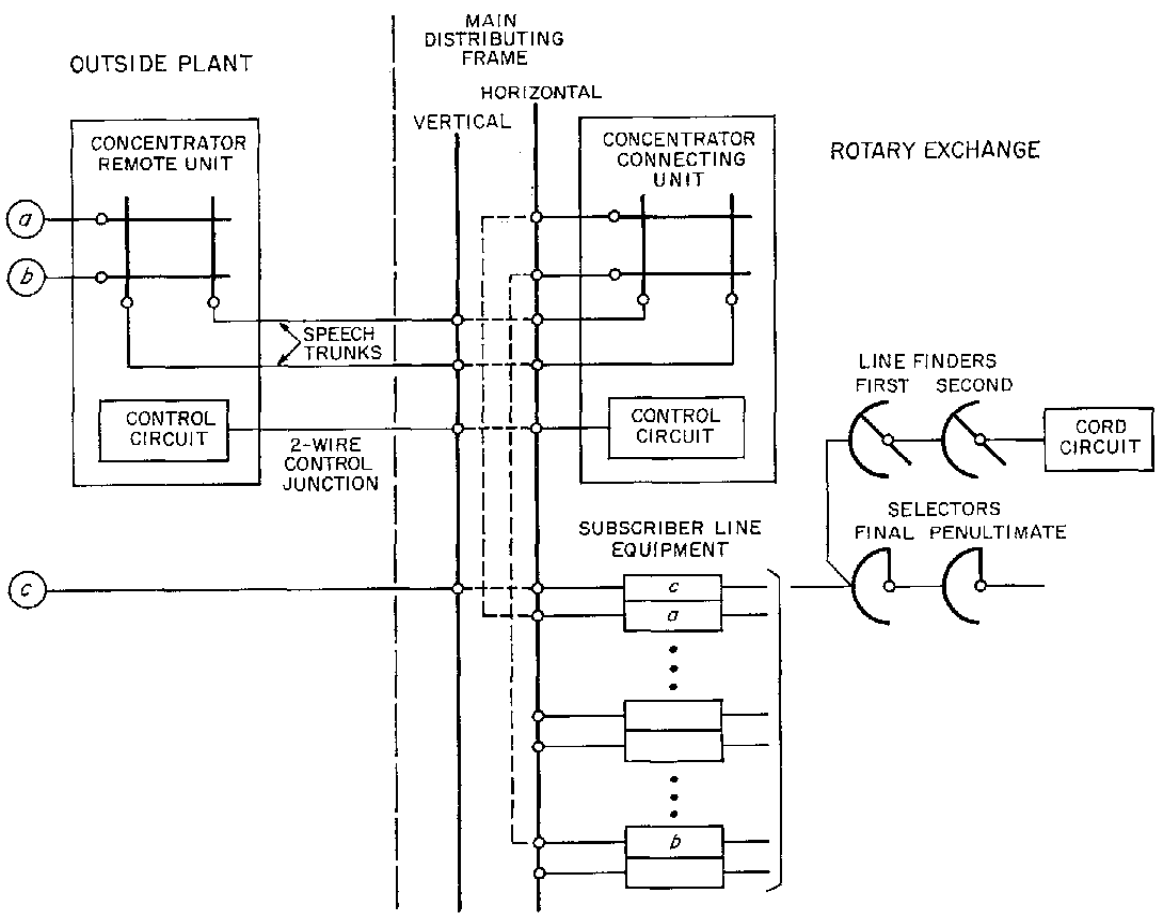


Figure 2—Junction diagram for Pentaconta concentrators working into a rotary switching system.

TABLE 2
TELEPHONE SUBSCRIBERS AND TRAFFIC

Type of Line	Percentage of Total Lines	Percentage of Total Traffic	Average Traffic per Line in Erlangs	Average Use per Day in Minutes
Residential Subscribers	61	15	0.01	7.5
Business, Single Lines	33	37	0.045	30
Private Branch Exchanges	6	48	0.3	100
Total	100	100	0.04	—

quality of service resulting therefrom is not noticeable. This number of trunks is usually larger than would be provided for a small rural exchange of the same capacity.

Finally, the time between the moment a subscriber lifts the receiver and the closing of the loop in the exchange is approximately 800 milliseconds, which also holds for extending a call from the exchange to the subscriber.

Subscribers connected to an exchange through a concentrator receive the same class and quality of service as other subscribers. There is no reason to apply a special subscription rate for this type of connection as the subscribers need not even know about it.

There are several characteristic differences between concentrators and small rural exchanges. In the latter, switching devices participate in the preselection and selection operations of the group to which they are connected. They provide for microphone current to the subscriber's set, for dial-pulse repeating, and for transmitting ringing current. When needed, message registers are generally located in the rural exchange. The subscribers are assigned consecutive directory numbers unless the rural exchange is further equipped with translators of subscriber numbers. Finally, if local connection circuits are utilized, registers or similar simplified circuits must be provided.

The concentrators do not perform any of these functions; their role is limited to extending a 2-wire line from the subscriber's set to the exchange for the duration of each call. Subscribers served by the same concentrator may be interconnected over 2 trunk lines in the concentrator.

3. Economic Factors

The question of under what conditions concentrators prove to be economic has been studied and discussed extensively without providing any definite answer. There would be no value in adding a new study to those already made; it will be sufficient to note that the use of concentrators tends to increase where they are already in service and to be introduced elsewhere.

There are several reasons for the increasing use of concentrators. In countries having high telephone development, the number of lines continues to increase as new categories of subscribers appear. Initially developed for business, the telephone has entered into social and family circles. The number of so-called residential subscribers is increasing and becoming preponderant. Consequently, the number of lines is increasing but, as a counterpart, their utilization rate is decreasing.

Figure 3, which has been prepared from European telephone statistics, illustrates this tendancy. Table 2 gives the breakdown of subscriber lines with regard to traffic and rate of utilization; the figures shown are averages taken from statistics; they can vary rather widely from one country to another.

However, it is relatively expensive to connect a subscriber to an exchange; as is evident from Table 3, which shows a distribution of costs

TABLE 3
BREAKDOWN OF CAPITAL INVESTED IN A TELEPHONE NETWORK

	Percent
Subscriber's Installation Including Subset	9
Subscriber Lines	60
Switching	25
Trunks	6
Total	100

for an automatic telephone exchange and its network.

The connecting of subscriber sets to the exchange represents the largest part of the investment. Considering the rate of utilization of residential lines, there should be intense interest in common connecting devices such as concentrators for groups of subscribers whose total traffic is small and who are located far from the exchange.

A rough estimate shows that if the distance between a group of 56 subscribers and the exchange is approximately 1.55 miles (2.5 kilometers) the cost of material used to connect them is the same, whether one uses 2 overhead distributing cables with 28 pairs of 0.5-millimeter diameter wires or one overhead cable of 14 similar pairs connected to a 52-line 8-trunk concentrator. In this example, means of supporting cables are assumed to exist and the cost of labor to install and lay the cables are considered to be of the same order. In the concentrator case, 4 subscribers have direct connections to the exchange.

Thus, beyond this distance, concentrators can provide an economical means of making connections. Such distances are encountered in rural or provincial areas, in which one exchange serves subscribers in a small town and in the surrounding country.

The usefulness of concentrators is clearly evident in suburban areas around large cities as well as in the urban areas where underground cables are mostly used. In some of these areas,

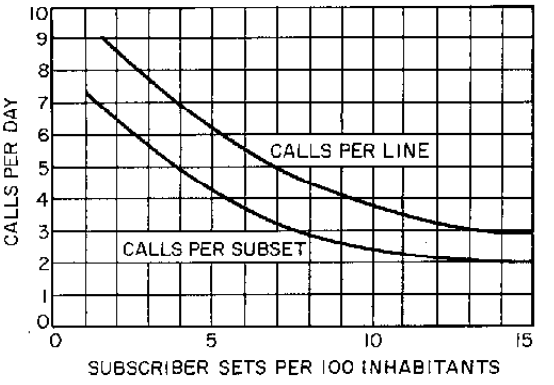


Figure 3—Average number of calls per day for each line and for each subscribers' set as a function of the number of subscribers' sets per 100 inhabitants.

population increases over a few years exceeds the telephone provisions previously thought to be adequate, residential districts rise and grow, and others similarly undergo unanticipated changes in their characters and population. It may not be possible to finance the extension of cables beyond the original plans, particularly if such needs arise at the same time in several places.

In such cases, concentrators give the possibility of creating new spare lines in overcrowded networks by regrouping the traffic of a group of existing residential subscribers on a part of their previous lines used as common trunks.

Savings through concentrators can be evaluated by referring to Table 3. The costs of certain items, switching for instance, are well known in every country. By applying the percentage shown in Table 3, it is easy to calculate the average cost of an average-length subscribers' line. It is interesting to note that the figures obtained in this way provide a good evaluation in France.

When a concentrator is used, the length of the line to be saved is generally greater than average. Thus, the cost of a concentrator and its installation is only a small part of the amount represented by 42 times the average cost. Moreover, the concentrator can be built immediately,

which may not always be possible for an extension of the cable network.

On the other hand, it seems difficult to justify the use of concentrators when setting up a new underground network or a new underground cable.

For such installations, the investment in technical studies and in laying cement ducts in the cluttered-up underground of cities or suburban areas exceeds the price of the cable itself. Therefore, the capacity of these ducts and cables is chosen to allow for expansion for many years to come. When these cables are new and all their spare lines are still available, concentrators would provide only additional spare lines.

To summarize, concentrators might economically be used in new networks employing overhead lines and cables in areas having low population density. This is an excellent and inexpensive way of developing new telephone networks in rural or provincial areas. In places having a high population density and where an underground cable network already exists, concentrators mainly are a means of providing immediately for new subscribers. This allows time for selecting the best plan under which the cable network can be extended, thus combining economic management with immediate service to new subscribers.

4. Other Uses for Concentrators

A judicious use of concentrators must take into consideration the total traffic, that is, the average incoming and outgoing traffic, of the subscriber group to be served.

In the preceding section, the expression "residential subscriber" does not mean that subscriber lines for professional use cannot be served by a concentrator; it means only that this type of connection should preferably be reserved for lines having a low utilization rate, that is, comparable to residential subscriber lines.

As there are 8 or 12 speech paths on a 52-subscriber-line Pentaconta concentrator, on the basis of a probability of lost calls equal to 0.005, the total traffic handled during the busy hour by the common trunk group can reach 2.73 or 5.28 erlangs, respectively, which corresponds to an average traffic rate per line of 0.05 or 0.1 erlang. These last figures are not absolute since they are average values and are meant to determine only an order of importance. The traffic on the different lines of the group can vary within rather large limits. However, too heavy a traffic on a line or group of lines may occasion some inconveniences.

These inconveniences can be illustrated by the following example. Let us imagine a group of 50 lines, 4 of which serve a private-branch-exchange group handling approximately 100 incoming and 100 outgoing calls each day. The total traffic of the 50-line group, including the 4 lines of the private-branch-exchange group, is supposed to remain below 5 erlangs, but the traffic of the 4 private-branch-exchange lines during busy hours is more than 2 erlangs. This 50-line group could be served by a 52-line 12-trunk Pentaconta concentrator, but it is much more advantageous to connect the 4 private-branch-exchange lines directly to the central office and to serve the other 46 subscribers by a 52-line 8-trunk concentrator. In fact, in both cases, 14 pairs are necessary for the connection, but in the second case, less money will have to be invested, current consumption and wear of equipment will be lower, and service, particularly that of the private-branch-exchange group, will be of better quality.

Therefore, it is recommended that subscriber lines served by a concentrator have a busy-hour traffic appreciably lower than that of the speech trunks of this concentrator, that is, appreciably lower than 0.34 erlang for the 52-line 8-trunk concentrators and 0.44 erlang for the 52-line 12-trunk concentrators.

The rate of utilization of subscriber lines is often not well known and a check of the number of calls recorded by message registers can give a false idea of their actual traffic. Therefore, if a concentrator is too frequently overloaded,

manual or automatic monitoring of the subscribers will determine which lines should preferably be connected directly.

The characteristics of the lines connecting subscriber sets to the remote unit of the concentrator are important. The line relay in the remote unit of the concentrators functions correctly even if the insulation of a subscriber line drops to 20 000 ohms. Theoretically, each subscriber line could present an insulation of 20 000 ohms without affecting the functioning of the concentrator. If this is the case, the battery in the remote unit of the concentrator would have to supply 2.16 more ampere-hours per day to cope with the current leakages occasioned by the poor condition of the lines. The capacity of cadmium-nickel watertight battery in the concentrator would not be sufficient and the charging current it receives through the idle speech trunks is not high enough to handle this additional drain.

If the subscriber connections to the remote unit of the concentrator cause significant leakage, it is necessary either to provide a special pair of conductors from the exchange to supply additional charging current to compensate for leakages, or to provide for a power supply independent of the concentrator, such as 3 motor-car type 12-volt batteries charged from the city electric mains.

5. Installation

The installation of the two units making up a concentrator cannot be handled in the same way. One unit is installed at the exchange and the remote unit is located so that the connections to the individual subscribers will be as short as possible.

In rural areas, it will often be possible to install the concentrator in a building in the community where the subscribers to be served are located and which is already utilized by the operating company. Otherwise, the concentrator will be installed outdoors in a metal cabinet to protect it from weather hazards. The cabinet will be mounted on a concrete base or, where open wire is used, on one of the poles of the network as is frequently done in the United States. Figure 4 shows an outdoor cabinet used by the French administration.

In urban and suburban areas, the pattern of subscriber cables often determines the position of the remote units. In France, for example, the area served by an exchange is divided into zones. Each zone is connected to the exchange by one or several cables with a capacity of 112 or 224 pairs. These cables lead to the entrance to the distributing zone called a junction manhole. Between the exchange and the junction manholes are the main cables. The cables beyond the manholes, called distribution cables, have a decreasing capacity starting from the junction manhole, the smallest cable utilized having a 14-pair capacity. All splices are lead covered.

With these arrangements, concentrators can be equipped near the junction manhole or at the end of the distribution cable.

In the first case, the concentrator makes it possible to delay the extension of the main cables. In the second case, the concentrator also increases the capacity of the distribution network.

When the remote unit of the concentrator is installed near the junction manhole, it can be located in the same building as the manhole if

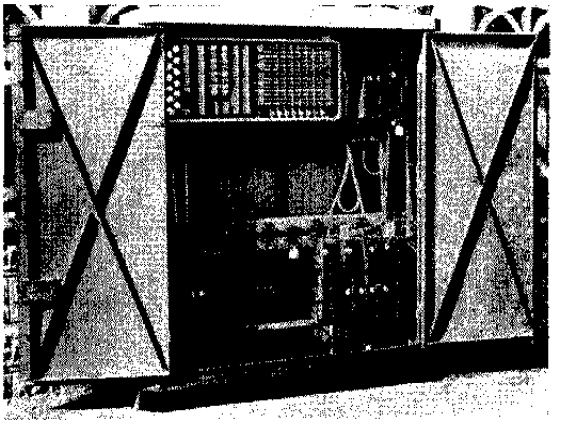


Figure 4—French outdoor cabinet for the remote unit of the concentrator.

the latter is located in a company building. If not, the concentrator is mounted in an outdoor metal cabinet.

When the concentrator is at the end of the distribution cable, it may be installed for example in the basement of the residential buildings where the subscribers reside.

At the exchange, the units are generally on bays in the room housing the main distributing frame. Figure 5 shows an example of this type of installation.

The necessity for connecting units in the exchange certainly limits the use of concentrators. There is usually not much space in the room of the main distributing frame for additional equipment. Unused space in an exchange is scarce or reserved for planned extensions.

6. Concentrators Can Lead to Remote Selection

One may gather that concentrators offer only a provisional solution to network problems. Several writers are of this opinion. Without being so categorical, one must admit that the connecting units increase to an important degree the amount of automatic equipment in the exchange. If only the remote unit of a concentrator were sufficient, this important objection would be removed and the economic advantages of concentrating the traffic of a number of subscribers on a common trunk group would remain. The selecting line unit of the Pentaconta system permits such operation of concentrators. It is possible by using special terminal selection frames to do away with the exchange connecting unit of the concentrators without changing the general operation of the selecting unit and particularly without modifying the marker or the control process.

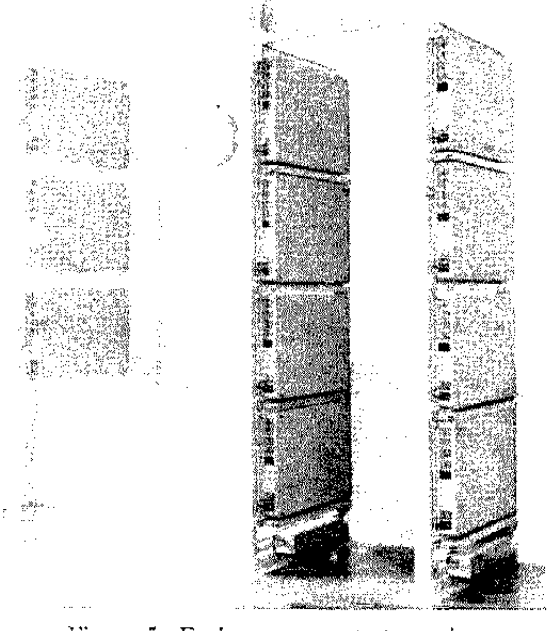


Figure 5—Exchange concentrator units.

In the Pentaconta line selection unit, the normal terminal selection frame includes, in addition to the multiswitch used as final selector, a certain number of relays; such as, line and cut-off relays of the 52 subscribers, selecting relays, and connecting relays to the marker. These are replaced by a special terminal selection frame that will also include the multiswitch used as final selector, the selecting relays and connecting relays to the marker, the cut-off relays of the 52 subscribers, but the line relays will be in the remote unit of the concentrator. Instead, this special frame will include the identifying matrix, common control circuit, pulse generator, and decoding relays that in conventional concentrators are part of the exchange unit.

As the number of relays on this special terminal selection frame is 50 percent larger than that for the normal frame, this new frame is slightly bulkier.

The multiswitch used as a final selector makes it possible to determine at the exchange whether subscribers' lines are busy or idle; it also allows for the connection of message registers and for dealing with the subscriber's category wire as used in the Pentaconta system. The line wires are normally multipled on the switches, but it seems useless to have them connected to the

main distributing frame. However, the presence of these line wires in the exchange solves a few problems such as listening in on a busy line, use of message registers at subscriber's premises, et cetera.

Where the 24-line concentrators are used in this arrangement, it is possible to make a special terminal selection frame working with 2 remote 24-line 6-trunk concentrator units.

Although this design of concentrators offers the same operating features as those described, the directory numbers are not so completely independent of the equipment location numbers; the subscriber can be given any number in the thousands group of the line selection unit.

Finally, the cost of such a connection will be half that for a concentrator using exchange connecting units.

A great flexibility results from the fact that the remote units are identical in both types of Pentaconta concentrators. Conventional concentrators can be utilized either to add new subscribers or to meet unforeseen expansion. They may be replaced later by remote selection units.

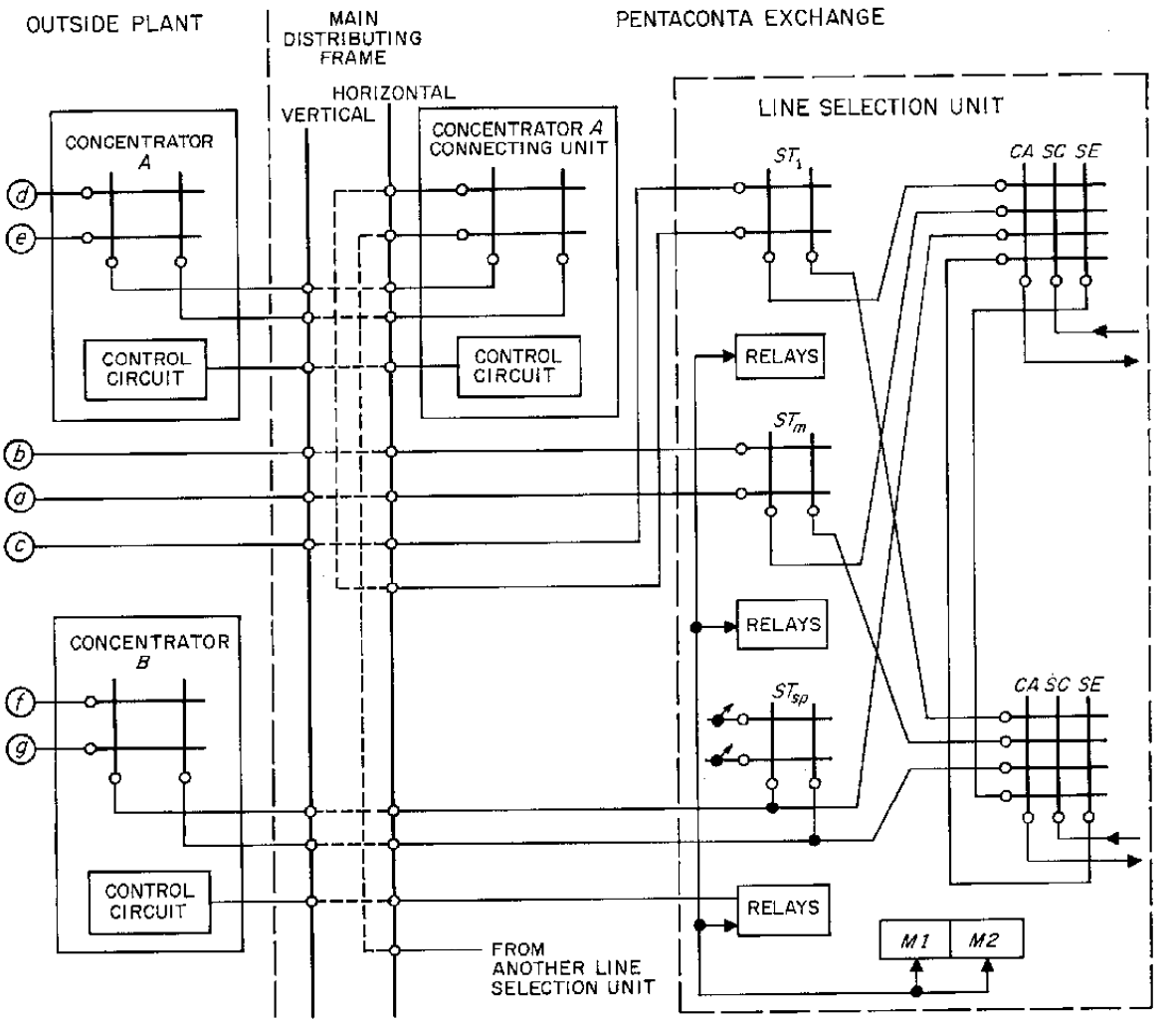


Figure 6—Conventional concentrator and remote terminal selector connected to a Pentaconta exchange.

The exchange connecting unit can then be used for another group.

Figure 6 shows the junction diagram of a Pentaconta line selection unit to which are connected normal subscriber lines *a*, *b*, and *c*; subscriber lines *d* and *e* served through the conventional concentrator *A*; and finally subscriber lines *f* and *g* served through concentrator *B* used as a remote terminal selector.

Further, Pentaconta concentrators of a capacity beyond 52 lines, capable of serving up to 1000 subscribers' lines, are available. These concentrators are remote line selecting units and are beyond the limits of this article.

7. New Uses of Concentrators

At the very beginning of the study on concentrators, their use with carrier systems was envisaged. It is tempting, of course, to reduce a common trunk group to the single pair of physical conductors needed by a carrier system.

Such utilization is rather far from the original concept of concentrators. The problem is no longer that of providing an entirely metallic path between the subscriber's set and the exchange. A method of supplying microphone and ringing currents, and of repeating dial pulses must be introduced in each speech trunk or channel between the remote end of the concentrator and that of the carrier system. These power supply and repeating bridges can, of course, be part of the terminal equipment of the carrier system. The signalling system used by the carrier current must be able, on the one hand, to transmit the ringing current for calls received by the subscribers served by the concentrator and, on the other hand, to repeat the dial pulses for calls originated by the subscriber. Finally a signalling system as simple as possible must be designed so that the necessary information for control of both ends of the concentrator may be transmitted in both directions.

This type of system, known as "rural carrier system," is rather expensive and its use can be justified only under special conditions; the association of a concentrator with this type of carrier system will increase its field of application.

Prototypes of adaptors based on such an association have been built and tested in the laboratory; field tests are now being undertaken by the International System in the United States and in Italy.

8. Conclusion

Pentaconta concentrators are extremely useful devices for the expansion of a subscriber network. They offer an economical solution to many problems mainly in the development of new networks and the rapid establishment of temporary telephone lines such as those used for exhibitions, meetings, et cetera. Above all in suburban areas they provide a rapid and economic way of meeting unforeseen demands for telephone service by improving the utilization rate of existing cables.

The separation of speech paths and control circuits simplifies the operation of Pentaconta concentrators in association with carrier systems. This leads to future applications of concentrators to connect existing subscribers with conventional subsets to possible time-division-multiplex electronic exchanges of the future.

9. References

1. J. P. Basset and P. Camus, "1000 B Pentaconta System," *Electrical Communication*, volume 38, number 2, pages 196–212; 1963.

2. "Distribution of Telephone Traffic in a German Town with 700 000 Inhabitants," Professor Lennertz, Technical University, Aachen, 1956.

3. "Telephone Policy: The Next Steps," Presented to Parliament by the Postmaster General, Her Majesty's Stationery Office, CMND 436; May 1958.

A. J. Henquet was born in Paris, France, on 15 December 1906. He graduated as a certified engineer in electromechanics in 1926 from the Breguet school.

In 1927, Mr. Henquet joined Le Matériel Téléphonique as a circuit designer and is presently chief engineer of switching developments for the telephone branch.

Applied Cryogenic Engineering

The University of California in Los Angeles presented a graduate-level series of lectures on the properties, characteristics, and utilization of liquefied gases as they are used in space systems. The contents of these lectures will be useful in the development of advanced propellant systems and in other fields involving cryogenics. A compilation of these lectures, edited by R. W. Vance of the technical staff of Aerospace Corporation and W. M. Duke, president of ITT Federal Laboratories, is included in this book. The book is divided into the following 15 chapters and 3 appendixes.

CHAPTERS

1. Basic Theory; R. W. Vance, Aerospace Corporation
2. Cryogenic Fluids—Their Properties and Technology; R. W. Vance and C. H. Reynales, Space Technology Laboratories
3. Mechanical Properties of Materials; R. H. Kropschot, National Bureau of Standards
4. Low-Temperature Thermometry; K. D. Timmerhaus, University of Colorado
5. Fluid Flow and Heat Transfer; K. D. Timmerhaus
6. Low-Temperature Insulation; R. H. Kropschot
7. Liquefaction of Oxygen, Nitrogen, and Hydrogen; A. J. Westbrock, Linde Company
8. Helium Liquefaction—Cryogenic Equipment Applications; A. J. Westbrock
9. Storage and Transfer of Cryogenic Fluids; P. D. Fuller and J. N. McLogan, both of Stearns-Roger Manufacturing Company
10. Safety Aspects of Cryogenic Systems; C. McKinley, Air Products and Chemicals, Inc.
11. Explosion Hazards in Liquid Bipropellants; M. A. Cook, University of Utah
12. Helium: V. Arp, Boulder, Colorado; and R. H. Kropschot
13. Storage of Cryogenic Fluids in Space; M. Adelberg, Space Technology Laboratories
14. Application of Cryogenic Rocket Propellants to Space Vehicles; S. G. Rumbold, Space-General Corporation
15. Future of Cryogenic Fuels for Space Systems; R. D. Long, Aerospace Corporation

APPENDIXES

A. Properties of Cryogenic Fluids
B. Processing Helium from Natural Gas
C. Contamination Control in Cryogenic Fluids and Systems; Major G. J. Murphy and Captain P. O. Pearce, Air Force Ballistic Systems Division

The book is 9 by 5¾ inches (23 by 15 centimeters) and contains 510 pages including an 11-page index. It is published by John Wiley and Sons, 440 Park Avenue, New York 16, New York, at $17.50 per copy. It is also available from John Wiley and Sons Limited, Glen House, Stag Place, London, S.W. 1, England at £6-11/9d.

Telegraph Transit Exchange Using Crossbar Pentaconta Design

PIERRE SAMUEL
Compagnie Générale de Constructions Téléphoniques; Paris, France

This transit exchange employing crossbar Pentaconta equipment[1,2] was produced for the Aerial Navigation Department and installed in the central telecommunications north office at Athis-Mons near the Orly airport that serves Paris. Cut over in June 1961, it is a part of the Aeronautical Fixed Telecommunications Network.

In this international network, the telegraph circuits ensure interconnection of the different airports of the world and communication between each airport and services such as for weather forecasting and regional control.

The messages are generally concerned with safety flight plans, meteorological information, and instructions from regional control centers.

Service information is also transmitted and all messages may be directed to more than one party.

In the aeronautical fixed telecommunications network, the circuits employ either voice-frequency wire or radio channels; they are generally long and high in cost. This requires that they handle a maximum of traffic. This is obtained by separating the two channels of each connection to provide for duplex operation.

The preceding considerations lead to the adoption, for a transit exchange, of a method of operation with registration of incoming messages and retransmission over the channel or channels concerned as soon as available. Each incoming channel will then be permanently connected to a receiving device.

1. Transit Exchange

The incoming lines of a transit exchange are connected to receiving devices such as printing reperforators and messages are transmitted over outgoing lines automatically.

In the Orly transit exchange, the tape is perforated without being torn, which avoids its possible mutilation and loss in being manually transferred to an outgoing-line transmitter.

The tape from each incoming reperforator goes directly to an automatic transmitter having access to all outgoing lines. Sufficient space between the reperforator and the transmitter permits an operator to read the routing indications and make the required selection of outgoing line. This is, therefore, semiautomatic continuous-tape operation.

2. Format and Procedure

The continuous-tape method requires a standard composition of each message, which must consist of a heading, routing, origin, text, and end signals.

The heading is provided as a check on the transmission between the two exchanges. Consequently, it is cancelled on reception and replaced by a new heading for the next transmission.

There are only two types of messages: normal messages, which are the business messages for which the system was installed, and checking messages, which are sent at set hours over each outgoing circuit if no normal message is being transmitted.

The normal message and procedure is as follows.

(A) Start Impulse—This causes the incoming reperforator in the receiving office to start.

(B) Heading—The identification element of a message, usually a serial number for that particular transmission.

(C) Routing—The address or addresses to which the message is directed.

(D) Origin—The indication of the origin of the message and the priority alarm.

[1] F. Gohorel, "Pentaconta Dial Telephone Switching System," *Electrical Communication*, volume 31, pages 74–106; June 1954.
[2] "M. R. Legare, 'Etude détaillée des circuits des autocommutateurs Crossbar Français: Premiere partie, LE PENTACONTA," Posts and Telecommunications, Paris; 1960.

(E) Text.

(F) End—Indicates the end of the transmission of a message and prepares the reception equipment for the following incoming message.

(G) Tape Advance—This permits separation of messages and is used only for transmission to a torn-tape transit exchange or a local exchange.

The checking messages are transmitted automatically by the outgoing line equipment at determined time periods, say, every 20 minutes, if there is no transmission in process. They are sent to check the correct functioning of the equipment and the condition of the channel. The message consists of the heading, the letters *CK*, and the end.

3. Operation of the Exchange

The operation of the exchange will be described for transit of a message to a single outgoing wire channel, to a single outgoing radio channel, to several outgoing channels, and the case of a message with a faulty address. Reference will be made to Figure 1.

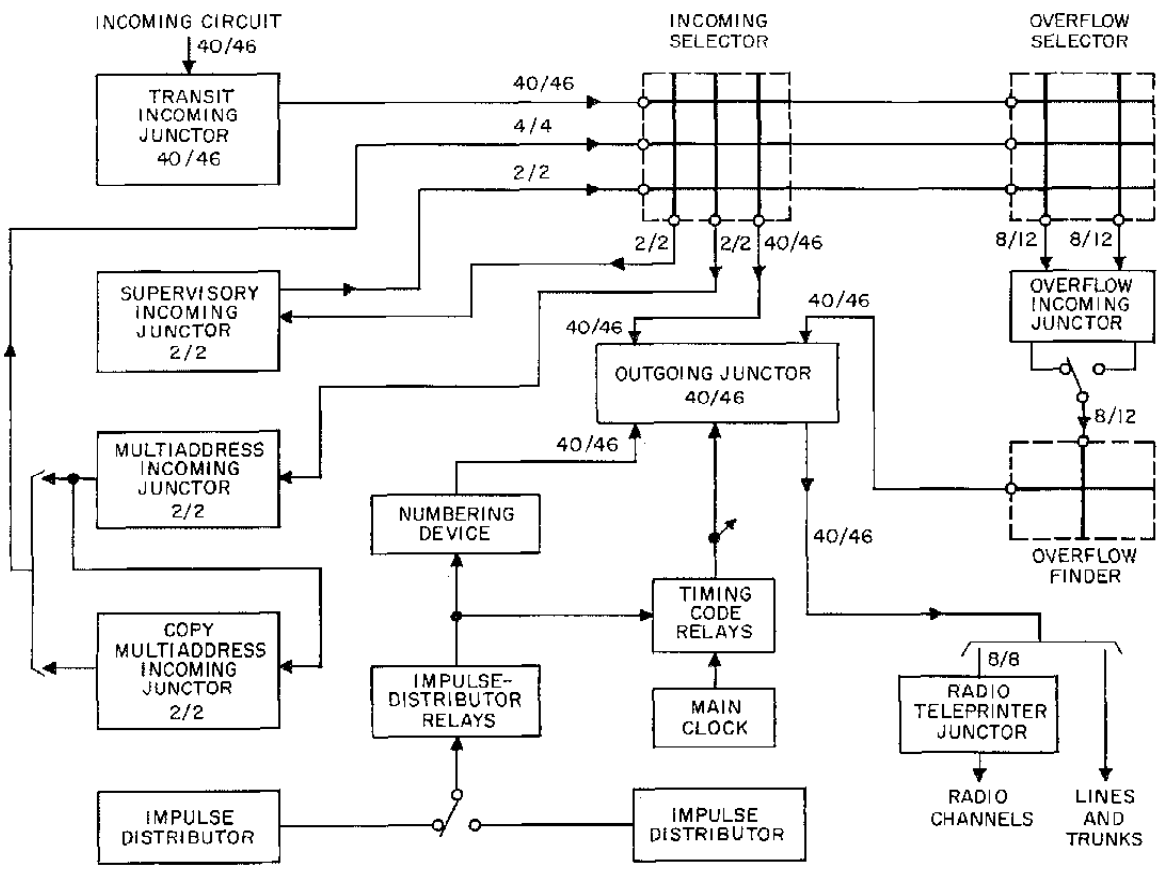


Figure 1—Trunking diagram for Orly telegraph transit exchange. The numbers indicate the initially equipped and ultimate capacities of the various elements. A typing reperforator and transmitting distributor is associated with each incoming junctor except that two such equipments are provided at the overflow incoming junctor.

3.1 Available Outgoing Wire Channel

The message is received on the typing reperforator at the incoming transit junctor, the perforated tape being threaded directly to the associated automatic transmitter. As the tape travels along, the operator checks the serial number in the heading, which under control of the operator is not transmitted.

The operator then reads the address and operates control buttons to select the outgoing channel. When the connection is made, the incoming transit position is connected through the incoming selector to the outgoing junctor. The outgoing junctor will send a seizure signal to the next exchange. After a timed-delay for starting the distant receiver motor, the outgoing junctor, with the aid of the numbering device, the timing-code relay set, and the impulse distributor relays, send a new heading for the message. The end of this transmission is signalled by the outgoing junctor to the incoming transit junctor as an order to transmit the message. The end-of-message signal is detected by the transmitter at the incoming transit position, which then initiates the release of the transit position and the outgoing junctor.

3.2 Available Outgoing Radio Channel

If the message must be routed over a radio channel, it is reperforated before going out of the exchange. Consequently, the heading sent by the outgoing junctor, as well as the message, are received on a typing reperforator at a radio teleprinter position, from which the associated transmitter distributor sends it to the radio transmitter. This storage at the exchange is necessary because of the uncertainties to which radio channels may be subjected.

3.3 Busy Outgoing Channel

If the selection ordered by the operator at the transit incoming junctor is not available through the incoming selector, the call will go to the overflow position through the overflow selector. This automatically operated position is equipped with two sets of typing reperforators and transmitter distributors under control of an overflow incoming junctor. The message is then transmitted from the transit incoming junctor to one of the overflow typing reperforators, which stores it until the required outgoing junctor is available. The overflow position is connected to the outgoing junctor through the overflow finder.

3.4 Fully Busy Outgoing Channel

If no overflow position is available or if the overflow position assigned to the outgoing junction is busy, after automatic reselection, the message is transmitted to a multiple-address position from the transit incoming position through the incoming selector. This is a form of storage to avoid overloading the transit incoming position. The message is then retransmitted by the operator of the multiple-address position.

3.5 Several Outgoing Channels

If the number of outgoing channels does not exceed 9, they can be selected by the operator of the transit incoming position. The message will be retransmitted simultaneously over the different outgoing channels or on overflow positions if the channels are busy. If the number of channels exceeds 9, the message must be sent to a multiple-address position.

3.6 Faulty Address

If the routing indications are incomplete or spurious when a message is received, it must be transmitted to a supervisory position through the incoming selector.

The supervisory position is operated with torn tape to permit reperforation of a message after the correct routing indications are determined. Retransmission over the outgoing channels is done through the incoming selector.

4. Operating Details

4.1 Selection

When the selection sought by a transit or supervisory position cannot be completed, because of

busy channels or an insufficient number of available overflow positions, the selection is automatically rerouted after 20 seconds. If this second attempt does not succeed, a multiple-address position is automatically selected. If the two multiple-address positions are busy, the message remains at the caller's position and the operator recommences the selection.

For a call from a multiple-address position, there is no automatic rerouting of the selection. If some selected lines are busy, the message is accepted by the associated copying equipment, thus avoiding the overloading of the normal equipment with messages in wait condition.

4.2 Outgoing Junctors

Each outgoing junctor is assigned to an outgoing channel. It can be connected to the different incoming positions either directly if it is available or through the intermediary of an overflow position that is temporarily allotted to it.

The function of the outgoing junctor is to send the heading before the transmission of a message and at the end of the message to send a sequence of tape unwinding signals if the correspondent is a torn-tape transit exchange or a local station.

Sending the impulses for the transmission of the heading is controlled by an impulse distributor common to all the junctors. There is a second emergency distributor with automatic switching. Identification of the message is done by the numbering device assigned to the junctor, by the common timing-code relay set operated from the clock, and by connections to be made according to other signals that must not change.

On the other hand, the outgoing junctors send the checking messages under control of the timing-code relay set, which requests the available outgoing junctors for this purpose every 20 minutes starting on the hour.

4.3 Timing-Code Relay Set

In addition to supplying identifying time signals and ordering checking messages, the timing-code relay set ensures the return of the numbering devices to zero at 00:00 o'clock.

4.4 Incoming and Outgoing Circuits

All outside circuits, incoming and outgoing, are connected to the exchange through telegraph relays so that internal transmission is totally independent of the outside circuits.

4.5 Overflow Positions

The functioning of an overflow position is entirely automatic. On the reception side, the overflow incoming junctor is connected by two selectors in the overflow selector and the two typing reperforators are seized alternately. If one equipment is engaged in receiving, the other can be seized, even repeatedly if necessary.

The two transmitter distributors alternately transmit to the outgoing junctor. The switching is controlled by the detection of the end-of-message signal by each apparatus.

5. Supervision

A page-type receiver monitors the transmission on each outgoing channel.

The supervisory signals from the different circuits operate in the equipment bays. For the outgoing junctors, these signals are also reproduced at the desk of the department chief.

A defect in a circuit engaged in transmission, such as in an outgoing junctor, is indicated to the transit position connected to it.

These signals are transmitted via the connecting selector.

6. Equipment Layout

Switching at the Orly telegraph transit exchange is based on the principles of the Pentaconta telephone switching system and uses its

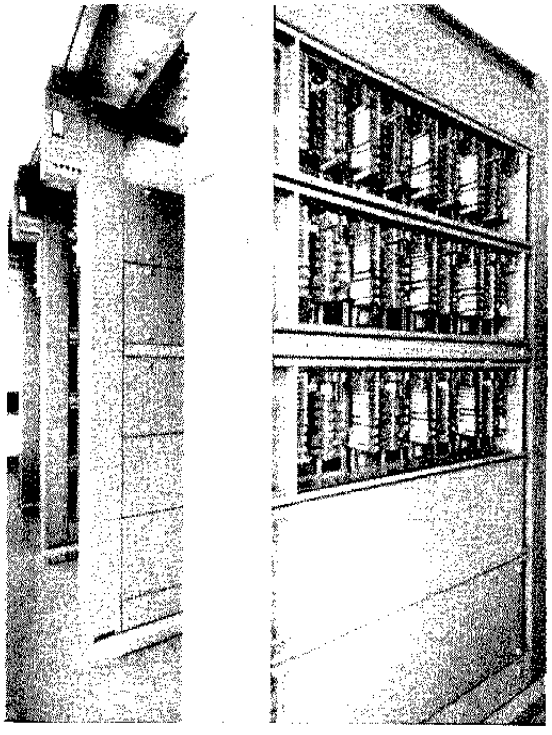


Figure 2—Partial view of the Orly exchange.

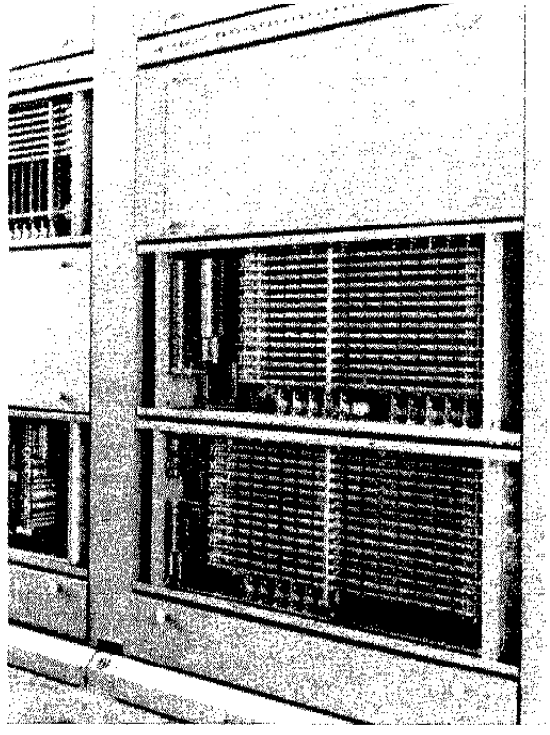


Figure 3—Pentaconta crossbar switches and control relays.

4- or 5-wire multiswitches and relays. The great flexibility of the Pentaconta design and equipment permits its easy adaptation to telegraph installations where its high operational speed and reliability are of prime importance. The exchange is equipped now for 40 incoming and 40 outgoing circuits and may be expanded to 46 circuits of both types.

Figures 2 and 3 are views of part of the equipment, which is installed in 4 rows of bays having a total length of 24 meters (79 feet) and a height of 2.65 meters (8.7 feet). The corresponding area of the room is 54 square meters (581 square feet). The apparatus is mounted in 62 frames.

Consideration is being given to replacing the electromechanical impulse distributors by electronic units to eliminate mechanical wear and reduce maintenance.

7. Conclusion

Since the cut over of the Orly exchange in June 1961, experience has proved that operators previously experienced in the torn-tape method adapt rapidly to semiautomatic continuous-tape operation. This is fundamental as the correct operation of the exchange depends on the total compliance with the message composing procedure.

Every day this transit exchange handles an average of 19 000 messages with peaks as high as 24 000. The mean duration of a message is 55 seconds although some meteorological messages require 300 seconds, which corresponds to 5 meters (16.4 feet) of perforated tape.

The outgoing traffic routed by the overflow positions varies between 28 and 34 percent of the total outgoing traffic.

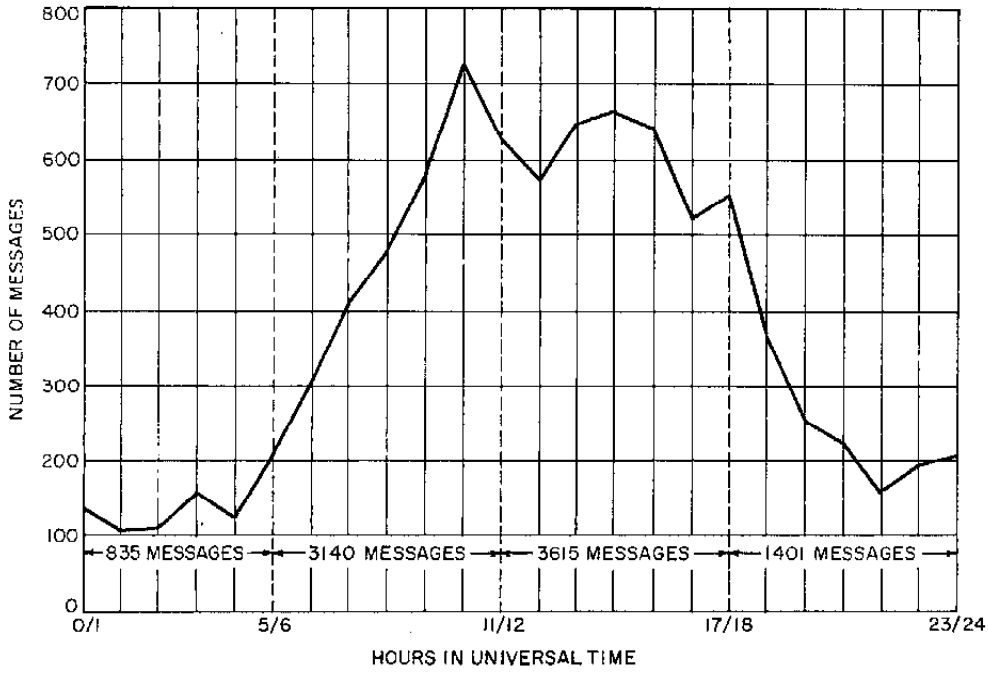
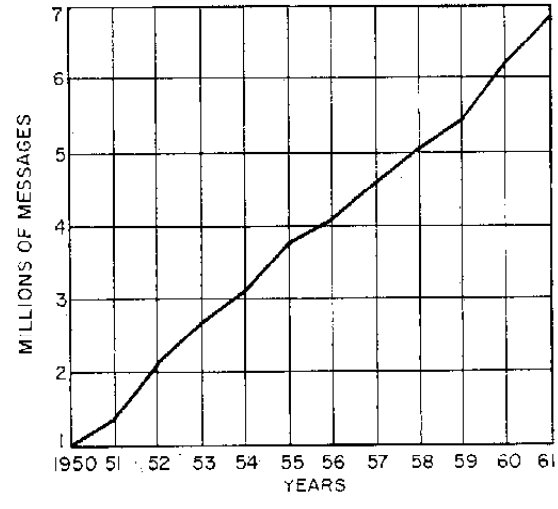


Figure 4—Hourly variations in traffic at the Orly exchange. Each value is for the number of messages handled in the preceding hour.

Figures 4 and 5 show the daily variations of traffic and the steady increase in annual traffic during the past 10 years.

As each transit position may transmit in as many as 9 directions, all messages may be immediately routed by the operators without concentration of traffic in a small number of positions.

Figure 5—Annual growth in traffic at the Orly exchange.



Pierre Samuel was born on 1 March 1928 in Paris, France. In 1952, he obtained a Licence-ès-Sciences Mathématiques from the Université de Paris.

In 1955, Mr. Samuel joined the technical department of Compagnie Générale de Constructions Téléphoniques; he has worked on private telephone equipment and on telegraph installations.

Assistant Type Telephone Set

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Standard Elektrik Lorenz AG; Stuttgart, Germany

H. VAN HOLST
Bell Telephone Manufacturing Company; Antwerp, Belgium

An entirely new series of subscriber telephone sets has been developed jointly by Standard Elektrik Lorenz and Bell Telephone Manufacturing Company. With the requirements of several European telephone administrations in mind, several models were designed and tested [1] by a large number of subscribers in various countries during 1956 and 1957. The majority favored the Assistant set shown in Figures 1 and 2.

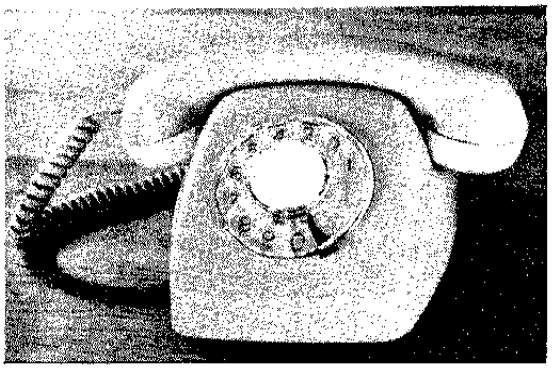


Figure 1—Assistant set.

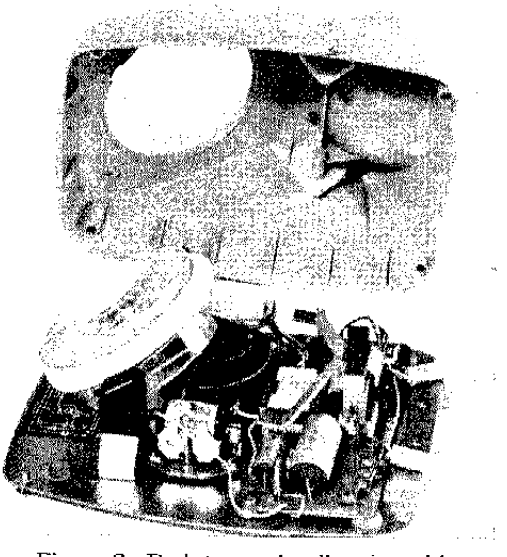


Figure 2—Desk-type subscriber set with cover removed.

1. General Design

For the case, a copolymer of styrene and acrylnitrile combines smart appearance, light weight, and an easily cleanable surface with good impact and scratch resistance. The standard color is grey with ivory mouthpiece and earpiece. The subscriber set is 115 millimeters (4.5 inches) high, 225 millimeters (9 inches) wide, and 200 millimeters (7.9 inches) deep. It weighs 1300 grams (2.9 pounds).

Miniature components are mounted on a printed-wiring board screwed to a metallic base plate, which is fastened to the case by screws. Only the ringer and dial switch do not form part of the printed wiring.

The circuits have been designed for improved performance on all existing feed lines connected to automatic or manual exchanges and to private branch exchanges.

2. Printed Circuit

Most of the wiring is on a rectangular sheet of high-grade glass fiber. The induction coil, gravity switch, capacitors, resistors, varistors, and terminals are automatically assembled and dip-soldered to this board. A 4-wire connector to the dial and a pair for the ringer are also dip-soldered.

The assembled printed circuit is screwed to 4 embossed threads in the steel base plate, which are positioned to support the gravity switch and the board in the region of the screw terminals.

3. Telephone Circuit

To meet the needs of various administrations, 4 different printed-circuit boards are provided.

(A) Circuit with automatic equalization (Western Electric 500 type).
(B) British Post Office Type.
(C) Circuit without equalization.
(D) Bridged circuit (German Post Office).

The printed-circuit board in Figure 3 is the equalized version.

The subscribers set with automatic equalization uses either a ring-armature or a rocking-armature receiver and a conventional transmitter capsule. The circuit includes 2 silicon-carbide varistors. A click-suppressor is mounted on the printed-circuit board.

The circuit that meets the requirements of the British Post Office differs from the equalized circuit in that the 2 matching varistors are replaced by a rectifier stack and a lamp.

4. Induction Coil

The miniature induction coil has about a quarter of the conventional volume. Its small size and weight permit it to be fastened directly to the printed circuit board by dip soldering to its wire-type terminals. A special nylon coil former is employed, and the core is of silicon-steel laminations. Its electrical characteristics are comparable to those of the conventional induction coil.

5. Gravity Switch

The gravity switch is also of miniature size and designed for automatic dip-soldering into the printed circuit. It is fixed directly to the base plate by 2 screws. The springs for the 2 make and the 2 change-over contacts are embedded in a plastic compound. A plastic device riveted to the gravity-switch operating arm actuates the spring and at the same time affords protection against dust. Each spring carries twin palladium contacts.

6. Ringer

The single-coil ringer of Figure 4 is a 2-polarity system. The permanent ferrite magnet and the soft-iron core are both contained within the coil.

The permanent-magnet flux path includes the soft-iron core, operating air gaps, armature, and the other pole area. The area of the armature opposite this pole area is large so as to reduce the reluctance of this air gap.

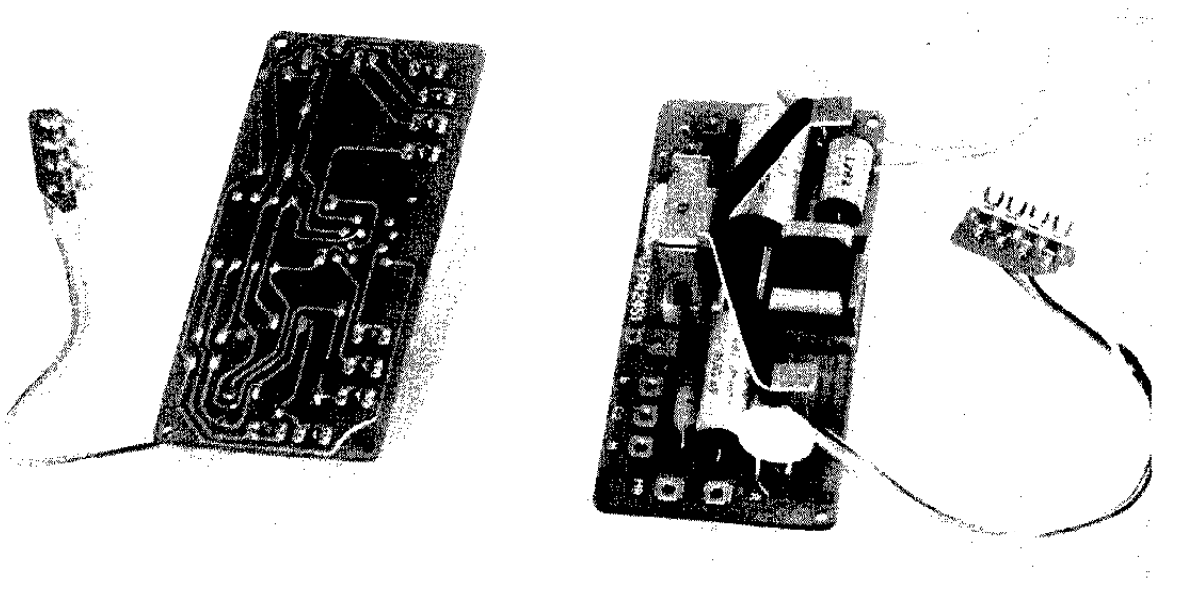


Figure 3—Printed-wiring board equipped with equalizer.

The clapper oscillates between the bell domes and may be laterally biased by an easily mounted spring. A simple knob regulates loudness over a range of about 7 decibels.

7. Dial

The dial, shown in Figure 5, consists of a brass punched base plate on which are mounted the main-spring holder, impulse mechanism, speed regulator, and transparent finger wheel, which are the rotating parts, and the spring nest, finger stop, number plate, and top plate, which are fixed.

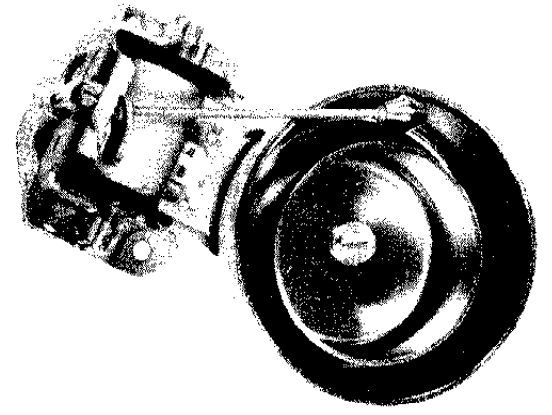


Figure 4—Single-coil 2-polarity ringer.

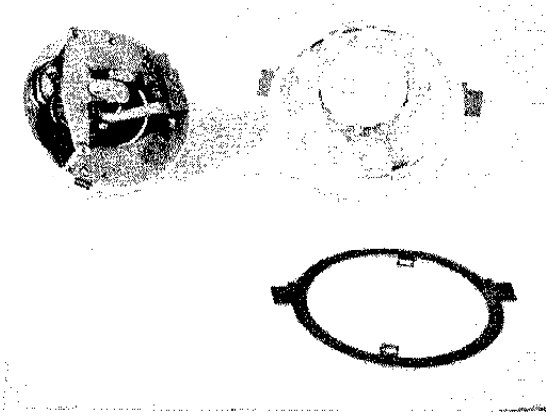


Figure 5—Dial mechanism.

The speed regulator is of the slow-motion type operating at 1000 revolutions per minute. Each of 2 centrifugal weights are controlled by a torsion spring, the tension of which can be adjusted by moving a slider over a graduated sector. Each graduation corresponds to a definite amount of speed variation, and this adjustment can be performed without tools. The regulating action of the centrifugal weights is obtained by the brake effect produced by a bakelised-canvas sector that is part of each weight. The weights rotate in a brass drum.

The impulse mechanism is designed so that the rotation of the speed regulator is started before the impulse cam starts. The impulse contacts are opened by 3 protruding sectors of the impulse cam, which is of the flutter type.

This impulse cam, containing 3 holes for operation with the ratchet springs, is made of polyamide to reduce noise that might be caused by the springs.

As in other designs, the impulse mechanism is provided with a spring clutch. The construction permits easy dismounting of the impulse mechanism.

The helical-type main spring is adjustable and a protecting plate, mounted on the top-plate, prevents any accidental disturbance of the adjustment. The number plate is integral with the outside cover and is of injected acryl.

The dial is acoustically insulated from the base plate by 2 rubber holders and from the case by a polyvinyl-chloride ring, so that noise radiation when dialing is greatly attenuated.

8. Handset

The handset dimensions correspond to recommendations of the Comité Consultatif International Télégraphique et Téléphonique. It is short, thus ensuring a lowered reference equivalent of the sending system and an increase in signal-to-noise ratio. It is shaped to suit the palm and does not tire the fingers. The light

weight of the handset causes no fatigue even during long conversations.

To make the transmission advantages of the shortened handset available to as many persons as possible, it is important that the distance between the transmitter and receiver capsules and their angular relations be properly chosen.

Years ago, extensive measurements of the human head were made in various countries [2]. The Comité Consultatif International Télégraphique et Téléphonique selected the head measurements [3] in Figure 6 as average values from which the position of the mouth center was derived. The frequency distribution of the values found in such head measurements [4] as a function of α and δ is shown in Figure 7.

The body of the handset is of a copolymer of styrene and acrylnitrile and is injection molded in one part. Two straight-sided cores that can be withdrawn directly after molding provide for the receiver and the transmitter cavities. The hollow connecting handle is filled with two core pieces that can be withdrawn through the receiver and transmitter cavities.

A disassembled handset is shown in Figure 8. The two capsule supports may differ to accommodate other types of capsules. They prevent acoustic feedback through the hollow handle. These capsules also form defined spaces behind the capsules to improve the acoustic properties of capsules having rear openings for coupling to these air volumes. The capsules are sealed by polyvinyl-chloride rings.

9. Transmitters

The *TMC2046* transmitter capsule in Figure 9 corresponds to the *T1* Western Electric

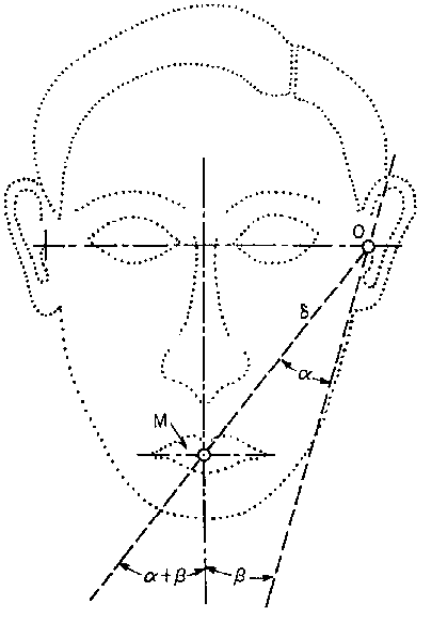


Figure 6—Head standard selected by Comité Consultatif International Télégraphique et Téléphonique. M is the center of the mouth, O is the center of the ear, $\alpha = 22$ degrees, $\alpha + \beta = 34.9$ degrees, and $\delta = 136$ millimeters (5.4 inches).

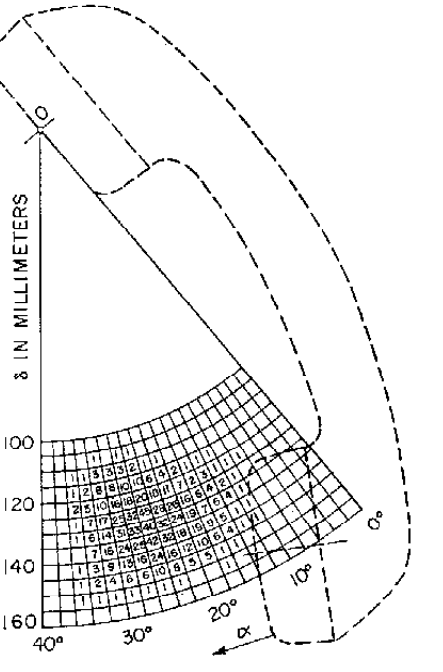


Figure 7—Shape of handset shown on statistical plot of mouth positions with respect to center of ear for 1000 persons.

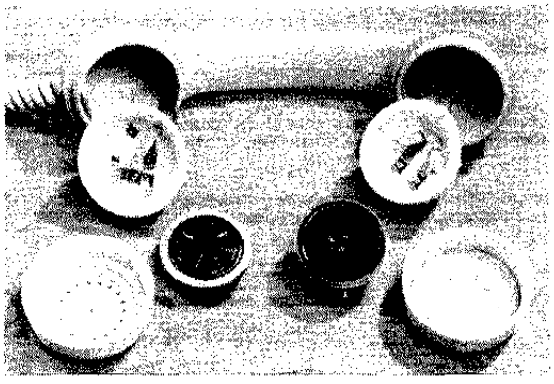


Figure 8—Disassembled handset.

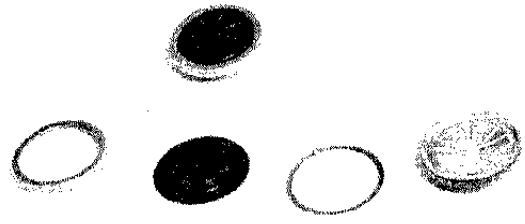


Figure 9—*TMC2046* transmitter capsule.

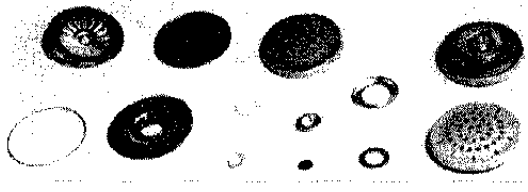


Figure 10—*TMC2047* transmitter capsule.

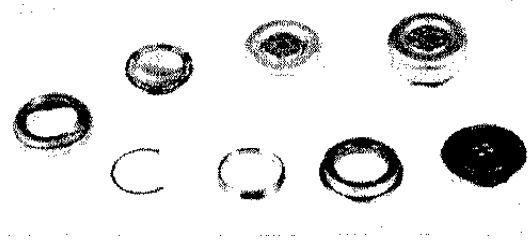


Figure 11—*RCC2687* receiver.

unit. Equalization of response with frequency is aided by three holes in the base that, covered with silk disks of acoustic resistance material, vent the air volume behind the diaphragm into a plastic connecting cup in the handset.

The *TMC2047* transmitter of Figure 10 corresponds to the Western Electric *F1* transmitter in which, however, the damping paper books have been replaced by a metallic separator provided with damping holes. This transmitter is completely sealed against humidity. All transmitters are well protected against corrosion by chromated outside surfaces, a vistanex protecting membrane in front of the aluminum diaphragm, and the use of gold-plated electrodes. Moreover, the electrodes are of spherical shape, thus providing stable performance at various inclinations of the handset.

For use with the Western Electric *500* equalized circuits, the capsules are filled with high-grade pre-aged granular carbon ensuring the high stability of resistance required for adequate operation of the automatic balancing network and for prevention of premature burning with life.

The response of the transmitter capsules are very well equalized with frequency, and approach, in combination with the responses of the receiver, the orthotelephonic response as measured over a 1-meter air path between the human mouth and ear.

The psophometric electromotive-force noise voltage, as measured with the psophometer standardized by the Comité Consultatif International Télégraphique et Téléphonique, the capsules being placed in a soundproof container and supplied with 200 milliamperes, does not exceed 0.02 millivolt.

10. Receivers

One of several suitable types is the *RCC2687A* receiver, which corresponds to the Western Electric U1 ring-armature receiver. As shown in Figure 11, it has a highly coercive alnico circular magnetic system and a dome-shaped

aluminum diaphragm that is well balanced between relatively large air gaps by a permendur ring providing a very stable magnetic circuit.

A composite diaphragm permits the central portion to move almost wholly like a piston, making it nearly 100-percent effective.

The shape and level of its response curve make it a high-quality receiver, since it and the transmitter give an over-all response that compensates to some extent for the high-frequency losses of long subscriber lines.

11. Microtelephone Cords

The microtelephone cords are of the coiled type. Their most obvious property is the high extensibility of the coiled part, which can be stretched, by exerting a very slight force, to 5 times its original length. As soon as the stretching force is released, the cord returns to its initial state.

The cords retain the above-mentioned properties, even after repeated pulling and under severe working conditions. The required resilience and elasticity of the material are obtained by a very special production process.

The conductors are made of 7 tinsel wires that are centrally reinforced by a nylon strand, which absorbs the stretching force that otherwise would be exerted on the wires. The conductors are covered with polyvinyl-chloride in such a way as to permit an appreciable sliding of the conductor inside the insulating sleeve. The sleeves are covered by an over-all spirally wound sheeting. The coiled part of the cord has a length of about 300 millimeters (1 foot).

12. Transmission Performance

The absence of automatic equalization permits the *SSB2900A* set to provide more-efficient transmission than the *SSB2900C* on short lines, whereas on long lines their transmission performances are about equal. The reception is weaker than that of the *SSB2900C* and the response is less smooth.

Optimum sidetone suppression is adapted to a 2.5-kilometer (1.6-mile) loop of 23 American Wire Gage cable, but it can be modified for any other loop.

13. Transmission Response

As shown in Figure 12, transmission is measured across a 600-ohm termination of a transmitting system consisting of the subscriber set, subscriber line of either zero or 5 kilometers (3.1 miles) of 26 American Wire Gage cable loop, and a bridged-impedance cord circuit of 2×250 ohms and 48 volts.

The response is referred to 1 volt per dyne per square centimeter as measured with a constant pressure of 30 dynes per square centimeter at the orifice of an artificial mouth before introducing the microphone into the sound field.

14. Receiving Response

The response plotted in Figure 13 is recorded as sound pressure delivered by the receiver capsule to an artificial ear consisting of a calibrated Western Electric *640AA* capacitor transmitter and a 6-cubic-centimeter coupler as recommended by the American Standards Association. The receiver is actuated from a 600-ohm termination across the receiving system, which includes a bridged-impedance cord circuit of 2×250 ohms and 48 volts, a subscriber line, and the subscriber set.

15. Over-All Response

The total or over-all response of two telephone sets connected over two subscribers lines and a central office represented by two bridged-impedance cord circuits, each of 2×250 ohms and 48 volts, is obtained by applying to the transmitting system a constant sound pressure of 30 dynes per square centimeter from the artificial mouth equipment and recording the sound pressure at the receiving system by the artificial ear equipment and level recorder. The

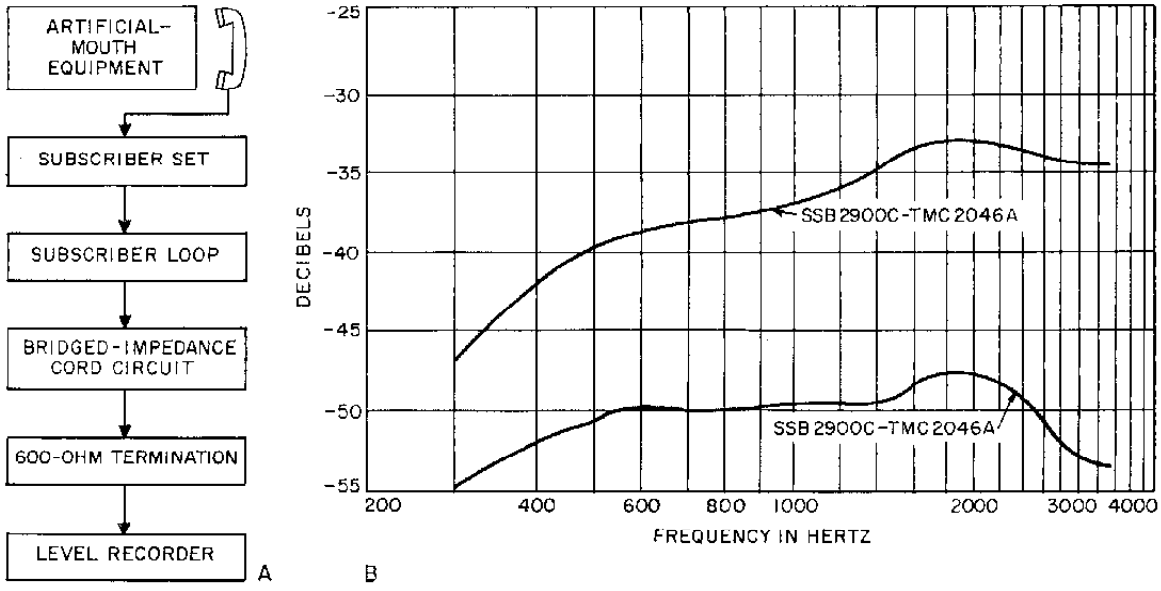
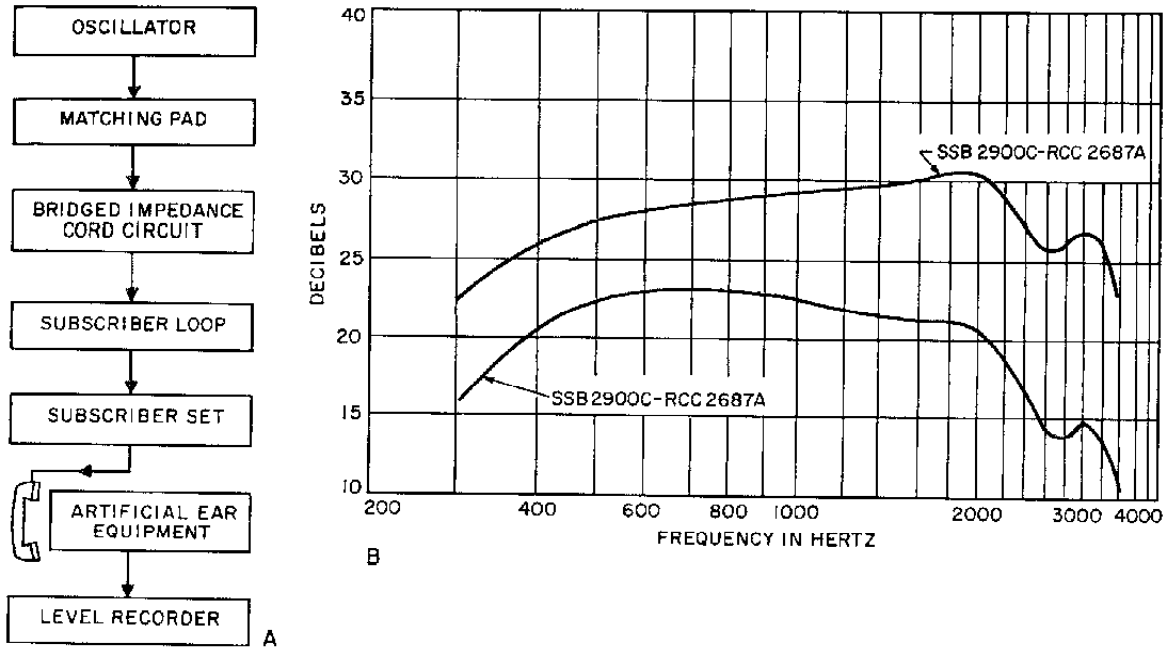


Figure 12—Transmitting response and test circuit. The upper curve is for zero subscriber loop and the lower curve is for an artificial subscriber loop equivalent to 5 kilometers (3.1 miles) of 26 American Wire Gage cable. Response is referred to 1 volt per dyne per square centimeter at the artificial mouth.

Figure 13—Receiving response and test circuit. The upper curve is with zero loop and the lower curve is with an artificial subscriber loop equivalent to 5 kilometers (3.1 miles) of 26 American Wire Gage cable. Response is referred to 1 dyne per square centimeter per square root of 1 milliwatt of available power.



over-all is expressed in decibels referred to 1 dyne per square centimeter per dyne per square centimeter.

16. Wall Set

The *SSB2901* Assistant wall set shown in Figure 15 fills a genuine need for a wall set similar in design to the Assistant desk set. Special attention has been given to standardizing the components of both sets.

The wall set is 185 millimeters (7.3 inches) high, 226 millimeters (9.9 inches) wide, and 76 millimeters (3 inches) deep. It weighs 1400 grams (3.1 pounds).

As in the case of the desk set, the miniature components make possible the use of a printed circuit, which is screwed to a supporting plate. This plate also serves as a support for the dial as is evident in Figure 16.

The housing has been designed so that the

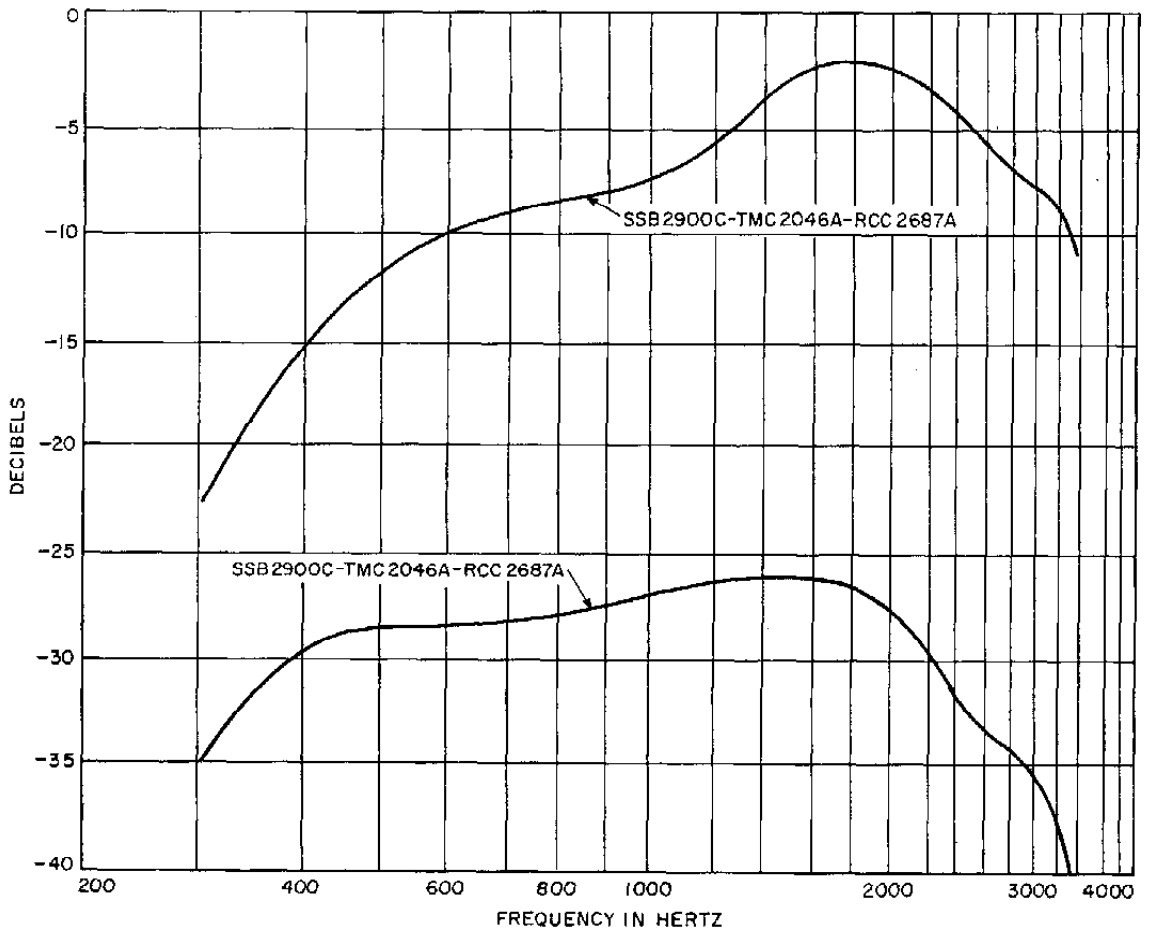


Figure 14—Over-all response using the transmitting and receiving test circuits joining the two bridged-impedance cord circuits to each other. The upper curve is for zero subscriber loops and the lower curve is for two loops or 10 kilometers (6.2 miles) of 26-gage cable. The response in decibels is referred to 1 dyne per square centimeter per dyne per square centimeter.



Figure 15—Wall set.

handset can be easily suspended from the left- or the right-hand side of the housing without cutting off the line. The indentation in the top of the case (left side in Figure 16) provides for this.

All the essential parts of the desk set, which are the induction coil, dial, ringer, capacitors, resistors, gravity switch, and handset are also used in the wall set.

The various circuits incorporated in the desk set have also been introduced in the wall set and the transmission performance data are identical for both sets.

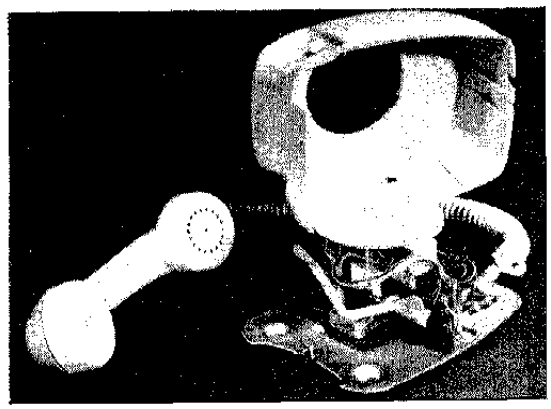


Figure 16—Case removed from the wall set.

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4. H. J. Lurk, "Ermittlung der Masse eines der Haeufigsten Kopfform angepassten Handapparates unter Zuhilfenahme einer neuartigen. Messapparatur," *Zeitschrift fuer Fernmeldetechnik*, volume 15, number 4, pages 49–52; 18 April 1934.

5. E. E. Mott and R. C. Miner, "Ring Armature Telephone Receiver," *Bell System Technical Journal*, volume 30, pages 110–140; January 1951.

W. Grüger was born on 28 December 1907 at Friedland, Sudetengau (at present under Czechoslovakian administration). In 1932, he graduated from the engineering school at Bodenbach.

After working for an electrical supply company, he joined the design department of Siemens & Halske in 1936. From 1944 to 1946, he was a prisoner of war.

In 1946, Mr. Grüger was employed by Standard Elektrik Lorenz and is engaged in the development of telephone sets.

Henri Van Holst was born on 27 August 1926 in Mainxe, France. He received the degree of civil engineer from the University at Ghent, Belgium, in 1954.

On graduation, Mr. Van Holst joined the acoustical laboratory of Bell Telephone Manufacturing Company. Since 1958, he has been in charge of the group developing subscriber sets.

Fehlerortungen, Ihre Messverfahren in Fernmelde und Starkstromkabeln (Fault-Location Measurements for Telecommunication and Power Cables)

Erwin Widl of Standard Elektrik Lorenz is the author of this book. It is divided into the 12 following chapters, a bibliography, and an index.

1. Introduction
2. Theoretical Principles
3. Types and Effects of Faults
4. Methods of Locating Faults
5. Location of Insulation Faults
6. Location of Breaks in Wires
7. Equipment for Conventional Measurements
8. Location of Crosstalk Faults
9. Location of Internal Reflections
10. Characteristics of Fault Location in Power Cables
11. Examples of Fault Location in a Large Municipal Power Cable Network
12. Comparison of Measuring Methods

The book is 17 by 23.5 centimeters (6.75 by 9.25 inches) and contains 166 pages and 130 illustrations. It is published by Dr. Alfred Hüthig Verlag, Heidelberg, Germany, at DM 28 per copy.

Intercommunication Telephone Sets

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Standard Telephon und Telegraphen, AG; Vienna, Austria

The standard telephone subscriber's set while amply fulfilling its intended functions, does not provide for all possible needs, particularly for certain types of intercommunication within offices, buildings, industrial plants, houses, and stores. Special intercommunication systems have, therefore, been designed both for connection to the public telephone networks and for private installations not so connected.

1. Intercommunication Sets Connected to Public Exchanges

The design of intercommunication telephone sets that are also connected to the public telephone network has been undertaken by Bell Telephone Manufacturing Company in Antwerp and Standard Electrik Lorenz in Stuttgart. These systems are characterized by their substantially larger number of switching elements compared with normal subscribers' sets. This emphasizes the need for small components and for combining components such as the equipping of a switching key with a signal lamp, which has the added advantage of unmistakably associating each key with its signal lamp.

To provide a full line of telephone equipments in the style and construction of the ASSISTANT subscriber's set, two larger housings capable of accommodating additional switching elements were developed. They permit the construction of any type of telephone set, as for example sets for intercommunication and secretarial systems and for answering stations for small private-branch-exchange systems incorporating call-metering facilities, monitoring, et cetera. An assortment of these will be illustrated by some examples.

1.1 INTERCOMMUNICATION SYSTEMS

An intercommunication system consists of a master station and a number of extensions. Each extension is equipped with the switching elements necessary to establish a connection. Units common to the entire system, as for example the feed choke and the ringer, are accommodated in the junction box that is associated with the master station. The power supply, which consists of a mains rectifier, is also at the master station. Such systems are economically attractive especially if all stations are within a relatively small area and thus do not require an extensive line network.

Small intercommunication systems for one exchange line and 2 to 5 extensions, coded $\frac{1}{5}$, share a common party-line connecting link. Each set has a red and a green key, for access to the exchange line, and 5 line keys for internal calls. When a set engages the exchange line, its green key is lighted and the red keys of all the other sets are lighted. If one of the line keys at the engaged set is depressed for a call-back, the exchange call is held, the green key is released but its green lamp stays lighted, and throughout the call-back, its red key also is lighted. If the green exchange key of the other extension is depressed to take over the exchange call, the green light on the originating set extinguishes and the handset may be replaced.

Each extension of an intercommunication system having several exchange lines, for example installations with 2 exchange lines and 5 to 10 extensions (coded $\frac{2}{5}$ to $\frac{2}{10}$), is provided with an individual internal connecting link. To enable interruption of an exchange call without replacing the handset, for example during a broker's call, a cutoff key is provided on each set. Figure 1 shows an extension set of the $\frac{2}{5}$ system.

In areas using time-and-zone metering with metering pulses transmitted to the subscriber, rate meters for toll calls are increasingly demanded to keep records of long-distance calls

and assign charges to the individual extensions. The master stations are equipped with one rate meter per exchange line. In systems without caller identification, every long-distance call has to be set up with the assistance of the master station which, notified by a visual or aural signal, records the rate at the termination of the call.

In systems with caller identification, each non-restricted toll-dialling extension is able to set up long-distance calls without requiring the assistance of the master station. On termination of the call, the toll supervision lamp for the exchange line and the identification lamp for the calling extension both light, thus permitting proper charging.

1.2 Secretarial System

Secretarial systems normally consist of an executive and a secretary station that are connected to 2 private-branch-exchange lines. Incoming calls are answered at the secretary station and transferred to the executive whenever necessary. Furthermore, outgoing calls for the executive may be set up by the secretary.

The only distinction between the executive and secretary stations is in the key arrangement. Both are similar to Figure 1 without the 3 center keys in the top row. The circuit of a



Figure 1—Extension set of system having 2 exchange lines and 5 extensions.

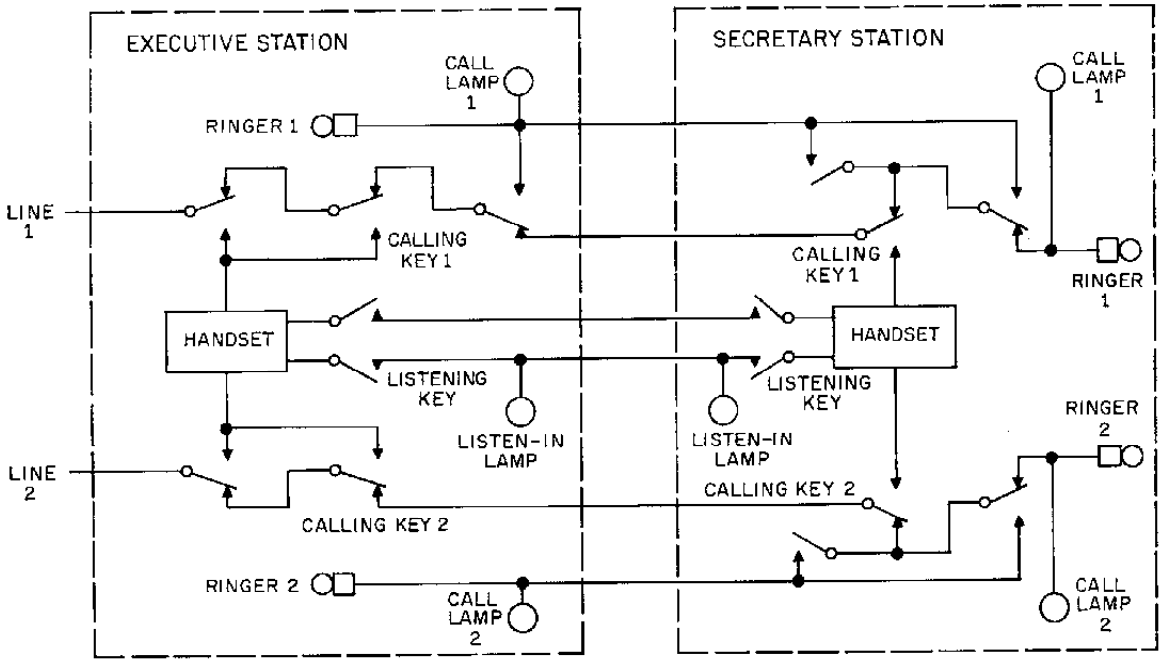


Figure 2—Circuit arrangement of secretarial system.



Figure 3—Intercommunication set having both handset and loudspeaker for connection to a private automatic exchange. The 5 buttons in the top row are for direct selection, camp-on-busy, transfer, staff locator, and loudspeaker. In the bottom row are buttons for microphone short-circuit, press-to-talk, and release.

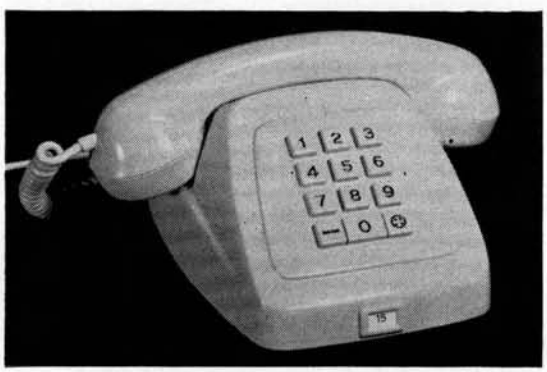


Figure 4—Set with extra facilities limited to press-to-talk and staff-locator buttons.

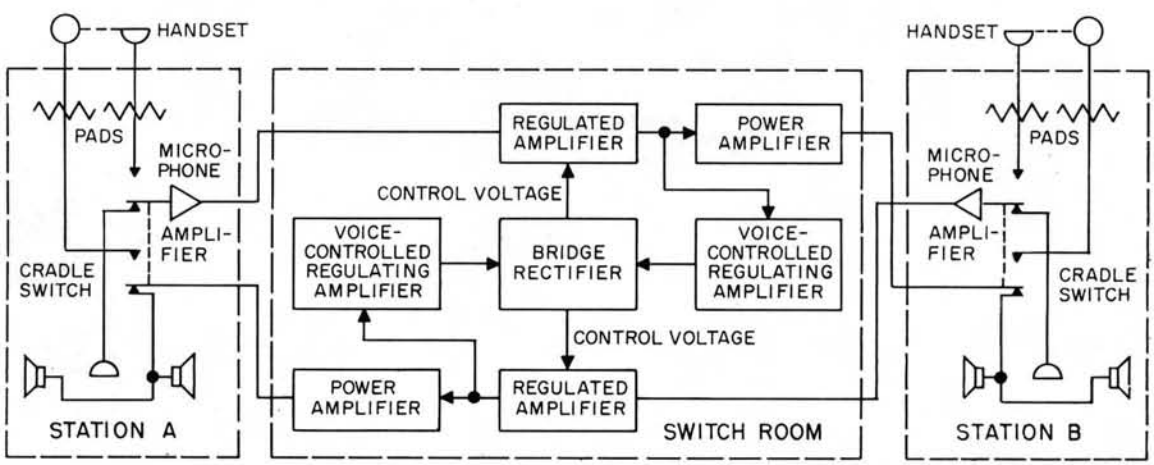
secretarial system is shown in Figure 2. Incoming lines 1 and 2 are looped over the executive station to the secretary station where they are terminated by the ringer. Both have equal access to the lines. However, it is possible to arrange that the secretary use one line predominantly and the executive the other. Finally, another variation is to loop only one line over both stations and provide each station with a separate second line. In this case, 3 private-automatic-branch-exchange lines are required.

2. Intercommunication Sets Not Connected to Public Exchanges

2.1 Large Capacity

Large-capacity intercommunication sets not connected to the public telephone network and using loudspeakers have been developed by Standard Radio & Telefon of Stockholm. They are designed for offices, industrial plants, buildings, and so on. The core of the system is a private crossbar exchange equipped with two-way amplifiers. Direct-current signalling is by push-buttons over 4 wires to the switching equipment.

Figure 5—Speech circuit showing method of controlling amplifiers to prevent howling.



The design is based on modern crossbar and semiconductor techniques and provides a maximum of conveniences, such as, handsfree operation, direct selection, staff location, transfer, camp-on busy, priority call, and group hunting.

2.1.1 *Intercommunication Sets*

One of 3 types of intercommunication sets is shown in Figure 3. It has both a handset and a loudspeaker for handsfree operation. A second design supply omits the loudspeaker. Both have signalling push buttons for all facilities, while certain facility buttons have been excluded from the set, in Figure 4, which has only a handset.

The loudspeaker systems consist of a rocking-armature microphone, a microphone amplifier, and a pair of 75-millimeter (3-inch) loudspeakers.

The axes of the loudspeakers are at right angles to the microphone axis, and the microphone is located midway between the loudspeakers. This arrangement greatly reduces the acoustic coupling between the loudspeakers and microphone.

The pilot lamp indicates that a connection is established. During conversation, the lamp shows a steady light. However, it also indicates the transfer condition by flashing, and the busy-for-called-subscriber condition when extended waiting is forced by operation of the camp-on busy button.

The system operates at 48 volts of direct current for both speech and switching.

2.1.2 *Speech Transmission System*

The speech transmission system is shown in Figure 5. The microphone amplifiers are incorporated in the sets and the main amplifier is in the switchroom. All amplifiers use transistors and are mounted on printed circuit boards.

The speech transmission path is from the microphone in station *A*, via its microphone amplifier, over the line to the main amplifier, and then to the loudspeakers in station *B*. The cradle switch disconnects the loudspeakers and set microphone and connects the handset when the handset is lifted. The set microphone is of the rocking-armature type with an impedance of 2000 ohms at 1000 hertz, which is suitable for the input transistor in the grounded-emitter connection.

The main amplifier is in the cord circuit of the exchange and is assembled on 5 printed circuit boards having connecting plugs. This amplifier has two identical channels. A voice-operated regulating circuit controls the speech direction by increasing the gain of the active channel and decreasing the gain of the passive channel by a larger amount. This prevents howling. Each channel has a 3-stage amplifier providing a maximum output of 0.5 watt. A high noise level makes use of the loudspeaker unsatisfactory. This is due to the masking effect of noise on speech, resulting in poor intelligibility. In this case, handset operation is recommended. However, in an occasionally noisy room, automatic voice-switching may be replaced by manual control using the press-to-talk button for switching.

2.2 Restricted Capacity

Loudspeaker sets for a limited number of stations in a network not connected to the public system have been developed by Standard Telephon und Telegraphen of Vienna. In addition to providing for normal telephony, their loudspeakers offer additional useful services.

There are two basic Dirigent systems: Dirigent in star connection has a master station connected individually to a number of substations, which cannot communicate with each other, and Dirigent universal, which provides unlimited intercommunication among all stations.

2.2.1 *Dirigent in Star Connection*

Dirigent master stations are available for operating with a maximum of 1, 5, or 10 substations. The latter being shown in Figure 6. Master stations are also available without handsets.



Figure 6—Dirigent master station for working with 10 substations in the star connection. The substations cannot communicate with each other.

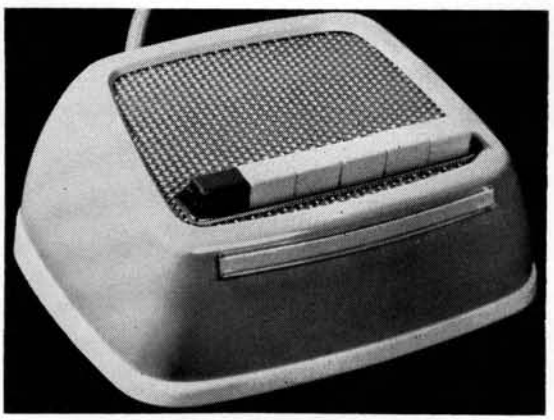


Figure 7—Dirigent universal station, which is one of 6 stations that can intercommunicate among each other without limit.

In addition to the transducer and amplifier, the master stations include a nonlocking button for talking and 1, 5, or 10 locking line buttons to the substations.

The substations are identical for all systems. They do not require an amplifier and in addition to the sound transducer they are provided with a nonlocking button for talking and a locking button for listening.

If there is more than one substation, the connection is made at the master station by pressing the line button of the desired substation. The substation can be called immediately by voice. For answering, the talking button of the called substation must be pressed and a two-way conversation can then be conducted with alternate actuation of the nonlocking talking buttons during speaking. When the conversation has been terminated, the locked line button at the master station is released.

When the self-locking listening button in the substation has been pressed, no further operation is required for the remote listening to the master station.

A call may be initiated from a substation by pressing the talking button and directly calling the master station by voice. If the master station is already busy, the call will not be received and thus disturb the established conversation.

2.2.2 *Dirigent Universal*

The Dirigent universal system may be assembled in networks of a maximum of 6 and 11 stations, each of which can be connected as desired to any other station in the system. A 6-station unit is shown in Figure 7. A single station unit may be connected as one of these substations but will be able to work with only the station to which it is connected.

The number of conversations is limited only by the number of stations. Conference conversations, whereby all stations connected for conference can hear each other, can also be conducted. In view of the necessary conversation discipline, it is recommended that conferences be limited to 4 stations.

Rudolf Bayer was born in Vienna, Austria, on 17 April 1907. After graduating from college in Vienna in 1925, he attended the Technical University in Vienna for 3 years.

He worked for some years in the broadcast receiver industry. In 1939, he joined the engineering staff of Standard Telephon und Telegraphen AG., Czeija, Nissl & Company where he developed carrier systems, public-address apparatus, and intercommunication equipment.

Rudolf Beith was born in Berlin, Germany, on 9 November 1920. He received a degree as a telecommunication engineer.

In 1950, he joined the Berlin works of Standard Elektrik Lorenz, where he developed small private automatic branch exchanges, telephone sets, and special switching equipment. He was head of the technical department from 1957 to 1962, when he was transferred to the Hanover office as sales manager.

Evert Ekbergh was born on 16 May 1928, at Härnösand, Sweden. He graduated from the Telecommunication Administration school for technical clerks in 1950 and received the degree of Master of Science in electrical engineering in 1954 from the Royal Institute of Technology of Stockholm.

After nearly 8 years with the Telecommunication Administration, where he was employed on administrative work, system evaluation and planning, and traffic analyzing, he joined the engineering staff of Standard Radio & Telefon in 1956 as assistant chief switching engineer. He became chief switching engineer in 1961.

Mr. Ekbergh is a member of the Swedish Association of Engineers and Architects.

David Eklöv was born near Uddeholm, Sweden, on 27 March 1918. He received a radio engineering degree in 1945. He was in the receiving laboratories of the Centrum Radio Company from 1939 to 1949 and was then transferred to the telecommunications department as head of the electronics laboratory. He joined the Sinus Company in 1952, where he was in charge of the electronics laboratory. He there invented the 2-way amplifier that made it possible to mass produce intercommunication equipment for handsfree operation. He also developed the Wegephone set.

He joined Standard Radio & Telefon in January 1962 as head of the electroacoustics department.

Matrix Multiplication in Search for Alternate Routes *

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This paper is concerned with routing in a line-switched communication system, that is, the problem of choosing a suitable path to connect on demand the originating switching point with the destination switching point. In more-mathematical language, the problem is to choose a path in a net consisting of nodes and links[1] that connects the originating node with the destination node.

The following two extreme approaches can be taken in connecting two switching points whenever a demand arises.

(A) Complete determination of the path at one location (either at a single central location or at the originating node).

(B) A link-by-link establishing of the path, where the decision as to which link of several specified choices should be taken next in the path[2] is made at the node where the last link in the tentative path terminated.

Approach (A) presupposes that either the control center, where each request for connection is sent, or each node is informed about the status of each link and each node in the net. This information has to be updated at very-short time intervals so that a link or a node that has become unavailable, even temporarily, is not traversed in the path proposed to serve a demand for the connection. A rather-frequent condition that a link is unavailable arises whenever all the trunks in the trunk group constituting the link are busy with traffic.

Maintaining such rapidly updated status information is a difficult undertaking in a net consisting of a large number of links and nodes. For this reason, approach (B) is usually taken in large systems. In the simplest version of this approach, control circuits (route translators) in any switching point, be it at the originating node or at an intermediate node, attempt to establish a path to a given destination along a specific link, determined by memorized instructions. If this link is busy or out of action, the route translators step through a predetermined sequence of attempts to establish the path along alternate links, until an available link is found. If the sequence of attempts is completed without success, a trunks-busy signal is returned.

This version is simplest, since no information is carried about the links already traversed from the originating node to the intermediate node. In other words, the route translators at a given switching point do not differentiate between a request originating at this switching point and a request originating elsewhere and for which this switching point is an intermediate node. This version is discussed in what follows, except for a modification described in the last paragraph of Section 4.

The sequences used by the route translators, taken for the whole net, constitute the alternate routing pattern. By virtue of the fact that connections can be made avoiding some unavailable links, the routing pattern need not be changed when some links become busy or if one or two links are not functioning. But when major changes occur in the communication system, such as large-scale damage, it is imperative to establish quickly a new routing pattern to obtain maximum use from the surviving system. The present paper gives a computational method that constitutes a useful tool in determining alternate routes in a communication net, in connection with setting up new alternate-routing tables. Updated information about the system is assumed to be available to the computer for the purposes of this computation.

* The work was carried out under United States Army Signal Corps Contract DA-36-039-SC-78806 while the author was resident-visitor at Bell Telephone Laboratories in Whippany, New Jersey.

[1] Still another terminology, which will not be used here, is vertices and edges.

[2] Or, if the system has back-out capacity, to subtract a link from the path when a dead-end is encountered.

1. Statement of the Problem

The following problem is discussed here. Given a communication net consisting of a number of nodes connected by links, find through the use of a computer procedure the preferred (first-choice) route and alternate routes (second-, et cetera, choice) between any pair of nodes.

Finding the route consists of determining

(A) The link on which the proposed path leaves node i.

(B) The total distance along the proposed path from node i to node j.

Each link has a weight (number) associated with it that denotes its length or its more-general figure of demerit, such as delay, noise, or cost of use. The first-choice route is to be understood as the *shortest* path measured along links between nodes i and j (here and in what follows, the word shortest means of least demerit quantity, which characterizes the links and which adds up for the links traversed). The second-, third- . . . choice means the shortest path between nodes i and j, *with the following conditions*:

(*A*) Provision that links that were used as initial links out of the node i in previous choices are not traversed.

(B) Provision, during the computation, that the path taken from the first intermediate node m to j *is* the shortest.[3] Note that under certain conditions the node m can coincide with the destination j.

Because of these provisions, the problem of determining alternate routes in general is quite different from the problem of determining the next-to-shortest, et cetera, paths between pairs of nodes. It should be realized that the above criterion, even though its definition is somewhat laborious, represents exactly the simple and intuitive approach to the problem of finding alternate routes under the principle of link-by-link establishing of the path in a communication net, where the choice of next-link-in-path is decided at the node i where the last link terminated. As a first choice, the link originating at i and corresponding to the most convenient path i-j, is taken. The first choice for any particular destination is memorized by the route translators. But when this link is busy or temporarily not functioning, the route translators take an alternate link, which corresponds to the most convenient path i-j not using the first-choice link.

The proposed computer procedure can be used in the central control facility of a communication net. This facility would, (A) receive periodically status information for the whole network, (B) compute, when necessary, the new optimum routing for the whole network, and (C) send the pertinent part[4] of the solution to each node, where the new route translators are simultaneously implemented for the whole network. It should be understood that once the route translators are given new memories, so to speak, each switching point is again in full control of routing any call that reaches it.

The automatic method proposed here for computing alternate routes, utilizes the approach of matrix multiplication developed by Shimbel.[5] To facilitate the presentation

[3] This provision is made even though some calls will overflow at some node along this shortest path m–j to the second-, et cetera, choice link. This is the simplest assumption to make, and of course a very large portion of the traffic will follow the shortest path.

An alternative criterion for a routing pattern could be along the following lines. Define an optimum operation of the system as when the sum of the products (link occupancy) × (demerit figure of the link), summed over all links, is a minimum. The problem of finding a routing pattern to achieve this optimum operation is extremely difficult. Solution in the form of a convenient computation procedure is not available.

[4] The ith rows of the computed routing matrices are sent to the node i. See later discussion.

[5] A. Shimbel, "Structure in Communication Nets," *Proceedings of the Symposium on Information Networks*, Polytechnic Institute of Brooklyn, New York, 1954.

of the method, Shimbel's dispersion-matrix calculation is first briefly presented.

2. Calculation of Shimbel's Dispersion Matrix

The matrix listing all the least-path lengths of a net is called by Shimbel the dispersion matrix of the net. Using his own rules, Shimbel calculates the dispersion matrix through successive multiplication of the structure matrix of the net.

The elements α_{ij} of the structure matrix as used by Shimbel denote the length, or some more-general property, of the link between the node i and the node j. When no such link exists $\alpha_{ij} = \infty$. The transmission demerits in opposite directions may differ, and then $\alpha_{ij} \neq \alpha_{ji}$. The diagonal terms α_{ii} represent the distance of the node from itself and all are therefore zeros. A small net and its structure matrix are shown in Figure 1.

Shimbel's procedure to find the dispersion matrix involves matrix multiplication under special rules in place of ordinary multiplication and addition of terms. Shimbel's arithmetic is as follows: For any arbitrary real finite or infinite numbers x and y

$$x \cdot y = \text{the algebraic sum of } x \text{ and } y \quad (1)$$

$$\infty \cdot x = x \cdot \infty = \infty \quad (2)$$

for any x, including infinity,[6] and

$$x + y = \min (x, y). \quad (3)$$

Definition (matrix product): If A and B are any two square matrices of the same order having only zero, positive, or infinite entries, then their product AB is defined as the ordinary matrix product with the proviso that the above-defined arithmetic be substituted for ordinary arithmetic.

[6] Infinity can be represented in an actual computer program by a fixed very-large number.

If S is the structure matrix of a net, then the product of S with itself, namely S^2, is a matrix giving the least distances between nodes, given that the path from one node to another *may* lead through a node. This, as Shimbel shows, holds true, because the term β_{ij} of S^2 is given in terms of α_{ij} of S as follows

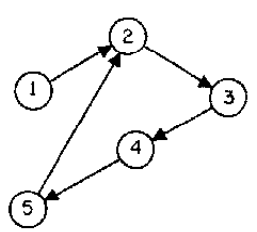
$$\beta_{ij} = \sum_{x=1}^{k} (\alpha_{ix} \cdot \alpha_{xj}) \quad (4)$$

where k is the order of the matrix, that is, the number of nodes in the net. In Shimbel's arithmetic $\alpha_{ix} \cdot \alpha_{xj}$ is the algebraic sum of α_{ix} and α_{xj}. The symbol Σ denotes the operation of choosing the least term. Thus β_{ij} is actually the shortest "one-link" or "two-link" path from node i to node j. The possibility of a one-link route arises since under the summation sign the term $\alpha_{ij} \cdot 0 = \alpha_{ij}$ appears. (This term appears actually twice, the second time in the form $0 \cdot \alpha_{ij} = \alpha_{ij}$.)

It follows that S^m is a matrix giving the shortest paths from node to node given that up to $m - 1$ intermediate nodes are traversed. For any S there exists an integer n such that $S^n = S^{n+1}$, where $n \leq k - 1$, k being the number of nodes. The dispersion matrix of S will be the matrix S^n such that $S^n = S^{n+1}$. The dispersion matrix is designated D in what follows.

3. Computer Procedure for the Alternate-Routing Problem

The procedure for computing alternate routes in connection with setting up new routing tables requires as the input to the



$$\begin{matrix} 0 & 1 & \infty & \infty & \infty \\ \infty & 0 & 1 & \infty & \infty \\ \infty & \infty & 0 & 1 & \infty \\ \infty & \infty & \infty & 0 & 1 \\ \infty & 1 & \infty & \infty & 0 \end{matrix}$$

Figure 1—Small net and its structure matrix.

computer an updated status report for the system, that is, the computer has knowledge of the pattern of interconnection of nodes and the weights of the links. This knowledge is in the form of a structure matrix of the system.

As the output, a set of routing matrices is obtained: first-choice routing matrix, second-choice routing matrix, et cetera. The matrices are square of order k. The index m, appearing as the element in row i and column j of the *first*-choice routing matrix, indicates that route translators at node i will be instructed, on a call to destination j, to attempt *first* the link i-m. Similarly, the second-choice routing matrix indicates the link for the *second* attempt if the first attempt is unsuccessful, et cetera.

As a second output, a set of distance matrices can be obtained: first-choice distance matrix (which is identical to the dispersion matrix), second-choice distance matrix, et cetera. These are again square matrices of order k. The significance of the dispersion matrix has been explained. The element ij in the second-choice distance matrix denotes the distance on a path from node i to node j provided the second attempt out of node i is successful and all first attempts out of subsequent nodes are successful. Similarly for the third-choice distance matrix, et cetera.

The procedure here consists of multiplying a modified structure matrix by the dispersion matrix. Computation of the dispersion matrix by Shimbel's technique is the initial step in the calculation, this matrix being designated by D with elements δ_{ij}. An order of procedure is now described.

First, modify the structure matrix, as used by Shimbel, replacing the zeros on the diagonal by ∞. The modified matrix will be referred to as the M matrix with elements γ_{ij}. Thus all elements $\gamma_{ij,\ i\neq j} = \alpha_{ij}$, and $\gamma_{ii} = \infty$. Next, examine the formation of the product $M \cdot D$. Multiply row i of M by column j of D. Shimbel's terms $\gamma_{ix} \cdot \delta_{xj}$, that is, γ_{ix} plus δ_{xj}, are formed, where x ranges over all nodes. Since the term $\gamma_{ii} \cdot \delta_{ij} = \infty \cdot \delta_{ij} = \infty$, it is eliminated from the competition for being the minimum, second-to-minimum, et cetera, term. Thus x is effectively limited to all nodes except i.

The terms $\gamma_{ix} \cdot \delta_{xj}$ are now compared, the minimum term $\gamma_{ix_1} \cdot \delta_{x_1 j}$ (which, of course, equals δ_{ij}) representing the length of path i–j on the first-choice route. The index x_1 corresponding to this smallest value of $\gamma_{ix} \cdot \delta_{xj}$ indicates that the first-choice route out of i on a call to j is on link ix_1. The index x_1 is entered as the ij element of the first-choice routing matrix, and the term $\gamma_{ix_1} \cdot \delta_{x_1 j}$ is entered as the ij element of the first-choice distance matrix.

The second-to-smallest term $\gamma_{ix_2} \cdot \delta_{x_2 j}$ is now determined. This term corresponds to the shortest path i–j in which the initial link i–x_2 is other than i–x_1. The distance $\delta_{x_2 j}$ corresponds to the shortest path from the first intermediate node, x_2, to j. Thus the two provisions specified for the alternate route are met, except in the case where the shortest path x_2–j transverses the node i. In this case, the path $i\ x_2\ j$ under consideration loops back to i and leaves node i for the second time on link $i\ x_1$, thus violating provision (A).

The shortest route from x_2 to node j passes through node i if

$$\delta_{x_2 j} = \delta_{x_2 i} \cdot \delta_{ij}. \tag{5}$$

The computer program checks for this condition and if it exists, x_2 is disqualified for the second-choice route and the next-smallest term $\gamma_{ix} \cdot \delta_{xj}$ is sought. In most instances

$$\delta_{x_2 j} < \delta_{x_2 i} \cdot \delta_{ij} \tag{6}$$

which indicates that the shortest path x_2–j does not pass through node i. The index x_2 is entered as the ij element of the second-choice routing matrix and the term $\gamma_{ix_2} \cdot \delta_{x_2 j}$ is entered as the ij element of the second-choice distance matrix. And so on for subsequent choices, up to the number of alternate routes desired. Of course, whether the desired number of alternate routes will be found

from all nodes depends entirely on the character of the net.

4. Route-Stability Examination

In link-by-link routing, the danger arises of route instability, which is often called "ring-around-the-rosie." This means that the connection loops back to a node previously traversed. Once back in this node, the call takes the same looping path and this process is continuously repeated until an all-trunks-busy condition is introduced on one of the links in the loop. In the routing scheme arrived at by the procedure described above, ring-around-the-rosie *may* occur if *more than once* in setting up a connection a first-choice link is busy.

It is suggested that inherent stability of the multialternate routing scheme can be achieved if the allowed connections will result in the call getting always closer to its destination in terms of distances (weights) along the links in the net. In the computations outlined above, it is easy to check whether the routing is an allowed routing or not. The necessary and sufficient conditions for an allowed routing on a call to destination j from i to an intermediate node x is that the distance along the shortest path i–j in the net be larger than the distance along the shortest path x–j

$$\delta_{ij} > \delta_{xj}. \tag{7}$$

The stability is achieved at the cost of eliminating some possible paths for establishing a connection. The resulting scheme is certainly not exhaustive; that is, a busy signal may be returned even if an available path i–j exists. The list of links on which attempts to connect are made can be extended beyond the allowed links without the danger of instability by resorting to some message coding. When all the allowed links are busy, route translators can be permitted to attempt setting up the connection to intermediate points x of "last resort" that violate the above equation. The permission to try such a link will be given to calls that have not previously tried last resorts in their quest for the final destination. Thus, virgin messages will have a bit in the address position that will be converted to zero the first time the route of last resort is taken. If a connection loops back to a node previously traversed, it is stopped there (call incomplete).

5. Actual Computer Program

The procedure for finding alternate routes has been programmed on an IBM 704 computer by Miss G. C. Watling. The program for a net consisting of 18 nodes took approximately 1 minute running time.

6. Conclusions

A procedure for determination of the alternate routing pattern in a communication net by a computer has been presented. The method seems to be especially useful when a communication network has been damaged. When the status of the surviving network is reported to the central supervisory facility, it is obviously important to compute the new routing pattern reliably and quickly so that the functioning parts can be put into optimum use with minimum delay.

7. Acknowledgment

Very helpful comments of Eric Wolman are gratefully acknowledged.

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Theoretical and Practical Aspects of Telephone Traffic

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1. Introduction

Calculation of the minimum quantities of telephone exchange equipment required for satisfactory service is possible because the traffic created by subscribers has certain characteristic features and shows regularities notwithstanding the fact that each subscriber station is used in an independent manner. Although telephone users have different occupations, interests, and habits, it has been observed that as a group they originate, within certain limits, about the same number of calls during corresponding periods of each working day and that the interoffice traffic flows show similar aspects.

Probability theory is applied to switching problems under certain generally accepted assumptions of a random arrival and duration of calls. Based on these assumptions and consideration of the manner in which traffic is handled by specific switching arrangements, equations have been developed to guide the telephone engineer. In evaluating these equations, the extent to which these assumptions agree with the corresponding types of traffic met with in practice must be determined.

It is equally important that the methods used to evaluate actual traffic yield quantities that, when inserted in the equations, give reliable results. The choice of a correct measurement period leads to a statistical concept of the busy hour, which subject will be considered later.

For practical purposes, the observation and evaluation of actual telephone traffic is generally restricted to the collection of data pertaining to the occurrence of numbers of calls during certain periods of time, to the number of calls in progress simultaneously, to the interoffice traffic flows, and to the holding times of the connections. They serve to establish the calling rate per line and all other data needed to calculate the quantities of telephone equipment required. It is implied that satisfactory agreement exists between the theoretical and the actual traffic flows. The number of studies made on the observation and verification of the characteristic features of actual telephone traffic is restricted.

The fact that most of the present telephone plant operates satisfactorily from a quantitative point of view during most of the time is insufficient proof of agreement between theory and practice as, in accordance with well-established engineering usage and to care for future growth, the basic traffic data are generally made to include a safety margin, and later plant extensions are based mostly on experience during years of operating the particular plant.

The object of this paper is to outline in general terms the means and methods by which actual traffic may be measured, to review the principal concepts of theoretical traffic, and to discuss their effects on the interpretation of the busy hour and on the choice of the busy-hour-traffic data. For the sake of clarity, the results of applying several well-known theories have been presented in concise form.

Finally some consideration is paid to the problem of deriving values for the traffic offered to a switching arrangement from observations of the traffic carried by it.

This paper does not deal with sampling studies that are possible from the observations made.

2. Theoretical Telephone Traffic

A flow of theoretical telephone traffic is fully defined by the distribution function of two independent random variables, one relating to the holding time of calls and the other to their arrival time or their coincidence.

2.1 Duration of Calls

For the duration of the calls, it is usually assumed that the holding time is either constant or it is variable and follows a negative exponential law.

The distribution function of the latter type of holding time expresses the probability that a specific call does not terminate within the

period t or, in other words, by the probability $p\ (> t)$ that the holding time exceeds a time t. If the average duration of the calls is denoted by h, the function reads

$$p(> t) = e^{-t/h} \qquad (1)$$

where $0 < t < \infty$.

The evaluation and verification of this holding-time distribution against that of actual telephone traffic requires appropriate testing equipment but otherwise it offers no difficulties.

The measurement of many thousands of telephone connections has demonstrated a persistent deviation from the negative exponential law given in (1). Figure 1 indicates the general shape of an observed distribution curve; the exponential curve has been added to show the general trend of the deviation. This phenomenon can be explained by the following considerations.

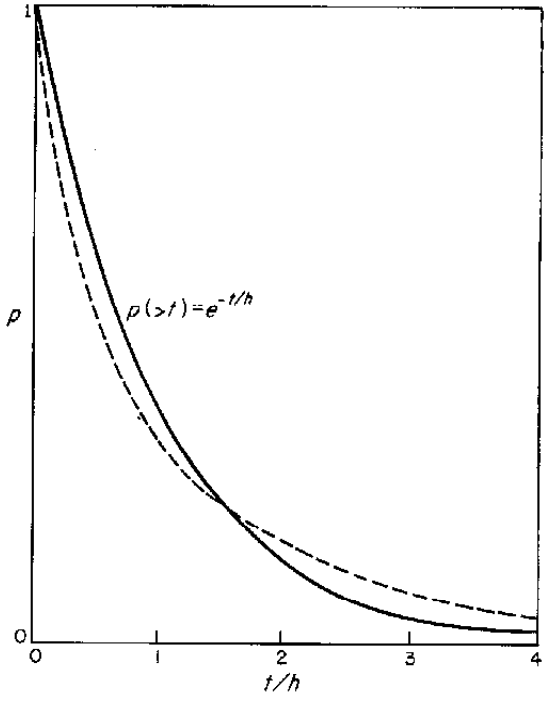


Figure 1—The broken line is an observed distribution of holding time and the solid line is the calculated exponential curve.

If the variation of the holding times of the calls of two traffic flows is expressed by the functions $p(> t) = e^{-t/h_1}$ and e^{-t/h_2}, h_1 and h_2 indicating the respective average holding times, and if the two flows combine or merge, the distribution function of the holding times of the merged calls is

$$p(> t) = \frac{n_1 e^{-t/h_1} + n_2 e^{-t/h_2}}{n_1 + n_2}. \qquad (2)$$

n_1 and n_2 represent the average number of calls of the flows.

A merged traffic flow comprises every call of the merging flows. It has been assumed that the merging process causes no change in the individual holding times of the calls. This is a simplification of what actually happens in a telephone exchange where dialling, selection, and hunting demand some time. As, in general, the difference is less than half a percent and of some random nature, this simplification does not affect our conclusions.

Consideration of the above equations reveals that the holding time of the combined traffic flow follows the simple negative exponential law comprising one term, only when $h_1 = h_2 = \cdots h_n$. Actual traffic rarely satisfies this condition as in many instances the average holding time depends on the destination of the flow. The group selectors of a telephone exchange direct their traffic to local subscribers, exchanges located in the suburbs, special services, and distant locations reached via rural and toll lines. Each of these classes of calls has a characteristic average holding time. Special-service calls distinguish themselves by a short average holding time, calls between residential subscribers tend to be longer than business connections, while the average duration of rural calls is rather long.

From the manner in which the distribution functions of holding times add and from the observation that different traffic flows often show distinctly different average holding times, it follows that the simple exponential

law, which is frequently adopted for theoretical studies, is applicable only in rare instances and that in general the actual distribution function of the holding time is better expressed by a function comprising a number of exponential terms.

This phenomenon is of special importance when dealing with delay problems in which, as is well known, the variation of holding time plays a critical role. The generally observed deviation is such that the theoretical delays based on a simple negative exponential distribution function are too optimistic.

It may be concluded from theoretical investigations made by Palm (1938) that the loss equations in general and the Erlang loss equation in particular are valid for any holding-time distribution. In Section 3, we shall revert again to this subject.

2.2 Arrival of Calls

For the variable describing the arrival of calls, a variety of distribution functions result depending on a number of assumptions made. For instance, the number of sources that create the calls may be assumed to be either finite or infinite. The latter assumption is generally preferred in practice as it leads to more-manageable equations.

When the traffic is unhampered by any switching arrangement causing loss or delay, when the calls may commence at any moment of the period considered, when no relation between the calls can be demonstrated, and with an infinite number of sources, the following Poisson function describes the probability of the arrival of precisely x calls during an observation period t.

$$P_x(n, t) = (nt)^x e^{-nt}/x!$$

where

$$x = 0, 1, 2, \cdots, \infty ; \qquad (3)$$

$0 < t < \infty$; and n is the average number of calls arriving during the unit of time. In the following section, we shall refer to this class of equations as time distribution functions.

The intervals between successive calls are characterised by the arrival of no new calls. Then (3) transforms to

$$P_0(n, t) = e^{-nt} = p(> t). \qquad (4)$$

The probability $P_0(n, t)$ indicates that no calls arrive during the period t and it is equal to the probability that the length of the interval between successive calls exceeds t.

Many thousands of intervals have been measured in a public and a private exchange with the consistent result that the interval between successive calls closely follows the negative exponential law. The minor deviations show no persistent tendency so that in this instance one may conclude that there is agreement between theory and practice. The influence of repeated trials appears to be negligible if it exists at all.

When independent flows of telephone calls that satisfy (4) merge, the distribution function for the merged traffic is obtained by the multiplication of the functions of the individual flows. If, for example, the variations of the two flows are characterized by $p(> t) = e^{-n_1 t}$ and $e^{-n_2 t}$, n_1 and n_2 being the average numbers of calls per hour, the combined flow will follow the function

$$p(> t) = e^{-(n_1+n_2)t}. \qquad (5)$$

Consequently the average number of calls of the combined flow is obtained by the straight addition of the averages of the merging flows. A similar rule holds good when traffic divides. There seems to be a discrepancy between actual and theoretical traffic in that the observed expectation n varies during the day whereas the n appearing in (3) is of a constant nature. This fact and the desire that theory and practice agree in the closest possible manner, leads to the concept of the busy hour, a problem discussed in Section 6.

Instead of (3), which pertains to a sequence of events, we may measure telephone traffic by call coincidence. The equations then express the probability of a specific number of calls

being simultaneously in progress. In the following equation of Poisson, y is the average of the variable x.

$$P_x(y) = y^x e^{-y}/x \qquad (6)$$

where $x = 0, 1, \cdots, \infty$ and $y = nh$.

Section 3 discusses the relationship between these two classes of equations, where the latter will be referred to as space distribution functions.

When two traffic flows that obey (6) merge, the distribution function of the combined flow is

$$\sum_{i=0}^{x} P_i(y_1)P_{x-1}(y_2)$$

where the suffixes 1 and 2 distinguish the two flows. Elaboration leads to

$$(y_1 + y_2)^x e^{-(y_1+y_2)}/x! \qquad (7)$$

which is $P_x(y_1 + y_2)$.

From this follows the previous conclusion that, when independent traffic flows merge, the average traffic value of the merged flow is obtained by the straight addition of the averages of the merging flows.

The multiplication theorem does not apply to interdependent traffic flows as will be shown by the following example. When hunting over a group of junctions in a sequential order always starts from a fixed home position, the distribution function of the intervals between successive calls picked by the first junction is

$$p(> t) = \frac{1}{1-y} e^{-nt} - \frac{y}{1-y} e^{-t/h}, \quad y \neq 1.$$

$$p(> t) = e^{-nt}(1 + nt), \quad y = 1$$

in which

$$y = nh \quad \text{and} \quad 0 < t < \infty.$$

The traffic offered to the group is assumed to obey laws (1) and (3). The corresponding function of the calls overflowing the first junction is, according to Palm (1943),

$$p(> t) = \frac{a^{\frac{1}{2}}}{a^{\frac{1}{2}} + b^{\frac{1}{2}}} e^{-bt/h} + \frac{b^{\frac{1}{2}}}{a^{\frac{1}{2}} + b^{\frac{1}{2}}} e^{-at/h}$$

in which

$$a^{\frac{1}{2}} = \tfrac{1}{2}[(4y + 1)^{\frac{1}{2}} + 1]$$

$$b^{\frac{1}{2}} = \tfrac{1}{2}[(4y + 1)^{\frac{1}{2}} - 1]$$

$$y = n \cdot h$$

$$0 < t < \infty.$$

If the multiplication theorem mentioned above applies, the product of these two functions should be equal to e^{-nt}, the distribution function of the interval between successive calls offered to the group. Evidently this is not the case, because the two functions relate to interdependent traffic. Calls flow over the first junction only when it is engaged.

Traffic changes its character after having passed through a switching arrangement. If, for example, the arrangement consists of a limited group of N junctions, the traffic peaks exceeding N calls are cut off. They may either be lost or delayed. O'Dell has introduced the term smooth traffic for that part of random traffic that has passed through such a limited group, while the traffic that has been rejected by the group during periods of congestion is called peak traffic. When such traffic is directed to another group of outlets it is called overflow traffic.

When any call has access to any of the outlets at any moment (full availability), the Erlang distribution function describing the number of calls in progress simultaneously is

$${}_1E_x = \frac{P_x(y)}{\sum_{i=0}^{N} P_i(y)} \qquad (8)$$

where $0 < x \leqslant N$.

Here x is the random variable associated with y, the average traffic offered to the group and expressed in erlangs. The calls that are not served are lost.

When the calls that are not served immediately are delayed, the distribution function of

the traffic carried by the group is expressed by the second equation of Erlang.

$$ {}_2E_x(y) = \frac{P_x(y)}{\sum_{i=0}^{N-1} P_i(y) + \frac{N}{N-y} P_N(y)}, $$

$$ x = 0, 1, \cdots, (N-1). \qquad (9A) $$

$$ {}_2E_N(y) = \frac{\frac{N}{N-y} P_N(y)}{\sum_{i=0}^{N-1} P_i(y) + \frac{N}{N-y} P_N(y)}, $$

$$ x = N. \qquad (9B) $$

The symbols in (8), (9A), and (9B) refer to (6). Equations (9A) and (9B) assume that the holding time follows the negative exponential law (1). Equations (8) and (9) may serve for the verification and evaluation of observed telephone traffic but (8) is apt to present difficulties in this respect, as explained in Section 7.

3. Relationship Between Time and Space Distribution Functions

In the preceding section, it has been explained that a traffic flow can be described by two distribution functions. The first of these relates to the holding time, for which no definite function is adopted in this section. The second relates to the calls and two expressions are used which pertain to

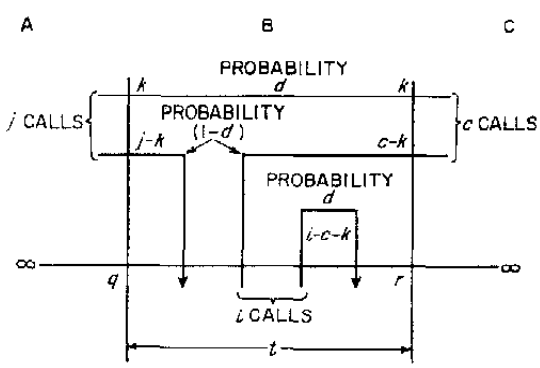


Figure 2—Events and probabilities.

(A) The probability that at a moment j calls are found simultaneously in progress, which we represent by the symbol ${}_2X_j(n, t)$.

(B) The probability that i calls arrive during a following period t, which we represent by ${}_1X_{i,j}(n, t)$, under the condition that j calls were in progress at the beginning of the period. i and j are positive numbers and range from 0 to ∞.

As these two functions relate to the same traffic flow, the average number of calls arriving during the average holding time must be equal to the average number of calls in progress simultaneously. In other words, the t's of both symbols represent equal periods, in the one instance t represents the average observation period and in the other the average holding time.

These two conceptions are distinguished here as time and space distribution functions. For their discussion, reference is made to Figure 2, in which the horizontal line indicates time, three periods being distinguished: A that extends from ∞ to the moment q; B that is limited by the moments q and r and has a duration t, the average holding time; and C that extends from the moment r to ∞.

A call in progress at r must have been originated either during the period A or B, to which events we attach the probabilities d and $(1 - d)$.

A call that started during B must end either during B or must continue past r. As the probability of the latter event equals $(1 - d)$, that of the former must be equal to d. As already stated, the average number of calls arriving during B is equal to the average number of calls in progress at the moment r.

By similar reasoning, we find that the probability $(1 - d)$ also applies to a call existing at the moment q and terminating during the period B. The latter can also be demonstrated by interchanging the moments of birth and death of the two kinds of calls that have the probabilities $(1 - d)$.

We suppose that at the moment r exactly c calls exist, k of which are included in the j calls existing at the moment q and $(c-k)$ are among the i calls originated during the period B. Consequently $(j-k)$ calls of the j calls terminate during B and $(i-c+k)$ calls of the i calls terminate during the same period.

The probability of finding c calls in progress simultaneously at r is consequently composed of an infinitely large number of partial probabilities as follows.

$$ {}_2X_c(n,t) = \sum_{k=0}^{c} \sum_{j=0}^{\infty} \sum_{i=0}^{\infty} \binom{j}{k} d^k (1-d)^{j-k} \times {}_2X_j(n,t) \binom{i}{c-k} (1-d)^{c-k} d^{i-c+k} {}_1X_{1,j}(n,t). $$

There is no need to consider the effect of the limit $i+j \geqslant c$ as combinations having negative lower numbers are equal to zero.

Using similar symbols throughout, we have

$$ {}_2X_c(n,t) = \sum_{k=0}^{c} \sum_{j=0}^{\infty} B_k(j,d) {}_2X_j(n,t) \times \sum_{i=0}^{\infty} B_{c-k}(i, 1-d) {}_1X_{i,j}(n,t). \quad (10) $$

In this equation, B stands for the Bernoulli law. It delineates the general relation that exists between the three distribution functions mentioned. One would deduce from this that the space distribution function depends in any case on the time and the holding-time distribution functions. There are, however, exceptions. If the time function obeys the Poisson law, the same law also applies to the space function without any postulate for the holding-time function.

When introducing

$$ {}_1X_{i,j}(n,t) = P_i(y) $$

and

$$ {}_2X_j(n,t) = P_j(y) $$

into (10), an identity appears and no condition exists in this instance that the holding time distribution function should satisfy. We then have for (10)

$$ {}_2P_c(y) = \sum_{k=0}^{c} \sum_{i=0}^{\infty} B_{c-k}(i, 1-d) \times P_i(y) \sum_{j=0}^{\infty} B_k(j,d) P_j(y). $$

It can be demonstrated that the following equation is of general validity.

$$ \sum_{j=0}^{\infty} B_k(j,d) P_j(y) = P_k(d \cdot y). \quad (11) $$

Introduced, we obtain

$$ {}_2P_c(y) = \sum_{k=0}^{c} P_k(d \cdot y) P_{c-k}[(1-d)y] $$

or

$$ {}_2P_c(y) = y^c e^{-y}/c! \times \sum_{k=0}^{c} \binom{c}{k} d^k (1-d)^{c-k} $$

which is an identity as the above $\Sigma = 1$.

The above method may be applied to other types of distribution functions, but care should be exercised when dealing with limited groups, as in such cases part of the i calls must necessarily be lost.

4. Artificial Traffic

Besides telephone traffic created by subscribers and the variety of theoretical forms of traffic discussed in Section 2, there exist various kinds of simulated traffic produced by mechanical or electrical means such as traffic machines or computing machines.

The complete simulation of traffic requires, as stated in Section 2, two appropriate random mechanisms, one for imitating the random arrival of calls and the other for the distribution function of holding time. Simulation of the arrival of calls presents no great difficulties and several equipments provide for this but a holding-time mechanism incorporating any specific random character has not yet been developed. The precision of such instrument should be good.

A reliable traffic machine should include means for checking the amount of traffic offered or the number of calls carried by the equipment with their average holding time. This check is required to verify correct operation of the machine during every test and the necessity for it increases with the accuracy to which measurements are made, which should range between 1 and 0.1 of 1 percent. Lacking such a check, there is no knowledge of the variation of traffic about the desired average and no possibility of eliminating faulty tests. With constant holding time, such a check is fairly easy as knowledge of the number of calls carried and lost suffices, but for variable holding time, matters become more complicated in that the durations of calls that may be simultaneously in progress have to be added.

With the roulette principle invented by Kosten, two independent random variables are used, one indicating the beginning and the other the ending of the calls. This method implies the adoption of the negative exponential distribution function for holding time with the result that machines based on this principle abandon the only means for checking for correct operation during the various tests.

The same principle is also employed in computing machines creating series of random numbers and placing the calls via some switching network in suitable memories. Such application has the inherent disadvantage, noted in the preceding paragraph, that no precise knowledge of the traffic offered or carried is produced so that faulty tests cannot be eliminated. As each test requires a new program, possible bias is difficult to discover and eliminate. The more so, as programming engineers are not supposed to possess sufficient experience to be aware of the pitfalls in the creation of random phenomena.

5. Traffic Measurement

Based on the description given for theoretical traffic, actual telephone traffic is measured in three distinct manners.

With the first method, lengths of periods are measured and such observations serve for gathering information on holding times of connections, intervals between successive calls, periods during which groups of circuits are blocked, et cetera. A contact-closing clock or a simple interrupter may be used as a timing device. The error committed thereby amounts to about half the interval of the basic timing period. (See Rabe, 1949.) The measuring equipment also includes a number of counters operated in series with a switch that steps in synchronism with the impulses produced by the timer.

With the second method, a count is made of the number of calls that arrive during consecutive equal periods. The test equipment may be simply a lamp that flashes each time a call arrives but there may also be included means of counting the number of call arrivals.

The drawback of this method is that it must be accompanied by observations relating to the holding times of the calls, whereas, as previously stated, no separate information on holding time is essential in case of lost-call switching system arrangements.

With the third method, the numbers of calls simultaneously in progress are noted. It has the advantage of being a direct measurement of the traffic expressed in erlangs. It dispenses with the tedious method of counting calls and of measuring their holding times.

The observations are generally made at equal time intervals that exceed the average holding time. Equal spacing, however, does not seem to be a stringent requirement as from a theoretical point of view the observation moments may also be chosen at random. Neither would it appear necessary to limit the interval to a minimum value as the numbers of calls in progress simultaneously can also be obtained, and even more precisely, by using a recording ammeter in series with an individual resistance per circuit. In this case the moments of observation may be considered to lie infinitely close together. However, in view

of the method described in the next section of determining the busy hour and the busy-hour traffic, a standard uniform interval between the successive observations is recommended.

Even the stipulation *simultaneous* need not be taken too strictly. Should an observer have no test equipment to indicate the instantaneous number of simultaneous calls, he may still proceed by observing in succession the condition of the circuits in the various groups but he should continue his busy count even if changes occur in the condition of the circuits already passed over.

By merely walking through an exchange and noting in a leisurely manner the number of outlets in each of the various groups and how many are busy, one can obtain a fairly accurate picture of the traffic the equipment carries and can conclude whether there is any danger of abnormal congestion.

6. Busy Hour and Busy-Hour Traffic

The subject of the busy hour was touched on in Section 2 in connection with the constant nature of the mathematical expectation n appearing in (3) and y in (6).

Busy hour was introduced in the early days of the telephone art to ensure that the traffic data underlying the computation of the equipment quantities were chosen sufficiently large to cover the traffic at any period of the day.

The definition of the busy hour contained in British Standard Specification 204 is in line with the above but does not recognize its statistical character. In the pertinent literature, the concept of the busy hour is but rarely referred to and is even sometimes called "vague."

This situation induced us in 1937 to propose a concept of the busy hour through which agreement between theory and practice would be ensured in the best possible manner. It was subsequently recommended by the Comité Consultatif International Télégraphique et Téléphonique (volume VI, page 16; 1954) for the evaluation of the international busy-hour traffic.

Realization of this agreement is aided by the application of a guided choice among the data observed and is based on three characteristic features of the theoretical traffic distribution function. These features are best explained by using the Poisson equation (6). The parameter y is of a constant nature and represents the average number of calls in progress simultaneously, taken over a period of unlimited duration. We may divide this period into H subperiods, each of 1 hour, as shown in Figure 3, and to make in every hour an arbitrary number of 10 equidistant observations.



Figure 3- Division of observation period into subperiods.

The three characteristic features referred to are:

(A) We cut the above period into H subperiods each of 1-hour duration and arrange the numbers of simultaneous calls observed in the following manner, (+1.9) representing the number of simultaneous calls observed at the moment +1.9.

.	.		.	.
.	.		.	.
.	.		.	.
(−2.9)	(−2.8)	···	(−2.1)	(−2.0)
(−1.9)	(−1.8)	···	(−1.1)	(−1.0)
(−0.9)	(−0.8)	···	(−0.1)	(0.0)
(+0.1)	(+0.2)	···	(+0.9)	(+1.0)
(+1.1)	(+1.2)	···	(+1.9)	(+2.0)
(+2.1)	(+2.2)	···	(+2.9)	(+3.0)
.	.		.	.
.	.		.	.
.	.		.	.

The observations are now arranged in 10 columns each containing an infinitely large

number of H data. The average for each column amounts to y, the average number of simultaneous calls. The subperiod of 1 hour is representative for the busy hour and the above demonstrates the constant nature of the mathematical expectation during the busy hour.

When considering the H observations of each of the above 10 columns, we would further find that they contain the same proportions of 0, 1, 2, $\cdots$, simultaneous calls. This demonstrates the principle of the statistical equilibrium introduced by Erlang, which may also be expressed by stating that the probability of finding a specific number of simultaneous calls is independent of the time. The principle of the statistical equilibrium is, therefore, directly associated with that of the constant nature of the mathematical expectation. If the latter is dropped for one reason or another, the principle of the statistical equilibrium must be revised.

(B) The variation of the observed numbers of calls in progress simultaneously is expressed by (6).

(C) Equation (7) demonstrates that arithmetical addition and subtraction of the average numbers of calls in progress simultaneously must be applied when unhampered traffic flows either merge or divide.

The features (A) and (C) pertain to theoretical traffic that satisfies (6). They also appear to be properties of traffic that satisfies the distribution function

$$p(x) = \binom{s}{x} t^x (1 - t)^{s-x} \qquad (12)$$

where $x = 0, 1, 2, \cdots, s$. The number of sources is denoted by s.

On these three properties of theoretical traffic that satisfy the functions (6) or (12) rests the following statistical method for determining the busy hour and the busy-hour traffic.

Suppose we have at our disposal extensive statistical data for a specific group of outlets operating on a full-availability basis and with a negligible loss or delay. These data pertain to the number of busies observed at the same moments of the day, for example, at the hours 0.0, 0.1, 0.2, $\cdots$, 23.8, 23.9, and for a sufficiently large number of working days. The data pertaining to exceptional days such as Sundays and national holidays are discarded. The larger the number of days, the greater the degree of accuracy will be. The number of 10 days mentioned in the recommendation of the Comité Consultatif International Télégraphique et Téléphonique referred to in the above, should, therefore, be considered as a minimum.

When plotting the averages for each of these times of the day, a well-known fluctuating curve will be obtained that remains fairly flat between the hours of 9:00 to 12:00 and 14:00 to 17:00 o'clock. The period located at the point of culmination is now chosen for the busy hour and its ordinate for the value y of the busy-hour traffic. Consequently it is assumed that the traffic flow is caused by a "calling potential" that remains stable at the same hours of the different days, which stability applies to the subscribers individually and collectively.

In so doing, the value obtained for y is in close correspondence with characteristic (A). A minor variation of y during the busy hour seems acceptable.

Characteristic (C) is met by choosing, possibly with a few minor exceptions, the same busy hour for all the traffic flows in an exchange or area. This seems quite normal as exactly the same calls that contribute to the busy-hour traffics of merging groups will also be found among the merged traffic of the combined group.

Practice shows that a displacement of the busy-hour period by a quarter of an hour causes no appreciable change in the average value of y, so that coincidence of the various busy hours of the groups can readily be arranged. The traffic peaks that possibly appear

outside the adopted busy-hour period contribute to the average of the hours outside this period and, of necessity, to a lower average traffic value.

Characteristic (B) leads to a simple check of the variation of the observed numbers of simultaneous calls around the average adopted for the busy hour chosen. Equation (6) or (12) may serve for this purpose.

A different and widely spread interpretation of the concept of the busy hour and the determination of the busy-hour traffic is to use the average of the individual traffic peaks of the days observed, which peaks may occur at different hours. The value y thus obtained will be greater than that found in the manner proposed by us and as random hours are used a theory was developed permitting subtractions when traffic flows merge or additions when they divide. In Section 2, however, we have demonstrated that for both (6) and (12) the theoretical traffic obeys the simple law of straight addition or subtraction of the average traffic values, so that the latter interpretation of busy hour and busy-hour traffic are not in agreement with the theoretical concepts underlying the well-known distribution functions.

Directly associated with the busy-hour problem is that of the quality of service. At the present moment, it is a common understanding that the grade of service determined by $P = 0.001$ is very good, by $P = 0.01$ permissible, and by $P = 0.10$ rather poor. It might be suggested that we completely abandon the busy-hour concept and base the switch calculations on traffic data extending over the full period of 24 hours. The consequence of this suggestion would be that new standards of service would have to be introduced and the probability equations now in general use would have to make place for others based on variable expectation. The principle of statistical equilibrium would no longer apply and some function expressing the variation of the mathematical expectation would have to be adopted. For normal circumstances, such a function would have to acknowledge that the value of expectation would be negligible for some 12 hours of the day, increase during 3 to 4 hours, and remain about constant for approximately 8 to 9 hours. We know however, that only the last group of hours contributes to delay or to lost traffic so that the adoption of a variable expectation would in fact lead right back to the equivalent of a busy period. Thus, the use of traffic data relating to the full period of the day does not appear to be very attractive, useful, or necessary, and no better solution than the present busy-hour concept seems to exist.

For the great majority of cases, the busy-hour traffic obtained as indicated satisfies the three characteristic features of the theoretical traffic concept. In several instances, however, and especially in automatic long-distance service, matters are complicated by the difference in standard time that may exist between the interconnected areas. As no guiding rule is available, the solution of such cases must remain with the traffic engineers who are acquainted with the full facts.

7. Deriving Traffic Offered from Traffic Carried

If the groups of outlets are sufficiently large so that the loss is of minor value, the traffic offered may be assumed to be equal to the traffic carried by the group. The error thus committed is negligible both for the purpose of verification and measurement. If on the contrary the switching arrangement causes an appreciable loss, special methods are needed because the observations are of the traffic carried by the group whereas the usual loss equations include traffic offered as a parameter. Consequently the traffic offered must first be derived from the observed data by checking the specific numbers of calls simultaneously in progress against the probabilities calculated by (8).

In the following paragraphs, four distinct methods that have some theoretical basis are described.

7.1 Method 1

From the lost-call equation of Erlang (8), the following proportions can be obtained

$$y = ip(i)/p(i-1)$$

where $0 < i \leqslant N$.

Each of these N proportions provides a theoretical value for the traffic offered to the group. By calculating these proportions from the observed data, we obtain N values for y. Their average is equal to

$$y' = \frac{1}{N}\sum_{1}^{N}\frac{ip'(i)}{p'(i-1)}.$$

The primes indicate the measured and derived values, which correspond with their theoretical counterparts. A disadvantage of this method is that the proportions calculated do not have the same degree of accuracy as the numbers that are found in the neighborhood of the average group occupation. These are larger and have greater likelihood of being correct. It seems therefore necessary to weight the observed proportions with a suitable factor that is a function of the various numbers observed, the sum of which must be equal to 1. A reasonable assumption is to adopt factors that are in direct proportion to the observed proportions $p'(0)$, $p'(1)$, et cetera. The manner of weighting used may be supplemented by others, perhaps more suitable, depending on the judgment of the analyzer.

An advantage of this first method is that all observed proportions are utilized for the computation of the traffic offered to the group.

7.2 Method 2

This method is based on a comparison of the observed and the theoretical values of $p(\leqslant x)$, the probability of finding a maximum of x calls in progress simultaneously. From the observed data, $N+1$ values for $p'(\leqslant x)$ are calculated while the corresponding theoretical values have to be computed for the necessary sequence of traffic values.

The comparison exists in finding among the theoretical traffic values the one that most closely approaches the observed $p'(\leqslant x)$. The $N+1$ values for the traffic offered found in this manner may be averaged with or without weighting.

7.3 Method 3

The lost traffic is in direct proportion to the time during which the group is fully engaged. The measured portion of the busy hour during which the group is fully engaged is compared with the corresponding data as, for example, those contained in the tables prepared by Palm (1947). In this manner, only one value for the offered traffic is obtained.

The disadvantage of this method is that only one of the numbers observed is utilized for the evaluation of the offered traffic. Application of this method to the other frequencies may lead to anomalies.

7.4 Method 4

This method is based on the comparison among the various values of $p'(x)$ calculated from the observations and their theoretical counterparts as these have been listed in the tables prepared by Palm (1947). For a sequence of traffic values estimated to cover the range of the traffic offered, the differences $p'(x) - p(x)$ are calculated. Calculation of the standard deviation and skewness from these differences provides two values for y', the one when the standard deviation is a minimum, the other when the skewness equals zero.

When the calls of the observed group of outlets that are not served immediately are delayed, the traffic carried by the group is equal to the traffic offered, at least from a theoretical point of view. It would appear, therefore, that the problem dealt with in this section does not exist for switching arrangements causing delays. In such instances, however, it is advisable to proceed with care as groups of outlets that are underswitched cause a strangling effect on the offered traffic. The

subscribers know by experience that such a situation exists and consequently use other means of communication instead, such as the telegraph or even the mail. Instances can be cited where with manual toll service the lines of a certain direction were doubled and redoubled while each time the delay, after a temporary drop, climbed back to its original value.

Several other influences may render a reasonable forecast difficult, such as a change in the structure of the rates, the changing from flat rate to measured rate, the introduction of cheaper night rates, and the change from 3- to 1-minute rate units for toll calls. Such changes exercise an unpredictable effect on the importance and the characteristic features of telephone traffic.

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Slow-Wave Structures for Millimetre-Wavelength Backward-Wave Oscillators

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1. List of Symbols

ω = angular frequency
v_p = phase velocity
v_g = group velocity, $d\omega/d\beta$
c = speed of light
β = propagation constant, ω/v_p
p = pitch of slow-wave circuit
α = attenuation constant
E_z = radio-frequency electric field in the direction of the beam velocity
P = power propagating along the circuit
K = interaction impedance, $E_z^2/2\beta^2 P$
I_0 = direct-current beam current
V_0 = direct-current beam voltage
C = Pierce's gain parameter, $\{K(I_0/4V_0)\}^{1/3}$.
η = efficiency

2. Introduction

Microwave electron-beam devices may be classified as to whether the interaction is confined to one or a series of short regions or whether it takes place over extended regions. In the first type, the interaction effect is enhanced by the use of resonant cavities, which may be regarded as effecting a repeated use of the same packet of electromagnetic energy. In the second type, a large effect is obtained by allowing the interaction to continue for a long time, corresponding to many periods of the wave. For this class, which includes the travelling-wave tube, backward-wave oscillator, carcinotron, and linear accelerator, the electromagnetic wave must be guided by a slow-wave structure in which the wave velocity closely matches the speed of the electrons.

Whereas in general it is a relatively simple matter to decide on the desiderata in the case of cavities, in which a single parameter, the shunt impedance, could in many instances be regarded as a figure of merit, in the case of slow-wave structures the situation is considerably more complex. One important characteristic is described by the interaction impedance, which is a measure of the intensity of the useful electric field per unit power propagating along the structure [1]. This impedance corresponds roughly to the shunt impedance of a resonant cavity. In addition one must, however, also consider the bandwidth over which the slow-wave circuit can operate, the attenuation, the ability of the circuit to dissipate large powers, and the ease with which it can be matched to a waveguide over its operating frequency band. Inevitably compromises must be sought, and the weight placed on any one of the desired properties depends on the application. At millimetre wavelengths, the relative importance of various factors is totally different from that which obtains at longer wavelengths. In particular, the minimization of the attenuation becomes of primary importance. To illustrate this point, consider a backward-wave oscillator as indicated in Figure 1. The radio-frequency current along the beam varies approximately sinusoidally, the amplitude of the wave on the circuit approximately cosinusoidally [2]. The radio-frequency power is generated primarily where both the current and the field are strong, that is, in the middle of the tube. This power must then pass through half the total attenuation of the tube before reaching the output guide. It is not uncommon at millimetre wavelengths for the attenuation to amount to 20 decibels whereas at a wavelength of a few centimetres it is not likely to be very significant. It can be shown [3] that if the attenuation of the slow-wave circuit is too high, it is not possible to obtain oscillation, irrespective of the length of the circuit.

The main purpose of this paper is to present an approach to the design of slow-wave circuits for the particular circumstances encountered at millimetric wavelengths and to describe some of the circuits that have been studied. For reasons of available space, we present the work

without the mathematical analysis on which it is based but wherever possible appropriate references have been included. Section 3 presents some of the considerations that govern the choice of the group velocity of the slow-wave circuit. Since both the interaction impedance and the attenuation are inversely proportional to this velocity, it is a very-critical design factor. An outline of an analysis aimed at the determination of the actual losses is also included. Section 4 deals with the possibilities of utilizing slow-wave structures that extend in three dimensions. In Section 5 some results relating to particular millimetric slow-wave circuits are discussed.

3. Choice of Group Velocity

In a backward-wave oscillator a certain relation must be satisfied between the circuit loss α and the gain parameter C before oscillations will commence. The larger α, the larger the required value of C. α depends on the material of the circuit, its geometry, and the group velocity v_g of the wave on the circuit. C depends on the beam impedance and the interaction impedance, which in turn depend on the geometry of the circuit, the group velocity of the wave on the circuit, and the phase velocity v_p of the wave component interacting with the beam. In designing a tube for a specific application, many of these factors may be largely prescribed; for example, the beam impedance and phase velocity may be determined by the power level required and by the highest perveance considered practicable for the electron gun. However, we have to decide whether to aim for a high or a low group velocity. If we require a tube that is tunable over a wide range, there is no choice; a high group velocity is inevitable. If, on the other hand, oscillation is required over only a narrow frequency band we are free to consider whether there is any advantage in reducing the group velocity. Since K is inversely proportional to v_g, an increase in the interaction impedance is effected and so C is increased (C^3 being inversely proportional to the group velocity). However, when we remember that the circuit loss α is also inversely proportional to the group velocity, it is not obvious whether we shall gain or lose by so doing.

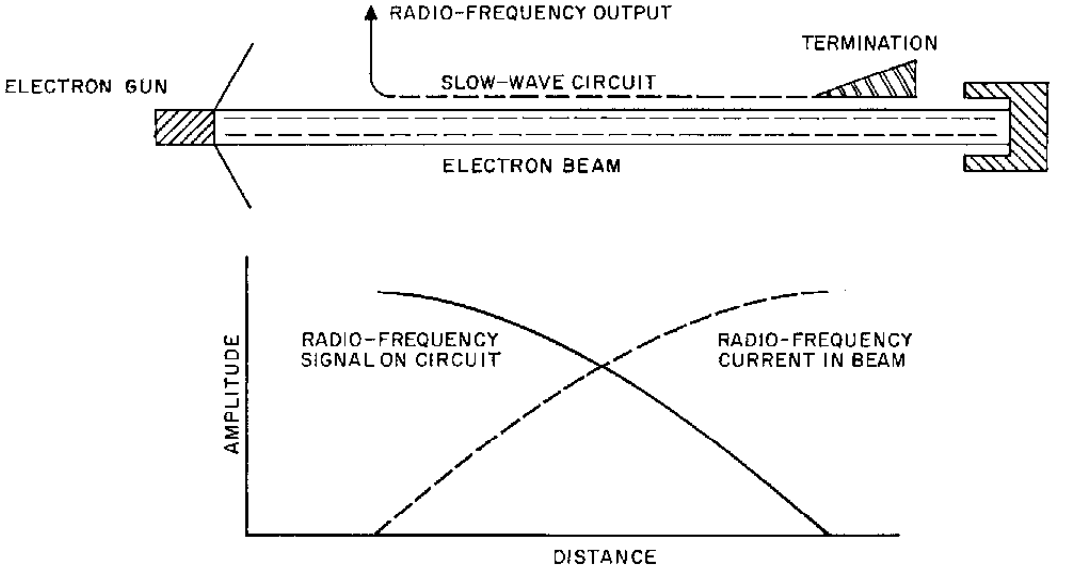


Figure 1—Variation of radio-frequency signal and beam current in a backward-wave oscillator.

If we let $\alpha = k_1/v_g$ and $C^3 = k_2/v_g$, we can summarize the situation with a single curve in which $N/(k_1/k_2)^{1/2}$ is plotted against $v_g(k_1^3/k_2)^{1/2}$, where N is the minimum length of circuit measured in guide wavelengths necessary to obtain oscillation. The numerical evaluation of the expression obtained required a knowledge of the starting conditions for a backward-wave oscillator in the presence of loss. This information is to be found in a paper by Johnson [4] and was used in the computation of the curve in Figure 2. It is seen that there is a minimum value for N and hence an optimum value for v_g. Although k_1 and k_2 do not contain v_g explicitly, they are dependent on the geometry of the circuit that in turn controls the group velocity. Thus if we attempt to change the group velocity, in order to optimize its value, the values of k_1 and k_2 will also change so that an iterative procedure is called for. Nevertheless, we have established a criterion that indicates in which direction the group velocity should be changed and how far it is from optimum.

So far we have only considered the conditions for start of oscillation and, when trying to work at the very-highest frequencies, this must be the first consideration. Eventually, however, we shall become interested in the efficiency and want to know how to optimize it with respect to group velocity. Grow and Watkins [5] have given an approximate expression for the efficiency of a backward-wave oscillator $\eta = C\exp[-0{\cdot}115\,N\alpha]$. In deriving this expression they have assumed that the interaction is sufficiently vigorous for the radio-frequency current in the beam to reach saturation and, provided this is achieved, there is no optimum value of N; the shorter the circuit the better. The group velocity that gives maximum efficiency can, however, be shown to be equal to $0{\cdot}345\,k_1N$, which is not the same as that giving the minimum starting length. To design for maximum efficiency, we must fix the starting current, relative to the working current, so that radio-frequency beam saturation is ensured. Then N and v_g must be selected to satisfy the condition for optimum efficiency and simultaneously the condition for start of oscillation given by Figure 2. The starting current is used to calculate k_2 so that N and v_g correspond to a point on the curve of Figure 2, although this will not now be the point of minimum N.

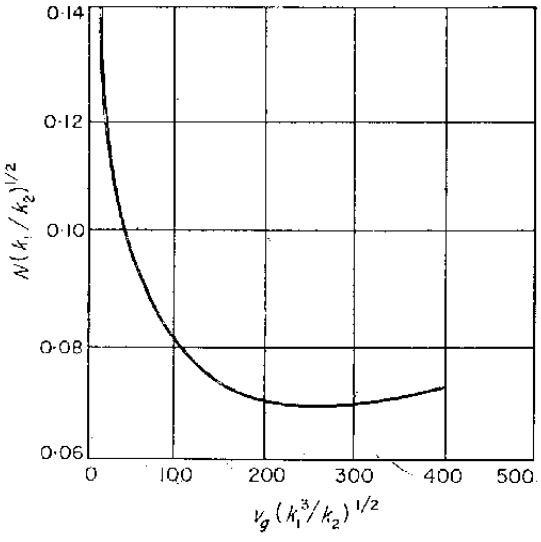


Figure 2—Relation between starting length, loss, and group velocity.

To use these expressions, we must be able to calculate α and the interaction impedance, which in turn give us k_1 and k_2. In general this is not an easy problem because of the difficulty of determining the field distribution in slow-wave structures. However, Fletcher [6] has provided a solution to this problem for structures that comprise (or may be approximated by) arrangements of parallel bars of rectangular cross-section, and interaction impedances may be readily obtained by this method. In principle it also permits the loss of such structures to be calculated, but in practice this is more difficult. Whereas in the calculation of impedance it is necessary to determine the total current flowing along a bar of the structure, to obtain the loss the square of the current has to be integrated over the surface of a bar. Difficulties arise at the corners of the bars where the field solutions are approximate and where in the

idealized model a current infinity occurs. The latter gives rise to some mathematical difficulties in the loss calculations but these can be overcome. However, since the loss is dependent on the way the current is distributed over the surface of the bar while the impedance depends only on the total current, the approximations in the field solution have a much greater effect on the losses than on the impedance. Nevertheless, values of α have been computed for a wide range of circuit geometries and it is considered that these are sufficiently accurate to provide a considerable measure of guidance in calculating the optimum group velocity.

The importance of this problem of optimum group velocity can be illustrated by showing how its value changes with wavelength for a typical circuit. In the case selected the circuit is of molybdenum, the beam current density is 1·5 amperes per square centimetre, and a fundamental backward wave is assumed. The optimum values of c/v_g and N to minimize the starting current at a particular frequency are given in Table 1. From the figures it can be

TABLE 1

λ in millimetres	c/v_0	N
5	70	12
2·5	20	29
1·25	6	69

seen that at a wavelength of 5 millimetres the optimum group velocity is embarrassingly low but at 1·25 millimetres it is very high so that for a non-dispersive circuit ($v_p = v_g$) the beam voltage for interaction would be 7 kilovolts.

Thus we can see that the design of a millimetric backward-wave oscillator for a narrow band requires a balance between the demands of high interaction impedance and low circuit loss. The optimum design can be established provided these two quantities can be calculated and this can be done for ladder structures by an extension of Fletcher's analysis.

4. Multiple Slow-Wave Circuits

In the previous section we discussed the choice of group velocity to optimize the efficiency of a backward-wave oscillator. Even with this optimum choice, and with a circuit configuration having the highest attainable interaction impedance, the maximum output power is still severely limited by the heat-dissipation ability of the circuit. At millimetre wavelengths, the fields decrease so rapidly with distance from the slow-wave circuit that, for example, for a tube operating at 6 millimetres wavelength and a beam voltage of 2·6 kilovolts, it is only the current within 0·1 millimetre of the circuit that makes any appreciable contribution to the power. The presence of thermal velocities in the beam and the fact that the available focussing field is limited, means that the beam edge must be ill defined; some interception of current is therefore inevitable. This problem is particularly acute at millimetre wavelengths for yet another reason. To obtain enough interaction, it is essential to use very-high current densities in the beam. To attain these, the current from the cathode must be compressed. This has the effect of raising the effective electron temperature, so that the edge of the beam is far less well defined than one would expect in a tube operating at longer wavelengths.

One way of obtaining greater powers would be to use several tubes in parallel. This would, however, involve considerable difficulties in coupling and probably insuperable difficulties in ensuring phase coherence between the separate oscillators. A far-more-elegant and promising approach is to combine several slow-wave circuits internally to form one multiple structure. A single electron-optical system can then provide the electron beam that interacts with the whole array. This principle is illustrated diagrammatically in Figure 3 for the case of ladder structures (to be discussed in the following section). In principle there is no limit to the number of circuits that can be combined in this way. In practice there are, of course, limitations arising from the difficulty of joining all

the circuits to a single output waveguide and also arising from the need to ensure that all the circuits oscillate in phase.

The conditions that must be met in a multiple structure to achieve phase coherence among all its constituent parts have been analysed and studied experimentally in a number of papers [7, 8, 9]. The conclusions reached in these studies may be summarized by saying that the tolerance errors on the various subcircuits must be held to very-tight limits, but that at wavelengths above about 3 millimetres these limits should be attainable. A second conclusion is that the problem of phase coherence is eased as the efficiency of the tube is increased. This means that the chance of successful multiple-circuit operation is greatest in a tube in which the current density and power output on each circuit are as high as technological limits will permit.

The question of greatest interest is naturally how much can be gained by resorting to the multiple-structure idea. If we incorporate M slow-wave circuits in a multiple array, we would expect to be able to get at least M times the power output obtained from a single circuit. In fact the possible improvement is much greater than suggested by this consideration. The improvement factor that can be anticipated has been investigated by a number of authors [10, 11] and the results depend to some extent on the assumptions that are made in the comparison. However, for the circumstances that apply in the case of a millimetre-wavelength backward-wave oscillator, the result [11] is as indicated in Figure 4. It is seen that the output that can be expected per circuit relative to that provided by a single isolated circuit depends on the distance between the circuits in the array. The improvement reaches a maximum value of just over 3 at a critical spacing and beyond this spacing falls rather rapidly to unity. On this basis one would therefore expect that the power output from M circuits would exceed that from a single circuit by a factor $3M$. Although the result, Figure 4, is based on a somewhat idealized model, a factor of around $2{\cdot}5M$ can be anticipated. We see therefore that quite a moderate number of circuits in a multiple array can lead to substantial increases in power output. So far, there have not been any reports of true multiple circuits operating at millimetre wavelengths. The advantages that can be anticipated are, however, sufficiently great to encourage attempts to realize such circuits. It is for this reason that our efforts to devise suitable circuits for the millimetre band have been concentrated on the ladder type, as this has a geometry that is particularly suitable for assembly into multiple arrays.

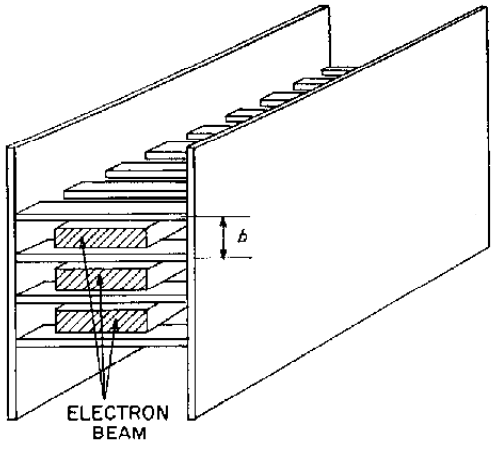


Figure 3—Multiple ladder structure.

5. Ladder Circuits

Figure 5A illustrates a simple ladder structure, the prototype of a class, which has a number of desirable features for use at millimetre wavelengths. In the first instance, it is a very-well-cooled structure. Under normal conditions, the spacing between the sidewalls is of the order of ½ free-space wavelength, so that no point on the circuit is more than ¼ wavelength away from a substantial metallic heat sink. The question of how much power can be safely dissipated on such a circuit has recently been analyzed [12] by solving the appropriate heat-flow equations. The problem is complicated by the fact that we are concerned with large temperature

differences, so that effects due both to radiation and to the variation of conductivity with temperature may be important. It was, however, possible to obtain useful approximations to the non-linear equations, and thus to derive quantitative results for the power dissipation capability. Applying these results to the case of a backward-wave oscillator centred at a 6-millimetre wavelength and making a number of plausible assumptions, one is led to the conclusion that beam power densities of 100 kilowatts per square centimetre should be attainable. This figure is several orders of magnitude higher than would apply to the case of a simple helix.

The possibility of stacking ladders to form a multiple array has already been mentioned; however, the ladder has yet a further appeal for the millimetre band; it is relatively easy to fabricate. It has already been mentioned that the ohmic loss of a slow-wave structure is inversely proportional to the group velocity, and that losses tend to be a dominant factor. There is another kind of loss that may become significant; this is a statistical random reflection loss due to irregularities in the structure. It can readily be shown that this loss also is inversely proportional to the group velocity. The avoidance of small errors is therefore of great importance at millimetre wavelengths, and this discourages any kind of complexity in the choice of structure. A completely planar circuit such as the ladder is amenable to a number of constructional techniques that work satisfactorily for small dimensions. Among these are precision winding, photoetching, and spark erosion machining. We have experimented mainly with the latter two and believe that good circuits can be made with their aid, at least down to 2 millimetres wavelength.

So far we have not mentioned one vital disadvantage of the ladder: in its simplest form, as shown in Figure 5A, it does not propagate a slow wave. If one regards each rung of the ladder as a resonator, it is at first sight surprising that an array of such resonators, which are certainly coupled to each other, do not give rise to a frequency band over which propagation is possible. The explanation that emerges from analysis is that the resonators are indeed coupled, but both magnetically and electrically, the effects being equal and exactly opposite. This also gives a clue to how a pass band can be obtained. It is necessary only to alter either the electric or magnetic coupling, in almost any manner, and propagation over some band will become possible. The simplest way of perturbing the basic ladder was first suggested by Karp [13], who added conductors adjacent to the ladder, either in the centre or at the edges (Figures 5B and 5C). These circuits have been described and analyzed by a number of authors [14, 15] and they have been remarkably successful in practice. They are used in many commercial tubes and have led to the first successful oscillator below 2 millimetres wavelength. They are not, however, suitable for the purpose that formed our primary aim, which is the realization of structures that can be stacked to form multiple arrays. The necessity of providing adjacent conductors would prevent the assembly of more than two ladders.

Our aim has then been to devise modified ladder structures that propagate by virtue of their

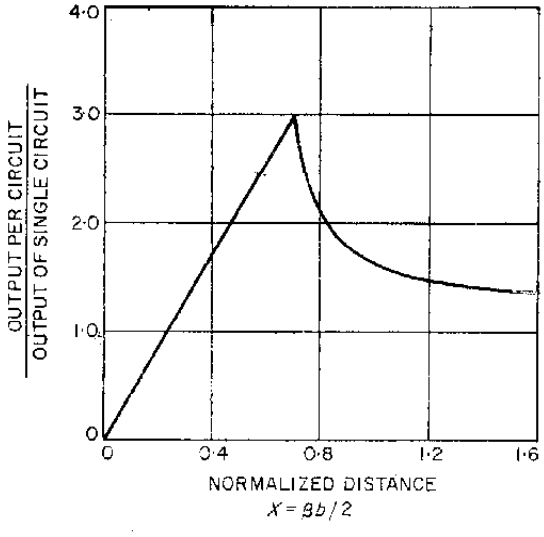


Figure 4—Normalized output and ladder separation.

own topology rather than by the addition of external conductors. The circuits may be divided into two groups according to whether they are suitable for broad- or narrow-band operation, where we will choose the dividing line at bandwidths of 20 per cent.

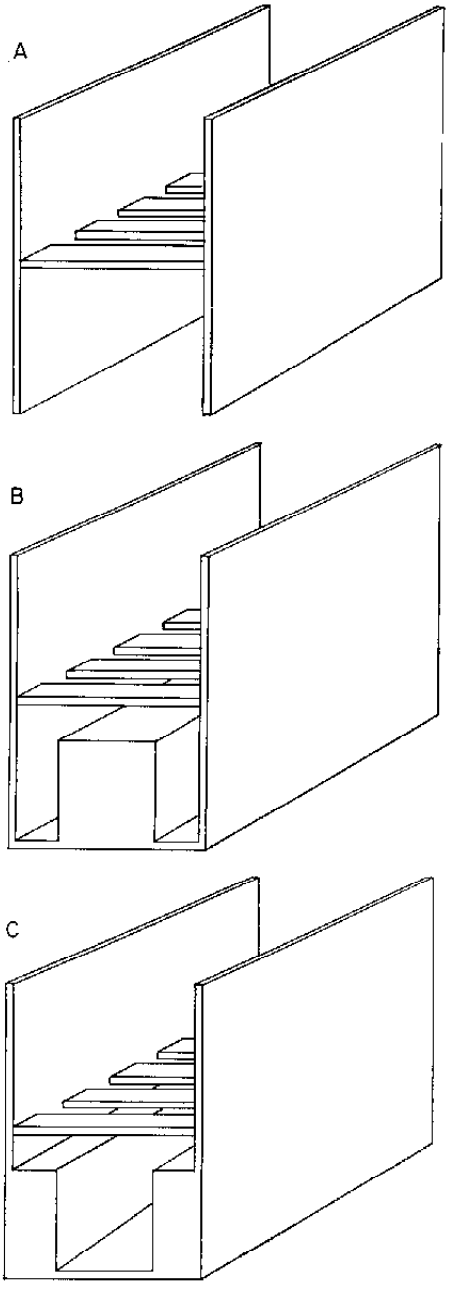


Figure 5—A, simple ladder structure; B, Karp structure; and C, Anti-Karp structure.

5.1. Narrow-Band Ladder Circuits

A very simple perturbation of the basic ladder has been described by White, et al. [16]. Here the slots extend over only a portion of the distance between the two planes, and this produces the dispersion curve *a* shown in Figure 6. The essential point is that a fundamental forward wave results and a negative space harmonic could be used in backward-wave operation. Unfortunately, the effective bandwidth is rather small, particularly at relatively low voltages (say below 10 kilovolts). This circuit is not therefore very attractive for our purpose, although it is probably a very good one for a narrow-band forward-wave tube operating at high voltages.

By far the most successful structure in the attainment of the highest powers at the shortest wavelengths has been that described by Mill-

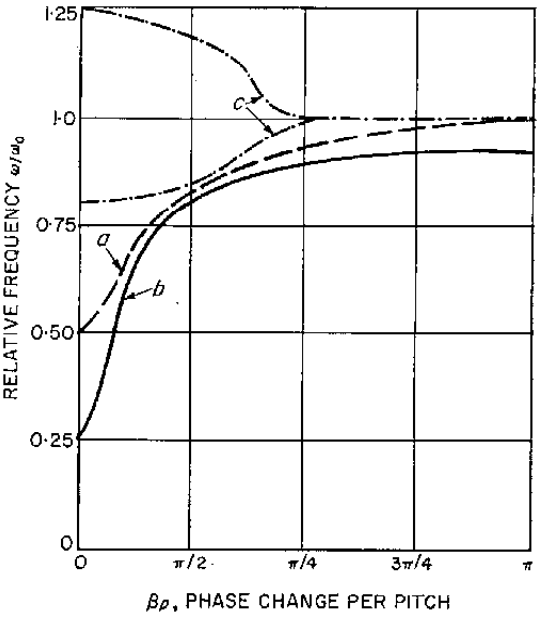


Figure 6—Dispersion curves for *a*, modified ladder; *b*, vane type; and *c*, crossed-wire circuit.

man [17], shown in Figure 7. This again is a very-simple modification of the basic ladder; in fact, it is half of such a ladder with the addition of a ground plane. Here it is the fringing field between the ends of the rungs and the ground plane that provides the perturbation that leads to the existence of a pass band. A typical dispersion curve is shown in *b* of Figure 6. Here again a backward space harmonic is required to provide backward-wave interaction. The available bandwidth is rather small —typically of the order of 10 per cent—but over a narrow band tubes using such a structure have given outputs of over 1 watt at a wavelength of 2 millimetres and have oscillated at wavelengths below 1 millimetre.

Yet another variant of the basic ladder is shown in Figure 8. Here the necessary perturbation has been obtained by skewing the rungs. If the rungs of a single ladder are skewed, very little pass band is obtained. However by utilizing two ladders with complementary angles of skew, a very-much-wider pass band is produced. This circuit was amenable to analysis, by using an approximate model in which the ladders are replaced by inhomogeneously conducting sheets. Using a variational method, the dispersion equation could be derived. A typical dispersion curve is shown in *c* of Figure 6. This circuit has both a forward as well as a backward fundamental mode. The existence of a backward fundamental mode means that the circuit will have a high interaction impedance when used in a backward-wave oscillator. Experiments at *X* band showed that oscillation over a 10-per-cent band was possible. These experiments were carried out with a very-weak electron beam in a demountable vacuum system. With a properly designed gun, somewhat greater bandwidths would result. The method of alternating the angle of skew can be extended to more than 2 circuits to form a larger multiple circuit. Analysis shows that this leads to some increase in the bandwidth. Experiments with up to 5 circuits and 3 beams carried out at *X* band demonstrated that multiple operation with this circuit was possible, and there appeared to be no difficulties with the excitation of spurious modes. These circuits have not yet been tried out at millimetre wavelengths, but it is felt that for a narrow-band multiple circuit they should provide a high performance.

5.2. Broad-Band Ladder Circuits

For a number of purposes, of which the provision of tubes suitable for a broadband waveguide communication system might be cited as an example, it would be very desirable to have an oscillator capable of covering a much-wider

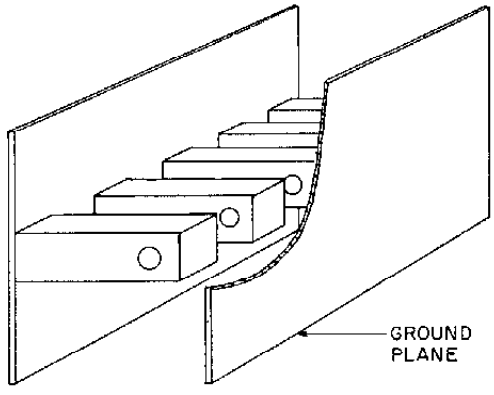


Figure 7—Vane-type circuit for narrow-band backward-wave oscillator.

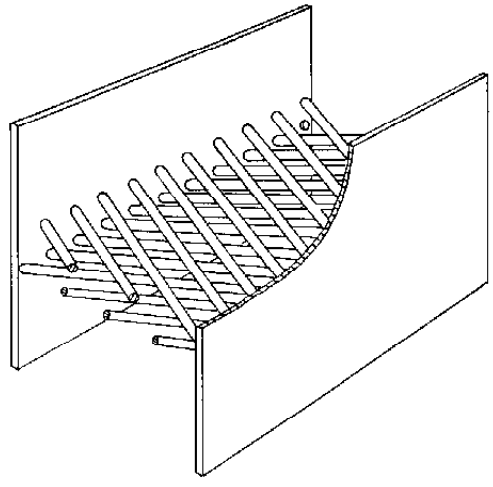


Figure 8—Crossed-wire circuit.

band than the circuits previously discussed. One circuit that will give such broad-band performance is the helix, but this has a very-poor thermal dissipation capability. A second possibility is the interdigital line, which is superior in this respect. However, this is mechanically very-much harder to make, and especially as there are free fingers that are not directly restrained, the necessary tolerances are difficult to attain.

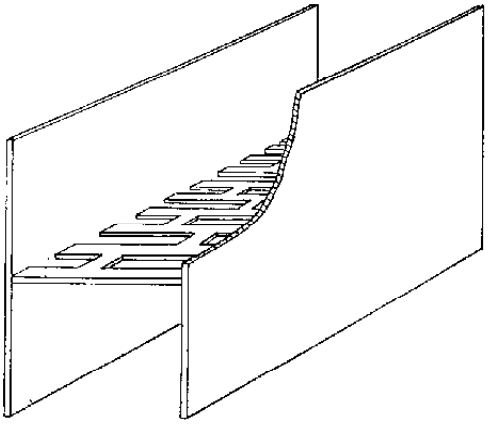


Figure 9—Stub-supported meander line.

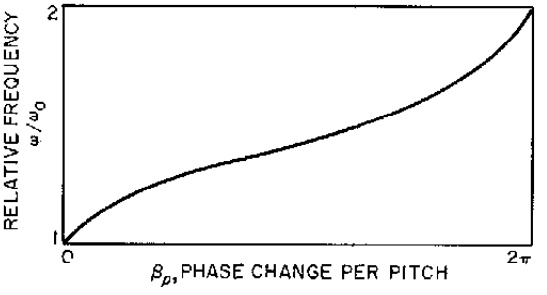


Figure 10—Dispersion curve of stub-supported meander line.

It therefore seemed desirable to devise an all-metal circuit in which all bars are directly connected to massive heat sinks and with no free ends. The circuit shown in Figure 9 was found to have many of the desirable properties. It was analyzed using the simplified field theory first suggested by Fletcher [6]. In this approach, the actual field is approximated by the superposition of a number of transverse electromagnetic waves travelling in the direction of the rungs, that is, at right angles to the direction of propagation of the slow wave. This analysis [18] led to dispersion curves such as that shown in Figure 10, which were in reasonable agreement with experiment except in the immediate vicinity of the cut-off frequencies. The discrepancies there are a result of the approximations in the theory.

The dispersion curve is seen to represent a forward fundamental mode, although here, as in the crossed-wire structure, a backward fundamental also exists. The most-important feature of this curve is that it extends without break over a complete octave. Naturally it is not possible to operate too near the cut-off frequencies, but nevertheless, in demountable experiments at X band, working on a backward space harmonic, an oscillation bandwidth of 1·7:1

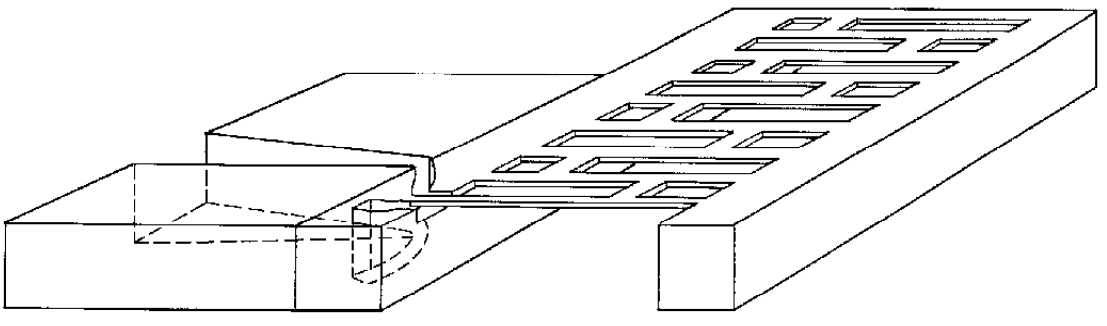
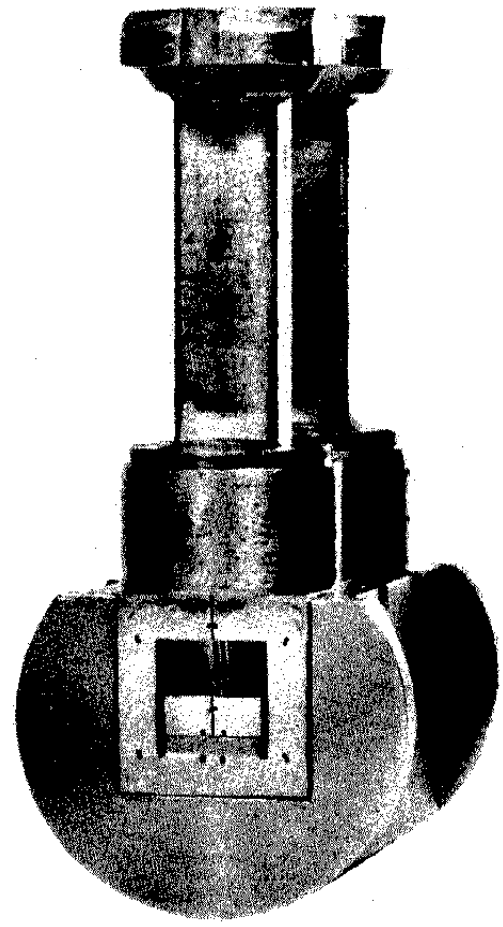


Figure 11—Stub-supported meander-line-to-waveguide transition.

Figure 13—6-millimetre meander-line assembly with two waveguide transitions.



was measured. Here again the electron beam was very weak; in a suitably designed tube an even-wider bandwidth might be obtained.

The analysis enables one to predict how the impedance of this structure will vary with the thickness of the rungs of the ladder. To choose this dimension, it is now necessary to take account of how the losses, the importance of which has been stressed in Section 3, vary with thickness, and also to estimate the required thermal dissipation. Finally there is the question of whether for a particular design the required thickness can be realized in the construction. The thickness is severely limited with photoetching methods. However, the spark erosion technique permits the use of very-much-thicker ladders, and in most cases the optimum thickness for a particular design can in fact be adopted.

One of the main problems in attaining large bandwidths is the difficulty of matching the circuit to a waveguide. The essence of the method adopted is indicated in Figure 11 and the results obtained are shown in Figure 12. This very-broad-band match was achieved by utilizing the rapid variation in impedance of the output guide close to its cut-off frequency.

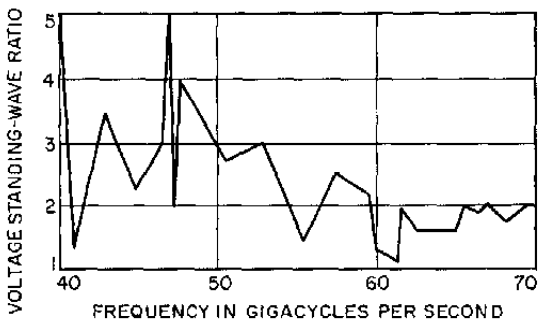


Figure 12—Typical measured performance of stub-supported meander-line structure.

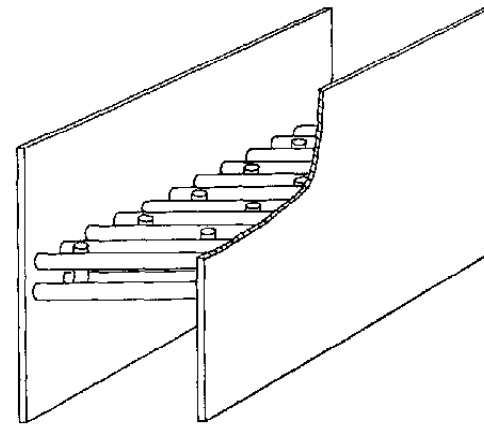


Figure 14—Double meander circuit showing stacked rod construction.

A circuit using this match has been incorporated in a backward-wave oscillator with a band centre at 6 millimetres. A photograph of parts of the assembly is shown in Figure 13.

The circuit of Figure 9 can be assembled into a multiple array in the manner indicated in Figure 3. There is, however, an alternative approach indicated in Figure 14 in which there is actual metallic contact between the various ladders. This will tend to increase the coupling, which is very desirable to ensure the phase coherence of the whole array. The circuit has a dispersion curve similar to that shown in Figure 10 but no hot experiments have so far been attempted.

6. Conclusions

The importance of attenuation as one characteristic that governs the choice of slow-wave circuits at millimetric wavelengths has been emphasized. Both the attenuation and the interaction impedance are inversely proportional to the group velocity. The choice of the group velocity is therefore a critical design decision; a method for choosing it so as to minimize the starting current, or alternatively so as to maximize the efficiency, is outlined.

The advantages of ladder structures for millimetric tubes are discussed, particularly with regard to their incorporation in a multiple array. Analysis has shown that a 4-ladder array should provide an output power at least an order of magnitude higher than that from a single ladder structure. The advantages of a number of specific ladder structures are compared. A particular variant, having an exceptionally wide pass band, has been studied in some detail, and a method of achieving a satisfactory match to an external waveguide is indicated.

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Mr. Froom is an Associate Member of the Institution of Electrical Engineers.

United States Patents Issued to International Telephone and Telegraph System; February–April 1962

Between February 1962 and April 1962, the United States Patent Office issued 54 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

B. Alexander, M. Press, and J. Murgio, ITT Laboratories, Collision Avoidance System, 3 025 514.

M. A. Argentieri and F. E. Lind, ITT Laboratories, Adjustable Delay Line, 3 020 497.

E. A. Ash and J. D. Pattenden, Standard Telecommunication Laboratories (London), Coupled Waveguides, 3 020 498.

R. W. Blanchard and G. M. Nonnemacher, ITT Laboratories, Automatic Visual Flight Information Display System, 3 029 304.

E. L. Bush, Standard Telecommunication Laboratories (London), Purification of Gases used in the Production of Silicon, 3 029 135.

K. W. Cattermole, Standard Telecommunication Laboratories (London), Electric Pulse Modulating and Demodulating Circuits, 3 020 349.

R. F. Chapman, ITT Laboratories, Automatic Gain Control System, 3 025 473.

P. Cheilik, Federal Telecommunication Laboratories, Transistor Multivibrator, 3 020 417.

W. Clement, Mix & Genest Werke (Stuttgart), Method of Sorting-Out Parts of Insulating Material, 3 020 409.

D. W. Davis, ITT Federal Laboratories, Information Storage Display Tube and Storage Screen Assembly Therefor, 3 031 597.

C. L. Day, Farnsworth Electronics Company, Method of Making Target Electrode for Barrier-Storage-Grid Tube, 3 020 622.

C. L. Day, ITT Laboratories, Method and Apparatus for Assembling Photo Tubes, 3 026 163.

K. De Brosse and E. Behun, ITT Laboratories, Cooling System, 3 025 680.

J. Donceel and M. Mus, Bell Telephone Manufacturing Company (Antwerp), Electrical Multiple Relay Unit Using Sealed Reed Contacts, 3 030 468.

J. J. Econom, L. F. Evans, and G. Reade, ITT Federal Laboratories, Feeding and Support Assembly, 3 028 943.

H. Endres, Mix & Genest Werke (Stuttgart), Automatic Character Recognition, 3 025 495.

R. C. Ferrar, Federal Telecommunication Laboratories, Signal-Level-Recording Receiver, 3 029 336.

A. N. Gulnick and R. K. Van Vechten, ITT Laboratories, Tape Transports, 3 029 006.

W. H. Heiser, Federal Telecommunication Laboratories, Frequency-Controlled Switch, 3 029 310.

R. A. Hill, Standard Telecommunication Laboratories (London), Electrolytic Capacitors, 3 029 370.

R. C. P. Hinton, Federal Telecommunication Laboratories, Party-Line Detector Connector, 3 029 314.

G. Hirschfeld, W. Hinz, and H. Fritzsche, Mix & Genest Werke (Stuttgart), Sorting Equipment, 3 029 944.

G. D. Hulst, Federal Telecommunication Laboratories, Reflected-Binary-Code Counter, 3 020 481.

S. M. Khanna and M. C. Vosburgh, ITT Laboratories, Radar Mapping Simulator, 3 028 684.

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H. E. Krefft, Kuthe Laboratories, Cathode Structure, 3 025 428.

E. J. Leonard, ITT Kellogg, Multidigit Electrical Door Lock, 3 024 452.

M. Losher, Federal Telecommunication Laboratories, Radio Navigation System, 3 020 545.

L. F. Mayle, ITT Federal Laboratories, Dynamic Focusing, 3 021 073.

A. E. Nashman, ITT Laboratories, Vehicle Suspension and Stabilizing System, 3 029 089.

S. B. Ost and B. Dzula, Federal Telecommunication Laboratories, Party-Line Substation Identification System, 3 025 354.

M. Padalino, ITT Laboratories, Signal Identification System, 3 025 352.

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W. H. P. Pouliart and G. Van Mechelen, Bell Telephone Manufacturing Company (Antwerp), Sorting Machine Using Markers, 3 028 961.

A. J. Radcliffe, Jr., M. Ribner, and W. V. Sayner, ITT Kellogg, Relayless Line Circuit, 3 029 315.

L. J. Regis and W. E. Richeson, Jr., Farnsworth Electronics Company, Voltage Comparator, 3 021 514.

A. G. Richardson, ITT Federal Laboratories, Magnetic Recording and Reproducing Apparatus, 3 020 358.

D. S. Ridler and A. D. Odell, Standard Telecommunication Laboratories (London), Intelligence Storage Equipment, 3 020 526.

W. L. Roberts and E. De Faymoreau, ITT Federal Laboratories, Strobing Circuit, 3 030 620.

M. Rogoff, Federal Telecommunication Laboratories, Phase-Measuring Device, 3 020 478.

H. W. G. Salinger, ITT Laboratories, Image Intensifier, 3 030 514.

P. C. Sandretto, Federal Telecommunication Laboratories, Loran Receiver, 3 025 518.

R. G. Schriefer, Capehart-Farnsworth, Automatic Tracking System, 3 020 537.

W. Sichak, ITT Federal Laboratories, Sideband Generator, 3 029 396.

W. Sichak, ITT Laboratories, Diversity Combining System, 3 029 338.

K. J. Staller, ITT Laboratories, High-Current Rectifier, 3 025 436.

F. Steiner, C. Lorenz, Doppler-Type Direction-Finder, 3 025 522.

R. K. Van Vechten and R. T. Adams, ITT Federal Laboratories, Signal Translating Device, 3 020 416.

E. P. G. Wright and J. Rice, Standard Telecommunication Laboratories, Equipment for Performing a Complex Sequence of Operations, 3 025 351.

E. P. G. Wright and J. Rice, Standard Telecommunication Laboratories, Intelligence Storage Equipment, 3 024 993.

E. P. G. Wright and J. D. Reynolds, Standard Telecommunication Laboratories, Static Electric Switches, 3 024 448.

E. P. G. Wright, Standard Telecommunication Laboratories, Data-Processing System, 3 020 336.

E. P. G. Wright, D. A. Weir, R. C. Price, and B. Dzula, Standard Telecommunication Laboratories and Federal Telecommunication Laboratories, Data Processing Systems, 3 025 341.

J. Young, Bell Telephone Manufacturing Company (Antwerp), Sorting Machine, 3 019 896.

Data-Processing Systems

3 020 336

E. P. G. Wright

This patent concerns the receiving matrix for the Strad switching system. It provides for a signal store for each line, common central apparatus for all lines, and a plurality of central stores. Apparatus controlled by a signal on a line causes a scanning of the stores and sequential transfer of information from an incoming line to an outgoing line.

Data-Processing Systems

3 025 341

E. P. G. Wright, D. A. Weir, R. C. Price, and B. Dzula

For the Apogee switching system, this patent covers the main storage equipment, the transfer of messages to the main store in accordance with an established precedence of the messages and in the order of their arrival, and their transfer to the outgoing store. Means are provided with control to transmit the messages over their selected outgoing lines in accordance with their precedence and order of their storage in the incoming line stores.

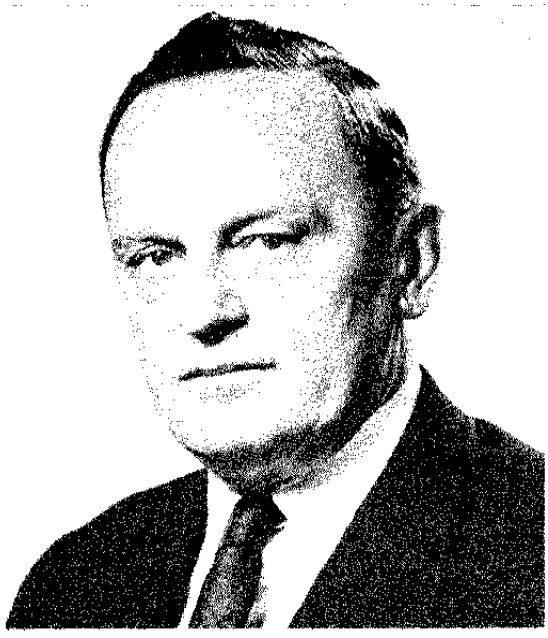
Sorting Machine

3 028 958

W. H. P. Pouliart and G. Van Mechelen

For collating documents, a sorting machine having only two sorting stages has been developed. The articles being sorted are compared and directed to one of two outputs in accordance with their rank. The outputs of each stage become the two separate inputs of the other stage. This enables the two stages to provide as many sorting passes as may be required.

In Memoriam



WILLI GRUBE

Willi Grube, member of the board of directors of Standard Elektrik Lorenz and general manager of the Mix & Genest Division, died unexpectedly on his 63rd birthday, 8 December 1962.

He was born in Bremen, Germany, received the Dr.-Ing. degree from the Berlin Institute of Technology, and then served at this Institute as an assistant for ten years, until 1934. In the years following, he was a member of the signal corps of the armed forces.

He joined Mix & Genest in 1947, during which year he became manager of the technical office in Hamburg and then manager of government sales. In 1950, he was promoted to director and two years later was elected to the board of directors of Mix & Genest and entrusted with the sales management of the company.

In 1954, he was appointed to the board of directors of Standard Elektrik Lorenz. In 1959, he became general manager of the Mix & Genest Division and manager of domestic sales of Standard Elektrik Lorenz. He contributed essentially to the rise of Standard Elektrik Lorenz to its present position in the telecommunications field.

His contributions were not limited to Standard Elektrik Lorenz. A large number of committees and organizations profited from his extensive experience and sound advice. He was for many years a member of the Board of Chairmen of the Association of Telecommunication Industries within the Central Federation of the Electrotechnical Industry (ZVEI), and was elected chairman at the last membership meeting.

Dr. Willi Grube will be remembered for his pre-eminent personality and his dedication to his work.

Principal ITT System Products

Telecommunication Equipment and Systems

Automatic telephone and telegraph central office switching systems
Private telephone and telegraph exchanges—PABX and PAX, electromechanical and electronic
Carrier systems: telephone, telegraph, power-line
Long-distance dialing and signaling equipment
Automatic message accounting and ticketing equipment
Switchboards: manual, central office, toll
Telephones: desk, wall, pay-station
Automatic answering and recording equipment
Microwave radio systems: line-of-sight, over-the-horizon
Radio multiplex equipment
Coaxial cable systems
Submarine cable systems, including repeaters
Data-transmission systems
Teleprinters and facsimile equipment

Military/Space Equipment and Systems

Aircraft weapon systems
Missile fuzing, launching, guidance, tracking, recording, and control systems
Missile-range control and instrumentation
Electronic countermeasures
Electronic navigation
Power systems: ground-support, aircraft, spacecraft, missile
Radar
Simulators: missile, aircraft, radar
Ground and environmental test equipment
Programmers, automatic
Infrared detection and guidance equipment
Global and space communication, control, and data systems
Nuclear instrumentation
Antisubmarine warfare systems
System management: worldwide, local

Industrial/Commercial Equipment and Systems

Distance-measuring and bearing systems: Tacan, DMET, Vortac, Loran
Instrument Landing Systems (ILS)
Air-traffic control systems
Direction finders: aircraft and marine
Ground and airborne communication
Data-link systems
Inverters: static, high-power
Power-supply systems
Altimeters
Flight systems
Railway and power control and signaling systems
Information-processing and document-handling systems
Analog-digital converters
Mail-handling systems
Pneumatic tube systems
Broadcast transmitters: AM, FM, TV
Studio equipment
Point-to-point radio communication
Marine radio
Mobile communication: air, ground, marine, portable
Closed-circuit television: industrial, aircraft, and nuclear radiation
Slow-scan television
Instruments: test, measuring
Oscilloscopes: large-screen, bar-graph
Vibration test equipment
Magnetic amplifiers and systems
Alarm and signaling systems
Telemetering
Intercommunication, paging, and public-address systems

Consumer Products

Television and radio receivers
High-fidelity phonographs and equipment
Tape recorders
Microphones and loudspeakers
Refrigerators, freezers
Air conditioners
Hearing aids
Incandescent lamps
Home intercommunication equipment
Electrical housewares

Cable and Wire Products

Multiconductor telephone cable
Telephone wire: bridle, distribution, drop
Switchboard and terminating cable
Telephone cords
Submarine cable
Coaxial cable, air and solid dielectric
Waveguides
Aircraft cable
Power cable
Domestic cord sets
Fuses and wiring devices
Wire, general-purpose

Components and Materials

Power rectifiers: selenium, silicon
Parametric amplifiers
Transistors
Diodes: tunnel, zener, parametric
Semiconductor materials: selenium, germanium, silicon
Capacitors: wet, dry, ceramic
Ferrites
Tubes: power, transmitting, traveling-wave, rectifier, receiving, thyratron
Picture tubes
Relays and switches: telephone, industrial
Magnetic counters
Resistors
Varistors
Fluorescent starters
Transformers
Quartz crystals
Crystal filters
Printed circuits
Hermetic seals
Magnetic cores

Probable Evolution of Telephony
1000B Pentaconta Crossbar Switching System
Pentaconta Line Concentrators
Telegraph Transit Exchange Using Crossbar Pentaconta Design
Assistant Type Telephone Set
Intercommunication Telephone Sets
Matrix Multiplication in Search for Alternate Routes
Theoretical and Practical Aspects of Telephone Traffic
Slow-Wave Structures for Millimetre-Wavelength Backward-Wave Oscillators

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This Issue in Brief

Trends in Component Developments—Much of the advancement in electronic systems has stemmed from progress in the development of essential components. For entertainment equipment, stress on price stimulated mechanization. The military market added severe environmental conditions and a lower failure rate. Later, special telecommunication and electronic systems demanded failure rates even lower than the military, closer tolerances, and lives of the order of tens of years. Presently, these and the military requirements have tended to coalesce.

Reliability can no longer be the process of selecting the best from commercial production but now requires special design and manufacture in clean areas by workers of high integrity who will not pass an inferior unit. Improved reliability has raised problems in life testing and quality control.

Interconnections between components and interfaces of different materials within components are additional sources of trouble. Two possible solutions are the use of a single chip of semiconductor material having both active and passive elements built into it and the use of a substrate of glass with passive elements deposited on it. These integrated circuits will be economical only if they are produced in large quantities and the difficulties in standardizing components indicate that it will not be easy to obtain acceptance of standardized circuits.

The new components with their demands for high purity of materials and close control of manufacturing techniques have brought the physicist and chemist into the components industry. Mechanization is no longer a cost-reduction tool but an essential means of attaining uniformity and reliability. Dust particles now are significant in size compared with the product, and dust-free regions, both temperature and humidity controlled, are found within the factory.

The epitaxial process of producing planar transistors has not only increased the operating range and power of these active components but has greatly improved life over other encapsulation methods.

More-accurate machining techniques, parametric amplifiers, lasers, vacuum deposition, molecular excitation, high-resolution cathode-ray tubes, and other techniques and components are continually extending the useful upper frequency limits of existing systems.

High-Reliability Testing and Assurance for Electronic Components—The reliability of an electronic component is the probability that it will perform a specific function for a specified length of time under defined environmental conditions. By nature, it is a quantity that is predicted rather than measured.

The determination of average levels of reliability does not present any exceptional difficulties. But as reliability is tending toward the increasingly higher values that are demanded, theoretical, practical, and economic difficulties are arising. To overcome them, various means such as accelerated tests, which also are debatable, are available. The problem becomes clearer and simpler if a separation is made between random unforeseeable failures and failures due to ageing, which are easier to predict and are of prime importance in long-term applications.

Finally, a broad organization of reliability, based on cooperation, has to be envisaged.

Microelectronic Devices—Three types of microelectronic devices are discussed. Integrated circuits consist of a number of interconnected components that perform a specific function on a common mount. Thin-film circuits use resistors and capacitors that are evaporated or sputtered on a common mount and the active elements are added separately. Semiconductor circuits have both active and passive elements processed on a semiconductor substrate by transistor techniques.

Integrated tunnel-diode circuits show spectacular improvements over lumped components.

Close matching of characteristics and low series inductance of leads and elements are important factors and may permit operation at speeds of several hundred megacycles per second. Smallness and compact arrangement of elements reduce propagation time over the signal path. Examples of logic, memory, and counting units are described but the techniques are not limited to tunnel diodes.

Frequently, transistors are used in matched sets and by mounting them in intimate thermal contact the matching will be improved over the full operating temperature range. The mounting in a single header of the two transistors that make up a Darlington amplifier increases reliability by reducing the number of connections between dissimilar materials as well as requiring less space and lower encapsulating costs.

Means of evaporating active devices on to the thin-film substrate of glass or ceramic to which the passive elements are applied are still under development. The Miniflake transistor is readily attached to the substrate, though not part of it.

In semiconductor circuits, the transistor diffusion techniques permit active and passive elements to be incorporated in the substrate. The semiconductor substrate may be used as a common electrode or alternatively isolation may be obtained by using a reverse-biased junction. Passive components may also be produced on the semiconductor circuit by thin-film techniques.

Semiconductor circuits should provide high reliability as all components are processed simultaneously of very pure materials under carefully controlled conditions. Packing density is limited by heat dissipation and low-power devices can be made very small. Tooling costs being high, production runs must be large. The active elements are most suited to digital operation.

Thin-film techniques have lower tooling costs and are simpler to design than semiconductor units. The active elements being added, the thin-film circuits may be better suited for linear operation. Semiconductor and thin-film devices are not necessarily competitive but have their own fields of usefulness.

Silring Mounting of Silicon Power Rectifiers—Based on the techniques of stacking selenium rectifiers, silicon units are mounted in doughnut-shaped structures that are alternately stacked with cooling plates on an insulated rod. The rectifier units serve as spacers between the cooling plates.

The rectifier element is soldered to a heavy copper ring and fits into an eccentric opening in a ceramic spacer. A second copper ring provides for the other connection to the rectifier and both plates and the ceramic insulator are soldered together to provide hermetic sealing.

The rectifier cells, cooling plates, and connection tabs may be mounted in any electrical arrangement on an insulated bolt passing through their centers. At an ambient temperature of 50 degrees centigrade, ratings between 2.5 amperes for the Silring unit alone to 15 amperes with cooling plates 125 millimeters (4.9 inches) square by 2 millimeters (0.08 inch) thick are obtained. Parallel operation of rectifiers of the same rating, but without sorting of units, permits direct-current outputs of 280 amperes or twice this current for forced cooling with an air speed of 6 meters (19.7 feet) per second. The relative sizes of stacks for approximately the same rating are 100 percent for selenium, 8 percent for conventional silicon mounting, and 4 percent for Silring.

Cause and Prevention of High Reverse Currents in Large-Area High-Voltage Diffused-Silicon Rectifiers—It is shown that the major factor responsible for "soft" reverse characteristics in silicon diodes is the presence of impurities, other than the elements of Groups III and V, distributed uniformly throughout the silicon. The impurities are present in both zone- and

crucible-grown silicon with inevitable further contamination during device processing.

An expression is derived relating the reverse current of such soft diodes to voltage and temperature, and a phenomenological explanation of the processes underlying this relationship is presented.

Two methods of removing the impurities from the vicinity of the junction are described. The first method, in which an acidic oxide is used as a getter, was first described by Goetzberger and Shockley. The second, a physical gettering process, has not been used previously to improve the quality of soft diodes.

The mechanism of these processes has been studied with a view to defining the optimum conditions under which they proceed. It is suggested that these techniques provide a means of stabilizing device yield at a high level, as well as the possibility of producing new devices by complete removal and subsequent controlled addition of the impurities.

Defects in Vapour-Grown Silicon—Single-crystal silicon may be deposited from a vapour on a single-crystal slice of the same material. Diffusion of impurities is reduced if the substrate is maintained at a temperature substantially lower than that of the melting point and, by suitable doping of the vapour, p-n junctions can be grown with predetermined impurity profiles.

Crystalline defects are reduced by annealing and cleaning the mechanically polished substrate on which the epitaxial layer is deposited, indicating that imperfections or surface impurities encourage such defects. Dislocation densities are not increased in deposited layers over those in the substrate with proper annealing and cleanliness.

Polycrystallinity and twinning are encouraged by contamination and by growth at too low a temperature. Stacking faults appear to originate at either a physical or a chemical imperfection but further study is needed of the nucleation process.

With adequate precautions, vapour-grown layers of silicon have defect concentrations similar to those of the substrate on which they are grown.

Silicon Epitaxial Planar Transistors—Two major shortcomings of normal diffused mesa transistors are high saturation voltage V_{CES} and long switching time, both being closely dependent on current levels and the latter also being influenced by the circuit arrangement. Noise and low current gain at low currents are additional problems. Epitaxial technology and gold diffusion have overcome the two major and the planar technology has overcome the other two shortcomings. They have also reduced manufacturing tolerances of characteristics, which are also retained reliably during life.

Epitaxial layers of silicon suitably doped to control resistivity have the same crystalline orientation as the substrate on which they are grown and may be extremely thin to reduce the series electric and thermal resistances. Epitaxy produces transistors having low saturation resistances that are independent of temperature, increased linearity of characteristics, and reduced minority-carrier storage time.

By diffusing gold into the base and collector regions, the storage time in switching circuits has been reduced from 100 to 10 nanoseconds in a typical low-level epitaxial transistor.

In the planar technique, an oxide layer less than a micron thick is grown on a clean silicon surface. A window is etched in this silicon dioxide and, exposed to a high-temperature boron-rich atmosphere, boron is diffused, producing a p-type layer underneath a new oxide layer formed over the window. Similarly another window is etched and an n-type layer is formed beneath another oxide coating by phosphorus. Thus, each junction is completely sealed by an oxide layer as it is formed, permanently preventing contamination. Leads are connected through small etched windows.

The epitaxial and planar techniques as well as

gold doping are combined, making the silicon epitaxial planar transistor. These are probably the most versatile and reliable transistors available. The production techniques are inexpensive and over a thousand units may be processed simultaneously. Tests at each stage of manufacture avoid further processing of rejects.

Characteristics of a number of typical planar and planar epitaxial silicon transistors are given. Epitaxial and non-epitaxial units are compared for direct-current, alternating-current, and switching characteristics.

Mixer Diodes for 6 Gigacycles Per Second—Two formed germanium diodes, both operating on nonlinear-resistance properties, were developed for down-converter and up-converter service between 70 megacycles per second and 6 gigacycles per second.

In the low-level down-converter, noise sets the sensitivity limit of the receiver. Noise, conversion efficiency, and output impedance at the intermediate frequency are inversely related to the local-oscillator drive, which determines the voltage swing and thus the extent of nonlinearity of the operating range.

Although silicon diodes are simpler to make, requiring only a pressure contact between the whisker and the polycrystalline semiconductor, in contrast to a formed contact in a germanium diode in which the whisker is welded by an electric current to a monocrystalline semiconductor, the germanium has lower series resistance and lower intermediate-frequency impedance, the latter being an important element for this design. The theoretical superiorities of gallium arsenide have not been obtained in production with assurance.

As the forming current and resulting heat is increased, the area of the junction also increases. The series resistance of the inactive part of the semiconductor varies as the reciprocal of the junction radius and the capacitance in shunt of the active region varies as the square of the junction radius. By putting two whiskers in parallel, the changes in series resistance and shunt capacitance are proportionate but the intermediate-frequency impedance is halved. Curves are given for noise, conversion loss, and intermediate-frequency impedance of the final design of diode.

An up-converter using nonlinear capacitance is more efficient then the nonlinear-resistance model. However, the series resistance of the latter can be used effectively as a means of limiting output amplitude. Also by making a large junction, its shunt capacitance is increased and harmonics are short-circuited. The *p-i-n* type, separating the *p* and *n* regions by an intrinsic layer, provides a larger junction the capacitance of which is invariant and not significantly sensitive to voltage.

If *n*-type germanium is heated sufficiently, it will pass through the intrinsic state and become *p* type. By using a high-resistivity *n*-type germanium, a *p*-type-impurity whisker, and the proper forming technique, a heavily doped *p* region occurs immediately under the whisker, a balanced intrinsic region of high resistivity is beyond it, and the original *n*-type germanium is beyond that.

The impedance change from $+1$ to -1.6 volts is predominantly resistive, the conversion loss at saturation drive is only 7.5 decibels, and the compression ratio is 5 : 1 for an input change of 5.6 decibels. All interfering harmonics were at least -117 decibels relative to the output signal.

Vacuum Tubes for Submerged Repeaters—The cost of an interruption in service and of replacing a submarine repeater on a multichannel telephone system is such as to justify almost any expense to insure reliability for a life of, say, 20 years.

The vacuum tubes, which are the essential active elements in the repeaters, must have high transconductance combined with low interelectrode capacitance to obtain the largest frequency bandwidth with the minimum number of repeaters. They must operate at low voltages and currents to impose the least demands on the

power supply that operates in series all the repeaters in a system. The design must balance these often-conflicting requirements.

The required reliability can be attained only by perfection in every stage of manufacture including the highest standards of cleanliness. The entire process from raw materials to life testing is done in a fully restricted area completely isolated from the rest of the plant.

Characteristics and some historical information are given on the following types: *5A/153, 5A/161, 5A/162, 5A/181, 5A/182, LS882,* and *VX7158.* Transconductance changes over many thousands of hours of operation indicate extremely small variations. About 18 million tube-hours have been accumulated in submarine repeaters from 1955 to 1962 and their reliability well justifies the faith placed in the design and production of these vacuum tubes.

Ceramic-Insulated Vacuum Tubes for Very-High-Frequency Industrial Heating—The production of larger powers at higher frequencies will expand the uses of dielectric heating in industry and has prompted the development of four new triodes producing up to 12 kilowatts at 220 megacycles per second and 40 kilowatts at 100 megacycles per second.

Tapered coaxial elements permit smaller structures having lower inductances than corresponding conventional pin-type tubes and give better performance at the higher frequencies. A high-alumina ceramic insulation between adjacent tube elements has lower dielectric losses than glass and its use also has the advantage of a resultant low radio-frequency loss at the seals. These ceramic seals may operate at 220 to 250 degrees centigrade against 180 degrees for glass-Kovar seals.

As exact matching of the coefficients of expansion of both ceramic and metal cannot be obtained from brazing to room temperature, the metal is selected to have the larger coefficient and it is brazed to the outside of the ceramic rings, thus operating the ceramic in compression.

The filament is operated at a relatively low temperature and with uniform optimum carburization will give a minimum life of 5000 hours with an unregulated filament voltage that may vary ± 5 per cent and will operate for short times at 90 per cent of rated voltage.

The grid design is particularly critical because if the tube is operated without a load, the grid dissipation can be several times the normal loaded value. A much-larger safety factor must be designed into the grid than would be required for a conventional radio transmitting tube.

Air, water, and vapour cooling are provided for in the several designs. The trend in medium-size equipments is to the use of a water-cooling coil hard soldered to the anode wall.

Travelling-Wave Tubes for 6-Gigacycle-Per-Second Radio Links—Two new travelling-wave tubes were developed for use as output amplifiers in 6-to-7-gigacycle-per-second radio links. One tube with a synchronous helix voltage of 3 kilovolts and a beam current of 40 milliamperes gives a saturated output power of over 10 watts. The other gives about 50-percent more power for use in the highest-capacity radio links and operates at 3.5 kilovolts and 50 milliamperes.

The tubes include the same cathode and electron guns used in earlier designs of tubes operating at 4 gigacycles per second as these have demonstrated an average life of over 20-000 hours. A novel design of mount, with a high degree of radio-frequency screening, incorporates the periodic permanent magnets used for focusing the electron beam, the waveguide feeds, and a natural-convection cooler. The tubes can be easily replaced in the mounts under field conditions and will give very good focusing.

Results of measurements of noise factor (around 26 decibels), modulation noise, gain (37 to 40 decibels), power output, and other characteristics are given.

Effect on an Electron Beam of Variations in Periodic Permanent-Magnet Focusing Systems —Theoretical solutions for electron beam focusing in uniform periodic permanent-magnet systems exist, but in practical cases it is important to know the effect of the variations and misalignments that inevitably occur.

In the paper the results of computer calculations are given for the behavior of an electron beam subject to random variations of magnetic field and half-period length. These show the magnetic field to be the more-critical parameter. The effect of a change in magnetic field has also been calculated, and it is shown that a gradual stepping-up of field will maintain a well-focused beam. Also a moderate increase in period length, such as may be necessary to accommodate a waveguide, can be compensated for by a slight change in magnetic field. If larger changes of period are necessary, the pole-pieces can be incorporated in the waveguide itself without adversely affecting the matching; details of this are given.

Another important variable will be the displacement of the various elements of the periodic focusing system from the desired axis. The resultant beam displacement is derived analytically leading to a value, for practical purposes, of $2sn^{\frac{1}{2}} \left|\sin (2\alpha)^{\frac{1}{2}}\right|$ where n is the number of elements, and s the standard deviation of the displacement of the elements.

Smithsonian Institution Receives Microwave Relics—Over thirty years ago, microwave communication was first publicly demonstrated by transmitting across the English Channel from Escalles, France, to Saint Margaret's Bay in England. Both telegraphy and telephony were tested. Operating at 1.7 gigacycles per second, a half-watt transmitter energized a half-wave dipole antenna mounted in a 10-foot (3-meter) reflector.

The Smithsonian Institution in Washington, District of Columbia, has accepted for display the original transmitter section with vacuum tubes and the antenna dipole assembly together with a replica of the 10-foot reflector used in the demonstration. The presentation, shown in Figure 1, was made by Mr. Andre G. Clavier, who led the engineering team that developed the equipment. The International Telephone and Telegraph Laboratories at Hendon, England, and the laboratories of Le Matériel Téléphonique in Paris cooperated in the project. The equipment was largely developed in Paris.

This startling demonstration of the effectiveness of centimeter waves for practical communication over substantial distances and the installations that soon followed it played a significant role in the development of this region of the electromagnetic spectrum.

ITT Federal Laboratories
United States of America

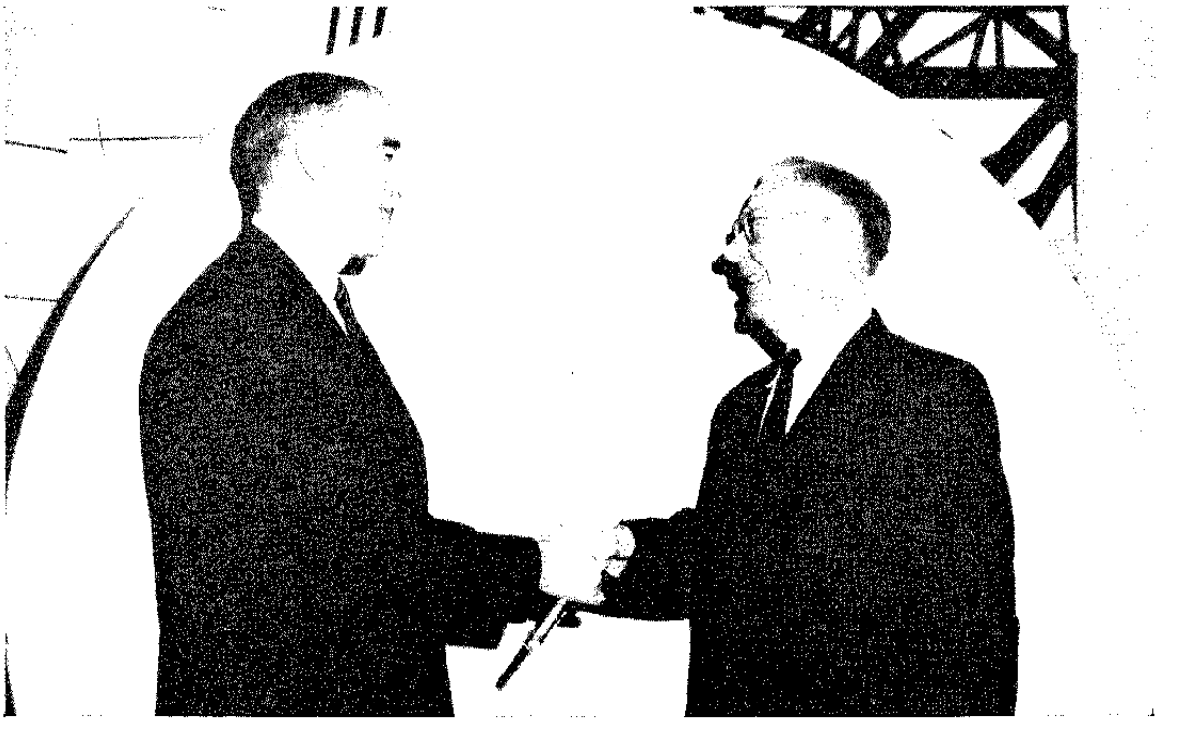


Figure 1—Presentation to Mr. Frank Taylor (at left), director of the United States National Museum, of historical microwave equipment used in the 1931 demonstration of 18-centimeter transmission across the English Channel. The presentation was made by Mr. Andre G. Clavier (at right), who led the engineering team responsible for this pioneering work.

Electronic Nurse—An "Intensive Treatment Telemonitor" was demonstrated at the Third Annual San Diego Symposium for Biomedical Engineering. It is capable of monitoring as many as 25 patients simultaneously for such physiological factors as heart and respiration rates, temperature, and blood pressure.

These factors are presented continually at the bedside and at a central location. They are also recorded permanently. The physician defines the upper and lower limits of each factor as determined by the condition of the patient. The observer will be immediately alerted by both visual and aural means to any significant changes in these factors.

As shown in Figure 2, the system elements at the bedside are the sensors attached to the patient and the electronic data-processing equipment. A transmission system carries the data to a central point, where a display unit keeps the observer informed of the condition of all the patients under supervision, and where a recorder makes a permanent record for each patient.

This system resulted from the development of a monitoring arrangement for continual checking of the physiological conditions of pilots and astronauts.

ITT Federal Laboratories
United States of America

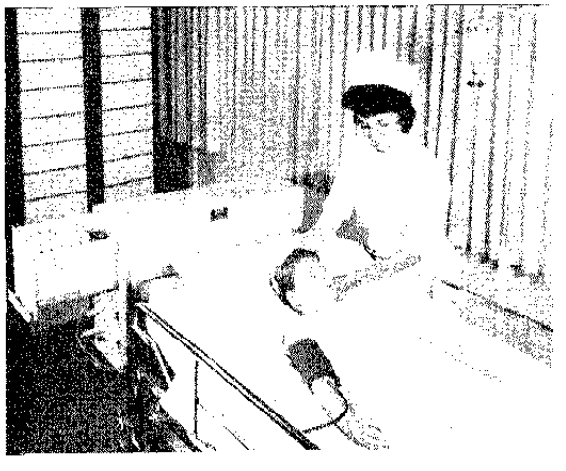


Figure 2—Patient "wired" for the Intensive Treatment Telemonitor. The data processor is in the cabinet at the head of the bed. Heart and respiration rates, temperature, and blood pressure are monitored continually.

Impulse Generator with Program Control—The impulse generator shown in Figure 3 will produce two independent series of impulses. In each series, both the pulse and space durations may be independently adjusted between 50 microseconds and 10 seconds.

Program control permits each series to be produced continually or to be limited to any set number of pulses up to 25. Up to 5 such groups of pulses, with desired spacing between the groups, may be repeated. Also, the two sources, using different patterns of pulses and spaces, may transmit such groups alternately.

Two output circuits, an electromechanical relay and a transistor, are each capable of handling 500 milliamperes at 48 volts. The sensitive polarized relay produces less than 2 milliseconds of distortion and is suitable for relatively low-speed switching, operating effectively to above 100 bauds.

The transistor output operates over the entire working range. Using it, circuit interruptions, vibration, and other disturbances can be simulated. Logic, memory, supervisory, automation,

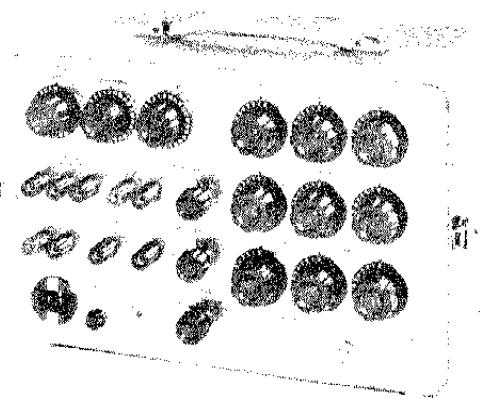


Figure 3—Impulse generator incorporating program control.

and remote-control systems may be tested with the generator.

Standard Téléphone et Radio
Switzerland

Spark Erosion Machine—The removal of metal by the eroding effects of an electric spark permits the shaping of pieces to highly precise dimensions. The machine shown in Figure 4 was developed originally for laboratory research but its effectiveness prompted its immediate use in manufacturing microwave components and in preparing masks for critical thin-film evaporation techniques.

The machine can cut slots as narrow as 10 microns in material 50 microns thick and slots of 50 microns in 1-millimeter-thick material. The accuracy and resolution is around ±2 microns.

Standard Telecommunication Laboratories
United Kingdom

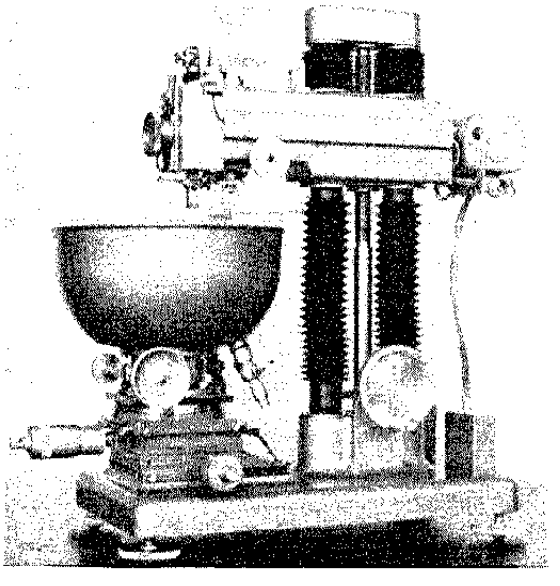


Figure 4—Experimental machine for precise shaping of metal by spark erosion.

Bank Signatures Transmitted by Videx—In 20 seconds, a slow-scan television system permits the central signature file of the Provident Tradesmens Bank and Trust Company in Philadelphia, Pennsylvania, to transmit over a regular telephone line a picture of the signature of a depositor who wishes to transact business at a distant office of the bank.

On receipt of a telephoned request to the central file, the card bearing the signature is placed in a Videx transmitting cabinet and on pressing a button is transmitted over the telephone line to the inquiring office. There, as shown in Figure 5, it appears as a still picture on the screen of a display tube, where it can be retained for several minutes. Operation of an erase button clears the tube immediately.

It is impracticable to ask every depositor in a large banking system having dozens of offices to sign a card for each office or to maintain in each office a file of signatures of depositors in all offices. Videx provides a simple, direct, and rapid means of verifying signatures so that every depositor may be protected in transacting his business at any office of the bank.

ITT Industrial Laboratories Division
United States of America



Figure 5—Bank teller verifying signature against original signature card reproduced over telephone line from central file using Videx system.

Reflex Klystrons for Millimeter Waves—A series of reflex klystrons has been developed to cover wavelengths from 3 to 12 millimeters for use as local oscillators and as pump oscillators for parametric amplifiers and masers. Figure 6 is a cutaway view showing the structure.

The 8-millimeter tube has produced 2.3 watts of output power although the production units will be rated between 0.5 and 1 watt. This output is substantially larger than is available from present commercial klystrons in this wavelength band. The sturdy structure results in good frequency stability. Life tests have now been run for only 2000 hours; a much-longer life is anticipated. The assembly is self-jigging, which reduces manufacturing costs.

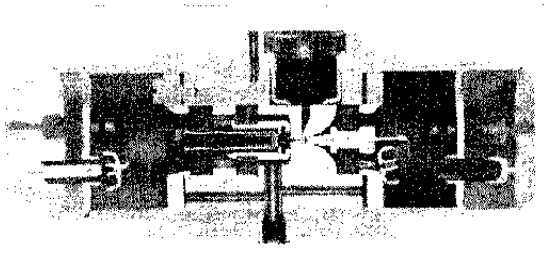


Figure 6—Reflex klystron for millimeter wavelengths. This tube, which operated near 3 millimeters, is about 4 inches (10 centimeters) long.

This klystron is tuned by an external cavity, the resonant frequency of which is controlled by an adjustable multilayer slab of dielectric material. Tuning ranges greater than 3.5 percent of the operating wavelength have been obtained.

Standard Telecommunication Laboratories
United Kingdom

Cheekphone for Noisy Places—To avoid the normal pickup of noise by the microphone of a telephone handset, a short-handled set has been devised that operates with the microphone pressed against the speaker's cheek. A 70-decibel transistor amplifier mounted inside the set amplifies the microphone output.

As shown in Figure 7, the shorter handset hangs on the gravity switch at the back of the pedestal. A shielded cord avoids crosstalk, which could be bothersome with the high amplification that is employed. Power is obtained from the alternating-current mains and a selenium rectifier assures proper polarity of the output voltage.

Fabbrica Apparecchiature per Comunicazioni Elettriche Standard
Italy

Steel-Jacketed Picture Tube—A television picture tube having the critical area of mechanical

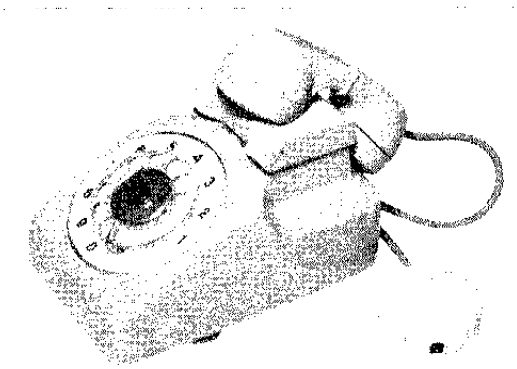


Figure 7—Cheekphone set using short handle to permit microphone to press against the cheek of the speaker and thus exclude ambient noise.

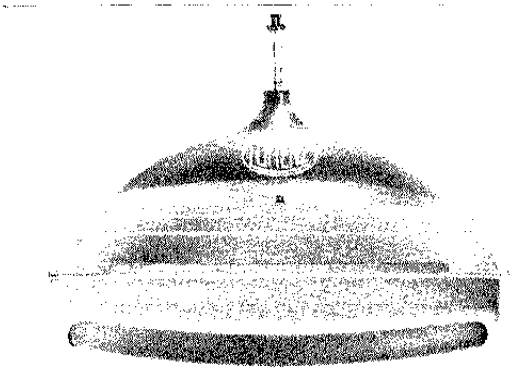


Figure 8—Steel-jacketed television picture tube that needs no safety glass and may project beyond the front of the cabinet.

stress of the glass bulb bonded by a filler material to a steel jacket avoids the need for a transparent safety shield.

The reflecting surfaces of the safety glass are avoided as is the accumulation of dust between it and the face of the tube. The protected outer edge of the bulb may safely extend beyond the cabinet front. The conventional mounting strap is eliminated by flanges on the steel envelope that facilitate both original installation and replacement. Figure 8 shows the *A59-12W* tube with 110-degree deflection and a screen diagonal length of 59 centimeters (23 inches). The over-all length is 36 centimeters ($14\frac{11}{64}$ inches).

Standard Elektrik Lorenz
Germany

Forest Fire Simulator—To train forest service personnel in combating large-scale fires, the simulator illustrated in Figure 9 has been developed. Designed to reproduce the appearance of a forest-fire command post, the system accommodates 12 trainees who face an 8-by-12-foot (2.4-by-3.7-meter) screen on which the fire scene is projected as if it were several miles away.

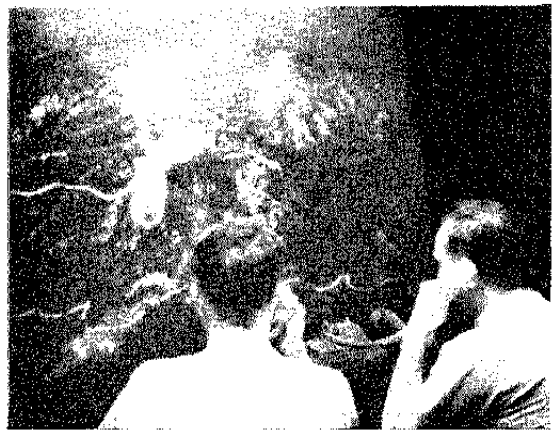


Figure 9—Simulator of a command post with projected image of a distant fire used to train forest service personnel in logistics and techniques of fighting large-scale fires.

The instruction staff consists of an effects operator, a role player, and an umpire, who sits in an isolated booth. The umpire establishes for each exercise the weather conditions, types of fuel in the path of the fire, and other pertinent and usually complicating factors. The effects operator controls the projected view of the fire area. The role player, taking his cues from the umpire, simulates the outside world of the weather bureau, aerial support, police, and other significant agencies.

The trainees must cope with each changing situation by controlling helicopters used for observation, fire-fighting aircraft and trucks, and personnel through orders sent over various communication systems at their disposal. The umpire evaluates the results of all factors and instructs the effects operator to change the projected image accordingly.



Figure 10—Combined phase-inverter triode and push-pull pentodes *ECLL 800*.

An exercise may, of course, include interference with the arrival of needed equipment by heavy road traffic of people interested in seeing the fire, the presence of summer campers in the fire area, and other complications.

International Electric Corporation
United States of America

ECLL 800 Twin-Pentode–Triode—A phase-inverter triode and two push-pull pentodes mounted in a single glass envelope with a 9-pin base simplify the audio system for low-priced radio receivers and phonograph amplifiers.

Class-*B* operation at 250 volts with only 28 milliamperes for the entire tube produces a power output of 6.5 watts with less than 1 percent of harmonic distortion and 9 watts at 5-percent distortion. In *AB* operation, 8.5 watts may be produced with 5-percent distortion and 8 watts with 3-percent distortion. The tube is shown in Figure 10.

Standard Elektrik Lorenz
Germany

7-Gigacycle-Per-Second Radio Link Equipment—Up to 300 voice channels are available over the radio transmitter and receiver, shown in Figure 11, which operates in the 7-gigacycle-per-second region.

Klystrons are used as oscillators in the transmitter and receiver and all other active elements are transistors. Printed boards and plug-in units conserve space and both equipments are mounted on a standard telephone bay 1.8 meters (5.9 feet) high. Working and stand-by sets, automatic change-over equipment, and the service telephone channel can all be accommodated in two bays.

There is provision for connecting the transmitter to the receiver through a frequency converter for adjustment purposes. The antenna filters are placed in the transmission lines and are not mounted in the bays.

Fabbrica Apparecchiature per
Comunicazioni Elettriche Standard
Italy

Color Data Display System—A modified teletypewriter at the Marshall Space Flight Center in Huntsville, Alabama, was used to interrogate an information processor in Paramus, New Jersey, 1000 miles (1600 kilometers) away. In reply, the requested data were sent over an ordinary telephone circuit and projected in seven vivid colors on a screen within half a minute.

The request for information was transmitted to an *ITT 7300 ADX* automatic data transmission system, a real-time stored-program information

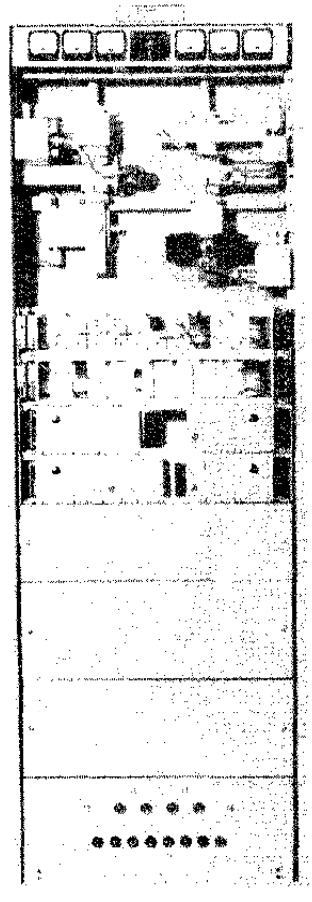


Figure 11—Radio link equipment operating at 7 gigacycles per second and providing for 300 voice channels.

processor suitable for on-line telegraph operation. The reply was transmitted to Huntsville over a Bell Telephone System newly developed *402* data set, which provides for 8 levels of binary information in parallel at 600 bits per second over a conventional voice-grade telephone channel.

The digital data were converted to visual alphanumeric form in seven colors in a Datachrome color projector. The Datachrome projector uses black-and-white 35-millimeter film. The seven colors are obtained by combining the three prime colors obtained by separating a white-light source with mirrors. The information may be projected as far as 50 feet (15 meters).

International Electric Corporation
ITT Federal Laboratories
United States of America

Electromechanical Filters—Electromechanical filters providing Butterworth response in the range from 50 to 500 kilohertz are useful in multiplex telephone and telegraph systems as well as in intermediate-frequency amplifiers and in single-sideband radio equipment.

A typical filter is shown in Figure 12. The design offers high performance in small size. Stability is assured by the choice of materials.

Standard Telecommunication Laboratories
United Kingdom

Recorder for Crossbar-Switch Utilization—Continual observations of the busy condition of crossbar selectors will expose overloads and permit such assignment of lines and switches as provides maximum service to all subscribers. Accurate estimates may also be made of needed expansion of an exchange.

An electronic analyzer permits each element to be checked during a 20-millisecond scan and its condition reported to a ferrite memory. The memories are then read and the number of seized elements in each frame is indicated on a counting relay or reported on a page teleprinter. A simultaneous operation reports on counting relays busy information on each of the 320 elements that may be observed. The observation periods can be controlled either manually or automatically.

The equipment has been used in Pentaconta public telephone exchanges to study traffic distribution in secondary selectors (links) and final selectors at the subscribers stage.

Compagnie Générale de Constructions Téléphoniques
France

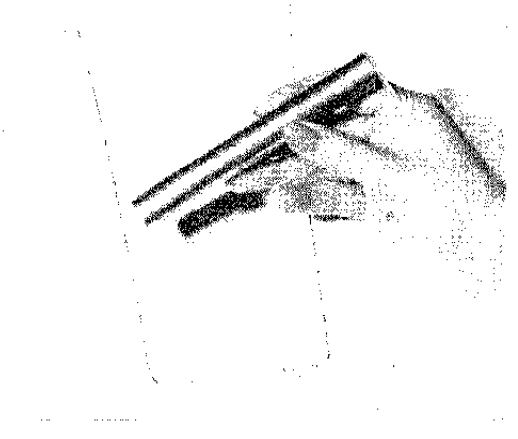


Figure 12—Electromechanical filter producing Butterworth response.

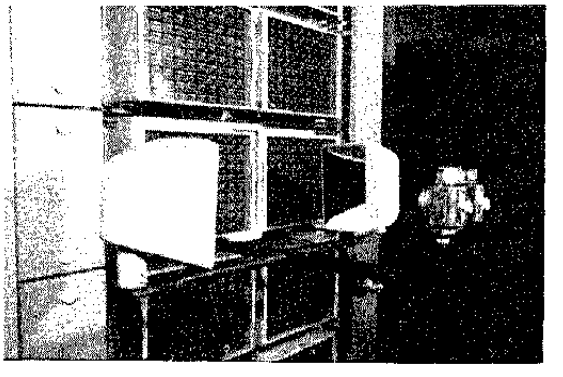


Figure 13—Subscriber meter frame with photographic equipment attached for recording calls for billing.

Telephone Subscriber Meter Frame—A new meter frame has been designed for Pentaconta exchanges. Two groups of 100 subscriber meters each are mounted on the frame and 10 such frames, containing 2000 meters, are accommodated in a standard Pentaconta bay. An identity finder frame is located in the lowest part of the bay.

Figure 13 shows part of a bay of meters to which photographic equipment has been attached for recording the readings for billing purposes.

Manufacture has begun for installation in the Pentaconta exchanges of the Compañia Telefonica Nacional de España in Spain.

Standard Eléctrica
Spain

Polyethylene-Terephthalate Dielectric Capacitors—Films of polyethylene-terephthalate (Mylar) as thin as 3.8 microns are available for use as dielectric in capacitors. Even a single layer may be used if the metal electrodes are applied by evaporation. They then exhibit the self-healing properties of other metallized paper capacitors. These polyester materials absorb very little moisture and capacitors need not be expensively sealed to ensure adequate life.



Figure 14—Multiconductor cable for use in submarines has all interstitial voids filled with a compound to prevent leakage through the length of a cable that may pass through bulkheads in the ship.

Series of capacitors are available in both simple tubular form and in metal cases. Capacitance ranges from 0.1 to 47 microfarads at direct-voltage ratings between 100 and 630 volts. The operating temperature range is from −55 to +125 degrees centigrade. At room temperature, the insulation resistance is between 5000 and 10 000 ohm-farads and at 1 kilocycle per second the loss angle, tan δ, is 0.8×10^{-3}.

Standard Elektrik Lorenz
Germany

Line Concentrator and Telephone Carrier—A *K-52* line concentrator that would normally serve 52 subscriber lines over 14 physical pairs of wire has been combined with a *K-31* rural carrier system to further reduce the required wire lines to 1 or 2 pairs.

Line concentrators provide simultaneous service to a number of randomly selected subscribers' lines that terminate in a relatively small geographical area over a much-smaller number of lines to the central office. This is feasible because residential subscriber lines average less than 10-minutes use per day.

The carrier system removes any limitation on the loop resistance between the concentrator and the central office. The limit for the *K-52* concentrator would be about 1400 ohms or 7 miles of a 22-gauge pair in a cable.

The new line concentrator and carrier has been successfully field tested by the Allied Telephone Company in Sheridan, Arkansas.

ITT Kellogg Telecommunications
United States of America

Radio Altimeter—The British Overseas Aircraft Corporation is installing the *STR.51* altimeter in its fleet of *VC-10* airliners. An essential element in all-weather landing operations, its accuracy at low altitudes is such as to make it useful over both the flare-out and touch-down phases of landing.

Duplicate units and automatic monitoring ensure prompt changeover in case of equipment

failure. The apparatus may be as much as 50 feet (15.3 meters) from its antenna system.

Standard Telephones and Cables
United Kingdom

Water-Blocked Multiconductor Cable—A multiconductor cable has been developed for use in submarines that, if cut, will not leak water through the interstices running parallel to the conductors in the cable. An impermeable viscous compound fills all these interstices without affecting either flexibility or electric properties. The cable will not leak even when subjected to very high hydraulic pressure.

A typical design is shown in Figure 14. The cables are available in coaxial form with copper braided shield and plastic jacket or in shielded individuals, pairs, and triads of up to 132 conductors.

Surprenant Manufacturing Company
United States of America

North Sea Telephone Cables—The British Post Office in conjunction with the German, Danish, and Netherlands administrations has ordered the 4 new submarine-cable systems shown in Figure 15. All are capable of providing 120 telephone channels, and the Germany and Denmark links will be the first direct high-capacity telephone cables to connect these countries with England. Two separate cables will be laid to Germany.

Over 900 nautical miles (1667 kilometers) of conventional armored cable will be required. Most of 0.62-inch (15.7-millimeter) diameter, a section between Borkum Island and the German mainland will be of 0.935-inch (23.7-millimeter) size to avoid the need for repeaters in very shallow water.

These will be the first installations of new low-cost shallow-water repeaters of which there will be 81 and 6 equalizers. The demountable housings for these units have been described [1].

Standard Telephones and Cables
United Kingdom

1. B. M. Dawidziuk and F. L. Jarvis, "Shipboard-Adjustable Submerged Equaliser," *Electrical Communication*, volume 38, number 1, pages 88–105; 1963.

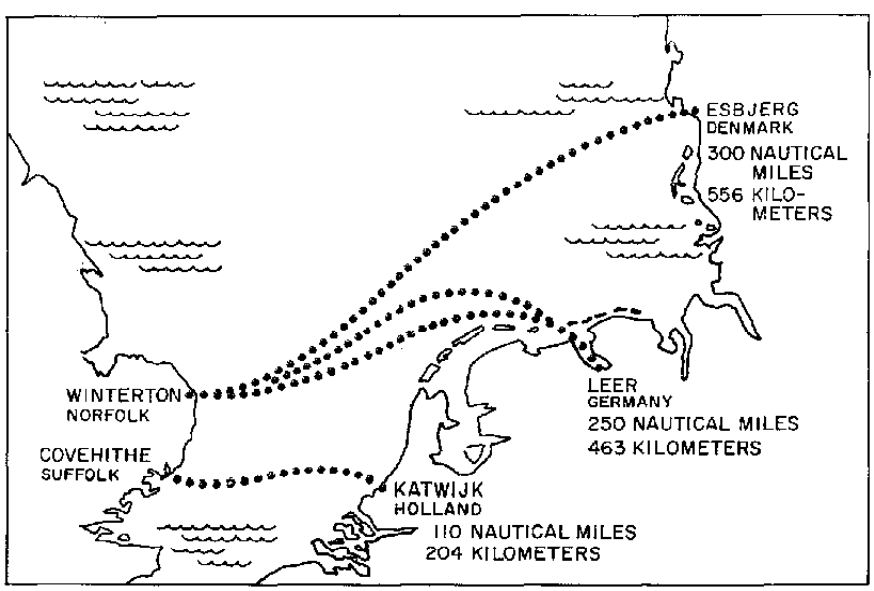


Figure 15—Map of 4 new submarine telephone cables to be laid across the North Sea.

Assistant Telephone Set for Call Assignment and Handsfree Operation—An executive's telephone station permitting handsfree operation is shown in Figure 16. It is based on the Assistant type of subscribers equipment. Provision is made for call assignment, which permits transfer of a call to the executive's set from that of the secretary.

Either handset or handsfree operation may be used, the latter employing a transistor amplifier and a suppressor to eliminate the damaging effects of electroacoustic feedback. The amplifier operates from the direct-current supply from the central office.

With the exception of the loudspeaker, which is in a separate housing, all components are mounted in the subscribers set and in an additional small box. A second microphone, also for handsfree operation, may be connected to the desk set.

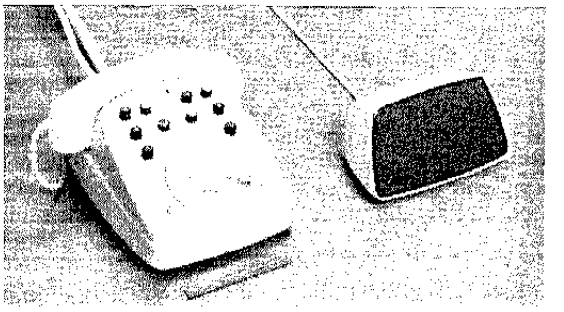


Figure 16—Assistant type 2-1 executive's subscriber set and separate loudspeaker for handsfree operation.

Standard Elektrik Lorenz
Germany

Trends in Component Developments

P. H. SPAGNOLETTI
Standard Telephones and Cables Limited; London, England

1. Introduction

In the advanced countries of the world, electronics is one of the fastest growing industries. A main spur behind this growth has been the continuous development of new products, many of which are available only because of progress in components development. Radar would not have been possible without high-power vacuum tubes operating in the centimetre band. Modern computers, with their high speed and capacity, are practical because of the low power consumption and small physical size of semiconductors. Microwave links owe a lot to the travelling-wave tube, and the extremely high reliability of special components has made feasible the use of submerged repeaters.

This short survey of electronic components is presented from a European point of view, although it is inevitably coloured by the experience and philosophy of the company with which the author is associated.

Since its early days, after the First World War, when it was concerned largely with radio broadcasting, the electronics industry has grown enormously; in the United States some 35 times since 1939. To-day, it is one of the largest in the country. Similar rates of growth have been experienced in the United Kingdom and other European countries. There have been three stimuli during the growth period. To some extent these have overlapped in time. Firstly, radio broadcasting and the following television development; secondly, military electronics and radar; and finally, professional electronics including communication.

The radio and entertainment industry has already lost its dominant role in the United States, where it now represents only a quarter of the total sales in electronics, and has similarly fallen in the United Kingdom to about one third of the total turnover. In both these markets, professional and defence products have now a greater importance.

The three phases or stimuli in the history of electronic components have had quite a marked influence on the products themselves. In the entertainment industry, price has always been the dominant feature, and, as a result of price competition, has come mechanization. The military market requiring, as it did during the Second World War, large quantities of components looked mainly to the entertainment-equipment factories for this production. However, the military superimposed its additional requirements for severe environmental conditions and for a lower failure rate than was acceptable in the entertainment market. Close tolerances were not essential nor was long life; the entertainment specification for the latter was traditionally 2000 hours, whilst the working life of components in missiles, including ground testing, could be as low as 200 hours; shelf life, of course, here had to be excellent.

At the same time as the military market was going through this most rapid phase of expansion, the telecommunications and electronics markets were requesting specialized components having failure rates as low as, or lower than, those in the military field, but with close tolerances to reduce the engineering costs of systems made in comparatively small quantities. The primary requirement of these components was, however, long life, of the order of 5 to 10 years in most systems; while, for submerged repeaters, lives had to exceed 20 years of continuous operation; environmental conditions, on the other hand, were less stringent than for military requirements. To-day, the tendency is for military and electronics requirements to coalesce. With the increased need for ground control equipment and the wider applications of computers, the longer-term aspects of reliability are increasing in importance. The techniques of submerged-repeater manufacture, where each component has its own pedigree, a punched-card record of its various stages of manufacture, and assembly is undertaken in scrupulously clean conditions, have been used for the components for the Minuteman project.

2. Growth

The high rate of growth of the electronic industry, as was mentioned earlier, would not have been possible without the introduction of new products. New products that not only fulfilled needs that were previously satisfied by other means, but also needs that had been awaiting technological advances before their satisfaction was a practicable proposition. For example, analogue computers for forecasting upper-atmosphere developments have made possible the complicated calculations needed to improve weather forecasting. The time element was sine qua non. Banks are now using electronic data processing for maintaining customer accounts and cheque sorting in place of clerical effort, whilst many production control processes are now handled by electronic equipment rather than by operators. In the military field, the importance of these new products may be shown by the increasing share of the defence dollar absorbed by electronics. In the United States, electronics accounted for 6·5 per cent of military procurements in 1951 but 18 per cent in 1961. That this growth is expected to continue is shown by estimates that expenditure on research and development in this field, which represented 35 per cent in 1959, will grow to 50 per cent in 1965 and 65 per cent in 1970.

For many years the growth of the components sector was in line with that of the equipments using these components. Recently, however, the components growth has exceeded equipment growth for two reasons. Firstly, as components become smaller, the supporting structures diminish in size and cost, and, with the use of the printed circuit board and multiple soldering with automatic assembly, the labour costs of equipments are reduced; components themselves, therefore, represent a larger part of the total cost. Secondly, these electronic components are also now being widely used outside the industry. Transistors and associated components are being used in watches, toys, and in automobiles for ignition and power systems. The development of semiconductors, the decrease in size of the active components, and the corresponding decreases in size brought about by the lower voltage and current requirements of the passive components has enabled highly complex equipments to be practicable. The manufacture and assembly of the vast numbers of components for such equipments has been aptly called the "tyranny of numbers" by Jack Morton of Bell Telephone Laboratories. As an example, the number of electronic components in military aircraft has increased 10 times between 1950 and 1960. Such large numbers bring home very forcibly the reliability problem, as a simple argument will show.

3. Reliability [1]

If one component in 10 000 fails within a given period, say 1000 hours, that is, a failure rate of 0·01 per cent or a reliability of 0·9999 per 1000 hours, an equipment of 100 components would have a reliability of 99 per cent; in an equipment of 1000 components, the reliability falls to 90 per cent; and with 10 000 components it is only 37 per cent. Thus, after 1000 hours, on only 37 per cent of the occasions when service is required would the equipment operate satisfactorily. Modern telecommunication equipment requires that the component failure rate shall not exceed 0·01 per cent per annum. In submerged repeaters, where the recovery and replacement cost of a component is prohibitively high, A. A. New [2] of the British Post Office quotes failure rates of 0·0003 per cent per annum as essential. Figures approaching this are also becoming vital for very-large installations.

Such reliability must inevitably raise the cost of these components, but they are still economic when weighed against the alternatives of redundancy or high maintenance costs. Even where the equipment may be said to be readily available, such as in aircraft, the Royal Air Force estimate maintenance costs to be 3 to 4 times the cost of the original

equipment. Component reliability has unquestionably improved over the years but is only just keeping pace with the needs. Fifteen years ago, reliable components were simply selections from the normal commercial production, principally for the entertainment industry. But it was very soon realized that you cannot "inspect quality into a product."

To-day, such components are specifically designed and manufactured in closely controlled conditions with inspection at every stage. Figure 1 shows a typical "extra-clean area." Mechanical or automatic processes, which are so essential for consistently higher reliability, have reduced the effects of human limitations. But where operators are necessary, experience has shown that they must be willing to make a positive contribution to reliability in manufacture; and, for this purpose, integrity is all-important: admitting carelessness or acknowledging limitations of knowledge will often prevent a faulty component from being made. As a measure of the improvements that have been achieved, the following typical figures are quoted from Mr. New's publication [2]. In British submerged repeaters, the failure rate per annum in 1951 was 0·08 per cent, whilst in the period 1954–1959 it was an order of magnitude better at 0·004 per cent. Valves alone had improved from 4·1 per cent to 0·20 per cent.

These low level rates create, obviously, a problem of measurement, which is examined in detail in another article [1] published in this issue. In this latter article, the example relating to the application of the Arrhenius equation to the ageing law of germanium transistors as a function of temperature is particularly illustrative. It results from an investigation published by J. M. Groocock and R. H. Chilton [3]. It may be seen that extrapolated times-to-failure for a temperature of 45 degrees centigrade have to be expressed in hundreds of thousands of hours. These results, incidentally, demonstrate the improvement obtainable by operating at reduced temperature.

Other life acceleration tests, using the step-stress technique, can produce similar life data under various electrical and mechanical stresses and provide as complete a picture as possible of the failure mechanism of a device. As a further example, the paper capacitors used in submerged repeaters are subjected to a stress of 6·5 kilovolts and an ionization discharge count is made to indicate the presence or otherwise of small voids in the impregnant: voids that contribute to a shortened life. For this particular application, it is essential that there should be no voids at all.

4. Next Phase in Circuits

The tyranny of numbers and the drive for reliability is responsible for the next phase in component developments. Interconnections or joints between components are fundamental causes of trouble: taking the earlier example again, if it is assumed that each component has two soldered joints, and if one joint in every thousand fails each 1000 hours (that is, a reliability of 0·1 per cent per 1000 hours—a figure often quoted) the reliability of the equipment with 100 components will drop to 81 per cent and that of an equipment using 1000 components to only 12 per cent. All connections or interfaces of different materials are potential trouble spots, and even the simplest components, such as cracked-carbon resistors, will have as many as 8 such connections or interfaces when they are mounted in a circuit.

It is obvious, therefore, how essential it is to reduce such interfaces to the absolute minimum. Two lines of attack are under intensive development throughout the world. The semiconductor circuit, where a single chip of silicon or other semiconductor material has built into it the active components required, as well as the resistors and capacitors; and the thin-film circuit, where a substrate of glass has deposited on it capacitors and resistors to which the active elements are assembled subsequently. Both of these approaches enor-

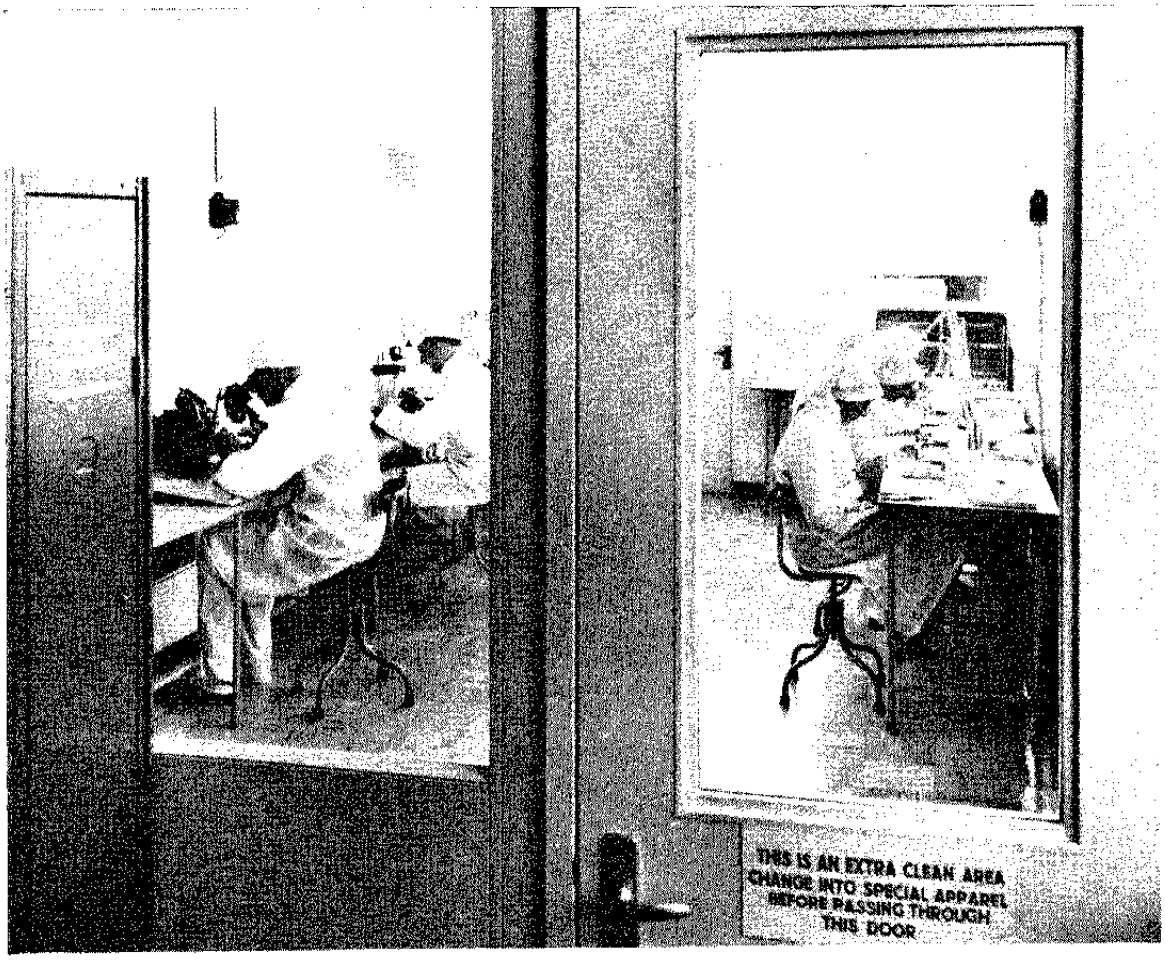


Figure 1—Typical extra-clean area for assembly of components.

mously reduce the number of interfaces and connections, and thus will contribute materially to reliability. This has, however, yet to be proved statistically. Apart, however, from the reliability advantages that such thin-film and semiconductor circuits give us, there is the additional requirement—for nanosecond computers; a signal will travel only 10 inches in this unit of time, so that the distance between the component parts becomes significant in equipment operation.

5. Semiconductor Circuits

Most of the semiconductor circuits available to-day perform specific digital functions, for it is in digital applications that one is likely to require large quantities of similar circuits. Only by the production of large quantities can the price fall to the economic levels that make these devices competitive with existing methods. However, a number of authors have questioned whether this is the right approach. Rice [4] mentions an attempt to build a

computer using 8 standard blocks; he reported that such a computer was feasible but that it would have better performance at lower cost if 486 circuits were used. Inevitably, standard blocks must involve compromises, in which case they are unlikely to be the optimum design for any particular application—so performance must be sacrificed. The performance of a system depends not only on the effective operation of the individual components but on the clever circuit design that makes optimum use of their characteristics.

The industry has been trying with very limited success to standardize its components for many years. It seems hard to believe, therefore, that it would be willing to accept standard circuits. J. M. Golday of Bell Telephone Laboratories suggests that better system performance, flexibility, and lower costs are achievable if only similar devices are fabricated on a common substrate; a number of substrates being put in a common encapsulation to form a circuit if desired. Golday maintains that "the major advantage of semiconductor circuits comes from the use of a common package, not a common substrate." Thin-film circuits, however, can be made economically in quantities and thus may overcome (at least for a period) the disadvantages imposed by the limitations of standard building blocks and the dislike of the average designer to be tied to standards, no matter how good they may be.



Figure 2—Equipment for producing complete thin-film circuits of passive components.

6. Manufacturing Techniques

The adoption of either semiconductor circuits or thin-film circuits raises two immediate problems. For component manufacturers, it breaks down the present divisions of products. For the equipment manufacturer, it means either increasing dependence on the suppliers of these two products with the concomitant loss of a further portion of a share of the value added in manufacture or a necessity to integrate vertically by setting up his own thin-film or solid-state factory. The expense and know-how required must, however, make this a risky and possibly undesirable process for many equipment manufacturers. It is obvious that the close liaison that exists to-day between the equipment and component designer will be even more essential in the future.

The new components, in whatever form, are inducing quite fundamental changes in the requirements for men and facilities in the production processes. Until recently component manufacture was basically an assembly operation. Close control of the properties of the parts used, such as metal parts in valves and paper and mica sheets in capacitors, was not essential to enable the product to function satisfactorily. Empirical knowledge and experience were often sufficient. To-day, however, the physicist and the chemist are far more important to the components industry. Semiconductors require materials of exceptional purity and close control or modification of their crystal structure for their satisfactory operation. The surface structure of substrates

for thin films and the effect of its life on the resulting assemblies is of great importance.

Most components can be made by thin-film techniques, and others are under development. Figure 2 shows equipment capable of producing a complete thin-film circuit of passive components without the necessity for removing the substrate. Metal and metal-oxide resistors are used both as individual components or as parts of circuits. Thin-film memory planes are challenging the toroidal memory devices.

Such changes and the overriding need for reliability are putting new requirements on the production floor. Mechanization, which used to be thought of mainly as a cost-reduction process, is to-day required to attain uniformity and reliability. H. H. Arnold [5] of the Western Electric Company says, "automatic control over manufacture can make it possible to produce a product at a higher level of reliability than would be attainable otherwise." He demonstrates the substantial improvements (of at least four



Figure 3—Cabinets providing small, clean, temperature-and-humidity-controlled enclosures in which transistors are encapsulated.

times) in the reliability of carbon-film resistors which he attributes to the use of computer-controlled automatic processing.

Again for reasons of reliability and also because of size reduction, modern component factories must be laid out to facilitate incorporation of dust-free temperature-controlled and humidity-controlled areas. The size of dust particles is now significant when compared with the product; measurements are now recorded in microns and Ångströms. It is not necessary, nor is it convenient, for the whole factory to be so treated, but small areas or even special cabinets must be readily available. Figure 3 shows a typical cabinet used to encapsulate transistors.

The rate of development in transistor fabrication techniques has given rise to its own manufacturing problems, in that a technique has often been superseded by the time production was in full swing. There seems to be general agreement, however, that some stabilization is likely with the planar epitaxial approach. The main drive leading to this development has, of course, been the desire for transistors of higher and higher performance, but in parallel has been the search for devices that lend themselves to lower-cost production and which are inherently reliable. Efforts to improve the reliability of germanium alloyed transistors have met with considerable success with the introduction of the molecular sieve and the cold-welded case. Figure 4 shows the improvement in life obtainable over earlier methods of encapsulation.

7. Future Developments

To-day's communication systems operate basically between direct current and 10 gigacycles per second. The extension of a further octave, therefore, gives us 9 times the present available bandwidth. It is in this octave, between 10 and 100 gigacycles per second, that the vacuum-tube engineers are working. Developments made possible by accurate machining techniques using, say, spark erosion or an electron beam, have extended the use of microwave tubes to these frequencies, and even beyond. Klystrons, backward-wave oscillators, and crossed-field devices have been developed, giving tens of milliwatts at these frequencies. Indications are that it may be possible to develop travelling-wave tubes capable of 10 to 15 watts of continuous-wave power at frequencies between 50 and 100 gigacycles per second.

Following the development of the parametric amplifier with its almost theoretical noise factor, making satellite communication a practical proposition, interest has jumped to the infra-red and light regions of the spectrum (300 000 gigacycles per second). Coherent light in the form of the laser is now practicable and ways of overcoming the problems of continuous-wave operation and modulation

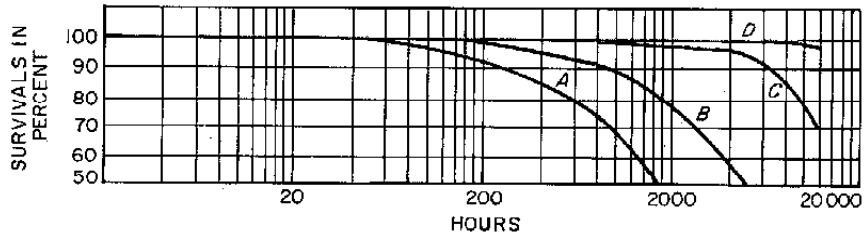


Figure 4—Survival curves for germanium alloy transistors. Curve *A* is for temperature-soldered cases and the semiconductor surface covered with silicon grease stored at 65 degrees centigrade. Curve *B* is in an araldite moulded case, temperature soldered, and stored at 65 degrees centigrade. Curve *C* is for a cold-welded case without an absorber and stored at 75 degrees centigrade. Curve *D* is for a cold-welded case, a molecular sieve built-in adsorber, and storage at 75 degrees centigrade.

are under development. One of the most promising approaches appears to be a further development of the gallium-arsenide light-emitting diode. This should enable very-much-higher efficiencies to be achieved in coherent light compared with the 1 or 2 per cent for ruby crystal or gas lasers.

Vacuum deposition will provide many new devices for the future. As mentioned previously, thin-film memories, thin-film circuits, and thin-film passive components will find increasing applications to-morrow. Because of the belief that such thin-film circuits, on account of their flexibility, may hold one of the best promises for the future, the development of the corresponding thin-film active devices is receiving a great amount of attention.

Another line, which is already producing some important devices, is the use of molecular excitation for controlling frequencies and for measuring magnetic effects. These have made possible accurate measurements of time, which could lead to new or improved navigation or guidance systems. Cathode-ray tubes with spot sizes of one third of a mil and capable of a resolution of 3000 lines per inch are being used for information-retrieval systems. The whole range of possible devices that are based on the low resistances of metals and alloys at temperatures near absolute zero are still only in the embryonic stage.

Further efforts are continuously being directed to extending the maximum frequency of operation of semiconductors and their power-handling capabilities. The upper frequency limit achievable to-day is of the same order as that reached by vacuum tubes using the conventional cathode, grid, and anode structures, although the outputs are somewhat lower. To overcome inherent limitations, vacuum-tube engineers had to employ new principles, the slow-wave structure and the cavity. It is likely that the semiconductors that will eventually compete with these vacuum-tube devices will also depart from the accepted emitter, base, and collector structure. Pierre Aigrain [6] has described active homogeneous semiconductor devices; that is, devices without junctions that do not depend on minority carriers for operation. These show great promise at high frequencies and high power. The encapsulation of components has advanced from the wax and paper era when radio and television provided the main market, through the glass, hermetic seal, and epoxy era of military components, and now to the planar transistor or diode that in many applications may not require encapsulation at all to protect it from its environment. Figure 11 in the article by B. D. Mills [7] demonstrates the stability of h_{fe} in a sample of 24 unencapsulated silicon planar transistors during a life test that has already lasted over 7000 hours. Other component developments, such as the silicon-dioxide capacitor and the metal-film resistor, place a lower requirement on the encapsulation, and are, therefore, potentially more reliable.

Various units of the International System in Europe and America have been and are in the forefront of many of these components developments. The impetus behind these has generally stemmed from the need and thus the desire to produce components with greater reliabilities, longer life expectations, and novel properties for the progress of the electrical communication art.

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He started his engineering career on a high-frequency world-wide radiotelephone system, working chiefly in England, Spain, and China. In 1931, he was responsible for the planning, engineering, and installation of the first two empire broadcasting stations for the British Broadcasting Corporation.

He was manager of the airborne radio communication equipment work of Kolster-Brandes throughout the war. In 1945, he became chief engineer and in 1947, general manager of the company, including manager of the valve activity. In 1958, he was appointed components group manager being responsible for all components within Standard Telephones and Cables. He serves also as a director of Standard Telecommunication Laboratories.

Mr. Spagnoletti is vice-president of the Television Society. From 1951 to 1954, he was chairman of the British Radio and Electronics Manufacturers Association and in this capacity represented the industry on technical and other discussions in connection with television standards and the formation of the new television authority.

Mr. Spagnoletti is an Officer of the Order of the British Empire. He is a Member of the Institution of Electrical Engineers.

High-Reliability Testing and Assurance for Electronic Components

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The necessarily restricted length of this article does not permit all definitions of terms used in reliability to be reviewed, and therefore only terms that are in current usage will be employed. The reader may refer to certain specialized publications [1]. The symbols used are defined in Section 8.

1. Failure Rates

It is usual to characterize the reliability of a component by a failure rate λ. If, among a very-large (infinite) population of identical components, N of them are still operative after a time t, by definition

$$\lambda = -\frac{1}{N}\frac{dN}{dt}. \qquad (1)$$

A priori, λ is a function of time. It is frequently expressed in the form x failures in percent per 1000 hours or also as $x \times 10^{-5}$ failure per component-hour or, abbreviated, $x \times 10^{-5}$. Thus, 0.1 percent per 1000 hours is equivalent to 10^{-6}.

This definition is interesting because in many practical cases and within certain limits the failure rate is a constant with respect to time. It is then possible to integrate (1) in the form

$$R = N/N_0 = \exp[-\lambda t] \qquad (2A)$$

where N_0 is the initial number of components. R is called the proportion of survivals at time t. If

$$S = 1 - R \qquad (2B)$$

S is the proportion of failures at time t. Again when λ is a constant, it may be demonstrated that the statistical mean life of a component is

$$\bar{t} = 1/\lambda. \qquad (3)$$

This is statistically the time at which 37 percent of the components are still operative.

Effectively, during the life of a group of components, there is a more-or-less-extended phase between the initial "infant mortality" and the wear-out* period where λ is practically constant. In principle, this is the phase referred to when nothing more explicit is indicated. But there is, here, a very regrettable risk of confusion in the suggestion that the rates thus defined could be utilized to design a large equipment in which components operate for such a long time that they are generally in the wear-out phase. Mortality is then much higher. Nonetheless, let us for the moment confine ourselves to the limits of the accepted usage of the term.

2. Direct Measurement of Failure Rates

2.1 Statistical Aspects, Sampling Method

When λ is a constant, it is said that we are dealing with a Poisson process. The following problem then arises: to determine how many samples n should undergo test during a period of h hours so that the observation of m failures assures a failure rate of λ_1 (expected) with a specified probability P also termed the confidence level. The *MIL-STD-105* sampling norm, well known, is inadequate in this case because it is based on attribute sampling (the objects being classified into good and bad). Here the time factor is involved. The mathematical base of a sampling plan in the case of a Poisson law is the following relation

$$P = 1 - \exp\left[-m\frac{\lambda_0}{\lambda_1}\right] \times \left\{1 + m\frac{\lambda_1}{\lambda_0} + \frac{\left(m\frac{\lambda_1}{\lambda_0}\right)^2}{2!} + \cdots + \frac{\left(m\frac{\lambda_1}{\lambda_0}\right)^m}{m!}\right\}. \qquad (4)$$

λ_0 is the observed failure rate for a sampling; in other words

$$\lambda_0 = m/nh \qquad (5)$$

and nh is called the number of component-hours of test. P then is the probability that the

* Wear-out as used here is not limited to mechanical deterioration but includes aging, degradation, and similar weakening of desired properties.

actual rate for the entire lot (supposed infinite) does not exceed the value λ_1 that has been specified.

From this, it is more convenient to deduce that, to assure a rate λ_1 with a confidence level P, the minimum value of nh must satisfy the equation

$$nh \cdot \lambda_1 = k \qquad (6)$$

where k is now a numerical coefficient that depends on P and on m. Its value is given in Table 1.

As an example, suppose we verify a failure rate of the order of 0.1 percent in 1000 hours, or more briefly 10^{-6}, with a 90-percent degree of confidence. The minimum value of nh, anticipating that no failure will be observed, corresponds to $k = 2.3$, and thus to 2.3 million component-hours. Since, by hypothesis, λ is a constant, in theory 230 samples can just as well be observed during 10 000 hours, 1000 samples during 2300 hours, or 2300 samples during 1000 hours, et cetera. Practically, this is obviously not irrelevant and we will come back to this question.

The above represents the minimum test. A slightly deeper analysis shows that it has not a high statistical significance. In fact, by means of Table 1, it may be determined that, having observed λ_0, there is 90-percent probability that

$$0.05\ \lambda_0 < \text{true } \lambda < 3\lambda_0. \qquad (7)$$

It thus appears advisable to tighten this interval by observing several failures. For 2 failures, $k = 5.3$, which leads to 5.3 million component-hours. The ratio between the extreme values of the interval then falls to 8, instead of 60.

It is intended here merely to point out certain aspects of these methods and to fix orders of magnitude. Usual sampling plans are even more complicated. Besides the *MIL-S-19500C* specification, which is known to specialists, the work already cited [1] may be consulted.

No serious difficulties, practical or economic, are encountered as yet with the number of samples determined above for an expected λ_1 of 10^{-6}, as far as carrying out the test is concerned. But the determination of failure rates of the order of 10^{-7} (0.01 percent in 1000 hours) requires 10 times the number of component-hours. It can be estimated that this represents the limit order of magnitude for *directly* testable reliability levels, leading to expensive tests, which would have to be justified by the importance of the envisaged application.

2.2 Physical Aspects

One should not overlook the complex nature of reliability in the sense that direct measurement of λ, as described above, assumes the following factors as defined: the component (including the *manufacturer*), the utilization circuit, and the environmental conditions. For the time being, the question of duration of operation is not being considered. It is obviously impossible to vary all these parameters in multiplying the tests. This leads to the

TABLE 1
Values of k in (6)

Confidence Level P	Number of Failures Observed m										
	0	1	2	3	4	5	10	15	20	30	40
0.95	3	4.7	6.3	7.7	9.1	10.6	17	23	29	41	52
0.90	2.3	3.9	5.3	6.7	8.0	9.3	15.4	21	27	38	49
0.10	0.1	0.52	1.1	1.7	2.4	3.1	7.0	11	15.4	24	33
0.05	0.05	0.35	0.82	1.35	2.0	2.6	6.2	10	14	22.5	31

separate consideration of two fundamental types of problems, which we are now going to examine.

2.3 Determination of λ for a Known Application

Test conditions are then well defined and constitute a single group. This is the problem treated above. There are indeed very few examples that really illustrate such a case. The following is among the most representative.

2.3.1 *Example 1*

This project to determine the long-term reliability of *2N357* and *2N396* transistors was undertaken jointly by General Electric Company and Radio Corporation of America as part of the Ballistic Missile Early Warning System (BMEWS) program [2]. Approximately 10 000 transistors of each type were tested for 10 000 hours, that is, 10^8 component-hours of testing for each type. The test conditions were varied, so that analysis of the results enabled a correlation to be found between the failure rates and the operating conditions. But the whole of the conditions applied was also considered as representative of the expected operating conditions and over-all failure rates were obtained. The example can be considered as an illustration of the above solely from this latter angle. Finally, these over-all failure rates were evaluated at 0.17 percent per 1000 hours (1.7×10^{-6}) for the *2N357*, and 0.024 percent per 1000 hours (0.24×10^{-6}) for the *2N396*, with a confidence level of 80 percent and a precision of 25 percent. The actual rates observed in operation were 0.071 percent and 0.027 percent per 1000 hours, respectively. The 100 000 perforated cards that recorded the measurements made on the *2N396* alone reveal that such a project is not a minor affair.

These tests of large quantities of electronic components should not be confused with similar ones made to preselect components for unusually critical applications, such as, transistors for artificial satellites and tubes for submerged repeaters. In these cases, the purpose of the tests is to *guarantee* the highest possible reliability through the choice of the most-stable components (for example, 5 or 10 percent of the initial lot) rather than to *measure* the reliability. These latter examples, therefore, are outside the scope of the present article.

2.4 A Priori Determination of Failure Rates

If the question, "What is the average failure rate for dry tantalum capacitors?," is posed, it would seem to have very little meaning in view of the above considerations. It is necessary, however, to try to give it some meaning for two useful purposes.

On the one hand, the user could require this information, even though approximate, to know, for example, whether long-life aluminum electrolytic capacitors, in view of their lower price, would also be as satisfactory for a particular application.

On the other hand, the time has come when the manufacturer must supply data on the reliability of his products, and even guarantees.

The current tendency is to present such information in the form of basic failure rates. These are established for somewhat severe operating conditions, for example, maximum temperature and maximum voltage, so that their value, thus increased, can be more easily and more economically determined.

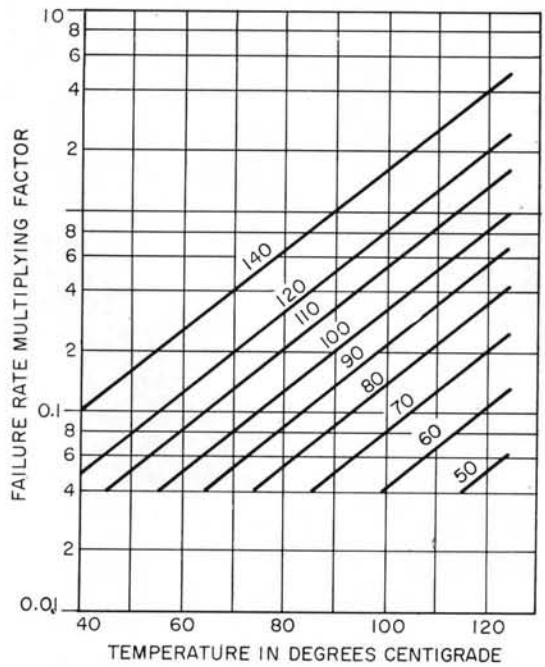
Naturally, the user is not entirely satisfied as he will usually operate the component under more-moderate conditions. Correction factors, which can be applied to the basic rate, thus have to be determined. We touch here on the principle of *accelerated tests* that either allow a smaller number of samples to be handled or the testing time to be shortened. This is of added interest, since the user cannot always wait for one or several years to pass before knowing the quality of a product he must use

now. We will consider this fundamental problem at greater length subsequently.

2.4.1 *Example 2*

It is nevertheless useful to cite an example from the very-important Parts Specification Management for Reliability Report [3]. It concerns paper or paper-polyester tubular capacitors for operation from 200 to 600 direct volts. Four classes of reliability are set: *M*, *N*, *O*, and *P*, corresponding, respectively, to basic failure rates of 3, 1, 0.1, and 0.01 percent per 1000 hours, defined with a confidence level of 90 percent for a temperature of 125 degrees centigrade and with nominal voltage applied. The manufacturer is supposed to classify his products in this manner after due verification. Figure 1 gives the correction factor to apply to the basic rate for other temperatures and other voltages. These correction factors are the result of an accumulation of observations carried out under various operating conditions and, in the present case, of very-long experience with this type of capacitor. It can be noted, for example, that if a capacitor of the type in question operates under its nominal voltage at only 90 degrees centigrade, the probable failure rate is then only 1/5 of the basic rate.

Figure 1—Correction factors for basic failure rates of paper or paper-polyester tubular capacitors for operation between 200 and 600 direct volts. The percentage of rated voltage is indicated for the various curves.



The very principle of accelerated tests continues to provoke a great deal of discussion, mainly among users who doubt the validity of the extrapolations involved. Although their contention may be valid, we have, nonetheless, little better to propose. Above all, a view that seems to be even more fundamental applies to this example: the failure rates thus predicted include both sudden failures and wear-out failures. The definition of the latter is to a large extent arbitrary because, for example, the lower limit agreed to for the insulation resistance cannot be the same for all applications. We are, therefore, now going to envisage a clearer distinction between these two types of failures.

3. Separation of Failures Due to Aging

It has (provisionally) been considered immaterial whether 500 components are tested for 10 000 hours or 5000 for 1000 hours; this implicitly assumes that λ is constant for such durations, in other words, that neither aging nor infant mortality should intervene in any appreciable manner.

There are ways of limiting the effects of infant mortality, either by preaging the components or by eliminating defective units during the final adjustment of an electronic equipment. This is not the case for wear-out failures, which occur late, so that the above hypothesis is only approximately suitable for short- and medium-term applications.

On the other hand, [1] states, "in the majority of cases, the components of an electronic computer operate in the wear-out phase." It

is thus even more strongly the same for large equipments, present or future, and in particular for the telephone exchanges of tomorrow. The reliable operation of these equipments is no longer determined by the failure rate λ established for the period when such rates are relatively constant, but by higher rates that incorporate wear-out failures.

Another aspect of the question has to be noted here. The failures that occur during the period when λ is constant have a random character. They are generally *total* (short circuit, open circuit, breakdown, et cetera) in the sense that the component is suddenly unable to fulfill *any* function. Such failures are definitely characterized and can be counted without too much ambiguity.

Wear-out, on the contrary, is generally characterized by a progressive alteration in the characteristics of the device (isolation resistance, current gain, et cetera). There is in general no a priori criteria for declaring the component out of use. It depends on the application and the choice of a criterion is essentially arbitrary.

The point of view that these two modes of failure are, in fact, two aspects of a single phenomenon has been expressed. But to avoid deviating from our subject matter through such somewhat theoretical considerations, we will limit ourselves to the phenomenological aspects, that is, there are two main types of failures, one sudden and the other gradual. The user needs information about both.

It is evidently not a question of furnishing such predictions for all the characteristics of various components. On the basis of experience and of common sense, we will limit ourselves to those that are fundamentally representative of the component, such as the capacitance of a capacitor or, particularly sensitive to aging, such as the current gain and I_{CBO} of a transistor.

Moreover, since wear-out phenomena do not intervene until after a time lapse, ideally as long as possible, it is generally necessary to resort to *accelerated* aging procedures to draw conclusions within a reasonable time. A few examples will demonstrate that the prediction of aging phenomena does not present, at least in theory, the same difficulties as the determination of sudden failure rates.

4. Definition of a Failure

4.1 Empirical Methods

A first group of methods of defining a failure follows in Section 4.1.1 and is based on the assumption that, despite what has been said, wear-out failures can be defined without being excessively arbitrary. In a second group (Section 4.1.2), it is assumed that such a decision is left to the user. But the decision is not rigidly governed by the nature of the component. In the following example, which concerns relays, notably, the criterion is based on the observation of the first fault. It might just as well have been left to the user to fix the criterion, on the basis of the evolution of the contact resistance, for instance.

4.1.1 *Cases Where a Failure Criterion May Be Defined*

4.1.1.1 *Example 3*

This example concerns sealed-contact relays. For precision applications, it is convenient and legitimate to consider that a relay has failed as soon as the first fault occurs. In addition, this example happens to be characterized by the discovery of a simple law for the appearance of wear-out failures.

Finally, as a first approximation, it may be accepted that the life of a relay may be expressed as the number of successful operations, rather than in hours. This allows a simple accelerated test consisting of operating the relay at a rapid pace on a given load. Strictly speaking, the operating speed is not totally immaterial for it can modify the heating of the contacts, among other things. But let us disregard this refinement.

Figure 2 represents the results of a life test on 44 sealed-contact *H80* relays of Standard Elektrik Lorenz. The test continued for 200 million operations at a frequency of 35 per second, and the relays switched 100 milliamperes of direct current in a resistive load of 500 ohms.

The diagram represents the percentage of survivals as a function of the number of operations. The ordinate scale is such that a normal, Laplacian, or Gaussian distribution as a function of the abscissa is represented by a straight line. This choice results from the fact that over the wear-out period the distribution of failures as a function of the logarithm of the number of operations is, precisely, Gaussian. The distribution is then qualified as log-normal. The bottom of the curve (dotted) corresponds to a constant λ of 1.4×10^{-9} failure per component operation and a statistical mean life $\bar{t} = 7 \times 10^{8}$ operations (if the process were continued indefinitely and without change).

A slightly more rigorous statistical analysis can be carried out by means of Table 1 and by substituting component operations for component-hours of test. There are 3 failures during the first 50 million operations per relay, that is, $50 \times 44 \times 10^{6} = 2.2 \times 10^{9}$ component-operations roughly. Since $k = 6.7$ for $P = 0.90$, the results are as follows

$$\lambda_{max} \text{ (90 percent confidence level)} = k/nh \approx 3 \times 10^{-9}.$$

This underlines the difference between an observed failure rate and a maximum probable failure rate with a given degree of confidence.

In summary, there is a superposition of two mechanisms: up to 50-million operations, the failures are random only and appear at the constant rates defined above. Beyond 50-million operations, wear-out failures preponderate, with a life duration of 130-million operations, defined as the moment when *50 percent of the devices are out of service.** The standard deviation is equal to the interval that separates the points for 84 percent and for 50 percent, which, taking into account the logarithmic scale in abscissa, are −46 and +86 percent. Naturally, if the law of failures could not be represented by a normal distribution, it would nonetheless be desirable to furnish the value of the standard deviation which, however, would have to be calculated instead of being so simply determined graphically.

The test can be accelerated by other than frequency of operation, such as by the influence of the switched current on life. This has been applied to the case of the *G29* type of sealed relays of the Western Electric Company [4]. We will not deal with this example, which permits derivation of the law for the average life as a function of the switched current.

Three important conclusions can be drawn from example 3.

(A) An ideal reliability test should furnish

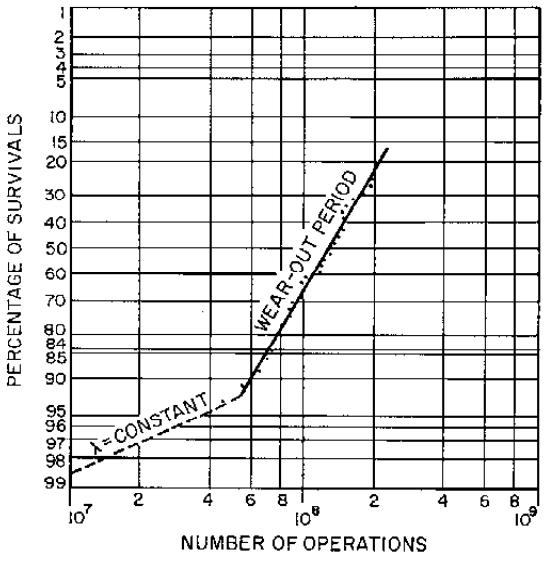


Figure 2—Life test on sealed-contact relays.

* τ is not identical to $\bar{t}$ as defined above except in particular cases, notably the present one of a normal distribution; but not in general.

information about both random failures and wear-out failures.

(B) Without entering into sophisticated analysis, it can be admitted on the basis of the good alignment of the values observed that probable life predictions have a high degree of confidence, even for as modest a sample as 44 devices.

(C) It can also be said that preliminary conclusions can be formulated long before the end of the test, for example, after 80 or 90 million operations in this case. This permits the test to be made either more economical or of shorter duration.

4.1.1.2 *Example 4*

In the preceding example, knowledge of the failure-distribution law permitted the use of an appropriate graph without any further consideration and with all the related advantages, in particular representation by a straight line. If a simple classical law cannot be uncovered a priori, the Weibull distribution [5], which is extremely general in application, can be used. It is expressed by a survival law in the form

$$R = \frac{N}{N_0} = \exp\left[- \frac{(x - \gamma)^\beta}{\alpha} \right]. \quad (8)$$

(Defined for $x \geq \gamma$.)

x is the parameter of evolution, such as time, number of cycles, et cetera, and α, β, and γ are constants.

Stange [6] and Kao [7], among others, have shown the extensive potentialities of such a function.

For $\beta = 1$ and $\gamma = 0$, it is identified with an exponential distribution (see (2A)). For $\beta = 3.25$, it represents a normal distribution,

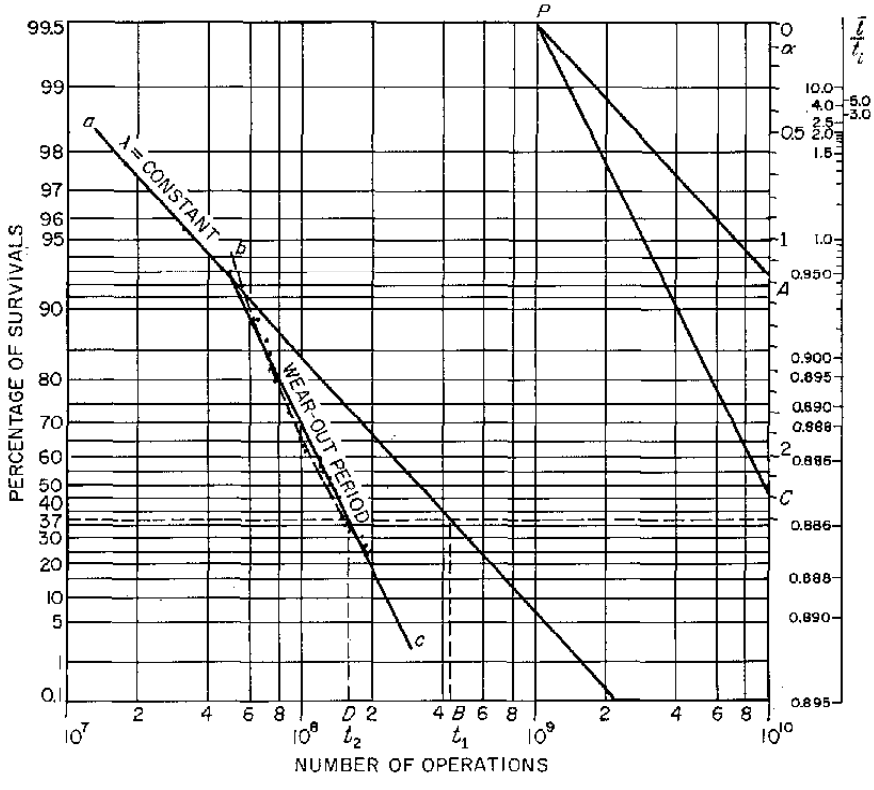


Figure 3—Weibull-Stange graph of life test in sealed-contact relays.

with an acceptable numerical approximation. Another advantage results from the relation

$$\log \ln (1/R) = -\log \alpha + \beta \log (x - \gamma). \quad (9)$$

It can be seen that on a graph where the ordinate scale is graded in (log log) and the abscissae in (log), a straight line with a slope proportional to β results. Making observations of R over a long period, we can thus expect to find two approximately straight lines intersecting at a certain angle, one corresponding to the quasi-exponential period (random failures) with $\beta \approx 1$, the other to the aging period with $\beta > 1$.

The use of Weibull's functions and graphs in itself implies precautions and refinements that are too lengthy to outline in this article. We will simply give an example, in the form of Figure 3, which is given the Stange presentation. To clearly delineate the differences with example 3, the same case of sealed-contact relays has been chosen.

It can be observed that the first 3 failures correspond to points situated exactly on a straight line, the slope of which corresponds to $\beta = 1.18$ (point A); that is to say, the hypothesis of an exponential law ($\beta = 1$) made in example 3 appears justified. A graphical treatment, on which we will not dwell otherwise, gives from points A and B the following statistical mean life duration

$$\bar{t} = 4.2 \times 10^8 \times 0.945 \approx 4 \times 10^8 \text{ operations.}$$

Analysis of the second part of curve bc is a little more delicate. $\beta = 2.2$ (point C) is found, corresponding to the fact that effectively a log-normal distribution is more nearly approximated than a normal distribution. The points C and D give an average life duration of

$$\bar{t}_2 = 1.6 \times 10^8 \times 0.885 \approx 1.4 \times 10^8 \text{ operations.}$$

In both cases, we find about the same orders of magnitude as in example 3. Strictly speaking, the points of the segment bc are better represented by the dotted curve than by the straight line bc. But this can lead to considerations that are unnecessary here. It may be said briefly that the Weibull-Stange representation appears to be, in the present case, better than the preceding one for segment ab and not as good for segment bc. The fact that it directly provides the statistical mean life is not a major advantage since the knowledge corresponding to x percent survivals is generally more useful, and can be seen directly on the diagram.

In sum, the Weibull-Stange representation is always useful as a last resort. A perfect representation is not at all basic theoretically, but only important in practice. In fact, the experimental points are generally scattered around the average curve. An improper choice of graph form thus can result in appreciable differences in the statistical parameters: average value (compare examples 3 and 4), standard deviation, et cetera. The resulting errors in the extrapolations that follow can become rather considerable.

4.1.2 *Cases Where No Failure Criterion Is Fixed*

Whereas in the preceding two examples, the failure (end-of-life) criterion, although arbitrary, was nonetheless very precise, it is now assumed that it is preferable for the user to determine that criterion according to the application envisaged. One then is limited to furnishing statistical data on the evolution of characteristics with time.

4.1.2.1 *Example 5*

The method will be applied below to dry tantalum capacitors. Because of the complexity of the problem, we will limit ourselves here to the characteristic that evolves most rapidly, leakage current. The capacitors were manufactured by Standard Elektrik Lorenz and were tested at the Laboratoire Central de Télécommunications by what one might call an accelerated test, although in this particular

example the operating conditions were not more severe than the limits authorized by the manufacturer. It would probably have been advantageous if they had been.

Capacitors rated at 4 microfarads and 35 volts were subdivided into 6 groups of 50 capacitors each and subjected to three temperatures (55, 70, and 85 degrees centigrade) and two voltages (17.5 and 35 volts, with a protective resistance of 100 ohms) for a maximum time of 8000 hours.

Figure 4 represents the evolution of the leakage current I_t for the group that was subjected to nominal voltage at an ambient temperature of 85 degrees centigrade. Curve A shows the leakage-current distribution at the start. The distribution turned out not to be strictly log-normal. By distorting the ordinate scale in an appropriate and empirical fashion, a nearly rectilinear average curve was obtained. This artifice is perhaps surprising, but it is nonetheless legitimate. The exact mathematical law representing the distribution is not of great importance, providing a single straight line is obtained; it is only a question of convenience and graphical precision. The thing to be carefully noted is that the scale selected is satisfying only between 2 and 98 percent. Consequently, the diagram has been truncated accordingly.

The straight lines B, C, and D show the situation at 1000, 4000, and 8000 hours, respectively. To avoid complicating the graph, the corresponding experimental points

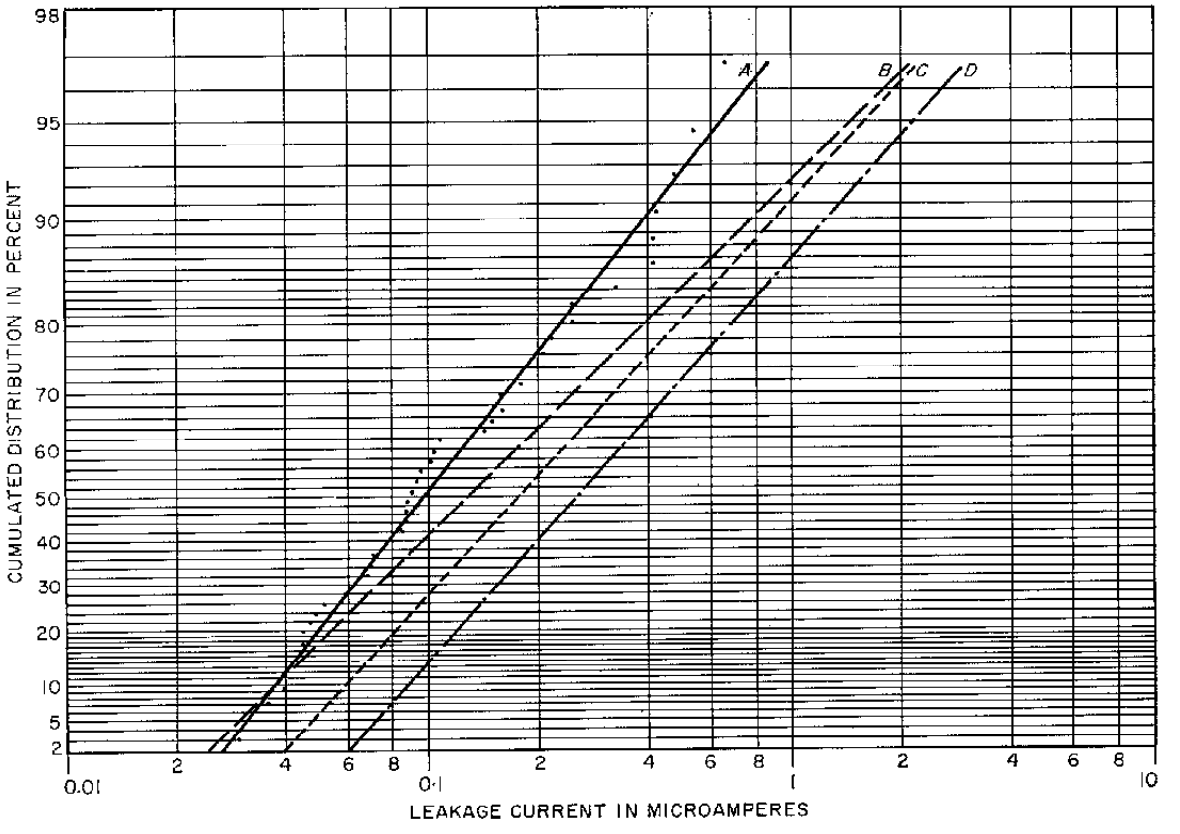


Figure 4—Reliability test on tantalum capacitors. Curve A is at the start of the test and curves B, C, and D, are after 1000, 4000, and 8000 hours, respectively.

were not plotted but their alignment was satisfactory.

Taking this diagram as a start, the evolution with time of any reference point whatever, for example the median value of 50 percent, can be studied. This is done on Figure 5, curve A. It is clear that an extrapolation up to 100 000 hours, although rather uncertain, can be essayed. The shaded area is purely figurative, but it can be considered unlikely that the median value of I_l will exceed 10 microamperes before 100 000 hours. Certainly an experimental point at 15 000 hours, even 20 000 hours, will greatly diminish the uncertainty when it is obtained.

Curve B shows the evolution of the 95th percentile. The uncertainty is a little bit greater because of the rather peculiar behavior of the capacitors up to the 1000th hour (curve B of Figure 4). It can be estimated, however, that 5 percent of the capacitors will have a leakage current exceeding 10 microamperes at around 30 000 hours.

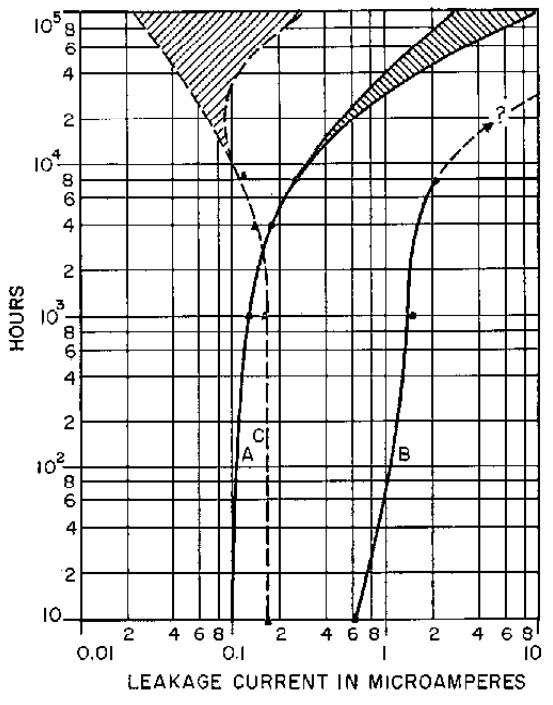


Figure 5—Extrapolations of reliability data on tantalum capacitors.

The other groups in the test can be similarly treated. Thus, curve C of Figure 5 represents the evolution of the I_l median value for the group submitted to nominal voltage at 55 degrees centigrade. Here, there is even an "improvement" up to 8000 hours. The extrapolation becomes very uncertain, and the shaded area is again purely figurative. But in all probability, there will not be any pronounced deterioration before 100 000 hours at 55 degrees centigrade.

4.2 Utilization of a Working Hypothesis or Model

In the preceding examples, the predictions were the result of the direct analysis of experimental observations which could either be represented by a classical mathematical law (Gaussian, Weibull, et cetera) or essentially exploited by graph. It goes without saying that if the mathematical *form* of the law of failures or wear-out is known, it is easier to devise the tests with a view to determining this law *quantitatively* (verifying it at the same time). When aging processes are basically physico-chemical in nature, it can be admitted that their speed with time v as a function of the absolute temperature T obeys Arrhenius' law.

$$v = A \exp [-(B/T)] \qquad (10)$$

where A and B are constants.

Passing to logarithms

$$\ln v = \ln A - (B/T). \qquad (11)$$

Thus, $\log v$ is a linear function of the inverse of the absolute temperature, which makes extrapolation easy. It can be admitted that the acceleration of a phenomenon is equivalent to changing the time scale, and therefore that the acceleration affects equally the random failures and the wear-out failures. However, this is not self-evident.

More-elaborate models have been proposed, for instance Eyring's. But the Arrhenius model is usually sufficient. It has been applied with success in numerous cases, notably that of semiconductors. As in Section 4.1, we should have made a distinction between cases where a failure criterion is fixed and those where this is not done. To limit the length of this article, we will consider only the case where a certain criterion is arbitrarily fixed.

The work of D. S. Peck of the Bell Telephone Laboratories is authoritative in this domain. Many other examples are described in the technical literature on the subject [8]. The example below is the result of work carried out at the Transistor Division of Standard Telephones and Cables.

4.2.1 *Example 6*

Germanium alloyed transistors were divided into groups, each subjected to a different temperature, that is 135, 115, 100, and 75 degrees centigrade. It had been previously recognized that the parameter most sensitive to degradation was the I_{CBO}, and the value of 10 microamperes was fixed as the failure criterion. The failures versus time were then plotted on a graph analogous to that used for the relays in Example 3. The alignment of points indicated a good fit with a log-normal distribution. It does not seem necessary to reproduce this diagram. After the examples already described, one can imagine that it is easy to find the moment when, for indicated temperatures, the total percentage of failures reaches whatever value one may fix: 10, 50, or 90 percent, for instance. We will limit ourselves to recording these values in Figure 6, where the abscissa scale is graded in $1/T$ and the ordinates are logarithmic.

Taking the inevitable graphical errors into account, the agreement between the theory and the actual experiment can be considered very satisfactory.

For the 50th percentile, the straight-line extrapolation is based on the points for 135 and 115 degrees centigrade. The point at 100 degrees is a short distance from this straight line, while the point at 75 degrees is too low. (As indicated, the point at 100 degrees was not observed but was calculated from other observations.) The situation is similar for the 10th percentile.

The discrepancy affecting the points at 75 degrees has since been explained. The corresponding transistors were from an earlier production lot and were not quite so good as the others. The test at 75 degrees was rerun with transistors from the same period as the

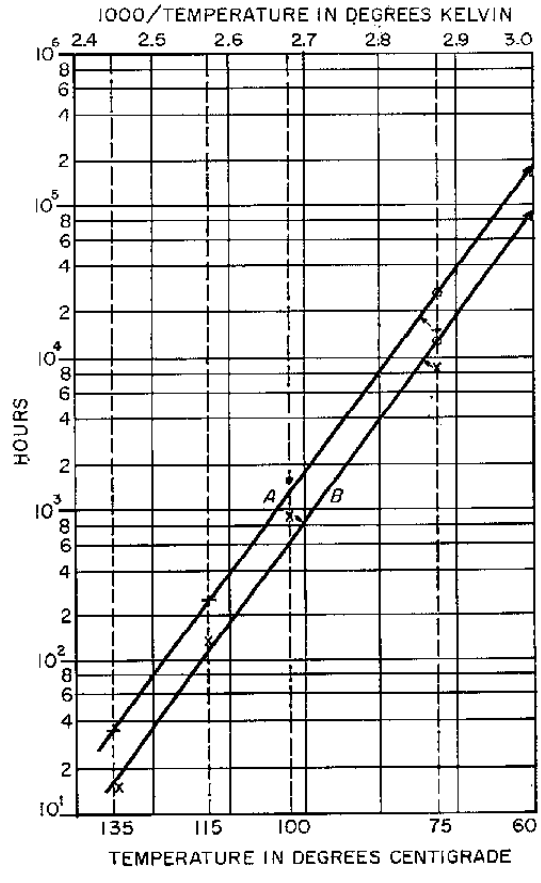


Figure 6—Reliability test on transistors based on the Arrhenius model. Curve A is for the 50th percentile and B is for the 10th percentile. The crosses are observed points, the dots are computed, and the circles are extrapolated.

others and the points now tend to be correctly aligned.

A relatively moderate extrapolation reveals that at 60 degrees centigrade (without dissipation of power in the transistors), 10 percent of the transistors will fail after 90 000 hours (10 years) and 50 percent after 180 000 hours (20 years). Whatever uncertainties may affect this result, it is extremely valuable, and better methods to establish such predictions are hardly conceivable.

Arrhenius' model can be applied with equal success to film resistors for variations in resistance as a function of surface temperature but not for failures as above.

A more-general character can be obtained by utilizing an effective temperature given by

$$T_{\text{effective}} = T_{\text{ambient}} + aP_d \qquad (12)$$

a being a constant and P_d the power dissipated in each device. There are therefore considerable potentialities in the Arrhenius model.

5. Synthesis and Conclusions

5.1 Recapitulation of Difficulties

The principal limiting factors involved in directly determining *very-small* failure rates can be summarized as follows.

(A) The hypothesis λ = a constant, which can be applied to only a limited period in the life of a component.

(B) The defining of both random failures and wear-out failures by a single parameter.

(C) The high number of component-hours of testing.

(D) The representativeness of a sample.

(E) The validity of extrapolations in accelerated tests.

(F) The often-disregarded multiplicity of factors, such as; humidity, shock, vibration, radiation, and human errors.

5.2 Proposed Solutions

We have considered at length the importance of separating random failures and wear-out failures. Whereas accelerated tests on moderate quantities of samples give sufficiently precise results insofar as the law of aging is concerned, difficulties are encountered in their use to determine very-small rates of *random* failures.

5.2.1 *Utilization of Sequential Tests*

We quote a passage from a very-significant editorial [9].

"Testing components to see how many in a group will fail within a certain time or testing components to destruction and determining why they failed is no longer satisfactory in many cases. The number of failures that will occur in a group of high-reliability components within a practical test period may be so insignificant that no conclusive evidence is found."

Reliability tests being long and *destructive*, it is not conceivable to test 100 percent of production; only sample quantities may be tested. This raises the question of the representativeness of any sample.

Thus, in view of the orders of magnitude we are now approaching, anxiety exists as to whether a sampling really represents the average quality of the production considered and also whether test results have not been distorted by a banal accident, clumsiness, or error, even in the absence of any suspicion of misrepresentation.

Difficulties (C) and (D) above, would seem to be in the process of solution along the general lines indicated in the reliability report already cited [3]. The cost and length of reliability tests will limit their application to average levels, which eventually might be the subject of guarantees on the part of the manufacturer. Beyond, reliability will no longer be determined by qualification or acceptance tests. A new spirit, based on

cooperation and reciprocal confidence between manufacturers and users, has to be developed. The manufacturers should set up a highly developed organization for quality assurance and control, not only to improve quality, but particularly to achieve uniformity of quality.

Then the determination of failure rates will assume an entirely different aspect. The manufacturer will be obliged to maintain accurate records of failures observed in all circumstances: manufacturing controls, tests for customers, et cetera. In accordance with the idea expressed above, only random and total failures should be taken into consideration. There are, then, several ways of statistically cumulating the corresponding information by putting the tests end to end, so to speak. This permits the accumulation of a large number of component-hours of tests, covering an extended period of the particular production.

5.2.1.1 *Example 7*

A first method is based on (6) and values from Table 1 for a confidence level of 90 percent, transferred to Figure 7. To demonstrate the principle, a series of tests carried out at Laboratoire Central de Télécommunications on tantalum capacitors from Standard Elektrik Lorenz has been analyzed. The temperature was not the same in all cases, but it is estimated that 80 degrees centigrade

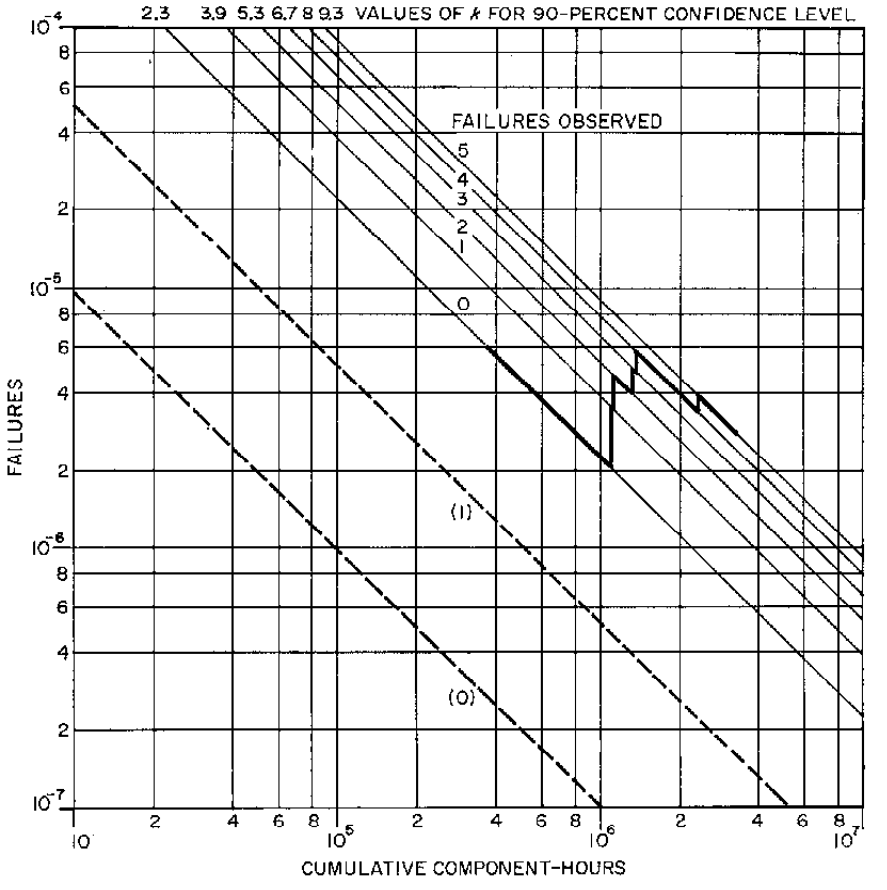


Figure 7—Cumulative reliability tests on tantalum capacitors based on (6) and Table 1.

TABLE 2
CUMULATIVE DATA FROM TESTS ON TANTALUM CAPACITORS

Test	Quantity	Failures	Accumulated Component-hours	
			Per Test	Total
1	10	0	220 000	220 000
2	30	0	720 000	940 000
3	25	1st at 7000 hours 2nd at 8000 hours end of test	175 000 +24 000 +92 000	1 115 000 1 139 000 1 231 000
4	50	1st at 2500 hours 2nd at 3000 hours end of test	125 000 +24 500 +336 000	1 356 000 1 380 500 1 716 500
5	50	0	500 000	2 216 500
6	50	1st at 3000 hours end of test	150 000 +343 000	2 366 500 2 709 500
7	50	0	500 000	3 209 500

represents a fair average. The results that were calculated are given in Table 2 and are recorded in Figure 7. Now, after a cumulation of 3 200 000 component-hours of test, it appears, with a 90-percent confidence level, that λ is less than 3 *total* failures per million component-hours at 80 degrees centigrade, which is very satisfactory. It is useful to compare this with the gross rate observed, 1.6 failures per million component-hours. The advantage of this presentation is that it furnishes an upper limit for λ, *with no preconceived idea* as to the value expected (similarly, the dotted straight lines give the lower limit). A disadvantage is, perhaps, compression of the divisions through the use of logarithmic scales.

In general, cumulative tests of this nature, called sequential, are handled in a different and much-more-elaborate fashion from the statistical point of view. The principles have been particularly developed by B. Epstein and M. Sobel [10], and A. Wald [11]. In these methods, a compromise is sought between the risk for the user of overestimation of life and the risk for the manufacturer of underestimation of life.

Complete analysis of these methods is known to statisticians but involves more-elaborate probability concepts, such as the chi-square distribution, and is decidedly outside the framework of this article.

5.2.1.2 *Example 8*

Poisson's law is a somewhat-different very-simple presentation that also derived from (4), (6), and Table 1. Figure 8 is the corresponding diagram. It is rendered more useful by grading the abscissae with $1/\lambda_1$ as unity, λ_1 being the target rate *to be verified*, which, being fixed a priori, is contrary to the preceding method. The significance of the curves is indicated in Figure 8.

Let us take example 7 for comparison and determine the situation for a failure rate of $\lambda_1 = 2 \times 10^{-6}$. As abscissa unit we then have $1/\lambda_1 = 0.5 \times 10^6$ component-hours. Let us record in ordinates the failures listed in Table 2 at the moments they occurred. This results in a stepped curve.

Let us arbitrarily assume that the production will be considered qualified for $\lambda = 2 \times 10^{-6}$

if this value is assured with a 95-percent confidence level (90 percent would be just as admissible). This would be the case if the stepped curve did not exceed curve A, a possible result which however is far from obtained in this case. At the same time, and just as arbitrarily, it can be considered that the production is disqualified for $\lambda = 10^{-6}$ if this rate is reached with only a 5-percent probability; thus corresponds to curve E and thus the production is effectively disqualified for $\lambda = 10^{-6}$. If $\lambda = 3 \times 10^{-6}$ is considered, curve F is the result, with a probability of 95 percent. It can be seen that the production has just been qualified after 1.1×10^6 component-hours. It is prudent to await confirmation by a new intersection, which seems likely to occur, without this being absolutely certain. A 90-percent probability that $\lambda = 3 \times 10^6$ is shown by curve G and is in agreement with Figure 7. Summing up, having agreed on acceptance conditions for a rate λ_1 and rejection conditions for a rate λ_2 (below λ_1, in general), intersection with the corresponding curve is awaited before a decision can be taken. This presentation, as can be seen, is somewhat different from the usual presentation for sequential tests, but it is simple in principle and in application. The reference curves can be traced on transparent sheets, which then can be placed over the diagram containing the stepped curve.

The report [3] already cited recommends that by means of sequential tests all manufacturers can achieve failure-rate qualifications that normally should improve with time. It is possible, however, to go backwards, through

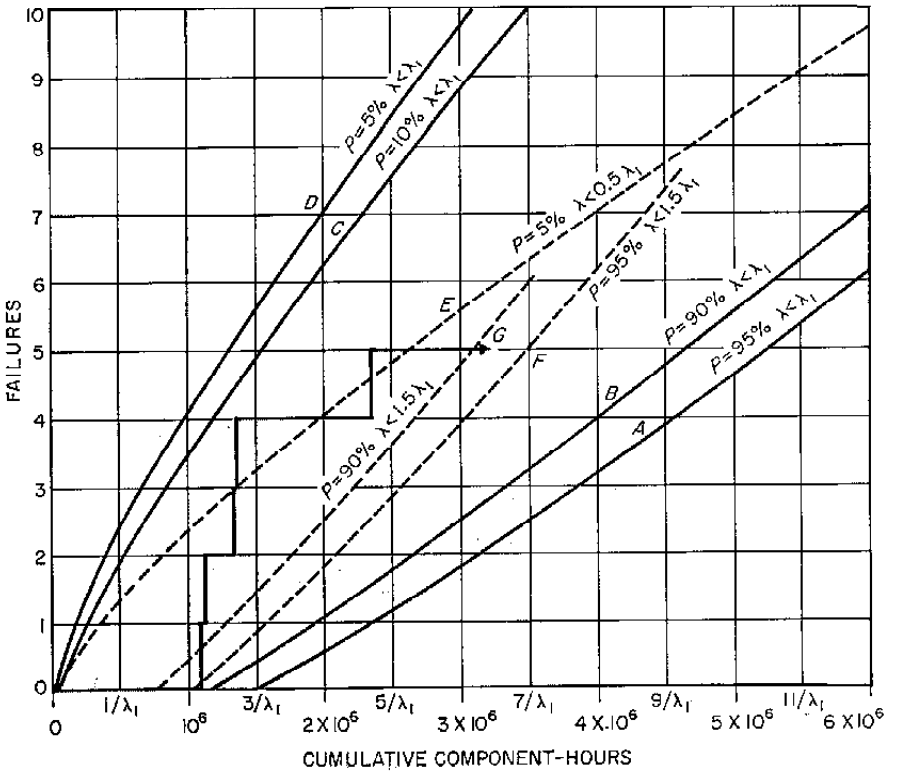


Figure 8—Sequential tests of tantalum capacitors. λ_1 is the target failure rate. In the first example, $1/\lambda_1$ equals 5×10^5 component-hours. The lettered curves are for various failure rates λ for the confidence levels P indicated.

the temporary loss of a qualification. A considerable discussion would be required on this subject.

An extension of this policy would lead to the pooling of confirmed information originating with one or several manufacturers and one or several users. There is a marked tendency in this direction in the form of setting up veritable testing "pools." The Electronic Component Reliability Center is an example of such. This group is sponsored by 13 important firms and it redistributes analyzed and synthetized information. But the publication of this information is private and it cannot be cited here.

Thus the tendency is toward a national or international professional reliability organization, which does not eliminate the need for independent testing by each company.

5.2.2 *Other Problems to Be Solved*

There are two further difficulties that we have not yet considered. One is the validity of the extrapolations connected with accelerated tests. The conditions for acceleration may introduce a profound alteration in aging mechanisms instead of a real acceleration. This is in part plausible and it is a physicist rather than a statistician who must penetrate the depths of knowledge of aging phenomena. It is obviously senseless to carry out accelerated tests at the melting point of solder.

On the other hand, an intransigent attitude does not appear justifiable. It is clear that aging is the result of a number of superimposed processes, which are different functions of exterior conditions, notably temperature. Thus a particular mechanism A can be preponderant at low temperature and another B preponderant at high temperature. It would seem that the essential thing is for the transition to be *continuous* and, if possible, *monotonous* (in the mathematical sense of the word). Then, it is always possible to discover a theoretical or empirical quantitative law for the whole phenomenon that will permit the extrapolation.

The latter considerations relate to the more-imponderable parameters such as: humidity, vibrations, human errors, et cetera. Unfortunately, no correlation has yet been established with a sufficient precision between the results of climatic tests and the behavior of a given component under normal conditions. This represents a tremendous gap, and attempts are currently being made to fill it. The same is true for shock, vibration, et cetera. For radiation, deterioration can in principle be accelerated by exposure to a very-dense radiation flux.

There remains as the second difficulty the question of human errors, which are very difficult to evaluate. We can only state that if a single failure will appreciably affect the results of a test, a test fault involving such a failure is no longer tolerable. The tests and notably the measurements and recording of parameters should be entirely automatic, as should the numerical calculations that follow.

6. General Conclusion

Today, the reliability of electronic components has reached such high levels that its determination is becoming more and more difficult. The obstacles to be met transcend the possibilities of even large companies and it is only by a pooling of efforts that a solution may be found. If this summary of the situation seems pessimistic, it is for the user's sake. The latter should constantly keep in mind the numerous uncertainties that influence measurements in this domain and consequently should provide for safety margins in accordance with the importance of the envisaged application. In particular, he will have to use with circumspection the failure-rate values published without sufficient information on the conditions under which they have been established.

7. Acknowledgments

The author wishes to thank the various testing laboratories of the ITT Europe companies that furnished an abundance of examples of methods and experimental results. His gratitude is addressed to the following persons in particular: Messrs. P. Morain and J. Simonet (relay testing at Le Matériel Téléphonique); M. B. Derjavitch (capacitor testing at Laboratoire Central de Télécommunications); Dr. J. M. Groocock (Transistor Division of Standard Telephones and Cables, London); and Dr. W. Ackmann (of Standard Elektrik Lorenz, concerning the Weibull-Stange methods).

8. Symbols

A, B = coefficients of Arrhenius' law
a = numerical coefficient
h = number of hours of test
$k = nh \times \lambda_1$
m = number of failures observed
N = number of survivals at time t
N_0 = number of components in the tested lot
n = number of samples
P = confidence level
P_d = dissipated power
R = proportion of survivals at time t
S = proportion of failures at time t
T = Absolute temperature
t = time
$\bar{t}(\bar{t}_1, \bar{t}_2 \cdots)$ = statistical mean life
v = speed of reaction
x = unknown or unspecified variable
α, β, γ = coefficients of Weibull's function
λ = failure rate (in general)
λ_0 = observed failure rate
$\lambda_1(\lambda_2 \cdots)$ = specified (expected) failure rate
τ = life at 50-percent survivals.

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Mr. Meuleau collaborated with Professor Goudet in preparing the book "Les Semiconducteurs—Diodes, Transistors et autres Applications."

Microelectronic Devices

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1. Introduction

The large electronic systems of to-day use circuits in which a particular assembly of components may appear hundreds of times. Designers are therefore thinking in terms of integrated circuit units, which may consist simply of two or three components in one package or which could be more-complex units equivalent, for example, to one or more stages of a shift register.

Although the impetus for the development of microelectronic components came mainly from the need for smallness, this is often not the main advantage of such devices. In this paper various microelectronic devices are described, and some of the reasons for using particular integrated circuits in place of the separate components are discussed.

2. Types of Microelectronic Devices

The term "microelectronic device" is used here when referring to a miniature component in general, and "integrated circuit" is used to refer to the particular case of microelectronic device having a number of interconnected components intended to perform a specific circuit function. Integrated circuits would normally have only the input, output, and supply terminations externally accessible.

The terms "passive components" (resistors, capacitors, et cetera) and "active components" (transistors, diodes, et cetera) are used to denote the circuit function and not the construction. Thus, a reverse-biased diode intended to act as a capacitor in a semiconductor circuit would be referred to as a passive component.

The various microelectronic devices can be divided into the groups shown in Table 1.

The first group, Integrated-Circuit Modules, is subdivided into standard components assembled in sealed units and into special components, such as multiple devices containing a number of active devices in only one package. Components in a special mounting intended for automatic assembly and interconnection would fall into the latter subdivision.

The second group, Thin-Film Circuits, consists mainly of two types, one having evaporated or sputtered resistors and capacitors with active elements that are subsequently mounted on the circuit separately. The other type, having evaporated active components, is still in a relatively early stage of development.

The third group, Semiconductor Circuits, has all the active and passive components fabricated on the semiconductor by processes similar to those used to manufacture transistors. The significance of the common-substrate connection and isolated-substrate connection will be discussed later.

The groups listed in Table 1 are intended as a general guide only and do not give a rigid

TABLE 1
MICROELECTRONIC DEVICES

Integrated-Circuit Modules		Thin-Film Circuits	Semiconductor Circuits
Modules with Standard Components	Modules with Specially Made Components	Evaporated Passive Components with Transistors and Diodes in Miniature Encapsulations Evaporated Active and Passive Components	Common-Substrate-Connection Circuits Isolated-Substrate Circuits
Transistor-Resistor Tunnel-Diode-Resistor Circuits Complex Logic and Linear Circuits	Multiple Devices Complex Micromodules Tunnel-Diode-Resistor Modules		

classification. There is considerable overlapping. For example, some modules could equally be classed as a multiple device or as a semiconductor circuit.

Figure 1—Matched pair of germanium tunnel diodes coded type *JK60A*.

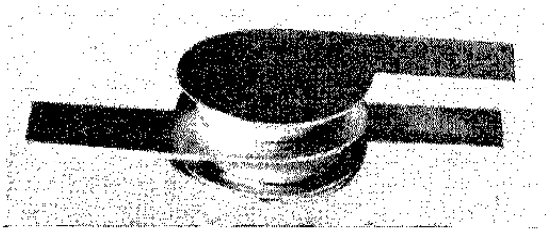


Figure 2—Dimensions of *JK60A*.

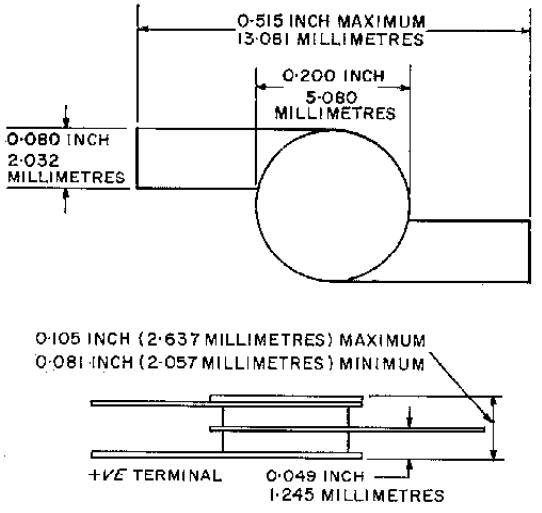
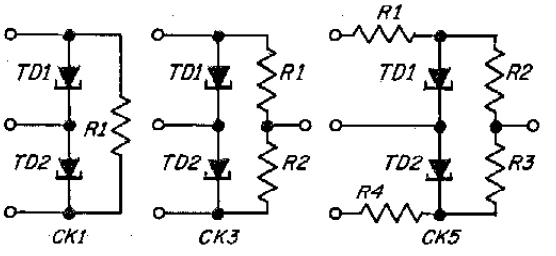


Figure 3—Circuit configurations for typical logic elements based on a Goto pair.



3. Integrated Tunnel-Diode Circuits

The tunnel diode is perhaps the most immediately rewarding device to use in integrated circuits. With relatively simple modules it is possible to achieve quite spectacular improvements over the performance of circuits using lumped components. To obtain the best performance with tunnel-diode circuits it is necessary to maintain close matching of certain device characteristics and to keep the circuit series-inductance values as low as possible. These conditions can best be met with integrated modules.

3.1 Tunnel Logic Elements

The Goto pair, two tunnel diodes connected in series, forms the basis of many high-speed tunnel-diode switching circuits. In this circuit the speed depends on close matching of the peak currents of the two diodes and also matching of the junction capacitances. By mounting two diodes in a microminiature dual package of the type shown in Figure 1, it is possible to obtain an over-all series inductance of less than 2 nanohenries. The dimensions of the package are shown in Figure 2. The method of manufacture enables the peak currents to be matched to within 2·5 per cent within the range of 5 milliamperes ±10 per cent and the junction capacitances matched to less than 10 picofarads about a nominal 25 picofarads, on the *JK60A* device illustrated.

For maximum speed in a bi-stable circuit it is necessary to integrate some of the other components with the matched pair to keep the over-all circuit inductance low. Circuits of the type shown in Figure 3 will operate with diodes having the characteristics given above, at clock frequencies of several hundred megacycles per second, provided the inductance of each resistor and its leads is less than 1 nanohenry. An example of one of these circuits is shown at the extreme right of Figure 4.

To obtain unidirectionality of the logic flow it is usual to use such devices in synchronous

systems. With 3-phase systems, decision rates of over 1 million per second are feasible.

The size of the circuit modules is approximately 3/8 by 5/16 by 3/16 inch (9·5 by 8 by 5 millimetres) plus tags. This is small enough to allow high packing density thus reducing propagation delay in the assembled system. Since tunnel diodes are low-level devices, there should be no difficulty with excessive heat dissipation due to high packing density.

The use of the technique of making fast integrated circuits for strip-line systems is of course not limited to tunnel-diode devices. Any system using ultra-high-speed majority- or minority-carrier devices must have very-low-inductance circuits and low-delay connections of the type obtained in the tunnel logic elements.

3.2 Tunnel Memory Elements

The fastest known storage systems can be made with tunnel-diode memory-circuit elements of the type shown in Figure 5. The speed of such stores, which have access times of around 100 nanoseconds, is not, in this instance, limited by the switching speed of the tunnel elements. It is therefore not so necessary to use low-inductance components. The speed is largely limited by the rewrite cycle, which in turn depends on the propagation delay along the read direction. Another very-important factor in the operation of tunnel-diode stores is the sum of the tolerances of the individual components that go to make up each bit.

The main reasons for integrating are therefore:—

(A) To provide suitably shaped small elements that can be stacked close together to reduce the propagation delay in the read direction.

(B) To allow the most-economical combination of component tolerances to be chosen. This can best be achieved during manufacture.

(C) To reduce the over-all size, weight, and complexity of the store. A memory element

Figure 4—Tunnel-diode integrated circuits.

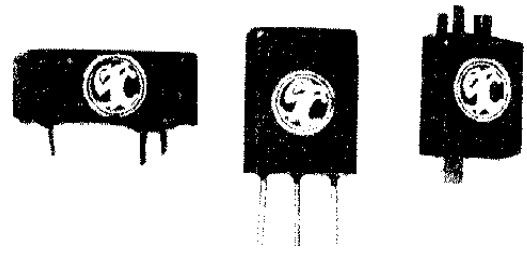


Figure 5—Tunnel-diode memory-circuit configurations.

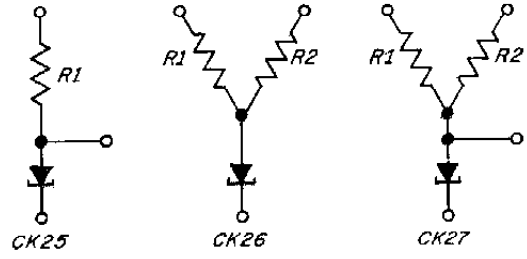
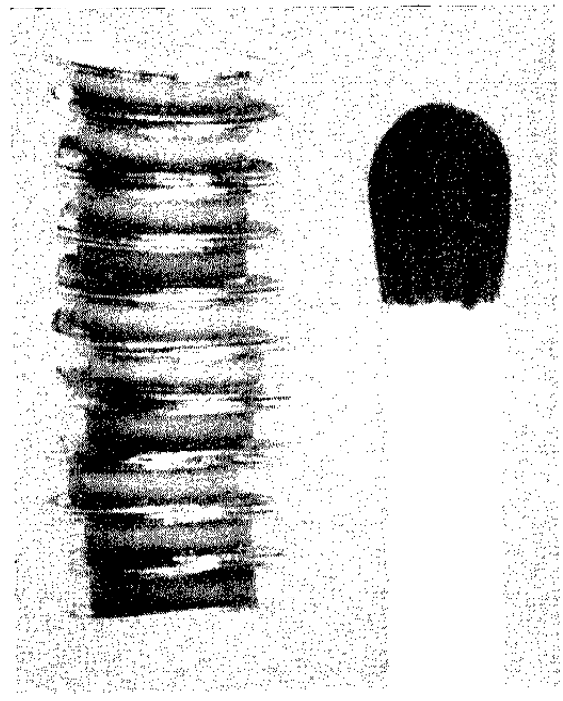


Figure 6—*CK50* before encapsulation.



type *CK26* is illustrated at the centre of Figure 4, having the circuit configuration shown in Figure 5. The modules have been made flat, so that they can be stacked closely together. A stack of 32 words in a 1000-bit store would be approximately 5 inches (127 millimetres) deep by 1 foot (0·305 metre) long.

3.3 Tunnel-Diode Counting Elements

By using a chain of tunnel diodes connected in series, it is possible to perform specialized counting operations. For example, *CK50* has 10 tunnel diodes with peak currents graded so that they switch sequentially in response to input pulses. This unit can be used as the basis of a simple decade scaler circuit. Units of this type may be readily integrated. Figure 6 shows the assembly of 10 tunnel diodes, before encapsulation. The finished device with terminals brought out from the ends of the chain and also from the penultimate diode, for counting and resetting, can be seen at the left of Figure 4.

Figure 7 is a reproduction of the voltage-current characteristic of the device. The peak currents of the diodes range from 7 to 13 milliamperes in equal steps. The maximum valley current is 2 milliamperes.

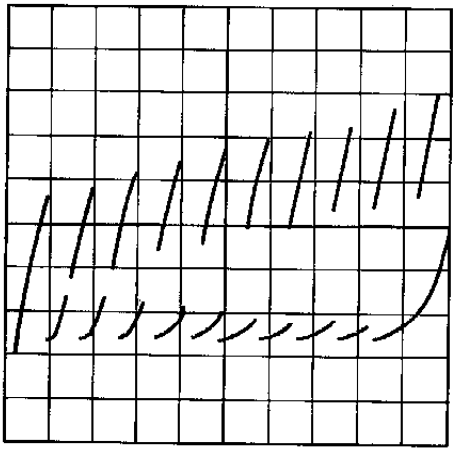


Figure 7—Typical current-voltage characteristic of *CK50*.

4. Multiple Devices

Multiple devices are packages containing more than one active device. Since the mounting and sealing processes for most low-level transistor and diode elements are common, it is advantageous to encapsulate widely used configurations of active devices together in one package. This represents a compromise between using separate components and complete semiconductor circuits. Although multiple devices do not have the advantage of completely integrated circuits, they are more flexible in application, and fit in readily with existing systems using established circuits. Multiple devices can be made either by mounting and connecting the individual elements on suitably isolated areas of the encapsulation or, alternatively, by using semiconductor circuit techniques so that more than one element may be produced in the semiconductor crystal mounted on the header. Multiple devices are frequently mounted in standard transistor outline packages having a suitably increased number of leads, depending on the complexity of the device. A commonly used header has the *TO5* outline with 8 leads, but even smaller headers with the *TO18* outline have been made with up to 10 leads.

The three configurations shown in Figure 8 have been chosen to illustrate some of the reasons for using multiple devices. The technique can, of course, be extended to more-complex configurations. For example, packages containing all the active components for diode-transistor logic gates have been made.

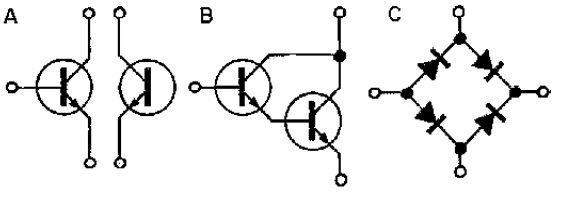


Figure 8—Circuit configurations for multiple devices. *A* is a double transistor, *B* is a Darlington amplifier, and *C* is a diode bridge.

4.1 Double Transistors

Transistors are frequently used in pairs. Where matching is required this can best be carried out during manufacture, and ensured by mounting the devices in a dual package. In direct-current amplifiers using balanced pairs it is particularly important to keep the operating temperatures of the two transistor elements as close as possible. This can best be achieved if the devices are in close thermal contact.

Figure 9 is a photograph of the double transistor type *BFY20*. This is in a *TO5* outline header. The two transistor elements are alloyed to the two blocks set in the glass base. This ensures that the elements are in good thermal contact whilst remaining electrically isolated.

4.2 Darlington Amplifier

The configuration of the Darlington amplifier is such that the terminations are the same as for a transistor. A standard transistor header can therefore be used, as shown in Figure 10. With devices of this type current gains of several thousand can be obtained. The advantages of using a multiple-device Darlington amplifier in place of two separate transistors are:—

(A) Increased reliability due to the over-all reduction in the number of connections between dissimilar alloys.

(B) Reduction in size.

(C) Saving in the cost of the additional transistor case and encapsulation process.

Functionally the Darlington amplifier is a simple semiconductor circuit, because the two transistors are terminated individually as are the transistors or diodes shown at *A* and *C* in Figure 8. The method described here is the multiple-substrate technique of making semiconductor circuits. Another method of making this amplifier is described in Section 6.1.

4.3 Diode Networks

Two or more diodes are used together in many logic and linear circuits. Among the examples are sets of diodes with common p or n regions. These can be mounted together on one chip, giving increased reliability and

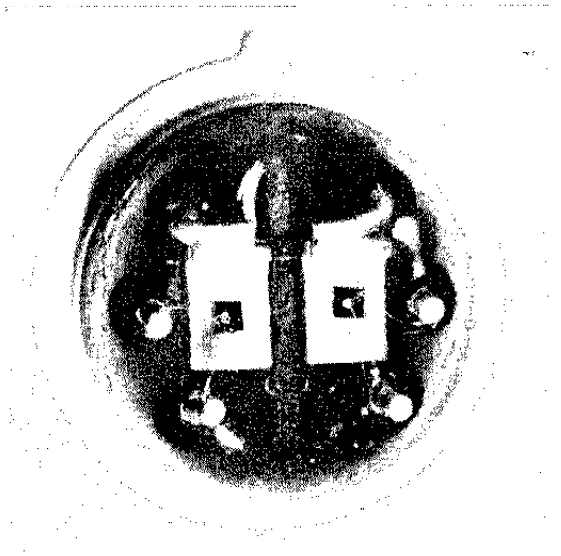


Figure 9—Double transistors in a single mounting type *BFY20*.

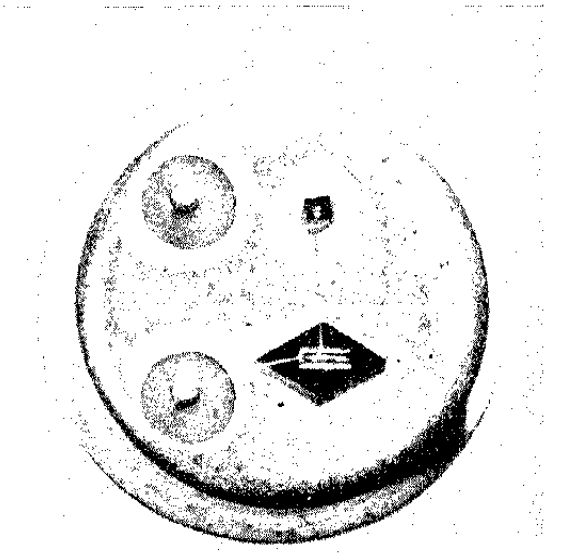


Figure 10—Darlington amplifier.

uniformity of characteristics. Diode bridge and ring networks can be pre-connected in small packages ready for direct use in the circuit boards.

5. Thin-Film Circuits

The two most-widely used techniques for fabricating the passive components on glass or ceramic substrate are:—

(A) Sputtered tantalum for resistors with an oxide of tantalum as dielectric for the capacitors.

(B) Evaporated nichrome for resistors with an oxide of silicon as dielectric for the capacitors.

Both methods enable components to be fabricated with tolerances better than 5 per cent. The nichrome resistors have a lower temperature coefficient than the tantalum resistors. The capacitors using silicon oxide have better power factors than those with tantalum oxide but they occupy a greater area of the substrate. Thin-film circuits and semiconductor circuits are regarded by many users as directly competitive. Since both techniques are in relatively early stages of commercial use it is too soon to predict their relative application or importance when they are in widespread use. However, certain fundamental technical differences suggest that where large electronic systems are considered, these two types of microelectronic devices are in fact not competitive, but compatible. This will be discussed later, when the characteristics of semiconductor circuits have been described.

5.1 Active Devices for Thin-Film Circuits

Experiments are in progress in many research laboratories on the development of evaporated active devices. Because of the difficulty of evaporating suitably doped single-crystal semiconductors, it has not yet been possible to make satisfactory transistors or diodes in this manner. The most promising experiments have been concerned with the manufacture of evaporated field-effect devices and tunnel triodes based on dielectrics such as cadmium-sulphide, which are readily evaporated with aligned crystallites. However, most practical thin-film circuits use separate transistors and diodes wired into the circuit.

One of the difficulties of producing active devices for thin-film circuits has been to provide a small package meeting the encapsulation requirements for the reliable performance of germanium alloy or silicon mesa transistors.

The advent of the planar transistor has created new possibilities in extending the scope and increasing the packing density of thin-film circuits.

The inherent surface passivation of the planar transistor suggests that it may be possible to use the devices mounted directly in thin-film circuits. Thin-film circuits are normally given some protection from the atmosphere in service, and the passivation of the transistor could protect the device until the circuit is finally sealed. An example of a thin-film circuit with unencapsulated transistors soldered directly to the contact areas is shown in Figure 11. However, this involves the handling, storing, and testing of minute devices having leads sometimes less than 0·0005 inch (0·0127 millimetre) in diameter. Where this is im-

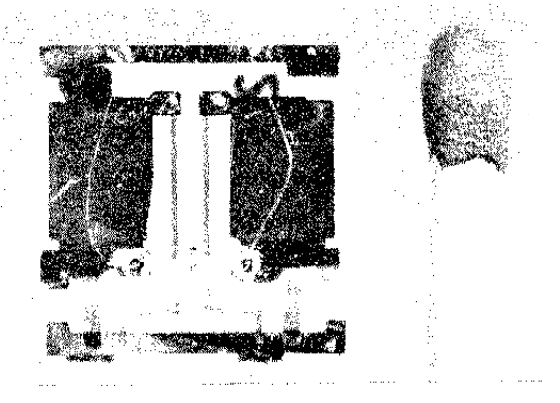


Figure 11—Thin-film circuit with directly wired unencapsulated planar transistors.

practicable, it is possible to take advantage of the fact that the planar device does not need to be encapsulated in a hermetically sealed envelope, but can be sealed in a solid package that can be made much smaller. The other important advantage of planar transistors when used with thin-film circuits is that they combine the properties of high speed with high gain at low currents. This enables efficient low-level circuits to be designed so that a high packing density can be used without the risk of overheating the equipment.

5.2 Solid Encapsulated Transistors

Figure 12 is a drawing of the "Miniflake" transistor. This device consists of a flat plate with part of the area occupied by the encapsulated device, the remainder being metallized areas to which the emitter, collector, and base contacts can be made. Such devices overcome the difficulties of handling and mounting the unencapsulated planar transistors; yet the over-all height of Miniflake transistors is less than 0·025 inch (0·635 millimetre).

A double-ended-contact pattern gives increased freedom in circuit layout; it enables the device (which has an insulating lower surface) to be used for crossovers if the crossing lead can be tied to one of the three transistor electrodes, and allows the device to be securely anchored to the circuit by its leads.

Figure 13 shows two Miniflake transistors mounted in a thin-film multivibrator circuit. Note that the transistors lie over the area occupied by the passive components, thus giving the maximum component packing density. Where the contact pattern on the thin-film circuit has been specially laid out to connect with the transistor, the device may be inverted and sweated directly to the circuit without intermediate leads.

6. Semiconductor Circuits

Semiconductor circuits have been made in germanium and also in silicon using mesa or planar construction. Research is being carried out in using the highly insulating properties

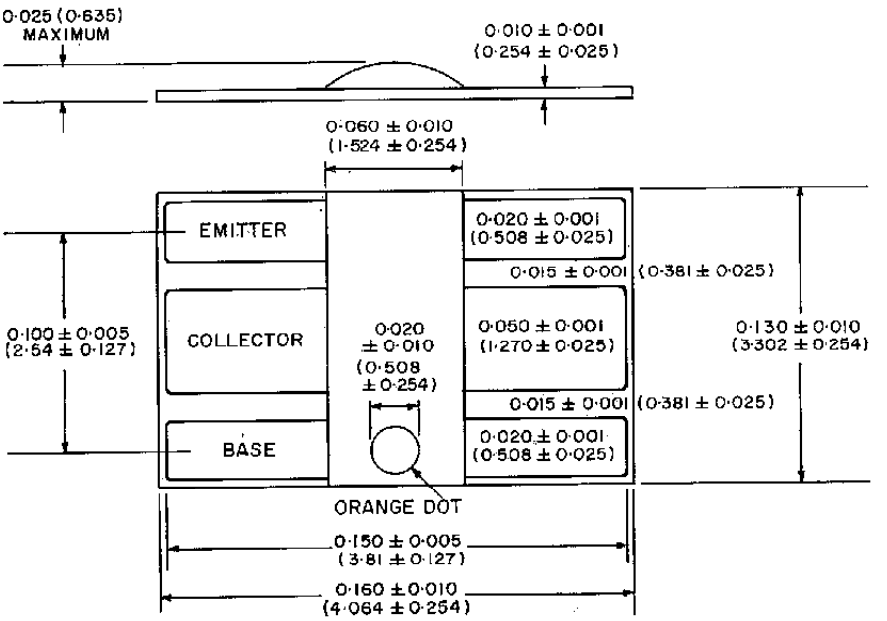


Figure 12—Dimensions in inches (millimetres) of a Miniflake transistor.

of intrinsic gallium arsenide as a substrate for semiconductor circuits. In this paper only the silicon planar technique will be described. By using essentially the same processes as for making transistors, it is possible at the same time to fabricate also diodes, resistors, and capacitors, but not inductors. It is possible to simulate resonant circuits with resistors and capacitors by using series-parallel connection, or by introducing piezoelectric elements into the circuit.

Figure 13—Thin-film circuit with Miniflake transistors.

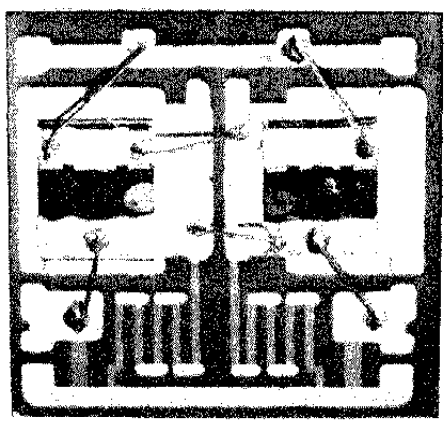
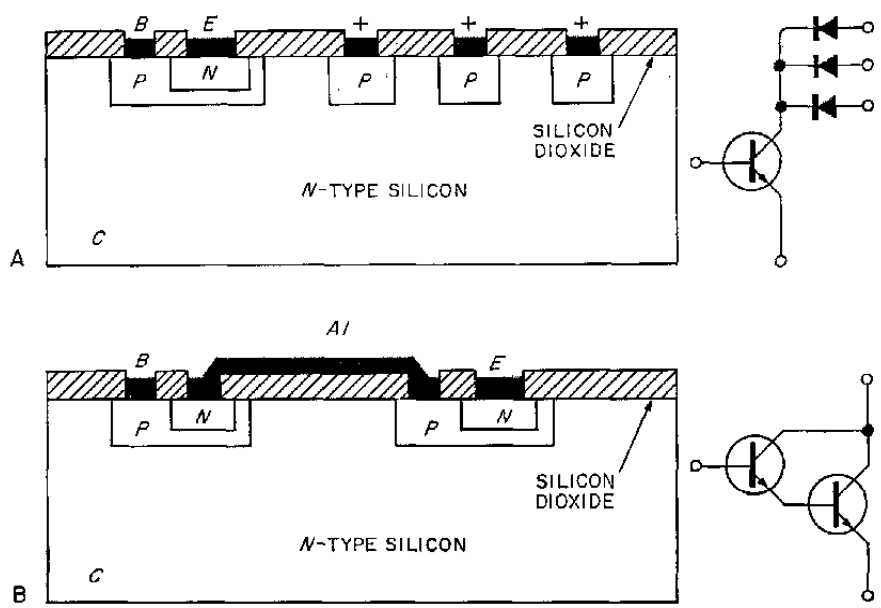


Figure 14—Semiconductor circuits with common substrate connections. *A* is a transistor diode gate and *B* is a Darlington amplifier.

6.1 Semiconductor Circuits with Common-Substrate Connection

Many useful circuit elements can be made with configurations of devices having all *n*-type or *p*-type electrodes common. It is thus possible to diffuse the devices in adjacent positions on the silicon, using the standard transistor-fabricating techniques. The semiconducting substrate provides the common electrode. This type of construction is the simplest form of semiconductor circuit.

Two circuits using common substrate connection are shown in Figure 14. A transistor gate with three output diodes is shown at *A*. Starting with an *n*-type substrate, the *p*-type base region of the transistor and the three diodes are diffused at the same time, using the silicon-diode masking techniques of the planar-transistor process. The position and geometry of the devices are controlled by photolithographic techniques to within 0·0001 inch (0·0025 millimetre). Whilst the diodes are protected by a new layer of oxide, the



emitter of the transistor is diffused through a window photo-etched in the oxide.

A semiconductor version of the Darlington amplifier can be made using the configuration shown in Figure 14B. This consists essentially of a low-level transistor adjacent to a high-level transistor. The collectors of the two transistors are joined by the common *n*-type substrate. The emitter of the low-level transistor is joined to the base of the high-level transistor by an evaporated aluminium film over the insulating silicon-dioxide layer.

It is possible to introduce passive components into semiconductor circuits with common substrate by using the bulk resistivity and junction capacitance.

Common-substrate-connection semiconductor circuits impose severe limitations on the type of circuit configuration that can be used, but their electrical performance resembles more closely the behaviour of the equivalent conventional circuit than that of the isolated-substrate semiconductor circuit.

6.2 Isolated-Substrate Semiconductor Circuits

6.2.1 *Technology*

To obtain the greatest freedom in circuit design it is necessary to have all the components electrically isolated. This is achieved by introducing an extra *p-n* junction into the semiconductor circuit as shown in Figure 15. Consider first the transistor at Figure 15A. Starting with a *p*-type substrate, for example, the silicon is oxidized as in the first stage of making a planar transistor. A hole is now etched corresponding to the total area to be occupied on the silicon by the transistor. A deep *n*-type diffusion is now carried out through the oxide mask resulting in a *p-n* junction to substrate. If this junction is reverse-biased, then the *n*-type region remains isolated from the substrate. The base region and emitter region are diffused using the standard planar transistor technology.

Contacts are alloyed to the three regions, thus giving access to all the electrodes from the top surface of the slice. A diode is made in a similar way but with only two diffusions. By maintaining the substrate at the most-negative potential of the circuit, all the device-to-substrate junctions will be reverse biased and all the components will remain electrically insulated. The silicon has a protective oxide covering all but the component contact areas so that components can be interconnected by evaporating aluminium leads across the top of the slice as in the case of the common-substrate-connection semiconductor circuit.

Resistors are made by cutting a narrow channel in the oxide layer during the first diffusion and making contact to the ends of the diffused path, as shown in Figure 15B. Capacitors are made by using large-area reverse-biased junctions. The dielectric properties of the silicon dioxide can also be used to form capacitors.

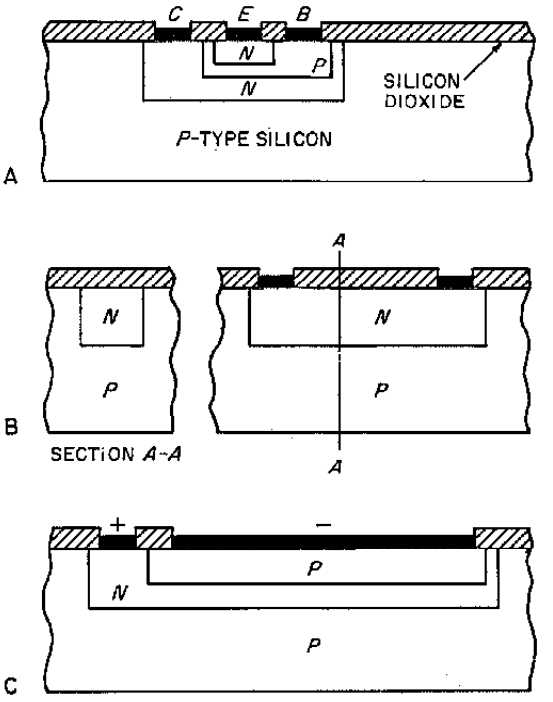


Figure 15—Isolated substrate semiconductor circuit components. *A* is a transistor, *B* is a resistor, and *C* is a capacitor.

Figure 16 shows the oxide patterns at the contact window-etching stage in the manufacture of a device comprising four transistors (centre), two resistors, and two capacitors. The area occupied by these components is approximately 1·5 by 2·5 millimetres.

Figure 17 shows the device with the evaporated-lead pattern mounted on a *TO5* header with thermo-compression bonding of the leads to the external termination of the circuit.

6.2.2 *Reliability*

Semiconductor circuits should provide the most-reliable circuit element because all the components are made at the same time by the minimum number of processes using very-pure materials and highly controlled techniques. By directly connecting the components on the silicon the possibility of failure of interconnecting is also reduced. From results quoted in the United States the reliability of semiconductor circuits equivalent for example to one complete stage of a shift register could be better than 0·01 per cent per 1000 hours. However, with this sort of reliability many devices must be tested and this is of course a much-more-expensive exercise than testing separate components of comparable reliability. Most of the significant results available to date are based on relatively small numbers of early samples made before manufacturers had established the processes. The results quoted for these circuits are one to two orders less reliable than the estimated figure given above. The performance and reliability of semiconductor circuits of the type described in this section will be the subject of a subsequent paper.

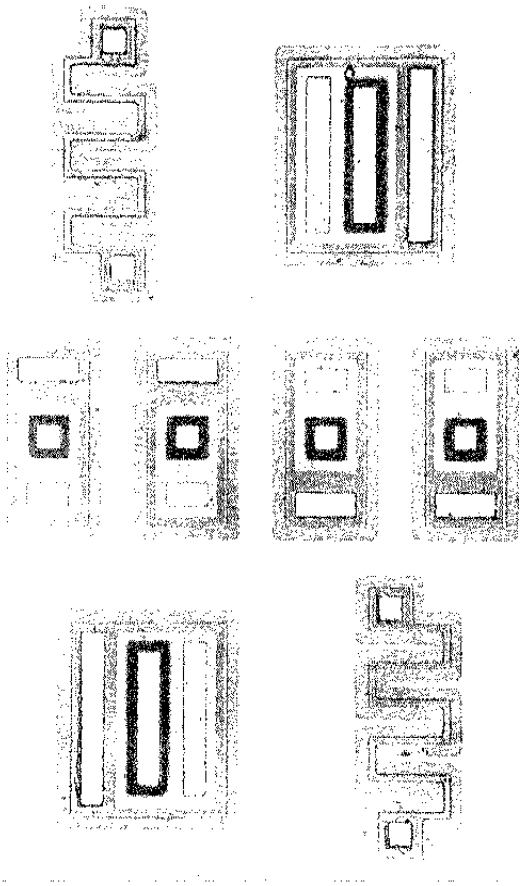


Figure 16—Contact window-etching stage in the manufacture of a semiconductor circuit.

6.2.3 *Packing Density*

The packing density possible in the semiconductor circuit is limited more by heat dissipation considerations than by device size. With low-level logic circuits, very-high packing density is possible. For example, a complete digital integrator using the equivalent of approximately 3000 conventional components per cubic inch has been designed.

6.2.4 *Comparison of Semiconductor Circuits and Thin-Film Circuits*

Returning now to the comparison with thin-film circuits it is clear that on the question of compactness and reliability the semiconductor circuit is to be preferred. The question of cost must be related to the number of devices required. An integrated circuit is obviously more specialized in application than a separate component. The tooling cost of producing a semiconductor circuit (mainly the cost of design and mask manufacture) is very-much higher than the cost of producing the equiva-

lent thin-film circuit. On the other hand, the final cost per circuit should be less for a complete mass-produced semiconductor circuit than for a thin-film circuit plus active devices. Thus on the question of cost it would seem that semiconductor circuits would be preferable where very large numbers are required and thin-film circuits where repetitive circuits are needed in quantities that justify an integrated circuit, but do not justify the cost of a semiconductor circuit. Taking a large digital system for example, leaving out the considerations of reliability and compactness, the shift registers might be made in semiconductor circuits and the read circuits in thin-film circuits.

It is further possible to draw some comparisons in terms of the electrical performance of semiconductor circuits and thin-film circuits. The tolerances on the passive components in semiconductor circuits are coarser than in thin-film circuits. There is also the coupling of components due to the distributed capacitance to substrate in semiconductor circuits whereas the capacitive coupling between components on the thin-film circuits is often no more than on a conventional printed-circuit board. To take these factors into account it may be necessary to carry out considerable re-design before a given circuit function can be performed in the semiconductor circuit, but the corresponding thin-film circuits can be made with essentially the same arrangement of components. Of course both techniques have limitations on the range of component values that can economically be fabricated but higher resistor and capacitor values can be specified in thin-film circuits.

The comparison of electrical performance can be summarized by saying that semiconductor circuits are more suited for applications where the active devices predominate in number and function, whereas thin-film circuits are more suited to applications where passive components predominate. In practice this means that the advantages of the techniques weigh more heavily in favour of digital circuits for semiconductor circuits, and linear circuits for thin-film circuits. This is because in digital circuits the active devices are normally brought from one extreme state to another so that tolerances are of less importance than with linear circuits where equilibrium conditions are maintained by the balance in passive components.

6.2.5 *Combination of Semiconductor Circuits and Thin-Film Circuits*

Naturally such generalizations do not apply in many cases. Calculations on performance or cost can often become meaningless because one essential component in a circuit is not compatible with the technique that is most suitable for all the others. The most-elegant way of solving such problems is to combine the two techniques. This can be done in two ways.

The first method is to use thin-film techniques in conjunction with the semiconductor circuit by using the silicon-dioxide film as substrate for evaporated resistors and as dielectric for capacitors. The circuit shown in Figure 16 has also been made with evaporated resistors in place of the diffused type shown in the

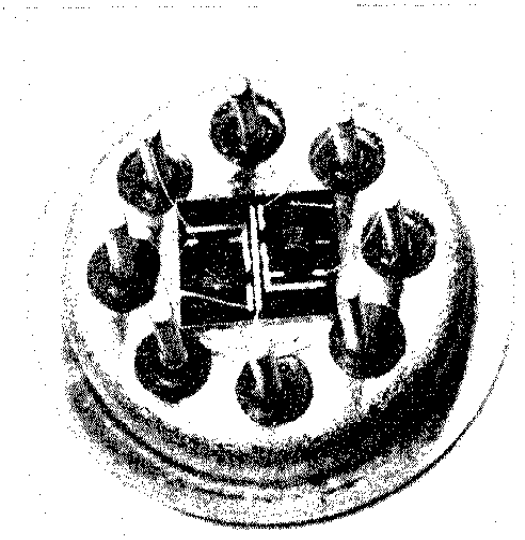


Figure 17—Semiconductor circuit mounted on a *TO5* header.

photograph. These were made with similar sheet resistivity so that in both cases the area occupied was the same. The evaporated resistors are more accurate and have a lower temperature coefficient but the semiconductor circuit made in this way is more expensive because of the extra process involved.

The second method of combining thin-film and semiconductor circuit techniques is to mount the semiconductor circuits on thin-film circuits. Semiconductor circuits having eight active components have been made in a form similar to the Miniflake transistor shown in Figure 12 so that they can be wired directly into thin-film circuits bearing additional associated passive components.

7. Conclusions

Integrated tunnel-diode circuits can be made to perform a variety of logic, memory, and counting functions. In many cases the speeds obtained with integrated tunnel-diode circuits are higher than those possible by any other means. The techniques of making high-speed integrated circuits for strip-line systems will apply to other devices beside tunnel diodes.

Multiple devices can give improved performance particularly in circuits where matched characteristics are important. They can also contribute to the reliability and compactness of circuits.

The use of planar transistors extends the scope of thin-film circuits. The devices can be connected into the circuits in a form that allows high packing density.

Thin-film circuits and semiconductor circuits are not necessarily competitive and will probably both find widespread application within the next few years. Thin-film circuits are more suited for linear and analogue functions whereas semiconductor circuits are at the moment more suited for digital applications. This pattern is likely to remain for some time although there is rapid progress in extending the capabilities in both techniques. In many cases it is advantageous to combine both types of device in a given system or even to combine both techniques in a single device.

Where large quantities are required, semiconductor circuits offer the ultimate advantages in terms of reliability, low cost, and compactness because all the components are made by the same processes as those used for making the transistors.

C. P. Sandbank was born in 1931. He obtained an Honours Degree in physics from London University and the Diploma of the Imperial College in electrical engineering.

In 1953, he joined Standard Telephones and Cables, working as a production engineer, and later as a development engineer. He is now responsible for the development of special semiconductor devices, which include integrated and semiconductor circuits.

Mr. Sandbank is an Associate Member of the Institution of Electrical Engineers.

Silring Mounting of Silicon Power Rectifiers

W. WEISS
Standard Elektrik Lorenz; Stuttgart, Germany

1. Introduction

The high current density in a silicon crystal used as a rectifier generates heat within a small area corresponding to approximately 1 watt per square millimeter of rectifying area. To prevent excessive temperatures that would harm the crystal, this heat is removed by soldering the silicon element to a metal of good electrical and thermal conductivity, usually copper, which is in turn connected thermally to a large-surface cooling unit.

Very-high reverse voltages are also encountered and glass-to-metal and ceramic-to-metal seals have proved to be advantageous for this application.

Permanent protection against deleterious ambient influences, particularly humidity, has required hermetically sealed cases. Manufacturers now use soldering, welding, and cold-welding processes to seal the case.

Basically, the cases of the various commercially available silicon rectifiers are of similar design, the only distinction being the way they are connected to their coolers. Some cases have threaded parts to be screwed to the cooling surfaces; some use flat surfaces for flange-type mounting; and finally pot-shaped cases (automobile rectifiers) with knurled outer walls are pressed into suitably prepared coolers.

Most applications call for combinations of several rectifier cells in bridge circuits, and therefore a space-saving assembly of stacks is very desirable. The Silring mounting provides effectively for compact grouping without sacrificing cooling ability.

2. General Design

Besides having to meet the basic requirements outlined above, it should be easy to assemble rectifier circuits out of individual cells. Unfortunately, this convenience of the well-known assembly principles of the selenium rectifier could not be applied to the conventional case and cooling fin of the silicon rectifier. It was the idea of using the spacing rings between the cooling fins of a stack of selenium rectifiers as cases for the rectifying silicon elements that finally led to the development of Silring.

2.1 Construction

Construction and dimensions of the silicon ring-type rectifier are shown in Figure 1. The case consists of a ceramic ring placed between two nickel-plated copper rings that serve as electric and heat contact areas of the rectifier cell. The silicon crystal is in an eccentric hole in the ceramic ring and soldered to the thicker of the two copper ring plates, connection with the other ring plate being established by a copper wire. The two copper plates are soldered to the partially metallized ceramic ring, thereby hermetically sealing the rectifying element.

The diffused silicon wafer may be soldered in with either polarity. To obtain a better transfer of the heat, however, the crystal element is always soldered to the thicker copper plate which, therefore, should be connected to the cooling plate.

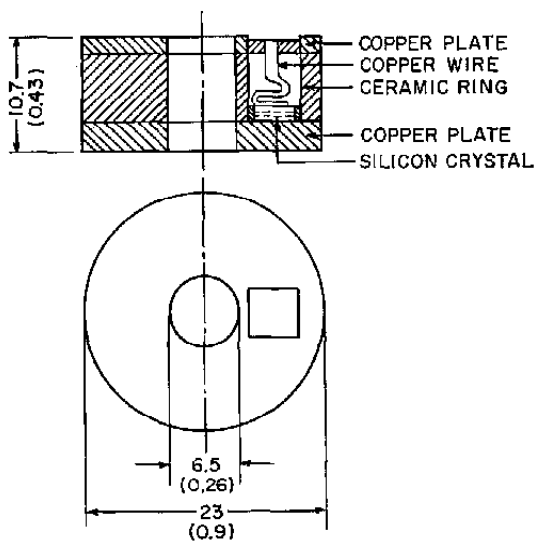
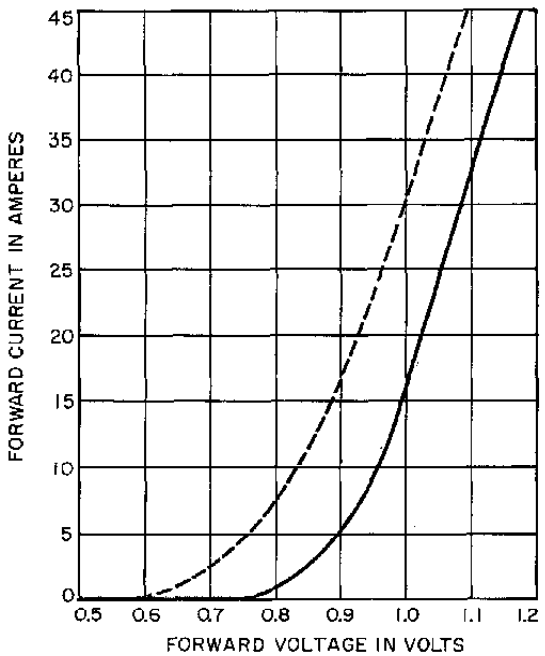


Figure 1—Construction of the Silring silicon rectifier element. Dimensions are given in millimeters (inches).

Figure 2—Mean forward characteristics of the Silring cell at a case temperature of 140 degrees centigrade (broken line) and 25 degrees centigrade (solid line).



2.2 Electrical Characteristics

The physical dimensions of the rectifying element influence the maximum current ratings. Small crystal areas necessitate too many parallel connections for useful ratings while large areas present contact problems and are liable to contain inhomogeneous crystal structures. Large crystal areas also encounter heat transfer difficulties because heat is bound to accumulate at the transition between the case and the cooling plate. This problem can be solved but only through expensive constructional measures.

The maximum current rating of the Silring, based on measurements, is 15 amperes per cell. Figure 2 shows the mean forward characteristics for two different case temperatures. The same ring-type cell may be used for smaller currents. The maximum rated current depends on the dimensions of the aluminum cooling plate. The data are shown in Table 1. The size of the cooling plates ensures that the rectifier case temperature will not exceed the maximum admissible value of 130 degrees centigrade.

The current ratings for an ambient temperature of 35 degrees centigrade, which are normally not given for silicon rectifiers, permit comparison with selenium rectifiers to determine in what particular cases silicon ring-type rectifier stacks are suitable for application under selenium-rectifier operating conditions.

As for conventional silicon rectifier cells, the rated crest working reverse voltage of the Silring depends on the type of crystal used and the manufacturing process employed and may vary between 50 and 600 volts.

3. Silring Stack

3.1 Assembly

The individual ring-type cells, aluminum cooling plates, and connecting tabs are assembled on a steel bolt, which is insulated by a heat-resisting plastic sleeve, and then tightened by a nut. Figure 3 shows a rectifier stack in bridge connection assembled in this manner, with a quarter of the front cooling plate removed to illustrate the arrangement.

Though similar in design and external appearance to the well-known selenium power-rectifier stacks, the ring and the cooling plate

TABLE 1
Ratings of Silring Rectifiers with Convection Cooling

Nominal Current in Amperes		Cooling-Plate Dimensions in Millimeters (Inches)
Ambient Temperatures in Degrees Centigrade		
50	35	
15	17	125 by 125 by 2 (4.9 by 4.9 by 0.08)
10	11.5	75 by 100 by 1 (3 by 3.9 by 0.04)
5	6	50 by 50 by 1 (2 by 2 by 0.04)
2.5	3	Without Cooling Plates

of the silicon ring-type rectifier stacks have distinctly different functions. In selenium rectifier stacks, the plate serves for both rectification and cooling while the ring is used merely as a spacer between the plates. In the silicon ring-type rectifier stack, the plate is used for cooling only and the ring assumes the duty of both rectification and spacing. The aluminum cooling plates are provided with an enamel coating to improve the heat radiation and prevent corrosion.

3.2 Electrical Data

By using suitably dimensioned cooling plates and by connecting the cells in parallel, it is possible to assemble, from only one Silring type, rectifier stacks for rated direct-current outputs of up to 280 amperes with convection cooling and up to 560 amperes with forced air cooling at an air speed of 6 meters (19.7 feet) per second.

Several stacks may be connected in parallel to handle even larger currents, the connecting terminals being designed to permit easy parallel connections.

4. Advantages of Silring

The novel ring-type design of the rectifier case permits an easy assembly of rectifier stacks, with any required internal circuit. The ease of assembly, in turn, permits a high

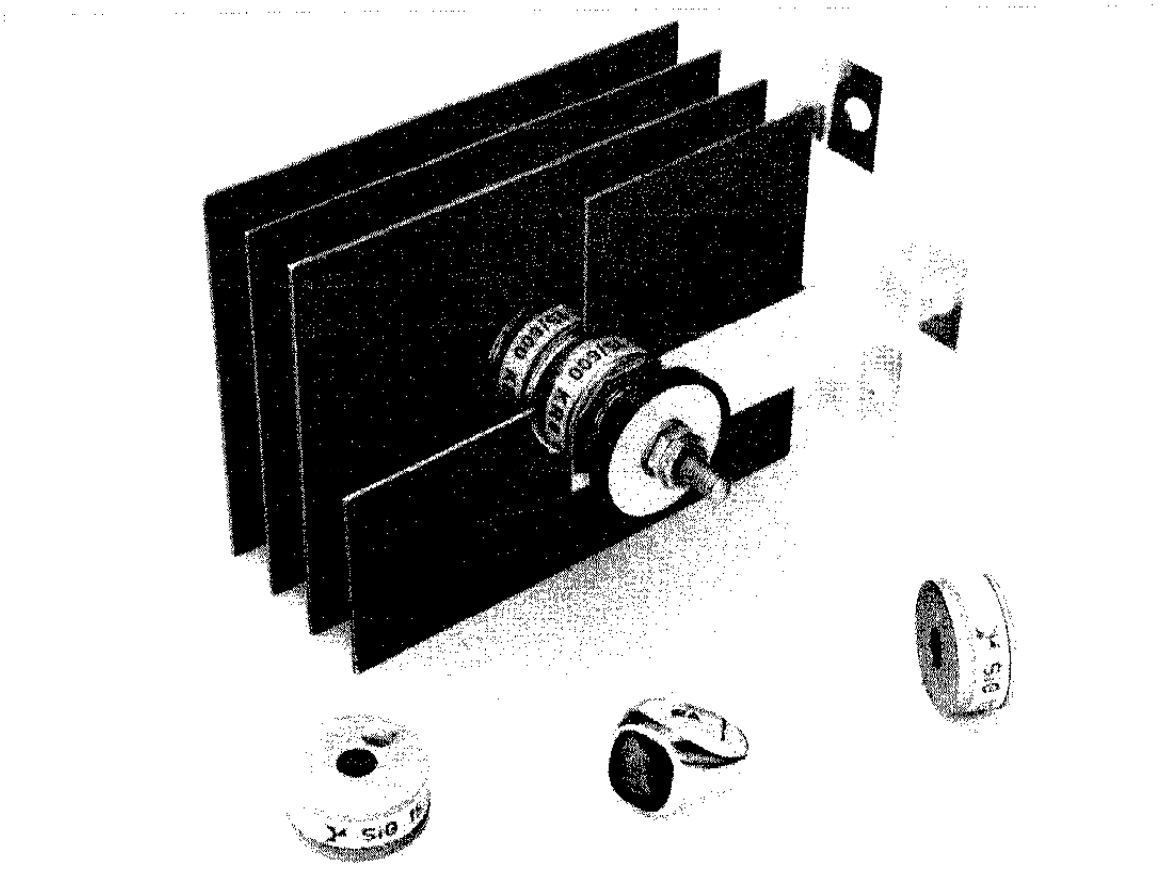


Figure 3—Rectifier stack of four cells in bridge connection with a part of one cooling plate cut away.

degree of standardization for the Silring elements. Even with only one type of Silring cell, rectifier stacks for large currents may be assembled on the building-block principle, necessitating merely different dimensioning of the cooling fins and different internal connections.

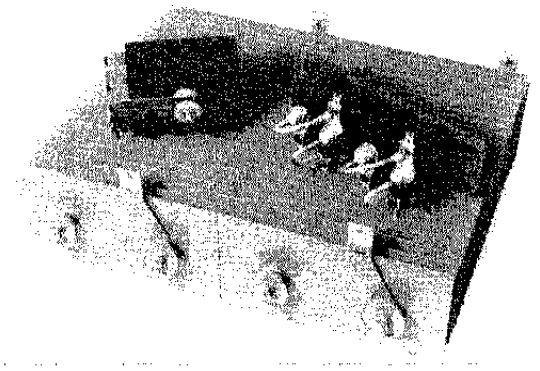


Figure 4—These three rectifiers have approximately the same electrical ratings. Dimensional details are given in Table 2.

The standard elements are dimensioned for reliable, easily controllable, low-cost manufacturing. Spreading of electrical and thermal values is so negligible that no sorting for parallel connection is required.

The flat form of the ring-type case is suitable for spacing of the cooling plates and facilitates mounting and internal circuit connections. Figure 4 shows a Silring, a conventional silicon rectifier, and a selenium unit all having approximately equal electrical ratings. Table 2 gives their dimensions.

During the development work on the silicon ring-type rectifier, close association of the rectifier cell and its cooling system was always envisioned.

TABLE 2
COMPARISON OF SILICON AND GERMANIUM STACKS

Rectifier	Type	Direct-Current Output in Amperes		Maximum Volume		
		Ambient Temperature in Degrees Centigrade		Percent	Cubic Centimeters	Cubic Inches
		35	50			
Selenium	B390/312–24	24	—	100	22 000	1342
Silicon, Conventional	B380/340–20	—	20	8	1 800	110
Silring	B380/340–20	23	20	4	900	55

Werner Weiss was born in Breslau, Germany, on 17 May 1921. He received his doctor's degree in physical chemistry in 1955 from Munich University.

In 1955, he joined the components division of Standard Elektrik Lorenz, where he was engaged in the development of tantalum capacitors, selenium rectifiers, and since 1958 silicon rectifiers. He is now head of the silicon laboratory.

Cause and Prevention of High Reverse Currents in Large-Area High-Voltage Diffused-Silicon Rectifiers

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Producers of electrical devices are subjected to a continuous pressure, generated by both economic and technological demands, to extend the working range of these components. For silicon devices, there is a three-fold demand; for lower reverse leakage current, for higher working voltages, and always for larger-area devices.

At each stage of technological development it is possible to isolate a particular problem the resolution of which will lead to a further extensive growth. Some years ago, the exact control of the absolute concentration and detailed distribution of significant impurities in silicon devices was the pivotal problem. This control is now a matter of simple routine, and a new restraint on further development has been exposed. The presence of so-called non-significant impurities—any elements other than those of Groups III and V in the periodic table—now determines the quality and yield of the more-sophisticated devices. The problem is not restricted to diffused structures; it arises equally in the production of epitaxially grown silicon devices. The work reported here has been directed towards the identification and classification of deleterious phenomena associated with the contamination of silicon by such non-significant impurities, and towards the development of methods for removing them. The future, however, may well see the effect of the controlled addition of these elements turned to the advantage of the device makers.

1. Reverse Leakage and Breakdown

The reverse current of most silicon diodes is orders of magnitude greater than the theoretical, and the breakdown voltage seldom reaches the design figure. Superimposed on the problem of excess current, is that of the spread of characteristics amongst devices from a production line. The variability is much greater than that attributable to variations in junction depth, surface concentration, or resistivity, and it may become manifest only at elevated temperatures. These effects, of minor importance in components working at less than 200 volts, become increasingly significant as the design voltage is increased, until they dominate the yield and performance of components for use in the 500-to-1000-volt range. Moreover an increase in the junction area often results in a catastrophic decrease in device yield.

1.1 Surface Current

The reverse current may conveniently be considered to consist of two components, one originating at the surface of the junction and the other within the bulk of the material. The surface current may arise from a number of causes. The junction may be partially short-circuited by dirt deposited on the surface, or localized surface breakdown can occur at regions of high electric stress. Both these effects can be readily identified by the sensitivity of the device to further etching and washing, and the effect can be eliminated without difficulty. Current associated with surface states can be distinguished from bulk leakage in two ways. If surface effects predominate, the total reverse current will be sensitive to the surrounding atmosphere and will not decrease proportionally as the area of the diode is reduced. In common with other authors [2–7, 12], we have found that with suitable surface treatment the diode performance is principally determined by bulk effects. The excess reverse current consequent on the bulk properties of the semiconductor will be the subject of the remainder of this discussion.

1.2 Bulk Current

In junctions with narrow space-charge regions operated at low voltages, the voltage-independent diffusion current is important. In high-voltage junctions, thermally excited generation current arising within the space-charge

region, and increasing as a fractional power of the voltage, is the more important at and near room temperature. The dependence of these components of the reverse current on the energy level of the responsible generation centre has been dealt with by Sah, Noyce, and Shockley [8] and by Chevychelov [10]. The effect of specific elements has been considered by many authors [2–5, 7, 9]. At voltages near the design limit, avalanche multiplication produces a rapid increase of current with voltage and the relevant data have been produced by Chynoweth [11].

These sources of reverse current adequately account for the characteristics of ideal or "hard" junctions, but much less attention has hitherto been given to "soft" junctions. The outstanding features of this type of junction are the rapid rise in current with voltage, and the anomalous temperature coefficients. An empirical relationship between current and voltage, $I = KV^n$, where n may vary from 2 to 9, has been generally accepted. Goetzberger and Shockley [12] have suggested that metallic precipitates in the space-charge region are responsible for this behaviour. Sandiford [7] suggested that the carrier emission probability

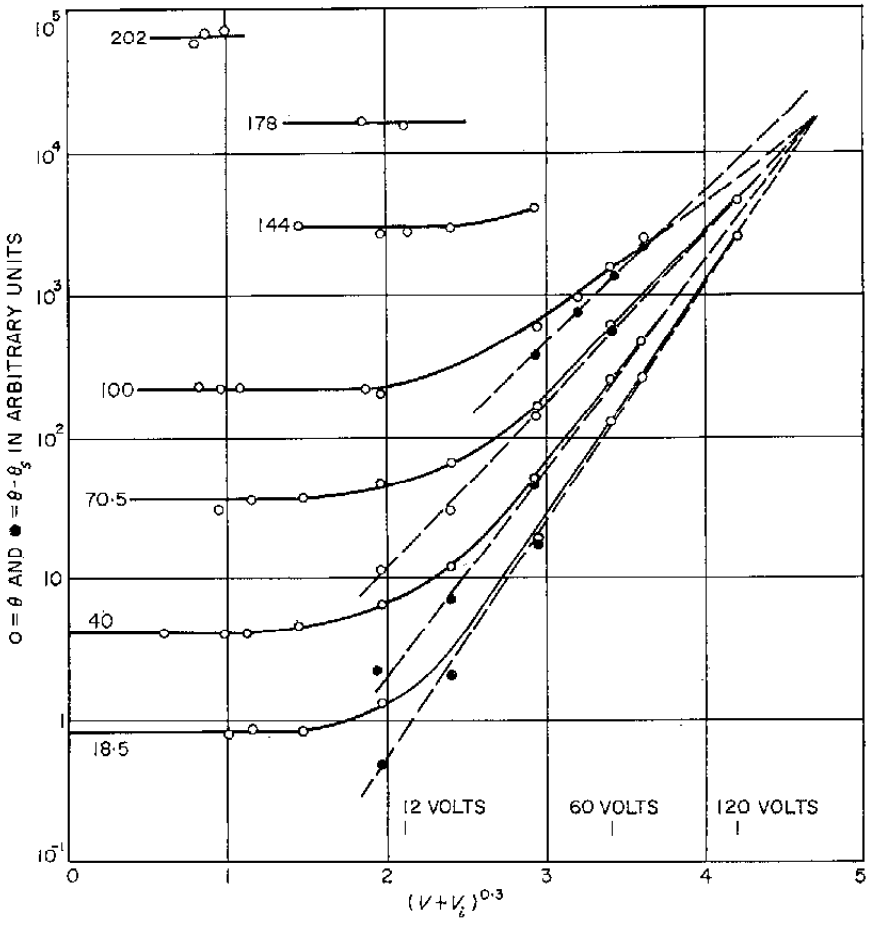


Figure 1—Reverse characteristics of typical soft junction. Gallium was diffused to a depth of 35 microns into 15 ohm-centimetre n-type silicon that had been polished on both faces by a schedule designed to minimize surface damage. The constants for this diode, obtained from this figure and from Figure 2, are given for diode 1 in Table 1. The temperature in degrees centigrade is given for each curve.

is field dependent; whilst Fistul and Orzhevskii [13, 14] have developed the idea of field-dependent emission rates in more detail.

Investigation of the source of the current in soft junctions is clearly a desirable preliminary to removing the cause.

2. Nature of Soft Junctions

Soft junctions were readily obtained by the diffusion of gallium in 10-ohm-centimetre *n* silicon to a depth of 35 microns. The general features of Figures 1 and 2 are typical of such junctions. Soft, epitaxially deposited Zener diodes, despite their lower operating voltage range, exhibit analagous behaviour in the reverse direction at voltages less than breakover.

As the current originates in the space-charge region a normalized current, $\theta = I/(V+V_i)^{1/2\cdot2}$ that is proportional to the carrier-generation rate per unit volume of space-charge region

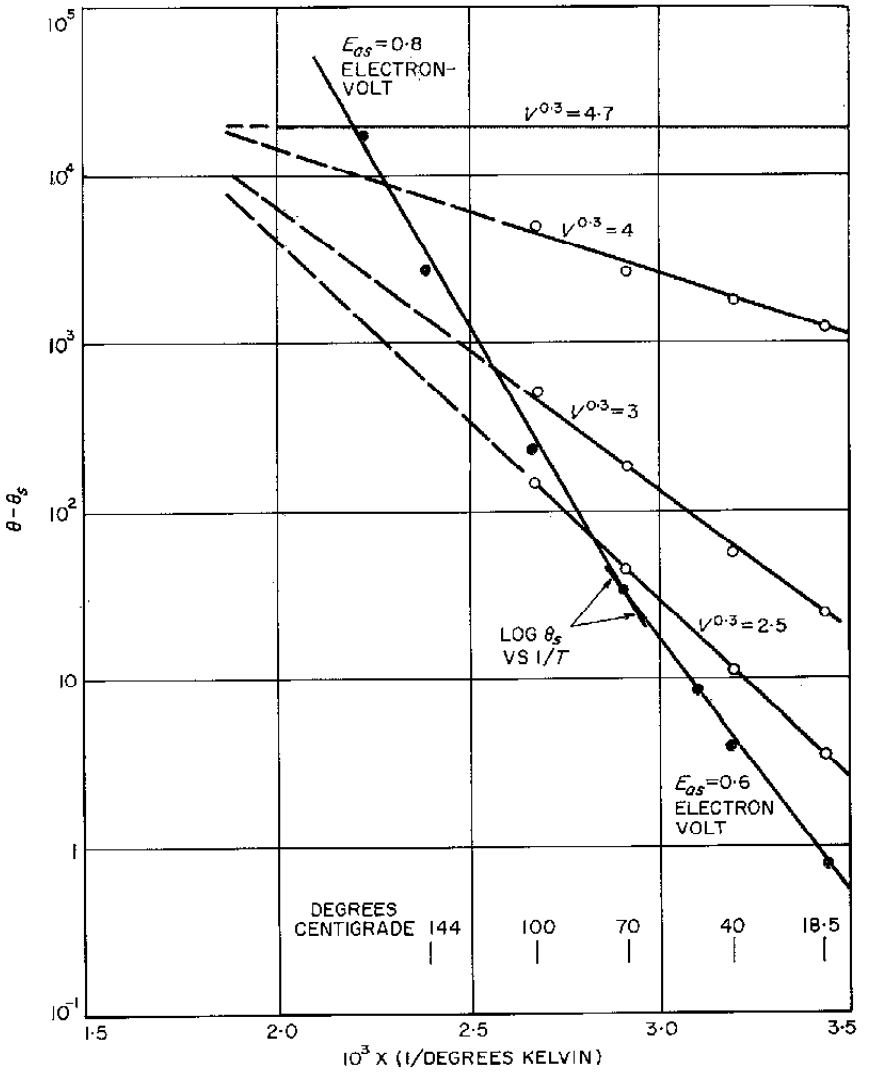


Figure 2—Reverse characteristic of soft junction. The plot of θ_s the saturation current at voltage gives the value of E_{as}.

is used hereafter. The $(V + V_i)^{1/2 \cdot 2}$ dependence of space-charge width on voltage found graphically and by capacitance measurements, agrees well with that found by other workers. The value of V_i, the built-in voltage, appropriate to the temperature, has been used.

2.1 Low-Voltage Region

At very-low reverse voltages (< 2 per cent of V_B) the classical voltage-independent space-charge current θ_s predominates. The temperature dependence is given by $\theta_s = A \exp [- E_{as}/kT]$ where the activation energy, E_{as}, is approximately equal to $E_G - E_{\theta S}$. $E_{\theta S}$ specifies the location of the centre in the band gap and is approximately the depth of the centre from the nearer of the conduction or valence bands. The plot of θ_s against $1/T$ (Figure 2) shows that at temperatures below 70 degrees centigrade the activation energy is 0·6 electron-volt; above this temperature it rises to 0·8 electron-volt. At low voltages, and between 20 and 70 degrees centigrade, the source of carriers is therefore generation centres located at or near the intrinsic level and located within the space-charge regions. It would appear that centres situated about 0·4 electron-volt from a band edge assume the significant role from 80 to 200 degrees centigrade, but the data do not permit an unequivocal distinction between this hypothesis and the alternative possibility that the combined contribution of generation centres within, and a diffusion length beyond, the space-charge region, are responsible for the apparent 0·8 electron-volt activation energy.

2.2 Voltage-Dependent Region

The current in the voltage-dependent region may be approximately represented by $I = KV^n$, but it is difficult to attach any physical significance to this relationship.

Fistul and Orzhevskii [13, 14] have suggested that the effective activation energy of deep generation centres is reduced by the space-charge field to a value $E_{eff} = E_a - K(E)^{1/2}$ leading to the semi-empirical expression

$$I = A \exp \left[- \frac{(E_a - BV^n)}{kT} \right] \qquad (1)$$

where $n = 0{\cdot}27$ for diffused junctions of this type.

A more-accurate expression can be obtained by replacing I by θ and subtracting the voltage-insensitive space-charge current θ_s when

$$(\theta - \theta_s) = A \exp \left[- \left(\frac{E - BV^{0 \cdot 3}}{kT} \right) \right]. \qquad (2)$$

Preliminary study of the data showed that $n = 0{\cdot}3$ for the diodes investigated. From (2) at constant temperature

$$\frac{\partial \ln (\theta - \theta_s)}{\partial (V^{0 \cdot 3})} = \frac{B}{kT} \qquad (3)$$

and at constant voltage

$$\frac{\partial \ln (\theta - \theta_s)}{\partial \left(\frac{1}{T} \right)} = \frac{- (E_a - BV^{0 \cdot 3})}{k}. \qquad (4)$$

A series of values of $- (E_a - BV^{0 \cdot 3})$ from the straight lines of Figure 2 give self-consistent values of E_a and B. From (3) the slope of the straight lines in Figure 1 should lead to a single value for B. Unfortunately this value of B is dependent on temperature and nowhere agrees with the value obtained from Figure 2.

An additional difficulty is that whilst (4) implies a critical voltage V_c, (such that $BV_c^{0 \cdot 3} = E_a$) at which θ is independent of temperature, the additional experimental observation that the produced straight lines of Figure 2 would converge at a critical temperature T_c at which θ is independent of voltage, is not a consequence of this expression.

The integration of (3) and (4), introducing the experimental observations as integration constants and preserving the observed relationship between V_c, T_c, and $(\theta - \theta_s)_c$, leads to

$$\theta - \theta_s = A \exp \left\{ - \left(\frac{E_a - BV^{0 \cdot 3}}{kT} \right) - CV^{0 \cdot 3} \right\} \qquad (5)$$

or, introducing the experimental constants and re-arranging

$$\theta-\theta_s=\theta_c \exp\left\{-\frac{E_a}{K}\left(1-\frac{V^{0\cdot3}}{V_c^{0\cdot3}}\right)\left(\frac{1}{T}-\frac{1}{T_c}\right)\right\} \quad (6)$$

where the constants

$$V_c^{0\cdot3}=\frac{E_a}{B}$$

$$T_c=\frac{B}{kC}$$

and

$$\theta_c = A \exp\left[-E_a/kT_c\right].$$

The values of E_a and B from Figure 2 are now consistent with the unique value of B from Figure 1. The values of the constants for three diodes, prepared from silicon from different sources and of different degrees of softness, are given in Table 1.

2.3 Discussion

It is clear that a new form of behaviour intervened in practice before the values V_c and T_c were reached. It is nevertheless necessary that a model should take cognizance of the result of extrapolation of the observed region of the V, θ, T characteristics. Phenomenologically, this required a model in which the generation rate initially increases with an increase in field and/or temperature. At a critical field F_c, corresponding to the voltage V_c, the temperature should cease to have appreciable effect on θ, and a similar argument applies to the effect of the field at T_c. It should be noted that the measurements have not been extended up to T_c and V_c as the current became unduly high. It is probable that the straight lines of Figures 1 and 2 should be continued, beyond the experimental points, as curves that approach θ_c tangentially.

The following model would appear to meet these requirements, at least qualitatively. Generation centres are assumed to be present, existing in successive stages of ionization and lying at levels E_1, E_2, E_3 in the band gap. The proportion of the total number of centres N existing in dynamic equilibrium in the depleted space-charge region, at a stage of ionization n_1, is a function of E_1, E_2, E_3, F, and T.

The generation rate will be a minimum when all are at the ionization stage n_1, the assumed state at low temperature and fields.

These centres at E_2, lying nearest to E_1, will provide the maximum contribution to the generation rate. As the temperature or field is increased from the lowest levels the effect will be to increase the proportion at the higher ionization levels. Thus at first the proportion of the centres lying at E_2 will be increased. However, a stage will be approached, as the field or temperature is further increased, at which an appreciable proportion of the centres will be further ionized to level E_3. At this stage

TABLE 1

Constants for Representative Soft Junctions

Diodes	1	2	3
E_{as} in Electron-Volts from 20 to 70 Degrees Centigrade	0·60	—	0·6₂
E_{as} in Electron-Volts from 70 to 205 Degrees Centigrade	0·8	—	—
E_a in Electron-Volts	0·9₂	0·8₅	0·9₅±0·05
$V_c^{0\cdot3}$	4·7	4·3	6·1
Critical Temperature in Degrees Kelvin	590	585	625
B	0·2	0·2	0·16
C	4·0	4·1	3·0
θ_c in Arbitrary Units	2×10^4	9×10^3	8×10^3
θ in Microamperes, 120 Volts, 20 Degrees Centigrade	40	100	0·3

The constants refer to the general equations

$$\theta-\theta_s = A \exp\left\{-\left(\frac{E_a-BV^{0\cdot3}}{kT}\right)-CV^{0\cdot3}\right\}$$

and

$$\theta-\theta_s=\theta_c \exp\left\{-\frac{E_a}{k}\left(1-\frac{V^{0\cdot3}}{V_c^{0\cdot3}}\right)\left(\frac{1}{T}-\frac{1}{T_c}\right)\right\}.$$

E_{as} was obtained from the slope $\partial \ln \theta_s/\partial(1/T)$ in the saturation region, E_a and B from $k\partial \ln (\theta-\theta_s/\partial(1/T) = -E_a+BV^{0\cdot3}$ (Figure 2), C from $\partial \ln (\theta-\theta_s)/\partial V^{0\cdot3} = B/kT-C$ (Figure 1), and T_c and V_c are read directly from Figures 1 and 2. Diodes 1 and 2 are for gallium diffusion into polished slices.

Diode 3 for comparison is a partially B_2O_3 gettered slice.

the normal increase in generation rate consequent upon the increase in temperature will be counterbalanced by the transfer of centres from the more-effective level E_2 to the less-effective level E_3.

The mathematical formulation of this model leads to expressions that can only be solved graphically, even for the grossly simplifying assumption of a uniform field in the space-charge region. The value of the model subsists in the indication it provides that centres with multiple ionization levels are responsible for the excess current in soft diodes.

3. Experimental

All the results reported here refer to p-n junctions obtained by diffusing gallium into 5-to-25-ohm-centimetre n-type silicon. The silicon consisted of both zone-grown and crucible-grown material from a variety of sources. The silicon slices were pre-oxidized in moist oxygen (25 millimetres of water) to produce a coherent silica film 0·4 micron thick. The slices were individually laid on a flat quartz boat and inserted rapidly into the hot zone of the furnace.

The diffusion was carried out in an open-tube quartz furnace with the source (specially pure Ga_2O_3) at 925 degrees centigrade and the slices at 1200 or 1245 degrees centigrade. The slices were withdrawn from the diffusion furnace at about 20 degrees centigrade per minute. Hydrogen plus water vapour (10-millimetres partial pressure) was used as a carrier gas.

Commercial gases, piped in copper, were dried by passing through activated alumina and humidified by bubbling through deionized water. An initial series of experiments in which the oxygen was passed through a silica-gel trap at −80 degrees centigrade, the hydrogen through activated charcoal at the same temperature, and both gases through porous metal-dust filters, gave precisely the same results as with the untreated gases. In the majority of the experiments therefore untreated gases were used.

The surface concentration of gallium was $2{\cdot}2 \pm 0{\cdot}2 \times 10^{18}$ atoms per cubic centimetre producing junctions 36 microns deep on both sides of the slice.

In the gettering experiments the boric oxide was applied to the surface of the gallium-diffused slice by vacuum evaporation of anhydrous B_2O_3 glass from a heated platinum strip. Uniformity, absence of dust, and thickness were investigated by visual inspection of the interference colour of the film. The slice was transferred rapidly from the evaporation plant to the hot furnace to avoid moisture pick-up. Raymond [1] has found that boric oxide applied by conventional paint-on techniques produces results consistent with those reported here.

Specimens were prepared for measurement by removing the oxide film from the slice in hydrofluoric acid and depositing nickel on both sides from an electrodeless nickel bath. In some runs the nickel was alloyed with the surface by heating for a short period (2 to 3 minutes) at 600 degrees centigrade in pure hydrogen. In a control run in which the nickel contact was not alloyed, there was no significant difference in the reverse characteristics of the diodes so formed. A pattern of mesas, each 2 millimetres in diameter, was made by masking with vacuum wax and etching in CP8. The wax was removed by trichlorethylene vapour, and the exposed junction surface was stabilized by heating in concentrated sulphuric acid for about 2 minutes at 200 degrees centigrade, washing in recirculated deionized water, and baking in air at 240 degrees centigrade for a minimum of 15 minutes. Measurements were made, unless otherwise stated, in dry nitrogen in the dark.

4. Chemical Gettering

Slices of gallium-diffused silicon, chemically or mechanically polished prior to diffusion,

were coated with boric oxide on one or both sides.

The slices were heated in an open-tube quartz furnace in a stream of oxygen. The effect of time, temperature, and moisture content of the oxygen, is shown in Figure 3 and Table 2. The data, with one exception, refer to measurements on mesas formed on the coated side of the slices. The breakdown voltage was arbitrarily specified as the voltage at which the reverse current of a 2-millimetre-diameter mesa reached 10 microamperes. In all cases the breakdown voltage of the junction prior to treatment was considerably less than 100 volts.

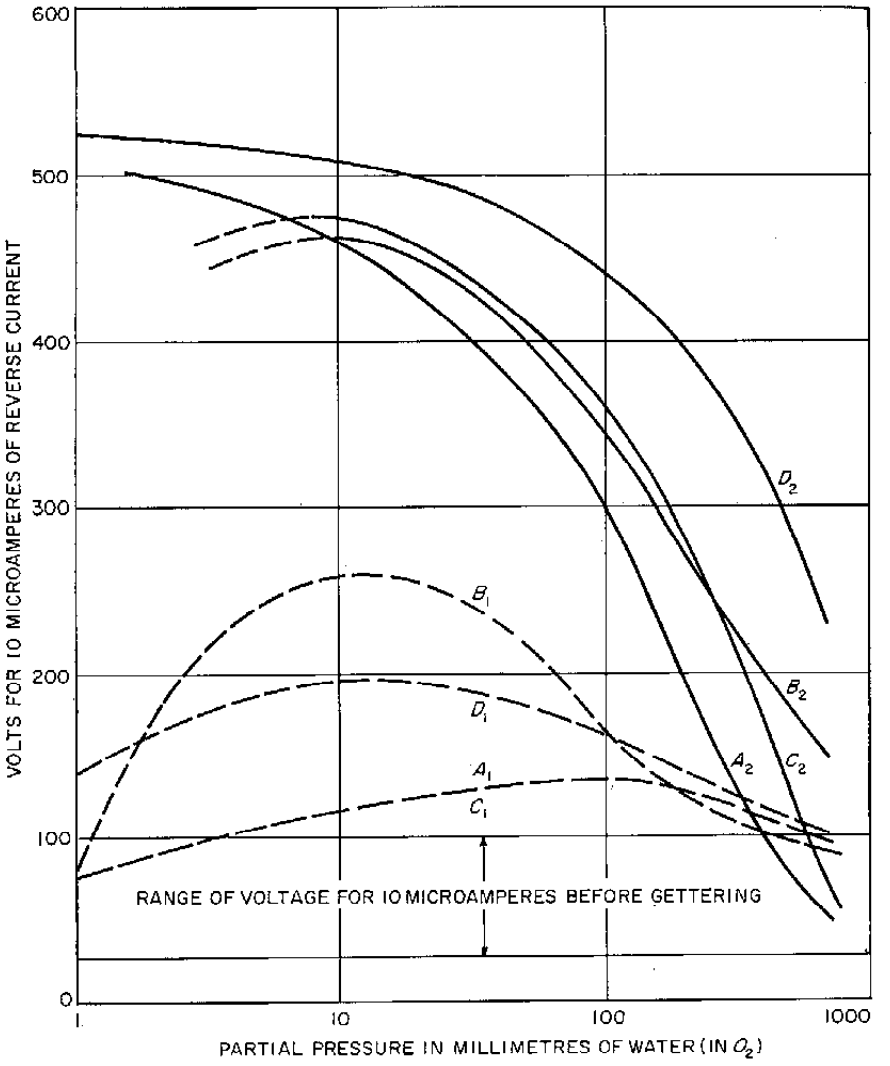


Figure 3—Junction quality as a function of humidity during boric-oxide gettering. Slices with soft junctions 35 microns deep on both sides, were coated on one side with boric oxide and gettered at 1200 degrees centigrade in moist oxygen. The letters on the curves indicate the intervals at which the boric-oxide glass was replaced: *A*, no replacement; *B*, after 3 hours; *C*, after 2 hours; and *D*, after 2 and 3 hours. The total gettering time was 4 hours in all cases. Subscript 2 indicates junctions on the boric-oxide side and 1 signifies junctions on the opposite face.

TABLE 2

QUALITY OF JUNCTION FOLLOWING BORIC-OXIDE GETTERING

Boric-Oxide Gettering		Voltage Range of All Tests	Proportion of Tests in Which Voltage for 10 Microamperes of Reverse Current Exceeded Specified Value	
Temperature in Degrees Centigrade	Time in Hours		Proportion	Volts
1200	1/4	100–400	5/8	300
1200	1/2	310–380	3/4	310
1200	3/4	324–420	15/15	324
1200	1	384–444	5/5	384
1200	1-1/2	456–540	5/5	456
1090	1/4	170–260	7/9	180
1090	1-1/2	450–480	8/8	480

For reference, gallium-diffused ungettered junctions were tested between 40 and 80 volts and none operated at 100 volts with the required 10 microamperes of reverse current.

The quality of the junction on the side of the slice opposite to that on which the boric oxide had been deposited varied from run to run. This was found to be so for slices ranging from 180 to 400 microns in thickness. Traces of contamination left on, or subsequently entering, the uncoated face were found to be responsible for this variability. It was shown that the impurities were deposited on the slice during dissolution of the silica film in hydrofluoric acid, and that further contamination during the subsequent heat treatment was virtually unavoidable. Coating the untreated face of the slice with thermally grown oxide provided no protection.

An experiment in which contamination was deliberately applied to the oxidized surface of a hard junction, which was subsequently reheated to 1200 degrees centigrade for a short period, showed that the oxide film was readily permeable to the likely contaminants.

Slices had both faces polished prior to the gallium diffusion. Boric oxide was applied on both sides of the slice, which was inserted straight into the hot zone of the furnace and then withdrawn rapidly. The atmosphere was of O_2 at a pressure of 25 millimetres of water. Measurements were made on 2-millimetre-diameter mesas.

4.1 MECHANISM OF CHEMICAL GETTERING

The results are consistent with the view of Goetzberger and Shockley [12] that a liquid phase of acidic oxide combines with the electro-positive impurity atoms. The surface concentration of the impurity remains very small until the boric oxide is excessively diluted with silica. It is unlikely that the formation of a solid film of silica is responsible for the cessation of gettering that occurs more rapidly the greater the water content of the oxygen. It has already been demonstrated that silica is readily penetrated by these impurities, and it is more likely that displacement of the equilibrium

$$M + O_2 \rightleftharpoons M \cdot \text{oxide} + B_2O_3 \rightleftharpoons M \text{ oxide} \cdot B_2O_3$$

towards the left, by the increased competition for the B_2O_3 by the silica is responsible.

The effectiveness of short times and low temperatures (Table 2) confirms the view that fast-diffusing impurities are being removed. Amongst such elements iron, copper, gold, and silver are known to have appropriate diffusion constants.

The difficulty of gettering a junction on the remote face of a slice, with boric oxide on the opposite face, was not a function of the thickness of the slice. It was evident that the rate of transport through the slice was adequate, as in those runs in which attempts to preserve the cleanliness of the remote face were successful, the quality of both junctions was excellent. However, the difficulty in completely removing any source of contamination made it clearly desirable to coat both faces of a specimen with the gettering agent.

A less-exhaustive series of experiments showed that a film of phosphorus glass behaved in a qualitatively similar fashion as a gettering agent. The glass was predeposited by passing a stream of dry oxygen carrier gas saturated

by phosphorus pentoxide maintained at 230 to 270 degrees centigrade over the oxide-free silicon slice at 1050 degrees centigrade. The subsequent heat-treatment was essentially the same as that used during boric-oxide gettering.

5. Physical Gettering

Chemical gettering provides an extremely effective method of obtaining hard junctions, but it is not always conveniently applicable to device production. Where the device can be so designed that the boric oxide or phosphorus pentoxide simultaneously produces the appropriate junctions, and acts as a getter, the chemical process is eminently suitable. But concomitant boron or phosphorus doping is unavoidable, and the whole surface of the slice must be coated if reliable gettering is to be obtained.

The new process of physical gettering described below is felt to be capable of more-general application.

The soft junctions discussed above were the result of gallium diffusion into mechanically or chemically polished slices. The mechanical polishing programme was designed to produce a surface with a damaged layer less than 1 micron thick, and the chemical polishing to remove completely the damaged surface layer.

A series of slices was so treated as to produce a surface layer of material of considerable thickness with a high density of lattice damage. The slices, gallium diffused after suitable cleaning treatments, had reverse characteristics similar to those of the chemically gettered slices.

Soft junctions that had received the same surface treatment (after diffusion) and were subsequently reheated for a time short compared with the diffusion time, showed improvements of the same order. Table 3 shows typical results for a single batch of slices which, after a standard gallium diffusion, were divided into 3 groups. Batch *A* was measured without

TABLE 3

EFFECT ON SOFT JUNCTIONS OF SURFACE DAMAGE AND REHEATING

Batch	Slice Preparation	Junction Quality
A	Gallium Arsenide Diffused	30 Volts
B	Gallium Arsenide Diffused and reheated	16 Volts
C	Surface Prepared for Physical Gettering and Reheated (as B)	480 to 540 Volts

further treatment, batch *B* was reheated without further treatment, and batch *C* was given a physical gettering treatment. Junction quality is defined by the voltage at which the reverse current of a 2-millimetre mesa reaches 10 microamperes.

Typical results of similar treatments are shown in Figure 4.

5.1 MECHANISM OF PHYSICAL GETTERING

The similarity between the requisite condition for, and the results of, chemical and physical

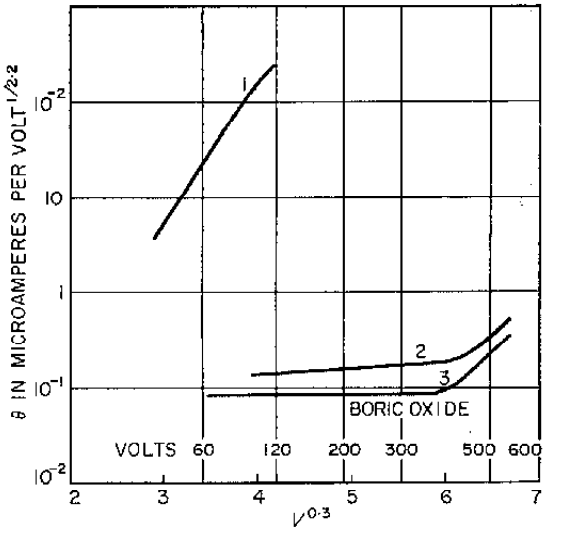


Figure 4—Effect on junction quality of surface preparation prior to gallium diffusion. Curve 1 is for a mechanically polished slice. Curve 2 is physically gettered and curve 3 is boric-oxide gettered. The gallium diffusion was 35 microns deep in all cases.

gettering, suggests that the mechanisms are themselves similar, involving in both cases the transfer of impurities from the bulk, to the vicinity of the surface where they become bound.

Kikuchi and Itzima [16] observed that grinding the surface of a germanium sample and re-heating for 1·5 hours at 650 degrees centigrade produced a substantial increase in the lifetime. They suggested that precipitation of copper at lattice defects was occurring. Pugh and Samuels [17] have shown that the primary result of abrading a germanium surface is the introduction of an extensive system of micro-cracks extending below the broken surface layer.

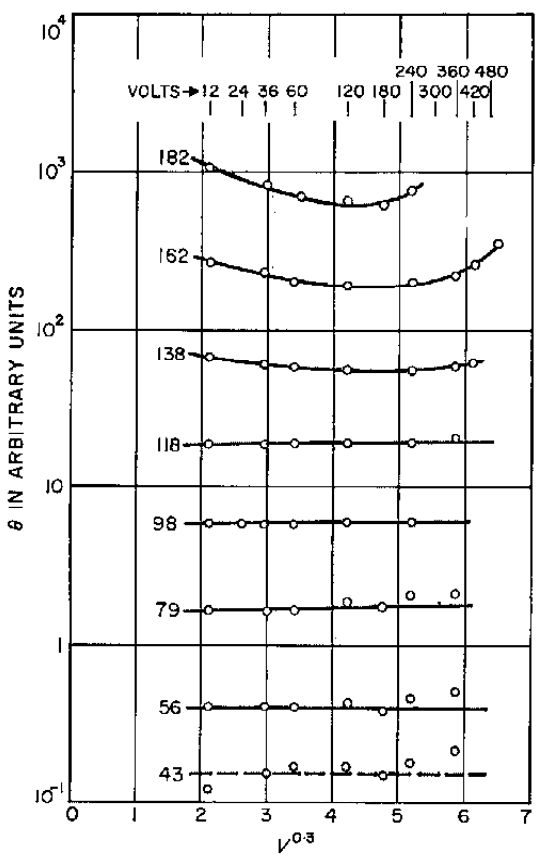


Figure 5—Reverse characteristics of typical gettered junction. The temperature in degrees centigrade is noted for each curve.

The following picture emerges: specimens that have been subjected to a long-term temperature treatment are rich in a variety of impurities. These impurities may be present in saturated or supersaturated solution, and as precipitates. From experiments in which soft diodes with undamaged surfaces were cooled from elevated temperatures rapidly or slowly, without significant change in their diode characteristics, it appears that further precipitation of the impurity cannot be induced in the absence of surface damage.

Thus, in soft diodes with undamaged surfaces, the impurities are distributed at room temperature in a manner that cannot readily be disturbed by further heat treatment. They may be present as precipitates that are dissolved and reprecipitated on heating and cooling, or they remain in solution, saturated at high temperature, supersaturated at room temperature.

On the other hand, a new process can occur if the surface of the specimen is damaged before reheating. Fuller and Ditzenberger [19] and others, have shown that the diffusion rate of many transition elements in germanium and silicon is very dependent on the dislocation density. In the presence of dislocations, the vacancy density permits equilibrium to be maintained between substitutional and interstitial atoms. Whilst interstitial solubility is normally an order of magnitude lower than substitutional, rapid diffusion ($D_{1000} = 10^{-5}$ centimetre^2 second^{-1}) can occur only interstitially.

During the cooling of a specimen with a damaged surface, precipitation will occur most readily at the broken surface, and around the micro-cracks. A concentration gradient is set up, in which the impurity will diffuse from the bulk to the surface until the flow is quenched by the drop in diffusion constant as the temperature falls. In view of the low activation energy of the interstitial diffusion process, flow will persist during the greater part of the cycle. Chemical gettering may also be partly

dependent on the defects present in silicon surfaces containing a high boron concentration. Only under such conditions can substitutional-interstitial equilibrium be maintained, permitting rapid interstitial diffusion to occur.

This view of the physical process is in keeping with an experiment in which two samples of silicon with soft gallium junctions 10 and 35 microns deep respectively have been treated to produce shallow surface damage.

After subsequent heat treatment, the deep junction had become hard while the shallow one was further degraded. Clearly the shallow junction was within the layer affected by surface damage. After reheating, further changes had occurred that were presumed to involve the removal of impurities from the region of the deep junction and precipitation in the vicinity of the shallow one.

Thomas [20] has shown that the mode of precipitation of copper in silicon depends on the surface treatment. Samples saturated at 1100 degrees centigrade with copper and cooled to room temperature in 10 seconds exhibit characteristic etch figures. The number and depth of the precipitates responsible for these etch figures depends on the surface preparation of the silicon. The thickness of the layer, in which precipitation at a density of 10^5 to 10^6 centimetre^{-2} occurred, was of the same order as that of the damaged layer. If the silicon surface had been previously chemically polished, the precipitation occurred uniformly throughout the material as a density of approximately 10^3 centimetre^{-2}. Thermal oxidation or prolonged annealing at elevated temperatures removed the worked layer present after mechanical polishing.

6. Nature of Hard Junctions

The characteristics of junctions produced by physical and chemical gettering are very similar. Over most of the range of reverse bias, θ_s is constant (Figure 5). The activation energy of the generation process is very close to $E_g/2$ below 70 degrees centigrade, and increases to about 0·75 electron-volt at higher temperatures (Figure 6). At high temperatures the initial decrease in θ_s with increasing voltage, is the result of a significant contribution by the voltage-independent diffusion current. The identification of a second process with an activation energy of 0·75 electron-volt must therefore be treated with reserve, as the diffusion current contribution with an activation energy of 1·2 electron-volts cannot be separately estimated.

The junctions have been designed to break down (avalanche mode) at about 600 volts. It is therefore not possible to separate avalanche effects in the high-voltage region from voltage-sensitive generation of the type found

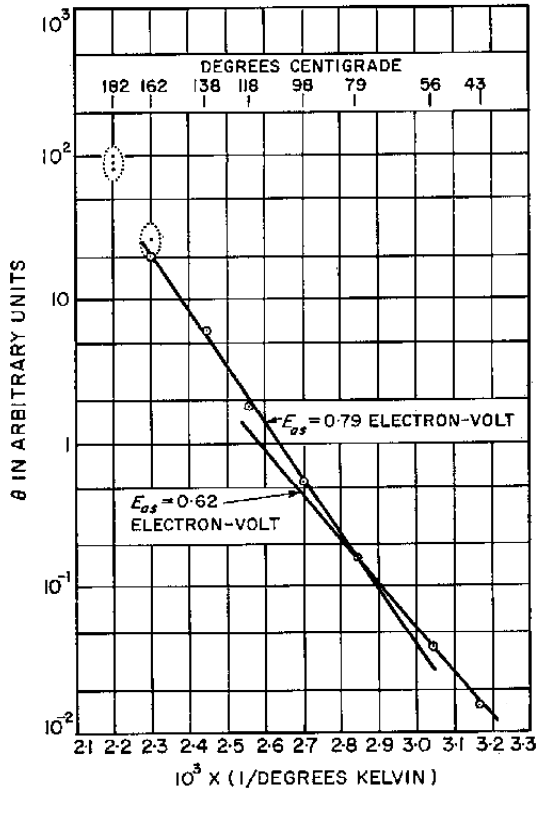


Figure 6—Reverse characteristics of the hard junction of Figure 4.

in soft junctions. It is clear however that any such effect had been largely eliminated by gettering.

The difference between the surfaces of hard and soft diodes is of interest. It is difficult to produce quantitative data relating to surface current, as diodes with bad surfaces are notoriously difficult to reproduce or even to measure. Comments must therefore be based on general experience. It has been noted that the harder the diode the smaller the absolute increase in the surface current as the surface conditions are degraded.

The reverse characteristics of hard junctions are therefore characterized by a current proportional to $V^{0.46}$, a temperature coefficient given by $(\partial \ln \theta_s/\partial(1/T) = 0{\cdot}6/k$ in the range from 20 to 70 degrees centigrade, and a relative insensitivity to surface contamination.

7. Conclusion

Crucible and float-zone silicon from a variety of sources has been investigated. No significant differences have been observed in the properties of well-gettered diodes made from these samples. Silicon treated in this way therefore provided a conveniently uniform starting material for other investigations.

The profound effect of apparently minor alterations in surface preparation, diffusion procedure, and cooling rate on lifetime and reverse junction current may well explain the diversity of the results reported by various investigators. The effect of heat treatment on recombination centre density is determined by the nature of the sample surface. Unresolved differences between results of such experiments can be explained by the differing amounts of surface damage initially present.

The extent to which chemical gettering occurs during the production of boron- and phosphorous-diffused structures will depend on many factors. Failure to take this into account in the design of the diffusion schedule can lead to a fluctuating yield of hard devices.

Further work is necessary to identify the elements responsible for the softness of some diodes and to discover the stage at which they enter the silicon. As long as the conventional methods of material preparation and diffusion are used it will probably be impossible to eliminate completely contamination of the silicon. Methods for removal of the contaminants are therefore essential.

It is fortunate that the two methods described herein are available. In the case of shallow-junction devices it may be necessary to adopt the chemical method, when a damaged surface layer cannot be tolerated. It is important to note, however, that whilst the impurities are removed from the system with the boric-oxide-silica glass, they remain in the surface after physical gettering. Long-term changes in lifetime, even at room temperature, reported for example by Gallagher [15] may be associated with release of the impurities from the surface layer. Precautions can be taken to avoid this contingency by removing the surface layer after physical gettering.

The additional restraint imposed on the designer if efficient gettering is to be included as an integral part of the production process, may demand some modification of this process. The gain in device quality and reproducibility would appear to warrant such a change.

8. Acknowledgments

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Defects in Vapour-Grown Silicon

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1. Introduction

A very-important development in device technology is the deposition from a vapour of materials, chiefly silicon, in single-crystal form. The substrate, usually a single crystal of the depositing material, is held at a temperature that may be several hundred degrees below the melting point of the deposit. Diffusion of impurities is considerably reduced at these low temperatures so that, by suitably doping the vapour prior to deposition, *p-n* junctions with predetermined impurity profiles may be grown.

Crystal imperfections play an important role in the mechanical and electrical characteristics of semiconductor materials and devices. In particular, dislocations may produce electrical effects in themselves and perhaps more importantly act as nuclei for the precipitation of foreign atoms that may be present in the parent matrix [1]. Such precipitates may have considerable effects on the properties of *p-n* junctions. Much information has been accumulated on the influence of growth conditions on the number and type of imperfections found in semiconductor crystals grown from the melt. In this paper an investigation into defects in vapour-grown silicon layers is reported. At the temperatures used for growth, the mobilities of silicon and impurity atoms, and of defects are much lower than at the melting point and this may well have an effect on both the type of the defect and its mode of formation. Several methods have been employed to investigate the defects including metallography, X-ray and electron diffraction, and transmission electron microscopy. The defects found in the layers have been studied and methods developed for controlling their numbers.

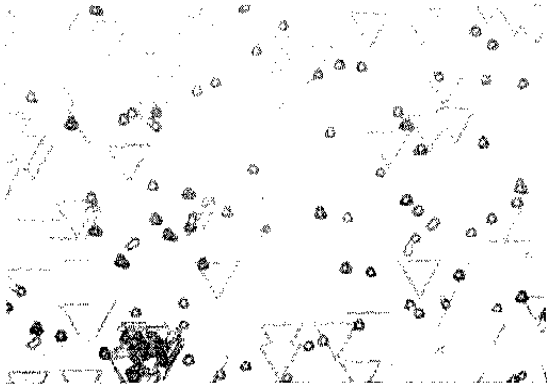


Figure 1—Etched surface of (111) epitaxial layer. (× 250)

2. Preparation of Samples

Most of the work has been carried out on layers grown on mechanically polished substrates. The samples were initially lapped with silicon-carbide abrasives and subsequently polished with alumina powders, the final particle size being 1 micron. The slices were then cleaned ultrasonically in mixtures of hydrofluoric acid and methanol after vapour degreasing.

The layers were grown at rates between 0·5 and 1 micron per minute at 1200 degrees centigrade by the decomposition of either silicon tetrachloride or silane. The hydrogen used as ambient was purified by diffusion through a palladium-silver alloy. The water and oxygen concentrations were below the detection limit of 0·1 part per million.

3. Examination of Layers

3.1 Line Defects

Most samples of epitaxial layers reveal, on etching in either Dash or iodine etch, line etch figures that in (111) surfaces often assume the form of equilateral triangles as may be seen in Figure 1. Normal dislocation etch pits can also be seen. The triangles on a particular sample are all the same size and their sides are parallel to ⟨110⟩ directions. The length of a side of the triangle was found to be about 1·2 times the layer thickness [2]. On (110) layers

the lines meet at angles of 70 degrees, on (100) layers the angles are 90 degrees and on (112) layers isosceles triangles with angles of 44 and 68 degrees are seen. On repeated etching of (111) surfaces, the centre of the triangle becomes progressively smaller until, when the depth of etching is the same as the layer thickness, the triangular shape become a point. On further etching the figures gradually disappear. See Figure 2.

From these facts, it is concluded that the etch figures are due to 2-dimensional defects that lie parallel to (111) planes and are introduced when growth commences. This is very well shown in Figure 3, where a sample has been cleaved to reveal the (111) surface and subsequently etched. The apex of the tetrahedron has not been exactly intersected, and the interface between substrate and deposit is well delineated as a p-n junction.

To investigate the origin of these defects, some annealing experiments were carried out. The samples were heated for various times and temperatures in pure hydrogen before growth was commenced at 1200 degrees centigrade.

As shown in Table 1, at the various growth temperatures, the defect density is seen to fall off, and under our normal growth conditions a 15-minute pre-anneal reduces the defect density below 100 per square centimetre.

TABLE 1

EFFECT OF PRE-ANNEALING ON DEFECT DENSITY

Annealing Time in Minutes	Number of Defects per Square Centimetre			
	Growth Temperature in Degrees Centigrade			
	1210	1150	1050	930
1	6×10^{4}	$2{\cdot}5\times10^{5}$		
2	8×10^{3}	8×10^{4}	2×10^{5}	
3	2×10^{3}			
4		5×10^{3}	3×10^{4}	
6			1×10^{4}	
10				7×10^{5}
15	1×10^{2}			1×10^{5}
30				9×10^{4}
60				8×10^{4}

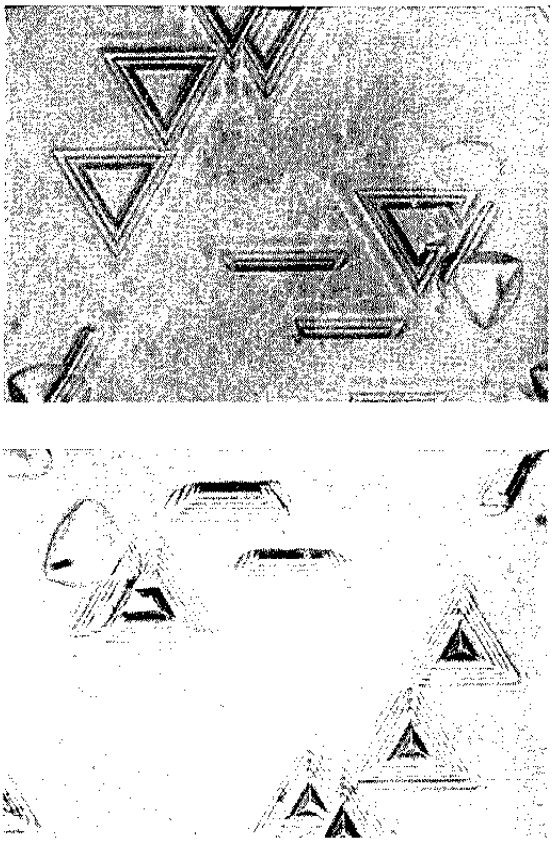


Figure 2—Effect of repeated etching on triangular structure. (× 1000)

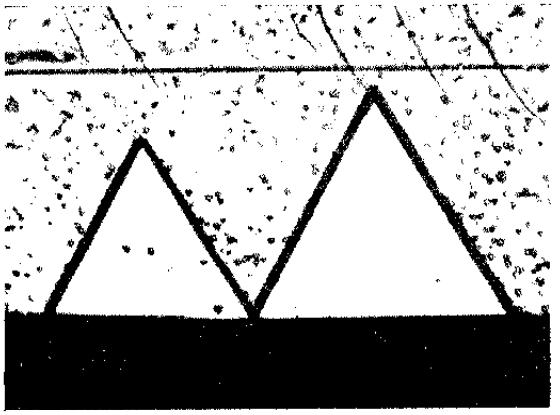


Figure 3—Etched (11$\bar{1}$) surface of epitaxial layer. (× 180)

These results suggest that the defects are caused either by an imperfection in or an impurity on the surface. Annealed samples that were covered with a thin (20-Ångström-unit) oxidized layer and also chemically polished samples, had about 100 defects per square centimetre when grown in conditions that would normally introduce about 10^5 per square centimetre. The activation energy of annealing for the defects, about 35 kilocalories per mole, is much lower than has been found for oxide removal in hydrogen (about 100 kilocalories per mole) [3]. This suggests that the cause of the line defects is physical rather than chemical in mechanically polished layers.

It was observed, however, that when chemically polished substrates were deliberately contaminated, such as by impure water or acetone, so that a visible film was deposited on the substrate, the grown layer showed large numbers of defects. Figure 4 shows the effect of a tap-water stain on a substrate before and after growth, and in Figure 5 the etched surface of 20-micron-thick silicon deposits at the edges of water and acetone stains respectively can be seen. By allowing acetone vapour to condense and re-evaporate from a substrate prior to growth, large numbers of line etch figures are generated, while the more-coarse water deposit gives rise to twinning, dislocations, and line figures.

3.2 Dislocations

It is immediately noticeable that the number of dislocation etch pits in layers grown on clean well-annealed substrates is substantially the same as in the substrate itself. It should be noted that dislocations lying in the (111) plane parallel to the surface, that is, not intersecting the surface, are not propagated and on etching $(11\bar{1})$ surfaces, fewerdisloc ations are

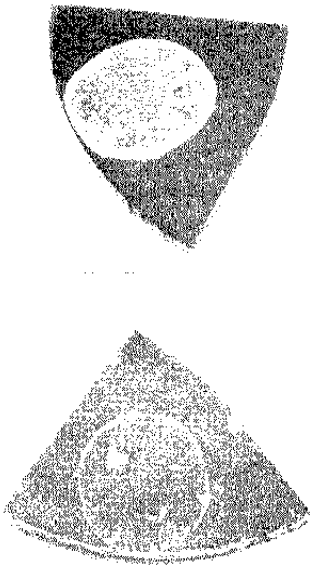


Figure 4—Effect of tap-water stain on substrate when applied before (top) and after (bottom) growth. (× 3)

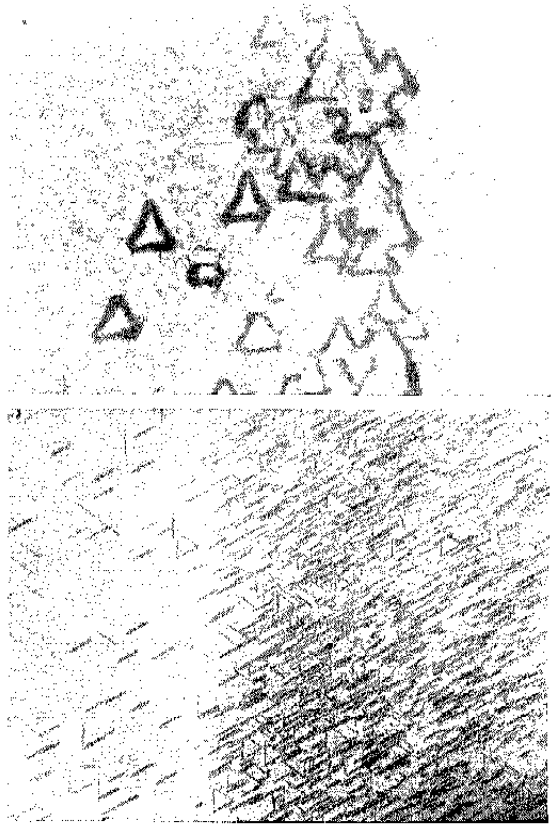


Figure 5—Effect on growth of water (× 500) (top) and acetone (× 330) (bottom).

seen in the layer than in the substrate. In contaminated layers the number of dislocations may be very large as was shown in Figure 5. In insufficiently cleaned substrates occasional bunches of dislocations are observed, which are also probably due to contamination.

Lineage boundaries are easily shown to propagate, and single dislocations can be shown to propagate by the following experiment. A slice was scratched with a diamond and annealed for 15 minutes at 1200 degrees centigrade before growing a layer on it. During the anneal, stresses in the scratch caused dislocations to move into the silicon along the {111} slip planes. These dislocations were propagated and considerable numbers of line defects were also introduced as is evident in Figure 6. In epitaxial layers that have been highly doped with boron or phosphorus during growth, dislocation nets, produced as a result of lattice mismatch and lying in (111) planes, have been observed. This is similar to the effect found after diffusion of these impurities into bulk silicon [4]. They may be observed either by the Lang transmission X-ray technique [5] or by etching as in Figure 7.

3.3 Polycrystallinity and Twinning

Polycrystallinity and twinning can be conveniently shown up by X-ray back reflection, and Figure 8 shows examples of random polycrystallinity, texture, microtwinning, and a twin lamella. These can be produced by unclean growth conditions or by growth at too low a temperature. Polycrystalline regions expose planes other than (111) and these, growing faster, appear as hillocks. Hillocks are also produced if the substrate is etched rather than polished chemically or mechanically, prior to deposition. From the patterns obtained, these imperfections are found to arise at the sites of etch pits, as shown in Figure 9. From etch-pitted regions X-ray photographs similar to Figure 8*C* are obtained. It is also noted that in the immediate etch-pit region, extra concentrations of triangles are found, and this gives a pointer to the reason for the formation of these structures. In thin layers that are smaller than 20 microns, no extra diffraction effects are obtained from regions with triangles, but on thick layers of 300 microns triangular areas of enhanced diffraction within the usual X-ray diffraction spots correlate on a 1:1 basis with the fault structures.

4. Stacking Faults

4.1 Nature of Line Defects

From etching studies, by X-ray, and by selected-area electron diffraction measurements it was found that the orientation inside



Figure 6—Propagation of dislocations. (× 500)

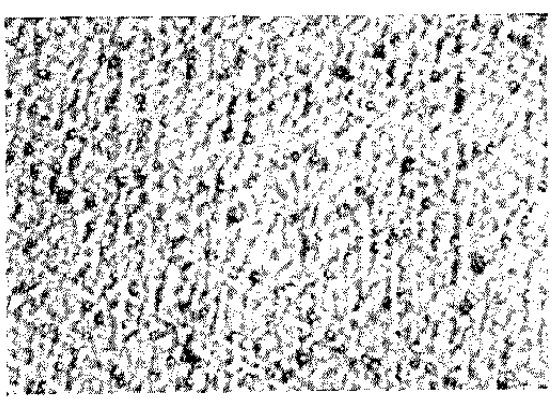


Figure 7—Dislocations in a layer highly doped with boron. (× 1000)

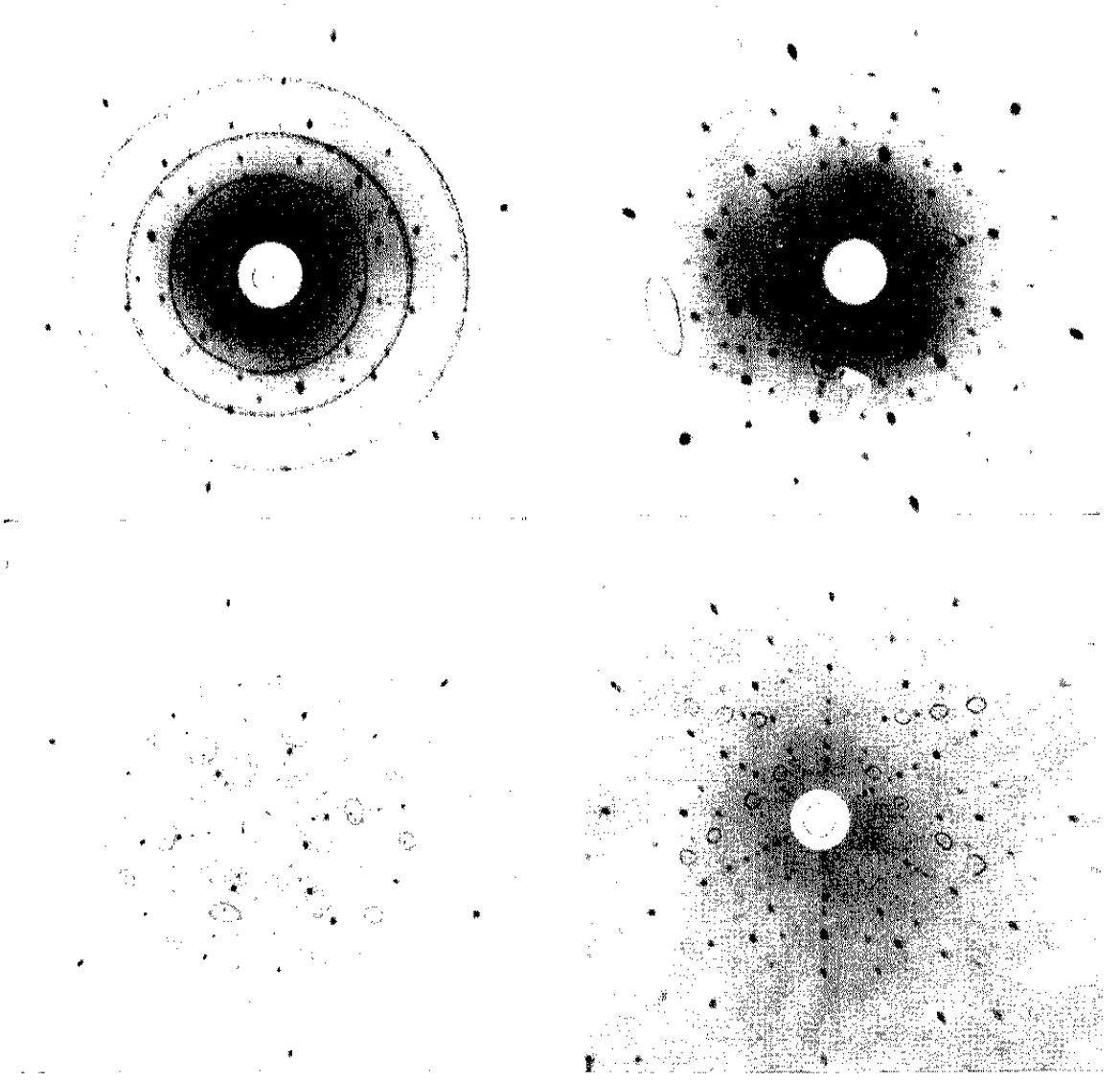


Figure 8—X-ray reflection photographs showing: (top left) polycrystallinity, (top right) texture, (bottom left) microtwinning, and (bottom right) twin lamella.

the triangles on (111) surfaces is the same as that on the outside. This leads to the conclusion that the two-dimensional defects that are parallel to the {111} planes must either be stacking faults or an even number of twins.

To investigate this further some experiments using transmission electron microscopy were carried out [6]. Defects in a crystal lattice produce diffraction contrast and there is a relationship between the Burgers or displacement vector and the Bragg reflection producing the image. Simply no diffraction contrast occurs if the Burgers vector lies in the reflecting plane. In Figure 10 the effect of tilting the specimen is seen. In A, the sides a of the triangles lie in ⟨110⟩ directions; at the ends of the

single faults *b* lines may be seen that lie in ⟨110⟩ directions; at *c* there are some individual lines. In *B* where the specimen has been tilted the diffraction contrast has changed. The contrast due to the faults at *a* has disappeared leaving only a line *b* at the end.

The fringe contrast is shown more clearly in Figure 11. The sample was at such an angle to the beam that faults *a* were not in the diffracting condition. The difference in the width of the various faults is due to changes in thickness of the sample. As in the previous figure there are a number of line defects, some of which are associated with fringes. There is also a line defect at *d* where two faults meet. Selected-area diffraction patterns taken adjacent to and across the faults are exactly similar and show no evidence of extra spots.

From this it is concluded that the defects are stacking faults for the following reasons.

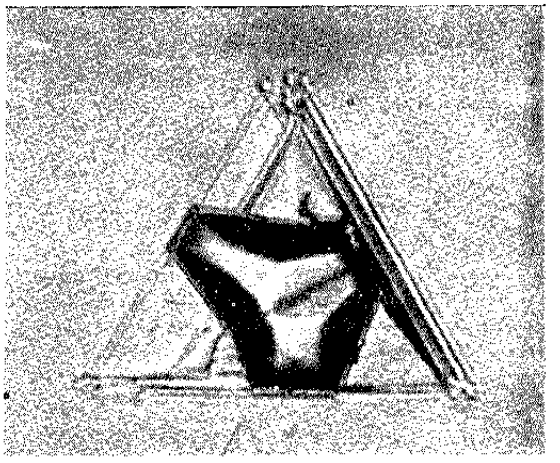


Figure 9—Defects produced at etch pits. (× 1300)

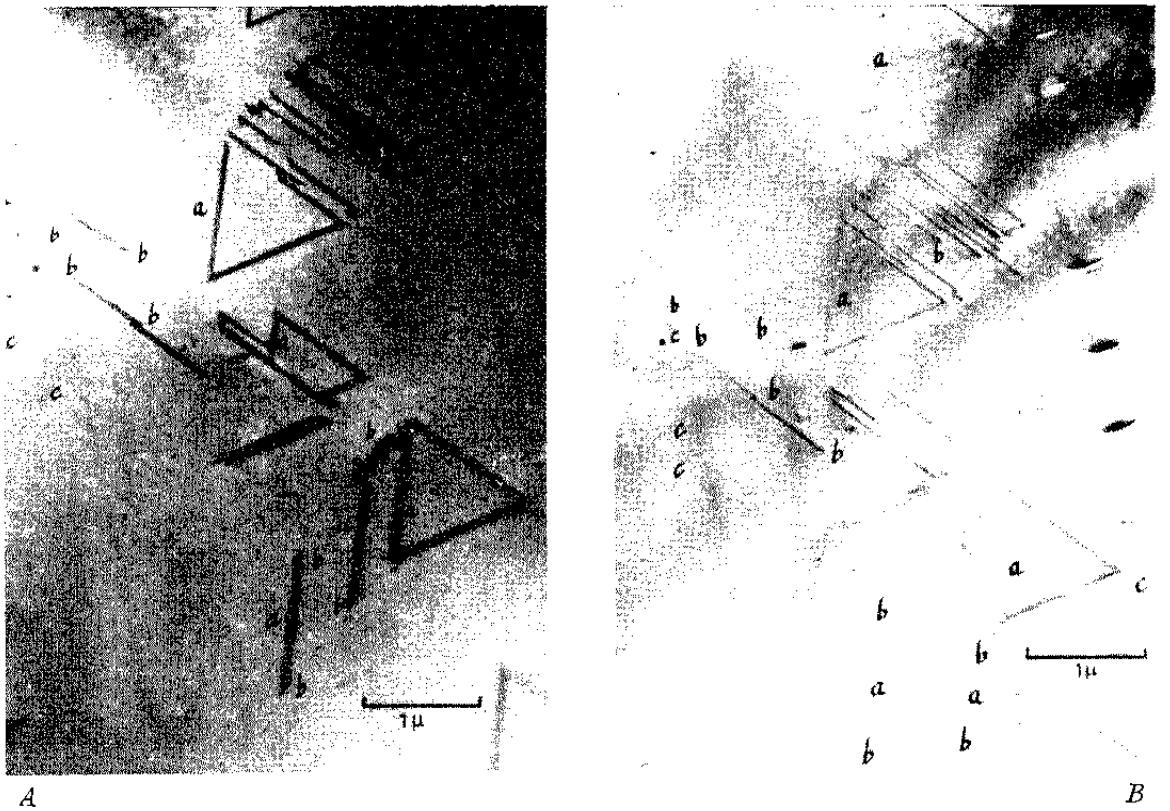


Figure 10—Transmission electron micrograph showing stacking faults and dislocations. (× 17 000)

Figure 11—Transmission electron micrograph showing stacking faults and dislocations. (× 10 000)



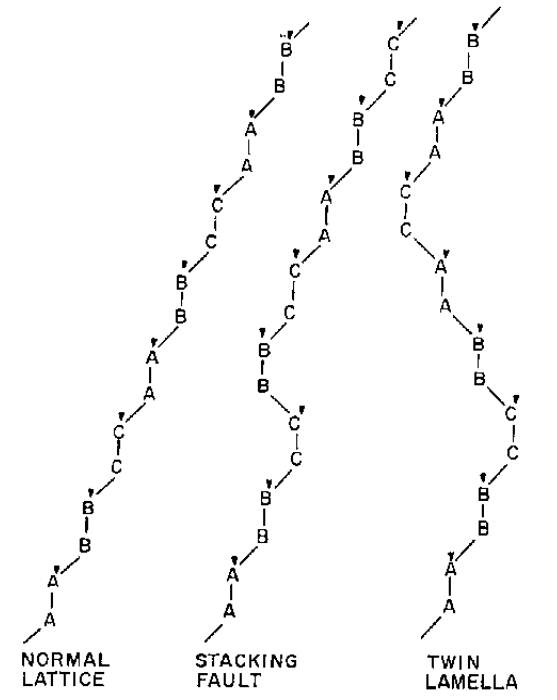


Figure 12—Diagram showing a stacking fault and twin lamella in the diamond structure.

Firstly, stacking faults give rise to diffraction contrast of the type observed. Secondly, single faults end at partial dislocations which are also observed at *b* in Figure 11 and 12. Thirdly, a stair-rod dislocation is produced when two stacking faults on different {111} planes meet and this too has been observed (*d* in Figure 11). Fourthly, multiple twins would produce extra spots on diffraction patterns and these have not been observed. The line contrast at *c* in Figures 10 and 11 is due to individual dislocations.

A stacking fault in the diamond lattice is produced when an extra (111) plane is inserted into or removed from a crystal and is shown schematically in Figure 12. This changes the stacking sequences of the planes from say *abcabcabc* to *abcbcabc*, that is, over the stacking fault region the symmetry is hexagonal. A stacking fault is also a pair of twins on adjacent planes and the boundary has surface energy about twice that of a single twin boundary. A stacking fault is also always associated with partial dislocations. Because of these properties stacking faults are expected to have effects on both dislocations and impurities.

Dislocations may be prevented from moving through the boundary due to the line tension of the latter and they may be anchored by it. The effect on dislocations is clearly shown in Figure 3 where the boundary has prevented dislocations passing into the tetrahedron during growth of the layer. This is analogous to the Dash technique for growing dislocation-free crystals from the melt by having a very thin seed [7]. In the present case the tetrahedron bounded by stacking faults starts from a point and grows outwards producing substantially perfect material. Dislocations may also be pinned by the partial dislocation at the end of a single stacking fault as shown in the transmission electron micrograph of Figure 13.

The stacking fault energy may be lowered by the presence of impurities as is the case in many metal alloy systems. The effect of

stacking faults on copper precipitation may be observed in Figure 14. The copper is preferentially precipitated at the stacking fault and the fault itself is broken up during this process. This type of precipitation in the fault may have a considerable effect on devices made from epitaxial layers.

4.2 Origin of Defects

It has been shown that stacking faults may be introduced into vapour-grown silicon layers and that the great majority are produced at or near the substrate interface. It has also been shown that the number of defects depends on both the physical perfection of the surface and the amount of contamination on the surface. It is suggested that the initial growth commences at specific nuclei and that the origin of a nucleus is either a physical or chemical imperfection. It is probable that these nuclei are slightly misorientated and that, when they meet, a stacking fault is produced. If the misorientation becomes too large other defects may be produced possibly as a result of the strain introduced when the growing nuclei meet. This is particularly the case when etch pits are present for, if the layers grow out from the side of the pit, considerable distortion will be produced when they meet and highly twinned or polycrystalline regions will result. It is interesting to note that polycrystalline regions grow faster than the (111) orientation of the layer as a whole.

Further study of the nucleation process is needed if a more-complete understanding of the mechanism of defect formation is to be obtained.

5. Conclusions

From the reported experiments it is clear that, provided adequate precautions are taken, vapour-grown layers of silicon have defect concentrations similar to the substrate material. To produce these highly perfect layers it is necessary to clean the substrate effectively, to anneal before growth for a sufficiently long time, and to have sufficiently pure conditions during growth. Growth mechanism has not been studied but it is clear from the highly perfect silicon found in the tetrahedra enclosed

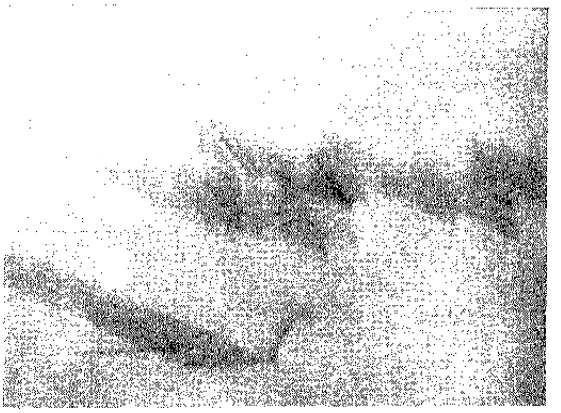


Figure 13—Transmission electron micrograph showing pinning of a dislocation. (× 45 000)

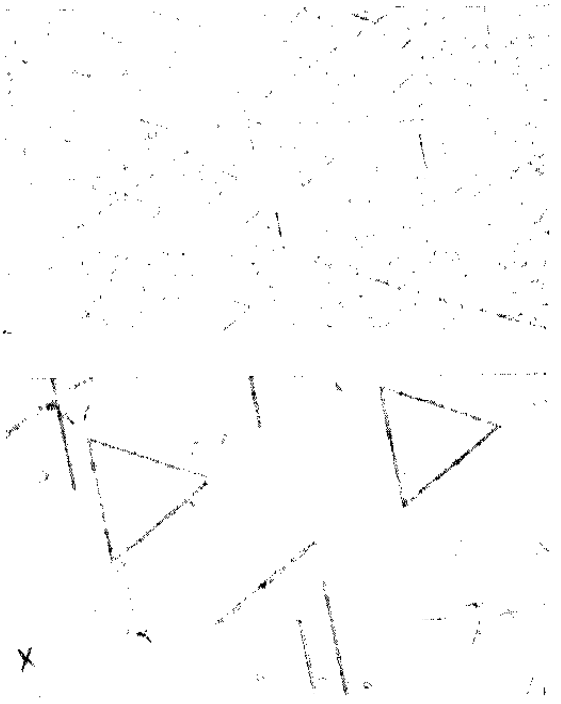


Figure 14—Preferential copper precipitation at stacking faults. Top print is × 300 and bottom is × 900.

by stacking faults that dislocations are not important. It is believed that the stacking faults found in the layers are primarily produced near the start of growth particularly when growing nuclei meet. There appears to be no good reason why layers free from both dislocations and stacking faults cannot be produced. The properties of vapour-grown silicon appear to be substantially the same as melt-grown material; for example, diffusion constants [8] and Hall mobilities [9] are similar.

6. References

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7. Acknowledgment

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In 1954, he joined the materials division at Standard Telecommunication Laboratories where he was concerned with X-ray and metallographic studies, particularly concerning the crystal perfection of semiconductors. Dr. Batsford is presently engaged in studying material parameters using X-ray, electron-microscope, low-temperature, and infra-red techniques.

D. J. D. Thomas was born in London on 2 June 1927. He obtained a degree in metallurgy at the University College of South Wales and Monmouthshire in 1946.

In 1949, after working in the tin-plate industry, he joined Standard Telecommunication Laboratories, where he is currently engaged with problems connected with semiconductor materials, such as, the purification and growth of germanium crystals, relationship between crystal defects and growth conditions in germanium and silicon, the behaviour of impurities in semiconductors, and defects in vapour-grown materials.

Silicon Epitaxial Planar Transistors Part 1—Appraisal

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Three major technologies are competing today for producing very-high-frequency transistors, which are described by the following designations:—

(A) Micro alloy diffused transistors (MADT)

(B) Post alloy diffused transistors (PADT)

(C) Diffused transistors.

(A) and (B) are available only in germanium, but the diffused transistor is available also in silicon. In silicon, the marvellous property of silicon dioxide to provide a selective mask against the diffusion of some elements, allows double-diffusion techniques to be used with localized emitter structures, thus giving low input capacitance. A cross-section of a diffused silicon transistor is shown in Figure 1. In germanium, this oxide property is not available, and the transistors have diffused bases but alloyed emitters.

Diffused mesa transistors have been available for many years and have represented a significant advance over preceding technologies. However, they had several shortcomings and various manufacturing difficulties.

1. Diffused Mesa Transistors

Table 1 summarizes the limiting characteristics of representative diffused transistors that were commercially available in January 1961. In considering these figures it must be appreciated that not all these characteristics were available simultaneously in a single transistor. It is unfortunately true that even now it is not possible to make a 150-volt 10-ampere 750-megacycle-per-second transistor with an input capacitance of 2 picofarads. This table is meant more as a guide to the circuit designer to help in deciding whether suitable transistors are available for his particular application. For instance, if the line voltage in an equipment is 50 volts, the designer will be obliged to use a silicon diffused transistor. Alternatively, the circuit designer may not have realized that current ratings up to 10 amperes in silicon were available; and so on. It will be noticed that in voltage, current, power dissipation, and leakage current, silicon offered a greater flexibility than germanium, whereas in frequency there was little to choose between them. Advances have since been made in the frequency capability of both silicon and germanium, and transistors that can operate at 1 gigacycle per second are available in silicon and at 2 to 3 gigacycles per second in germanium.

TABLE 1

LIMITATIONS OF DIFFUSED MESA TRANSISTORS AS OF JANUARY 1961

Characteristics	Measurement Unit	Diffused Silicon	Diffused Germanium
V_{CB0}	Volts	150	40
V_{EB0}	Volts	12	4
I_C, Mean Maximum	Amperes	10	0·5
Power Dissipation	Watts	125	6
f_{hb}	Megacycles per Second	750	750
C_{ob}	Picofarads	2·0 – 3·5	1·2–1·5
I_{CB0}	Microamperes	0·01–0·1	1

It will be noticed that in Table 1 no values are listed for saturation voltage V_{CES}, which depends greatly on the current levels involved. To quote a typical figure, a circuit required to switch 0·5 ampere would involve a bottoming

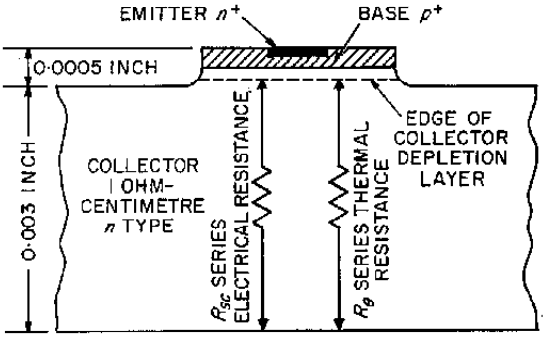
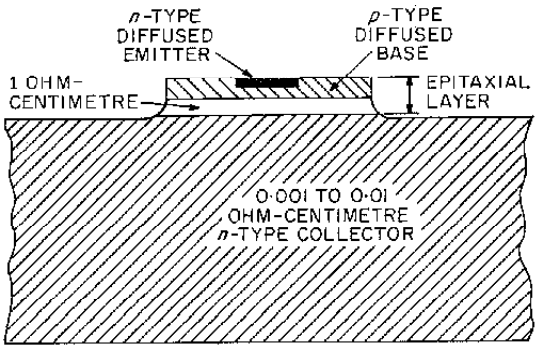


Figure 1—Double-diffused silicon mesa transistor.

Figure 2—Cross-section of an epitaxial double-diffused silicon transistor.



voltage in the range of 1·5 to 3·5 volts. Similarly, switching times are not given as they depend greatly both on the current levels involved and on the actual circuit used. A given transistor may produce quite different storage times in two different circuits. It is unfortunate that it is not possible to present realistic figures for these two characteristics of diffused transistors, because undoubtedly they represent the two major shortcomings of normal diffused mesa transistors. There are three other minor shortcomings of diffused transistors; firstly, they offer disappointing noise figures; secondly, they have low current gain at low currents; and, finally, they often show a spread in characteristics, particularly in output capacitance and in ON voltage (V_{BES}). The two major shortcomings have been overcome by the epitaxial technology and gold diffusion, and the other three by the planar technology which, in so doing, has introduced many other advantages, not least of which is its very high order of reliability.

2. Epitaxial Transistors

The reason for the high saturation voltage can be seen in Figure 1. This shows a cross-section of a double-diffused silicon mesa transistor, but the argument applies equally well to the germanium mesa transistor. The base layer in a diffused transistor may be some 4 or 5 microns below the surface, and at maximum applied collector voltage the depletion layer may extend into the collector region a further 4 or 5 microns. Thus, the operating part of a transistor is in fact only about 9 or 10 microns thick; that is, less than 0·0005 inch. Unfortunately, silicon being very brittle, it is not possible to handle large pieces of silicon of this thickness in mass production. Indeed, for mass production of diffused transistors, the minimum practical thickness of semiconductor is 0·003 inch (76 microns). To obtain the desirable characteristics of high collector breakdown voltage and low output capacitance, a rather high resistivity (1 to 6 ohm-centimetres) is required at the collector-to-base junction. The unwanted thickness of silicon introduces a series electric resistance and a series thermal resistance, both of which degrade the inherent properties of the mesa transistor. This series electric resistance is the source of the high saturation voltage of normal diffused mesa transistors.

The epitaxial transistor has provided a brilliant solution to this problem. An epitaxial layer is one that has the same crystalline orientation as the substrate on which it is grown. If silicon is exposed at 1200 degrees centigrade to hydrogen containing 1 percent of silicon tetrachloride, the tetrachloride will decompose and silicon will be deposited on the silicon substrate. At these temperatures the silicon atoms condensing on the surface of the substrate are quite mobile and are able to take up the orientation of the substrate lattice. By adding minute amounts of phosphorus trichloride or boron tribromide to the silicon tetrachloride it is possible to control the resistivity of the deposited layer.

For the epitaxial transistor a high-resistivity (1-to-6-ohm-centimetre) layer of silicon about 12 microns thick is deposited epitaxially on a low-resistivity (0·001-to-0·01-ohm-centimetre) piece of silicon some 0·005 inch (127 microns) thick. A transistor is then fabricated in the normal way by oxide masking and double-diffusion techniques, as shown in Figure 2.

The advantages of the epitaxial transistor are

(A) Low saturation resistance

(B) Saturation resistance independent of temperature

(C) Increased linearity of characteristics

(D) Reduced minority-carrier storage time.

The improved saturation characteristics are apparent from Figure 3, which shows the $I_C - V_{CE}$ characteristics of two 100-megacycle-per-second silicon power transistors, identical except that one used epitaxial material.

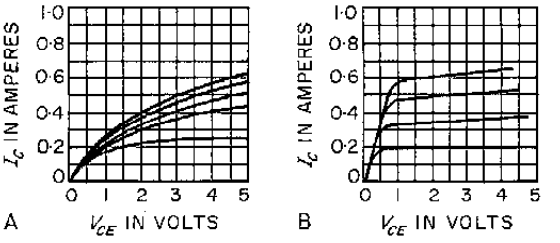
The implications of the lower saturation voltages will be obvious to circuit engineers. In switching circuits greater efficiencies are achievable, leading to lower dissipations; or for a given transistor design, greater power-handling capacity. In amplifier and oscillator applications, greater voltage swings are possible, giving increased output powers to the load. An amplifier using two non-epitaxial *BLY10* transistors in push-pull from an 18-volt supply gave 4 watts output power at 5 megacycles per second. By merely substituting epitaxial transistors, the output power was increased to 9 watts at 55-per-cent efficiency. By redesigning the circuit slightly to take advantage of the lower saturation resistance, an output power of 10 watts was obtained at 5 megacycles per second.

The second advantage of epitaxial transistors shown above derives from the fact that the collector series component of total saturation resistance has been virtually eliminated by the epitaxial construction. The actual degree of improvement depends very much on the resistivity of the collector material used in the non-epitaxial transistor; but one typical comparison shows an increase in saturation resistance between 25 and 150 degrees centigrade of 2·15 times for a *BLY10* non-epitaxial unit and only 1·15 times for a *BUY10* epitaxial transistor.

Thus, on the epitaxial transistor, the saturation resistance increased only 15 percent from room temperature to 150 degrees centigrade.

The improved linearity (and higher gain) is also obvious from Figure 3. Thus, with epitaxial transistors lower distortion levels can be obtained, which make them attractive for amplitude-modulated transmitter applications.

Figure 3—Typical characteristic curves for two silicon power transistors, that at the right being for a *BUY11* epitaxial design and that at the left is for a *BLY11* non-epitaxial unit.



Finally, the improvement in switching performance offered by epitaxial transistors is due to the location of a low-resistivity layer close to the collector-to-base junction. In saturated switching circuits the collector is forward-biased in the ON condition, and at turn-off the minority carriers that have been injected from the base into the collector have to be removed before the transistor will switch off. Circuit conditions limit the rate at which these carriers can be removed, and the turn-off time is dependent to a large extent on the lifetime of the minority carriers in this region. The presence of a very-low-resistivity region near the collector junction limits the number and lifetime of the carriers that are injected, thus reducing the storage time in switching circuits.

In a specific circuit, a low-level switching transistor gave an average storage time t_s of 550 nanoseconds. The epitaxial version of this transistor gave a t_s value of 100 nanoseconds.

3. Gold-Doped Transistors

To switching-circuit engineers, the storage times quoted above will not seem very attractive.

There is, fortunately, another technology available with the diffused transistor. This is gold doping, by which it is possible to control

the lifetime of minority carriers. Gold is diffused into the base and collector regions of the transistor; this reduces the storage time in switching circuits. For example, in the low-level epitaxial transistor quoted above, the storage time of 100 nanoseconds was reduced to 10 nanoseconds and less. This speed should be acceptable for most switching circuits.

4. Silicon Planar Transistors

In the mesa transistor, the area of the collector-to-base junction is defined by etching a mesa. The planar transistor is made by countersinking the base into the silicon, using photolithographic techniques to establish a suitable oxide mask. The reason for the high reliability and the other attractive properties of the silicon planar transistor will be better understood by considering Figure 4, which shows the most significant steps in the fabrication of the planar transistor. It should be appreciated that the transistor shown in this figure is just one of up to 1200 similar transistors formed on a single slice of silicon about 1 inch (25 millimetres) in diameter.

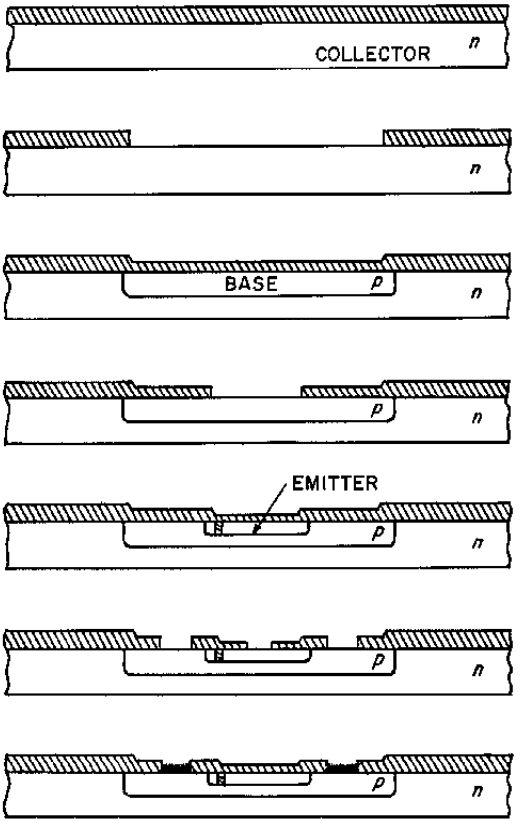


Figure 4—Steps in fabricating planar transistors.

The first step is to grow an oxide layer of 0·7-micron thickness on the silicon surface under scrupulously clean conditions. The area of the base is then defined by a photolithographic process, and a window etched in the silicon dioxide. This process is very precise and repeatable. The slice is now exposed at high temperatures to a boron-rich atmosphere. Silicon dioxide masks the diffusion of boron and, hence, a *p*-type layer is formed only where the window was etched in the silicon dioxide. Fortunately, boron diffuses isotropically in silicon, so that lateral diffusion of boron takes place as well as diffusion normal to the surface. This might not appear so from Figure 4, which is not drawn to scale in order to show all the information in one figure. Thus, the junction is not formed opposite the edge of the layers, but 4 or 5 microns *underneath* the oxide layer. This process gives control of the surface states at the *p-n* junction, and it is from this control that all the advantages of the planar transistor derive. Once a *p-n* junction is formed, an electric field exists across it, and if this junction is exposed it will attract ionic contaminants to it, thus degrading its properties. In the planar transistor the sealed junction prevents this from happening, and the properties of the junction, once made, remain unchanged throughout the fabrication and life of the transistor. During boron diffusion more oxide is grown on the surface, so that we are able to repeat the photolithography and diffusion processes with phosphorus to produce an *n*-type emitter. Note that the emitter junctions are also underneath the oxide layer. Finally, during phosphorus diffusion more oxide is grown to give a completely sealed transistor. Windows are etched in appropriate places in this oxide to allow contacts to be made to the operative areas of the transistor.

This transistor is now complete; it has its final characteristics and requires no etching or other treatment to improve its characteristics.

A 400-megacycle-per-second silicon planar transistor is shown in Figure 5. Its diameter is 0·007 inch (175 microns).

The advantages of planar transistors are summarized below:—

(A) Very-low leakage currents

(B) High gain at low currents

(C) Low noise figure

(D) Very-consistent characteristics

(E) Very-stable characteristics

(F) Very-high order of reliability.

As a result of the sealed junctions, the leakage currents in silicon planar transistors are very low indeed. On a 10-watt power transistor, which is rated at 1·5 amperes peak current, the typical leakage currents are in the range 1 to 3 nanoamperes. Low-level devices give typical leakage currents in the range 0·1 to 1 nanoampere. This makes the planar transistor unquestionably the most suitable for direct-current amplifiers.

Due to surface effects, mesa transistors have very-low gains at currents in the microampere range. At 1 microampere, the small-signal gain might be 0·3 to 3. With planar transistors, gains as high as 100 at 1 microampere have been achieved, and transistors are sold with minimum gain of 60 at 100 microamperes. This feature of planar transistors could be important in compact designs where power densities must be minimized.

Planar transistors offer considerably lower noise figures than equivalent mesa transistors; this results from the control of the surface states of the p-n junctions. Low-frequency noise figures in the range 3 to 10 decibels at a few hundred microamperes are typical. The low-frequency noise figure is current-dependent and increases with increasing collector current beyond 1 milliampere. An interesting experiment is to measure the noise on a planar transistor, and then remove the oxide with a weak acid. The noise figure increases by 8 to 12 decibels. The low noise figure is an advantage in input stages of direct-current amplifiers, as well as in many other circuits. At 1 megacycle per second, noise figures of 1 to 2 decibels are available with planar transistors.

The greater consistency of characteristics offered by planar transistors comes from two sources. Firstly, the base area, which defines the output capacitance, is determined with almost mathematical precision by a photolithographic process. Secondly, and by far the more important, is the fact that there is no final etching required with this transistor. Before encapsulation of any exposed-junction transistor, final etching is necessary to remove the ionic contaminants that have been attracted by the electric field during the

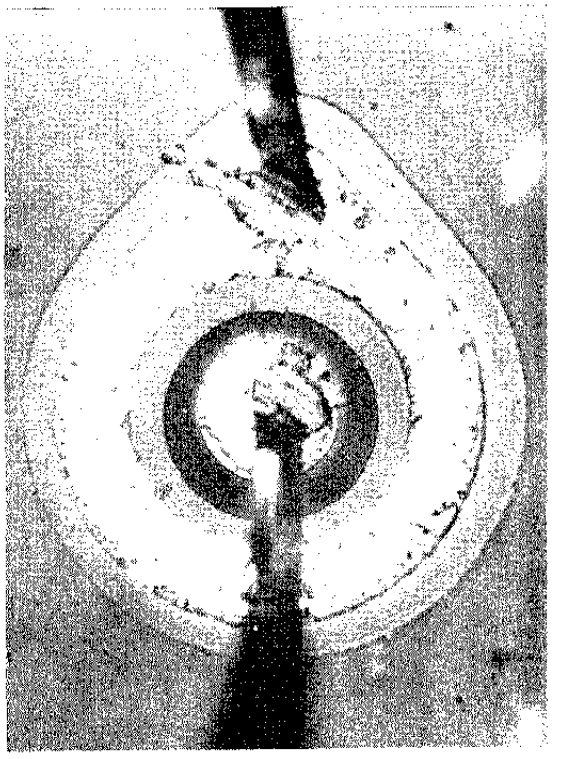


Figure 5—A 400-megacycle-per-second silicon planar transistor.

fabrication process. This restores the original characteristics. Variable amounts of such etching can affect the area of the mesa and, hence, the output capacitance. An even-more-serious effect of this etching is on the ON voltage or V_{BES}. The V_{BES} is determined mainly by the extrinsic base resistance which, in a diffused transistor, depends on the resistance between the aluminium base contacts and the operative area of the base. Base current flows through the path of least resistance along the surface to the edge of the emitter and thence to the operative part of the base. The surface areas are of very-low resistivity due to the diffusion profile, and when the collector-base junction is etched, the surface layers are attacked very rapidly; this can result in a considerable spread in extrinsic base resistance and, hence, in the ON voltage.

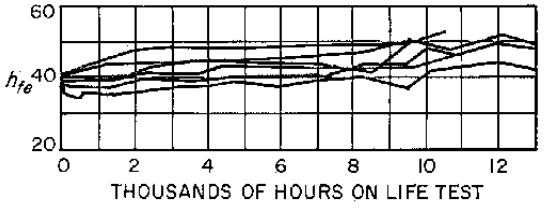


Figure 6—Values of h_{fe} for 6 *BLY11* silicon planar transistors measured at 1 kilocycle per second with $V_{CE} = +9$ volts, $I_C = +20$ milliamperes, and 3 watts dissipation.

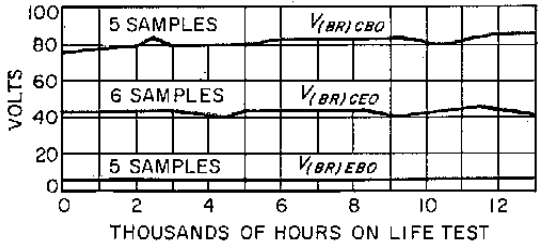


Figure 7—Type *BLY11* silicon planar transistors life tested at 3 watts dissipation. $V_{(BR)CB0}$ was measured with $I_C = +1$ milliampere, $V_{(BR)CE0}$ with $I_C = +2$ milliamperes and $I_B = 0$, and $V_{(BR)EB0}$ with $I_C = 0$ and $I_E = +1$ milliampere.

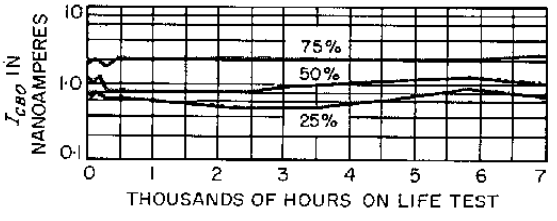


Figure 8—Unencapsulated silicon planar transistors, type *BFY15*, stored at 175 degrees centigrade and measured at $V_{CB} = +9$ volts and $I_E = 0$. The percentage of the 24 samples on test having measured values not exceeding those of the curve is indicated on each curve.

The planar transistor offers a high order of characteristic stability due to the control of the surface states at the emitter-base and collector-base junctions. Figures 6 and 7 show stability results on some very early experimental *BLY11* silicon planar transistors; h_{fe} characteristics may leave something to be desired, but it will be seen from later results that this performance has been considerably improved. These transistors have been dissipating 3 watts in free air throughout their operating life. A sample of 6 units showed an unvarying maximum value of I_{CB0} of 0·003 microampere over a life test of 13 000 hours with V_{CB} of 9 volts, I_E of zero, and dissipation of 3 watts.

The very-high order of reliability offered by planar transistors is not dependent on the quality of their encapsulation or the control of the ambient during life, but is inherent in the design and fabrication of the transistor. Firstly, the key areas of the transistor, the *p-n* junctions, are sealed beneath an impervious layer of silicon dioxide. That this layer is impervious can be illustrated by the fact that unencapsulated planar transistors can be operated under water without any change in their gain or breakdown voltage characteristics! Secondly, a high order of mechanical reliability is introduced by the absence of etching during the fabrication process. Very strong acids are necessary to etch silicon junctions so that while clean-up etching is in progress in mesa transistors, chemical attack also proceeds on the metal parts. In particular, the gold-alumin-

ium-silicon bonds of the transistor can be severely attacked and weakened. Planar transistors are sufficiently new so that reliability results are of limited duration in terms of total operational hours. As further evidence of this reliability, however, storage on these transistors for 7000 hours at room temperature and at 150 degrees centigrade have produced no changes in their characteristics. Storage life tests at 175 degrees centigrade have passed 8000 hours to date without any appreciable change in gain, breakdown, or leakage characteristics.

A series of experiments has established that the reliability of the transistor is inherent in the device and not dependent on its encapsulation. Unencapsulated transistors have been stored at room temperature and at 115, 135, and 175 degrees centigrade. Storage at room temperature has reached some 10 000 hours without any apparent deterioration in gain, breakdown voltage, or leakage current. Storage tests under the other three temperatures have now reached 7000 hours and similarly show no apparent changes in characteristics. Figures 8, 9, and 10 show the results at 175 degrees centigrade, which are typical of all temperatures. We have also subjected unencapsulated transistors to a range of severe environmental tests that include humidity cycling, salt-fog tests, and boiling in deionized water for prolonged periods. Unencapsulated transistors have successfully withstood 100 cycles of the standard United Kingdom Government humidity cycling test. Rust from the metal parts of the mounting completely covers the transistor and causes an increase in the total leakage current; but the breakdown voltage and gain remain unaffected. Similar results are achieved with the salt-fog test and the boiling-water test. The last two tests have been applied for over 500 hours without producing any change in the gain or breakdown voltage. Figures 11, 12, and 13 show the results on the salt-fog test; these results are typical of the others. After about 550 hours the metal parts of the transistor including the leads have also rusted away, and it is this fact that brings the test to an end, *rather than any deterioration in the transistor*. To prove that the increased leakage currents were due only

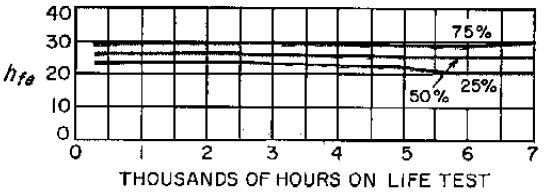


Figure 9—Unencapsulated *BFY15* transistors stored at 175 degrees centigrade and measured at $V_{CE} = +9$ volts and $I_C = +20$ milliamperes. Percentage of the 24 samples not exceeding the values of each curve are indicated.

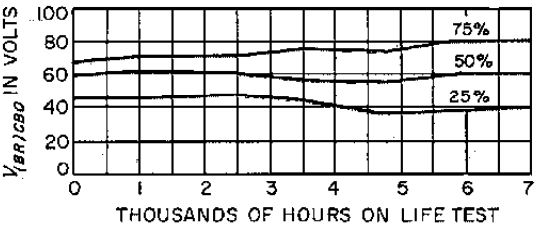


Figure 10—Unencapsulated *BFY15* transistors stored at 175 degrees centigrade and measured at $I_C = +10$ microamperes and $I_E = 0$. Percentage of the 25 samples not exceeding the values of each curve are indicated.

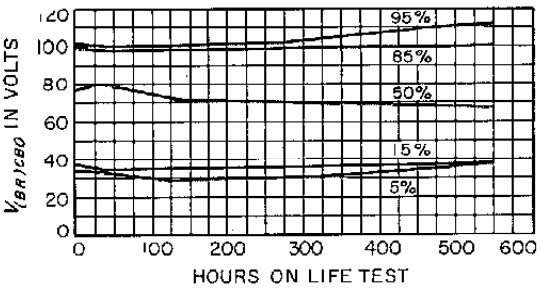


Figure 11—Unencapsulated silicon planar transistors, type *BFY16*, subjected to salt spray at room temperature. $V_{(BR)BC0}$ was measured at $I_C = +10$ microamperes and $I_E = 0$. Percentage of 24 samples having measured values not exceeding those of the curves is indicated on each curve.

to the parallel path of the rust, the transistors were dipped in a 50-percent hydrochloric-acid solution that removed the rust and returned the leakage currents to their initial values. This can be seen clearly in Figure 13.

Similarly, unencapsulated transistors have been boiled in saturated brine solutions for several hours without any deleterious effects.

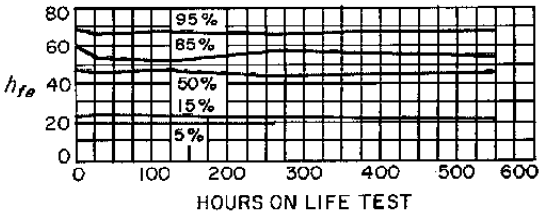


Figure 12—Unencapsulated *BFY16* transistors subjected to salt spray at room temperature. Measurements of h_{fe} were made at $V_{CE} = +9$ volts and $I_C = +20$ milliamperes.

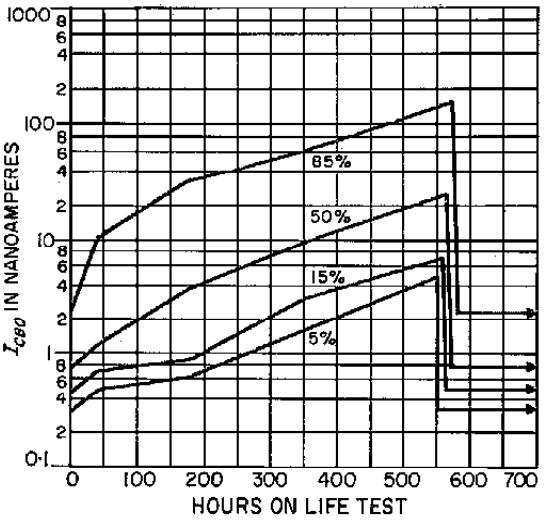


Figure 13—Unencapsulated *BFY16* transistors subjected to salt spray at room temperature. Measurements of I_{CB0} were made at $V_{CB} = +9$ volts and $I_E = +0$. At 550 hours, the corrosion products, which included the leads, were removed in hydrochloric acid and the initial values reappeared indicating that the deterioration was a superficial effect of corrosion.

Thermal-shock tests have also been carried out on these transistors to ensure that no troubles could arise from possible differences in expansion coefficients between the silicon and the silicon dioxide layer. Transistors have been stabilized at 150 degrees centigrade and then plunged into liquids at −70 degrees centigrade. This process has been repeated many times without any apparent effect on the transistor characteristics.

Finally, a whole series of sensitive ambient tests has been conducted on these transistors to determine any trace of surface instability. No such instability was found.

5. Silicon Epitaxial Planar Transistors

The planar design can, of course, be combined with the epitaxial processing and the gold-doping technology to produce the silicon epitaxial planar transistor. This is probably the most versatile and reliable transistor presently available. It may be thought that such reliable and versatile transistors would of necessity be expensive. This is completely invalid, as diffused transistors are made by production techniques that render them inherently cheap. Firstly, all the diffusion and photolithographic processes are performed simultaneously on up to 1200 transistors. Secondly, with the planar process it is possible to test the transistors at every stage in manufacture so that a poor slice can be thrown away before large costs have been involved. Finally, each individual transistor die can be fully tested before assembly, thus ensuring high assembly yields. Thus, silicon epitaxial planar transistors have many advantages from a manufacturing point of view, as well as from the technical aspect. It is confidently predicted that their prices will reach the level of germanium alloy transistors in due course.

6. Conclusions

The epitaxial planar technology is a very versatile one, which can be used over a wide range

of frequency, current, and power levels. Gigacycle-per-second amplifiers and 20-watt 10-ampere devices can both be made by planar techniques. It is felt that silicon planar epitaxial transistors can fulfil a large percentage of all transistor applications, with an order of reliability that has not previously been achieved.

This technology is now also well established in solid-state circuits, where its flexibility and reliability have led to significant advances.

Bernard Douglas Mills was born in Charters Towers, Queensland, Australia, in 1928. He received the degree of Bachelor of Electrical Engineering in 1950 and Bachelor of Science in 1951 from Queensland University in Brisbane.

After 18 months in the power industry, he served as a development engineer on radio receiving valves.

He joined Standard Telephones and Cables in 1957, and was engaged in work on semiconductor devices. He is now chief engineer of the Transistor Division.

Silicon Epitaxial Planar Transistors
Part 2—Characteristics

J. BICKLEY
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The range of silicon planar transistors manufactured by Standard Telephones and Cables contains more than 15 codes, and it must first be made clear why this number has been brought into existence, and why there is not a universal transistor.

Any such device represents a compromise between cost and performance over the wide range of applications for which transistors are used. Although a low-cost transistor can be made to give good performance over a wide range of applications, for example, the *BSY27*, better performance over a narrower range of applications, probably at a higher cost, can always be obtained. Consequently, although there are general-purpose transistors, there is never likely to be a universal transistor.

The main characteristics of the range of silicon planar transistors are given in Table 1.

1. Direct-Current

Figure 1 shows the variation of large-signal current gain h_{FE} with collector current I_C for representative specimens from each family.

The highest gain device is the *BFY19*, the 400-megacycle-per-second non-epitaxial amplifier. However, the gain falls sharply at collector currents beyond 30 milliamperes. The epitaxial version *BSY29*, on the other hand, has a lower maximum gain, but maintains its gain at a useful level to 200 milliamperes. It has the widest range of useful gain of all these devices.

The *BSY27* has a peak gain similar to the *BSY29* but, being of larger area, the peak occurs at a higher value of collector current. The non-epitaxial version of the *BSY27*, the *BFY18*, peaks at the same collector current, but as in the case of the *BFY19*, the gain falls sharply at higher currents due to the collector series resistance.

Figure 1 also shows the excellent gains of the large-area *BUY11* and *BSY25* at collector currents up to the peak rating of 1·50 amperes.

TABLE 1
CHARACTERISTICS OF CODED SILICON TRANSISTORS

Family	Type	Encapsulation	Dissipation in Watts	Typical f_T in Megacycles per Second	Typical h_{FE} or h_{fe}	Structure	Typical Application
A	*BLY10*	TO-3	10 With	100	30	Planar	High-Power High-Frequency Oscillators and Amplifiers
	BLY11	TO-3	Heat Sink	150	60		
	BFY15	TO-5	0·6 in Free Air	100	30	Planar	Medium-High-Frequency Power Oscillators and Amplifiers
	BFY16	TO-5		150	60		
	BUY10	TO-3	10 With	100	30	Planar Epitaxial	High-Level Switches, High-Efficiency Power Oscillators and Amplifiers
	BUY11	TO-3	Heat Sink	150	60		
	BSY24	TO-5	0·6 in Free Air	100	30	Planar Epitaxial	High-Current Switches, Medium-Power Oscillators and Amplifiers
	BSY25	TO-5		150	60		
B	*BFY17*	TO-5	0·6 in Free Air	300	60	Planar	High-Frequency Amplifiers
	BFY18	TO-18	0·3 in Free Air	300	60		
	BSY26	TO-18	0·3 in Free Air	300	30	Planar Epitaxial	General-Purpose Fast Switches and Amplifiers
	BSY27	TO-18		300	60		
C	*BFY19*	TO-18	0·3 in Free Air	400	100	Planar	Very-High-Frequency Amplifiers
	BSY28	TO-18	0·3 in Free Air	400	30	Planar Epitaxial	Very-Fast Switches
	BSY29	TO-18		400	60		

The curves for the non-epitaxial versions of these devices have not been shown on the diagram to avoid confusion, but their drop in gain occurs at currents beyond 0·50 ampere.

Both large- and small-signal current gain rise with temperature throughout the range from −55 to +175 degrees centigrade. The rate of rise varies somewhat for different types of device, a temperature rise between 100 and 150 degrees centigrade being needed to double the gain.

The advantages of epitaxy are shown most markedly in the variation of V_{CES} with collector current. This is shown in Figure 2 for a typical sample of each family. These curves are for a ratio of I_c/I_B of 10. At low currents V_{CES} varies only slightly, but beyond a certain collector current, which is dependent on the area of the active element and its collector structure and resistivity, the V_{CES} rises as a function of current. It is interesting to note that for the three non-epitaxial types, V_{CES} rises linearly with collector current; whereas for the epitaxial devices V_{CES} rises approximately with $I_C^{\frac{1}{2}}$. This difference is probably due to conductivity modulation of the epitaxial layer in the collector region, stemming from the much-higher current density for the collector currents at which V_{CES} begins to rise with current.

Figure 3 shows the variation of V_{BE} and V_{CES} with temperature. V_{BE} falls at 1·85 millivolts per degree centigrade for both epitaxial and non-epitaxial devices, both absolutely and as a percentage of the room-temperature value.

2. Alternating-Current

The low-frequency gain h_{fe} reaches a maximum at a value of I_C determined roughly by the area of the device in question. In the case of the non-epitaxial devices, for instance, the maximum occurs at 10 milliamperes in the case of the *BFY17* and *18*, and 5 milliamperes for the *BFY19*, compared with 80 milliamperes for the power transistors *BLY10* and *11*. For the smaller-area devices the gain remains at a

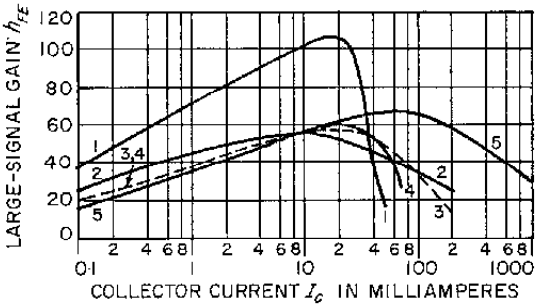


Figure 1—Typical variations of large-signal current gain h_{FE} as a function of collector current I_C. The type of transistor and the measurement V_{CE} (in volts) are as follows: curve 1, *BFY19* (6); curve 2, *BSY29* (2); curve 3, *BSY27* (2); curve 4, *BFY17/BFY18* (6); and curve 5, *BUY11/BSY25* (2).

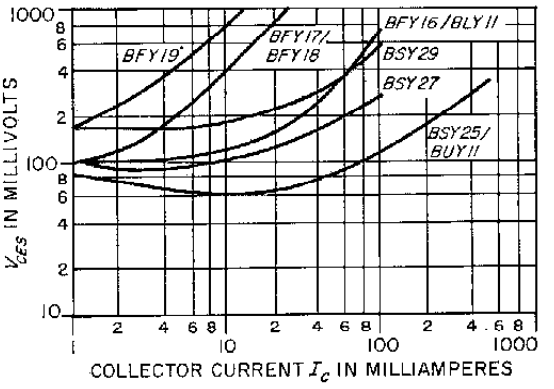


Figure 2—Typical variations of the collector-emitter saturation voltage V_{CES} with collector current I_C for $I_C/I_B = 10$.

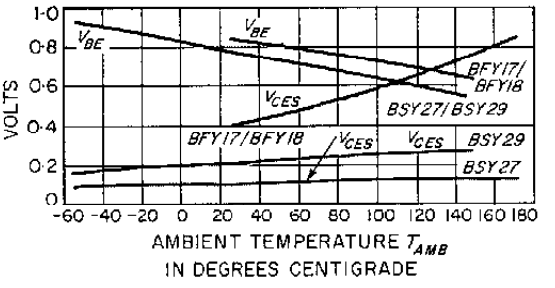


Figure 3—Typical variations of input voltage V_{BE} and collector-emitter saturation voltage V_{CES} with ambient temperature for $I_C = 10$ milliamperes and $I_C/I_B = 10$.

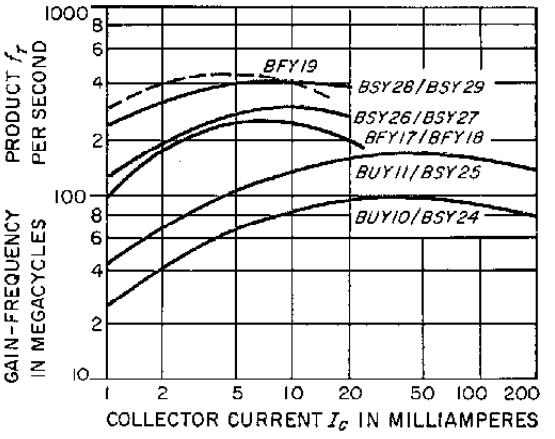


Figure 4—Typical variations of the common-emitter gain-frequency product f_T with collector current I_C.

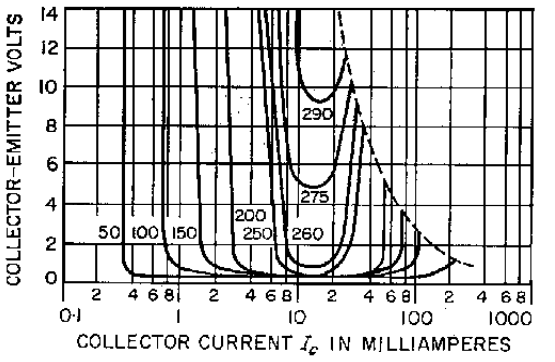


Figure 5—Typical variation for epitaxial *BSY27* of gain-frequency product f_T witsh collector current I_C and collector-emitter voltage V_{CE}. The values of f_T in megacycles per second are indicated for each curve.

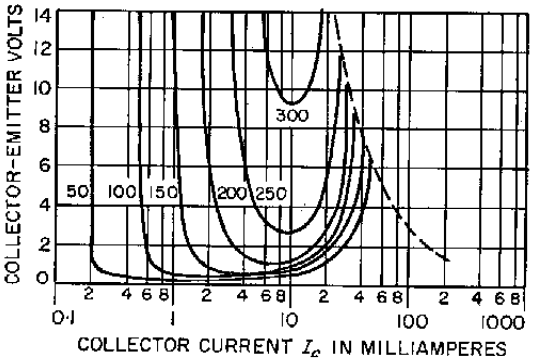


Figure 6—Typical variations for non-epitaxial *BFY17* of gain-frequency product f_T with collector current I_C, and collector-emitter voltage V_{CE}. The values of f_T in megacycles per second are indicated for each curve.

useful value for collector currents down to the microampere region.

The variation of the gain-frequency product f_T as a function of collector current is shown in Figure 4 for representative samples of the three main groups of devices. As in the case of low-frequency gain, f_T reaches a maximum value at a current that becomes smaller for smaller-area devices. The effect of epitaxy is to improve the value of f_T at high currents by preventing the collector-base junction voltage from falling at high currents. However, it can be seen from the curves that the improvement is slight at currents around 5 milliamperes, which are commonly used for small-signal amplifiers.

The effect on the f_T of a particular device of varying the bias conditions can most conveniently be illustrated by plotting contours of constant f_T in the I_C, V_{CE} plane. This has been done in Figures 5 and 6 for the *BSY27* and *BFY17*. These two devices have basically the same structure, but the *BSY27* is epitaxial, unlike the *BFY17*. It will be noticed that the contours of the epitaxial device are more "flat bottomed" than those of the non-epitaxial, that is, the f_T of the epitaxial device varies less with current near its peak value for a given collector voltage and maintains its f_T more nearly constant to lower voltage levels than the non-epitaxial device for currents above 1 milliampere.

The effect of epitaxy is much more marked at high collector currents. For the epitaxial device the contours are nearly symmetrical on either side of the peak f_T axis, but for the non-epitaxial device the contours are very much compressed and distorted towards the higher-voltage and lower-current region. This implies an inferior switching performance at high currents for the non-epitaxial device, even in a non-saturating circuit.

The variations of the common-base output capacitance C_{OB} at zero emitter current with collector voltage for the three basic families of devices are shown in Figure 7. At voltages high enough for the built-in voltage to be

neglected, the capacitance is approximately inversely proportional to the cube root of the applied voltage, as would be expected for a graded junction. The measured capacitance includes about 0·5 picofarad due to the stray capacitances inside the case.

The emitter depletion capacitance C_{Te} is important in switching applications because it influences the delay time, that is, the time between the application of a drive voltage to the base of the transistor and the beginning of the rise in collector current.

In amplifier applications the emitter depletion capacitance causes the fall in f_T at low emitter currents. However, in this case it is the value of C_{Te} when the emitter junction is forward biased that is required, and this cannot be measured directly.

For individual devices an effective value of C_{Te} appropriate to this application can be deduced from measurements of the variation of f_T with emitter current; but for both applications the value of C_{Te} at low reverse bias is useful for comparison purposes.

Typical values of C_{Te} at an emitter-base reverse bias of 0·5 volt for the three main families are: *BSY25* family, 45 picofarads; *BSY27* family, 6·5 picofarads; and *BSY29* family, 3·2 picofarads.

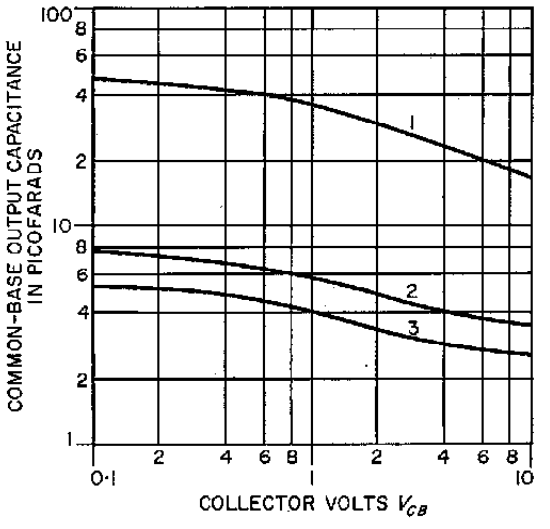


Figure 7—Typical variations of common-base output capacitance C_{OB} with collector-base voltage V_{CB} for $I_E = 0$. Curve 1 is *BSY24/25*, *BFY15/16*, *BUY10/11*, and *BLY10/11*. Curve 2 is for *BSY26/27* and *BFY17/18*. Curve 3 is for *BSY28/29* and *BFY19*.

3. Switching Performance

The differences between devices can be seen by comparing measurements of the times for delay, rise, storage delay, and fall in a particular circuit. Typical results are shown in Table 2.

TABLE 2
TYPICAL SWITCHING TIMES OF EPITAXIAL DEVICES

Times in Nanoseconds	*BSY24*	*BSY25*	*BSY26*	*BSY27*	*BSY28*	*BSY29*
t_d	9	7	8	9	7	7
t_r	10	9	12	10	7	5
t_{sd}	350	310	15	19	8	7
t_f	25	15	16	16	14	13
I_C in Milliamperes	100	100	10	10	10	10

4. Conclusions

This survey has been too brief to describe more than a few of the important parameters of silicon transistors. It is hoped, however, that it will help circuit designers to find the type most suited to their requirements.

J. Bickley received an honours degree in electrical engineering from Queen Mary College of the University of London in 1954. After three years of research in liquid-dielectric breakdown, he joined the Royal Naval Scientific Service and worked on measuring techniques for high-frequency silicon mesa transistors at the Services Electronic Research Laboratories at Baldock.

He joined the Transistor Division of Standard Telephones and Cables in May 1960 to develop measuring techniques for the diffused silicon high-frequency transistors then being developed. At present he is in charge of the device-characterization laboratory.

Mixer Diodes for 6 Gigacycles Per Second

J. B. SETCHFIELD

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1. Introduction

It is almost universal practice in microwave communication systems to amplify the signal to be passed on from one repeater to the next at a lower more-convenient frequency. This involves superheterodyne frequency conversion, first down to the chosen intermediate frequency and then, after amplification, up to the original or to a new microwave frequency. The accepted non-linear elements used for conversion in this frequency range are semiconductor diodes. For down-conversion the non-linear-resistance property of these diodes is invariably used while for up-conversion there is a choice between non-linear-resistance and non-linear-capacitance action. Each has its own merits, and while the latter might be the popular choice today on the grounds of greater conversion efficiency, the former can be designed to exhibit certain worthwhile advantages that can outweigh the disadvantage of a slightly higher conversion loss. In the present case non-linear resistance action for the up-converter was chosen to achieve a controlled intermediate-frequency limiting level.

Below are detailed the target specifications for the two diodes, a low-level down-converter and a medium-level up-converter. It will be seen that in addition to qualities that compare with the best mixer diodes available, certain unusual features were demanded. For the down-converter a very-low intermediate-frequency impedance was required and for the up-converter an envelope- or amplitude-limiting characteristic had to occur at a prescribed intermediate-frequency drive level.

Further justification for the choice of non-linear-resistance action for the up-converter lay in the request for an exceptionally low level of harmonics of the intermediate frequency in the super-high-frequency band. It is well known that variable-capacitance action results in the efficient production of harmonics whereas variable-resistance diodes, besides being inherently less efficient in this respect, can be designed to suppress this tendency still further. There may be ways of obtaining similar suppressions in reactance modulators while emphasizing desired modulation products, but so far no exhaustive investigations are known that have explored this in the depth done with resistance modulators.

1.1 Target Specifications

Low-Level Down-Converter

Super-High-Frequency Range = 5·8 to 6·5 Gigacycles Per Second

Local-Oscillator Level for Optimum Noise Figure = 1 Milliwatt

Noise-Temperature Ratio = 1·5, Maximum

Conversion Loss = 4·5 Decibels, Maximum

Intermediate-Frequency Source Impedance at the Above Local-Oscillator Level = 100 to 150 Ohms

Intermediate Frequency = 70 Megacycles Per Second

Burnout Level (Continuous) = + 8 Decibels Referred to 1 Milliwatt, Minimum

Ambient Temperature Range = 10 to 60 Degrees Centigrade

Spread in Radio-Frequency Impedance Match = 20 Per Cent, Maximum

Encapsulation = Hermetic

Medium-Level Up-Converter

Super-High-Frequency Range = 5·8 to 6·5 Gigacycles Per Second

Intermediate-Frequency Impedance of Two Diodes in Parallel and in Reverse Polarity Conducting 8 Milliamperes Each = 100 ± 15 Ohms

Conversion Loss (Super-High to Super-High Frequency) at Super-High-Frequency Input of ±15 Decibels Referred to 1 Milliwatt and Rectified Intermediate-Frequency Current of 8 Milliamperes = 11 Decibels, Maximum

Direct Harmonics of 70-Megacycle-Per-Second Signal at 10 Milliamperes of Rectified Intermediate-Frequency Current = − 120 Decibels Referred to 1 Milliwatt, Maximum

Flatness of Rectification Response from 40 to 100 Megacycles Per Second = 0·5 Decibel with 5-Per-Cent Maximum Spread

Super-High-Frequency Burnout Level with 8 Milliamperes of Rectified Current (Per Diode) = 100 Milliwatts, Minimum

Intermediate-Frequency Burnout Level with ±15 Decibels Referred to 1 Milliwatt Super-High-Frequency Input = 15 Milliamperes of Rectified Current, Minimum

Ambient Temperature Range = 10 to 60 Degrees Centigrade

Envelope Limiting (Intermediate- to Super-High-Frequency Compression of Amplitude Modulation) at 8 Milliamperes Rectified Current = 5 to 1, Minimum

Spread in Radio-Frequency Impedance = 20 Per Cent, Maximum

Encapsulation = Hermetic

2. Low-Level Down-Converter

It is assumed that the reader is familiar with the principles of frequency mixing and conversion and the peculiar requirements at microwave frequencies that dictate the use of the point-contact semiconductor diode.

2.1 Design Parameters

The most-revealing property of any radio receiver is its over-all noise figure, which defines the sensitivity limit. In a superheterodyne down-converter employing a crystal diode the parameters of the diode are related to the receiver over-all noise figure thus:—

$$F_R = L_X(t_x + F_{IF} - 1)$$

where F_R = receiver over-all noise figure, expressed as a ratio.

L_X = conversion loss of the crystal mixer, which is defined as the ratio of the available power at the signal input to the available power at the intermediate-frequency terminals.

t_x = noise-temperature ratio of the crystal diode, which is defined as the ratio of the available noise power output of the diode to that of an equivalent resistor at room temperature.

F_{IF} = noise figure of the intermediate-frequency amplifier, expressed as a ratio.

Conversion loss and noise-temperature ratio are seen to be the significant diode parameters with the former playing the dominant role since F_{IF} is always greater than unity.

2.1.1 *Conversion Loss*

The microwave mixer is usually represented as a time-varying conductance between the radio-frequency input terminals and the intermediate-frequency output terminals. As can be seen from Figure 1, a diagrammatic representation of a crystal diode, this time-varying conductance corresponding to R_B, the barrier resistance, cannot be realized in semiconductor diodes without the accompanying degrading elements C_B, the barrier capacitance and R_S, the bulk series resistance of the inactive part of the semiconductor. It can be said that the conversion loss is determined by the non-linear barrier resistance alone and that the degradation above the theoretical minimum is due to the parasitic elements C_B and R_S.

Considering the effect of R_B alone, the current-voltage relationship for the junction is

$$I = I_o \exp[(qV/kT) - 1].$$

The conversion loss depends on the degree of non-linearity which in turn depends on the voltage swing V and thus is a function of the local-oscillator drive. As the local-oscillator

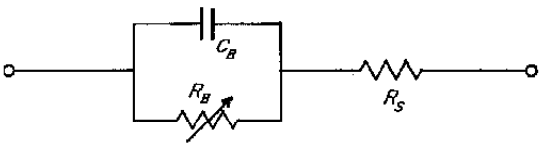


Figure 1—Network of principal elements of a crystal diode.

drive is increased from zero, conversion loss decreases from infinity to a theoretical minimum of 2. When the effects of the parasitics C_B and R_S are added it is seen that for a weak local-oscillator drive the value of R_B is high and C_B shunts a large proportion of input signal power into R_S where it is dissipated. For a large local-oscillator drive R_B becomes smaller, the shunting effect of C_B is reduced, but in the limit the value of R_B approaches that of R_S and the power is shared.

Combining these effects it is obvious that for any values of R_B, C_B, and R_S there is an optimum local-oscillator drive for minimum conversion loss and that minimization of the product $C_B R_S$ leads to the lowest conversion loss.

The ratio of power available to the nonlinear barrier resistance P_B to the total power available P can be calculated.

$$\frac{P_B}{P} = \left(1 + \frac{R_S}{R_B} + \omega^2 C_B^2 R_B R_S\right)^{-1}$$

where $\omega = 2\pi f$.

The local-oscillator power and the bias applied to the diode are adjusted to optimize R_B. The local-oscillator drive should be sufficiently large to provide the necessary degree of nonlinearity but not so large that R_B becomes comparable or smaller than R_S.

2.1.2 *Noise*

When excited by a direct or alternating voltage, a mixer crystal produces more noise

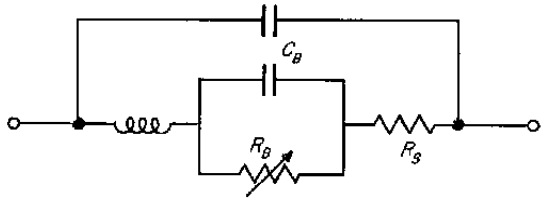


Figure 2—Equivalent circuit of a crystal diode including capacitance of case and inductance of connections and whisker.

than an equivalent resistor at the same temperature. This noise has three main sources: (A) thermal or Johnson noise, (B) shot noise, and (C) flicker noise. The last is small enough at high frequencies to be neglected. Thermal noise is generated mainly in the total series resistance. Uhlir [1] has postulated that shot noise is associated with minority-carrier flow across the barrier and may happen in three main ways: (A) a thermally generated minority carrier may diffuse to the barrier where it falls down the potential hill, (B) a majority carrier may possess sufficient energy to overcome the potential barrier to become a minority carrier, and (C) this minority carrier, if it does not recombine, may recross the barrier. The total barrier noise is the sum of these three components, the origins being principally the emission admittances of both types of carrier and the flow of saturation current. To reduce, as much as possible, the incidence of recrossing the barrier by minority carriers, the carrier lifetime in the semiconductor material should be as short as possible. It can be shown that both Johnson noise and shot noise can be minimized by holding R_S at a low level and that small saturation current or a good reverse characteristic is consistent with low shot noise. It is obvious that a reduction of Johnson noise will follow a reduced R_S but the bearing of this factor on the shot-noise power is not so obvious. Uhlir showed that for an ideal nonlinear resistance (barrier only) lowest shot noise occurs at infinite local-oscillator swing. This, however, is modified by the presence of series resistance. An increase in local-oscillator swing results in a reduced R_B, an increase in shot-noise current, but a decreased R_B shot-noise power contribution. When R_B falls to a value comparable with that of R_S, the R_S and R_B contributions become comparable and because the noise power contribution of R_S increases with shot-noise current while that of R_B decreases, a situation is reached where further increase of local-oscillator drive results in increased noise. As in the case of conversion loss, therefore, there is an optimum value of local-oscillator drive

for minimum noise and it is fortunate that the two requirements are not incompatible. The importance of a low value of series resistance is again emphasized. Shot noise associated with high reverse saturation current has been noted in practice. Lindell and Oxley [2] have reported that when a crystal diode is alternately and equally biased forward and backward the noise output for current in the reverse direction is considerably greater than for current in the forward direction, although the reverse current is very much smaller. Further, the greater the reverse current the greater is the noise produced.

2.1.3 *Radio-Frequency Impedance*

In considering the radio-frequency impedance of a crystal mixer the equivalent circuit of Figure 1 must be modified to include the overall case capacitance and the inductances of the whisker and internal connections. These are included in Figure 2. These additional factors can be tuned out in the crystal-holder design but have to be held constant from sample to sample since it is the spread in radio-frequency impedance in a given design of holder that is of significance. In fact, a consistent radio-frequency impedance is chiefly a matter of precision in manufacture. This precision applies not only to the mechanical tolerances of component parts, but accurate assembly to a pattern, control of semiconductor properties, surface treatment, whisker-point shape, whisker pressure, and forming.

2.1.4 *Intermediate-Frequency Impedance*

The reactive components considered in 2.1.3 are too small to be very significant at 70 megacycles per second, the intermediate frequency, and the same is true of the effects of the radio-frequency system in the low-Q case as implied in the present broad-band application. Thus the intermediate-frequency impedance is almost wholly resistive and corresponds closely to the average resistance when the characteristic is traversed by an alternating voltage. R_S, being only a few ohms, has a negligible effect so that the intermediate-frequency impedance is clearly dependent on R_B the barrier resistance.

It will be remembered that the value of R_B decreases with increasing local-oscillator drive and this must also be true of the intermediate-frequency impedance. It is necessary, therefore, to arrange in design that there is correspondence between optimum conversion loss, minimum noise, and the required intermediate-frequency impedance at the chosen local-oscillator drive level. Junction area obviously also bears on intermediate-frequency impedance but there is a very distinct limit to the permissible increase in C_B resulting from any effort to reduce the intermediate-frequency impedance by an increase of junction area.

2.1.5 *Burnout Level*

Burnout is a condition due to overload resulting in a significant deterioration of mixer performance. It has become customary in radar usage to specify the maximum spike energy (measured in ergs) to which the crystal diode may be submitted. This is because the radar crystal is subjected to such spikes due to leakage past the transmit-receive cell at each transmitted pulse. This kind of rating is not relevant to crystals used in communication systems, a continuous-wave overload rating being more appropriate.

Burnout is entirely a function of heat dissipation at the junction and depends on the semiconductor material, the area of the junction, and whether or not the junction was subjected to prior electrical forming.

2.2 Practical Mixer

The structure and mode of operation of the point-contact microwave mixer diode are not, even now, unambiguously understood so that, except for the guiding principles outlined in the previous section and the experience of

other workers, the ultimate approach is empirical to an appreciable extent. It is evident that a small $R_S C_B$ product and particularly a small R_S should be significantly important.

2.2.1 *Choice of Semiconductor*

Messenger and McCoy [3] in an excellent treatise on microwave mixer diodes have shown that maximization of the quantity $n^{1/2}b$, where n is the impurity concentration and b the carrier mobility, is desirable. The quantity bears strongly on the bulk series resistance R_S which, as has been shown, must be minimized.

Microwave mixers in the past have been made from p-type silicon and n-type germanium. Experimental units have been made from n-type gallium arsenide. From considerations of mobility the choice would be n-type since electron mobility is always higher than hole mobility. Silicon of n-type, however, turns out to be poor for reasons not well understood. Electron mobilities in germanium and gallium arsenide are quoted as 3800 and 5000 square centimetres per volt-second, respectively, while hole mobility in silicon is 500 square centimetres per volt-second. This is why series resistance in silicon microwave diodes ranges from 15 to 75 ohms compared with 3 to 5 ohms for germanium and gallium arsenide. On this evidence it would appear that, in this respect at least, silicon is inferior. On the question of intermediate-frequency impedance too, a special problem in our case, silicon diodes are at a disadvantage, showing higher values than those of germanium and gallium arsenide. The popularity of the silicon diode can perhaps be explained by its simplicity in manufacture consisting, as it does, of a tungsten whisker merely "shaken down" on to a die of polycrystalline silicon. In contrast to this, germanium and gallium arsenide are invariably monocrystalline and the whisker has to be welded or formed by the passage of an electric current to the semiconductor surface. A mechanical advantage of this forming is that it confers a degree of ruggedness not present in the silicon diode.

Other factors for comparison are lifetime, energy gap, and dielectric constant. Lifetime should be short to avoid charge storage and transit-time effects, energy gap should be large to ensure high working temperature, and the dielectric constant should be small to minimize barrier capacitance. At the high impurity concentration levels involved lifetime is so short, due to the density of recombination centres, that it is difficult to measure.

On energy gap and dielectric constant gallium arsenide is seen from Table 1 to score over both germanium and silicon.

Reasons for the rejection of silicon have been given but the fact that germanium was chosen in face of the apparent superiority of gallium arsenide requires justification. The reasons are as follows.

(A) The much-greater background of metallurgical experience with germanium.

(B) The apparent inability to achieve the promised electron-mobility advantage in heavily doped gallium arsenide.

(C) The very-marginal superiority of gallium-arsenide diodes reported by other workers and then only in selected laboratory samples.

TABLE 1
PROPERTIES OF SEMICONDUCTORS

Material	Energy Gap in Electron-Volts	Dielectric Constant
Germanium	0·68	16·0
Silicon	1·1	11·8
Gallium Arsenide	1·35	11·1

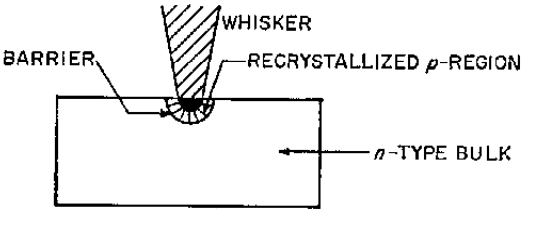


Figure 3—Model of a diode.

(D) The report that one reputable American manufacturer had withdrawn from the market its gallium-arsenide microwave diodes.

(E) Considerable doubts about the long-term stability of other heavily doped abrupt-junction gallium-arsenide devices.

(F) Our own inability to achieve results superior to germanium in our early investigations coupled with the doubts above.

2.2.2 *Whisker Material*

The function of the cat whisker in the formed germanium diode differs from that in the unformed silicon diode. The latter is an example of a metal-to-semiconductor contact diode while in the former, the semiconductor immediately under the whisker undergoes a change due to the heat developed by the passage of the forming current. In the extreme case (heavy forming), the conductivity type is changed and a *p-n* junction results. In the case of the microwave diode where the forming has to be extremely light to preserve a small low-capacitance junction and where the doping level of the *n*-type bulk semiconductor is high, it is doubtful whether complete conversion takes place. A diode still results but of the contact type with probably a region of higher-resistivity *n*-type material immediately under the point. This is not a desirable state of affairs since this region of higher resistivity adds to the unwanted series resistance R_S. The conversion to *p*-type can, however, be enhanced by the inclusion in the whisker material of a metal of Group III in the periodic table, for example indium, gallium, or aluminium, which are metals commonly used as *p*-type doping agents in germanium. When such a metal is included in sufficient quantity a minute *p*-type region of low resistivity results and an abrupt *p-n* junction is formed. The increased series resistance due to alloying an undoped whisker is largely avoided.

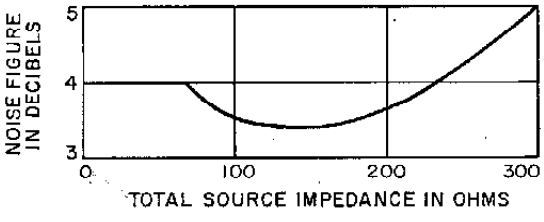


Figure 4—Noise figure at an intermediate frequency of 70 megacycles per second.

Mechanically a whisker wire has to be resilient, springy, and capable of supporting a sharp point under some pressure. A wire having the requisite electrical and mechanical properties is composed of 90-per-cent copper and 10-per-cent aluminium. Such a composition is known as aluminium-bronze and forms the whisker wire in this diode.

A model of this diode is depicted in Figure 3. It is assumed that a convergent flow of electrons takes place from the highly doped *n*-type region to the less-heavily doped *p*-region. It is probable that lifetime is very short in the aluminium-copper-doped *p*-region and recombination can be speeded by providing a germanium surface of high recombination velocity, that is, mechanically polished.

2.2.3 *Intermediate-Frequency-Impedance Realization*

The necessity for matching the intermediate-frequency terminals of the mixer to the intermediate-frequency amplifier is of prime importance. Modern grounded-grid low-input-impedance amplifiers demand either mixers of correspondingly low intermediate-frequency impedance or matching transformers or networks. Such matching transformers are undesirable in broad-band working, hence the request for low intermediate-frequency impedance in the mixer. The value of 100 to 150 ohms is the optimum for minimum noise in the present case as is seen in Figure 4.

It was shown in Section 2.1.4 that, apart from effects of the radio-frequency circuit which are fixed by other considerations, the intermediate-frequency impedance is largely

dependent on the barrier resistance at the working local-oscillator drive. The barrier resistance also depends on the resistivity of the semiconductor and the junction area or, as is equivalent, on the forward conductance. Most germanium microwave mixers have intermediate-frequency impedances of about 200 ohms and this was confirmed in our early work. Occasionally a diode of 150 ohms would appear but generally speaking the trend of the curves of Figure 5 was confirmed. These curves show rectification efficiency and intermediate-frequency impedance against forming current for 1 milliwatt drive at 6 gigacycles per second (increased forming current means larger junction area). The intermediate-frequency impedance is seen to continue to fall with increased forming but rectification efficiency (and presumably smallest conversion loss and optimum junction area) passes through a well-defined maximum.

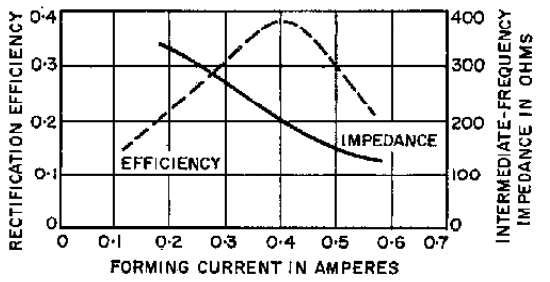


Figure 5—Rectification efficiency and intermediate-frequency impedance plotted against forming current for a single-junction germanium diode for 1 milliwatt drive at 6 gigacycles per second.

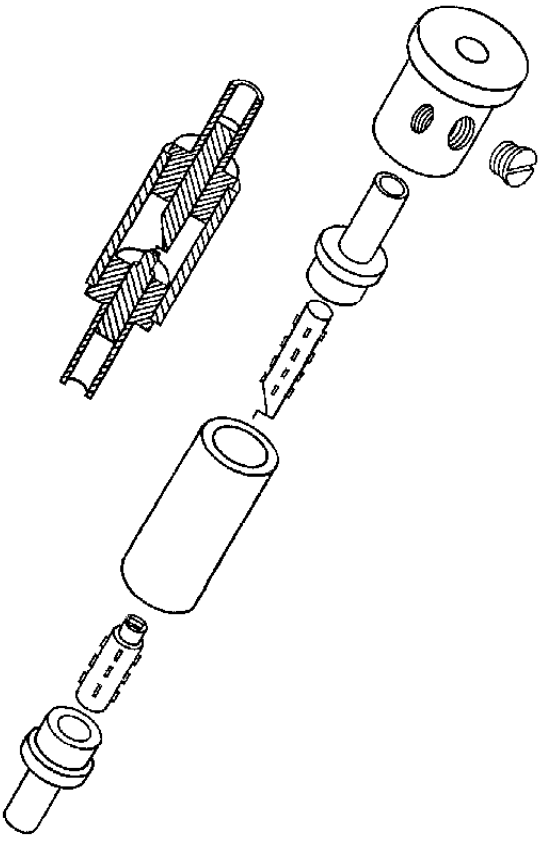


Figure 6—Final form of diode.

It should be noted that to keep R_S low the germanium resistivity was as low as would support efficient rectification.

The reason for the degradation of conversion loss with increase of junction area can be seen from the following relations, which apply to a diode of this geometrical form:—

$$R_S \propto 1/a$$
$$C_B \propto a^2$$

where a is the radius of the junction.

The disproportionate change in R_S and C_B for an increase in junction area means a higher R_SC_B product and an increased conversion loss.

If now the effective junction area is, instead, increased by putting down two points on to the germanium surface, the changes in R_S and C_B are proportionate, the product R_SC_B is unchanged but the impedance is halved.

2.2.4 *Final Diode*

Figure 6 shows the final form of the diode together with its component parts. The semiconductor die is of n-type germanium, having a resistivity of 0·004 ohm-centimeter and the twin cat whisker is of aluminium-bronze (90-per-cent copper and 10-per-cent aluminium). The cartridge is of the standard form adopted

by most diode manufacturers, is symmetrical, and has the facility of polarity reversal by virtue of a separate end cap.

Tests at microwave frequencies are complicated and would warrant a separate report. Figures 7, 8, and 9 show typical results for conversion loss, over-all noise figure, and intermediate-frequency impedance respectively. These are for an average diode; conversion losses of 3·9 decibels have been measured.

The curves show that the target specifications for conversion loss, over-all noise figure, and intermediate-frequency impedance have been met for a local-oscillator drive of 1·2 milliwatts and a direct-current forward bias of 0·1 volt. The actual over-all noise figure is degraded by the poor noise figure of 3·3 decibels of the amplifier used for the tests.

Substitution of the figures for conversion loss and over-all noise figure in the equation in Section 2.1 will reveal a rather larger value for the noise-temperature ratio t_x than that specified. The overriding effect of conversion loss has to be taken into account in this connection and the possibility of a slightly optimistic conversion-loss figure has to be admitted. No independent measurement of t_x has been made.

The double-whisker structure ensures a burn-out figure much better than the specification.

3. Medium-Level Up-Converter

It is recognized that up-converters using modern non-linear-capacitance diodes are more efficient than the older converters employing non-linear-resistance action. It is pertinent, however, to weigh this disadvantage against certain other desirable effects that can more-easily be realized in the non-linear resistance. The destructive influence of the parasitic bulk series resistance noted in the down-converter can be used to good effect in the up-converter to bring about an early saturation condition, which is useful for its amplitude-limiting action. Degrading the rectifying action at the higher frequencies results in lower levels of interfering harmonic production.

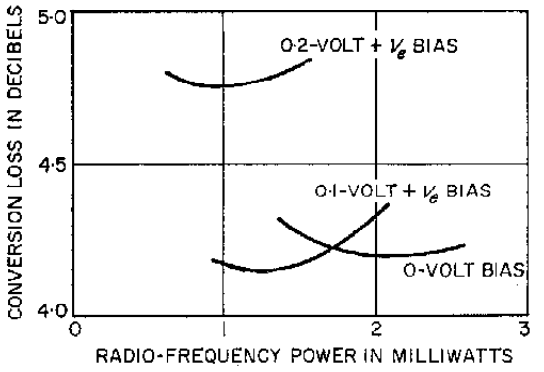


Figure 7—Conversion loss versus radio-frequency power input as a function of bias.

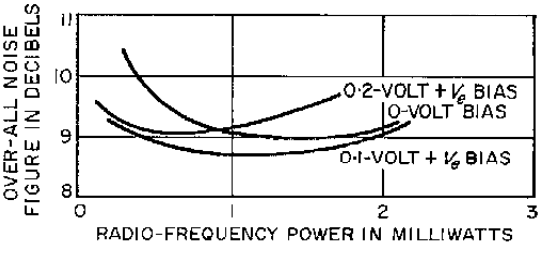


Figure 8—Over-all noise figure plotted against input power. The amplifier had a noise figure of 3·3 decibels. Three bias values are shown.

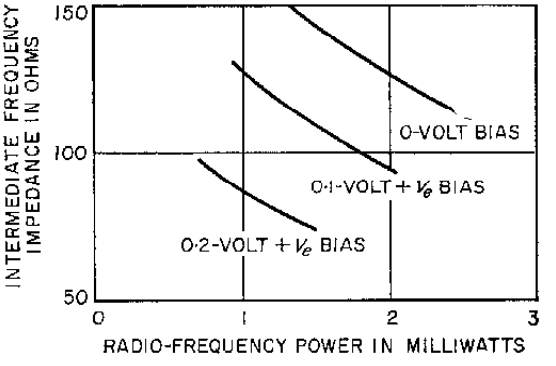
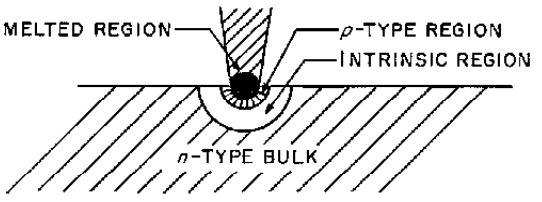


Figure 9—Intermediate-frequency impedance as a function of input power for the indicated bias values.

Figure 10—Formed point-contact germanium diode.



3.1 Design Parameters

The up-converter may better be regarded as a modulator in which the impedance of the diode to the high frequency is varied by the incident intermediate-frequency signal. If the diode resistance could be swung from open circuit to short circuit then the conversion could be made nearly lossless. It is important to note here that we are considering the loss from in to out at super-high frequencies, that is, from unmodulated carrier in to frequency-modulated carrier out.

In a typical circuit arrangement a hybrid T is used as a balanced mixer, the diodes are mounted in the side arms at points equidistant from the centre of the junction and are energized in antiphase at the intermediate frequency. The carrier is introduced in the H arm and the wanted sideband is taken out of the E arm through a filter.

3.1.1 *Conversion Loss*

The model of Figure 1 is equally applicable to the up-converter. If pure-resistance action is desired it is evident that the change in R_B must be maximized while C_B should preferably be as invariant as possible. This means that both the forward conductance and the reverse resistance should be as high as possible and that these extreme conditions should be achieved at the required intermediate-frequency swing. The junction capacitance can be larger than in the down-converter but should not be so large as to shunt seriously the barrier resistance in the reverse-biased condition.

3.1.2 *Limiting*

A saturation effect occurs when the intermediate-frequency swing encroaches on the linear region of the forward characteristic. As in the down-converter R_B reduces with increase of drive until a point is reached where R_S takes over and further increases in drive level produce little effect on the impedance. The incidence of this saturation can be controlled to a large degree by choice of semiconductor resistivity, and hence R_B, and by designing deliberately for a high value of R_S. In this way the rectifying property of the diode at high frequencies is also degraded.

3.2 Practical Up-Converter

The *p-i-n* structure, in which an intrinsic layer separates the p and n regions, has all the attributes for a variable-resistance modulator.

The intrinsic layer ensures a high reverse resistance and permits a larger junction area for a given capacitance. The capacitance, too, is not so voltage dependent and is largely invariant. During the forward half-cycles of the intermediate-frequency drive the intrinsic region is flooded with carriers and its conductance is high. At the super-high frequencies transit time across the intrinsic region precludes efficient rectification.

3.2.1 *Realization of p-i-n Structure*

In the formed point-contact germanium diode a variety of effects can be obtained by choice of the parameters, starting type, resistivity, whisker composition, and degree of forming. It is well known that when n-type germanium is heated sufficiently the conductivity decreases and, depending on the temperature to which it is raised, eventually goes through an intrinsic state to the opposite p conductivity type (thermal conversion). By correct choice of the parameters mentioned the structure of Figure 10 can be realized. For this case a comparatively high-resistivity n-type germanium is chosen in combination with a whisker carrying a p-type impurity metal. The degree

of forming is chosen to produce a heavily doped p region immediately under the whisker tapering into a region where the conductivity types are balanced and therefore of high resistivity. The transition can be made relatively sharp if a square pulse of forming current is applied having a rapid fall time. In this model, as distinct from that of the down converter, the relative resistivities of the p and n regions would suggest a divergent flow of holes from the p region into the bulk.

3.2.2 *Results*

From the Smith chart of Figure 11 the impedance change from +1 volt to −1·6 volts is seen to be predominantly resistive as expected. The graph showing conversion loss against intermediate-frequency drive, Figure 12, demonstrates that the conversion loss of 7·5 decibels at the required saturation level is better than the specification by 3·5 decibels. The required compression of 5:1 is maintained for an input power change of 5·6 decibels which is considered adequate.

No interfering harmonics of the intermediate frequency were detected at −117 decibels, the limit of the measuring equipment.

4. Acknowledgements

The author wishes to recognize the valuable assistance of A. B. Kaiser in making the diodes, Mrs. E. B. Evans and G. M. Meathrel

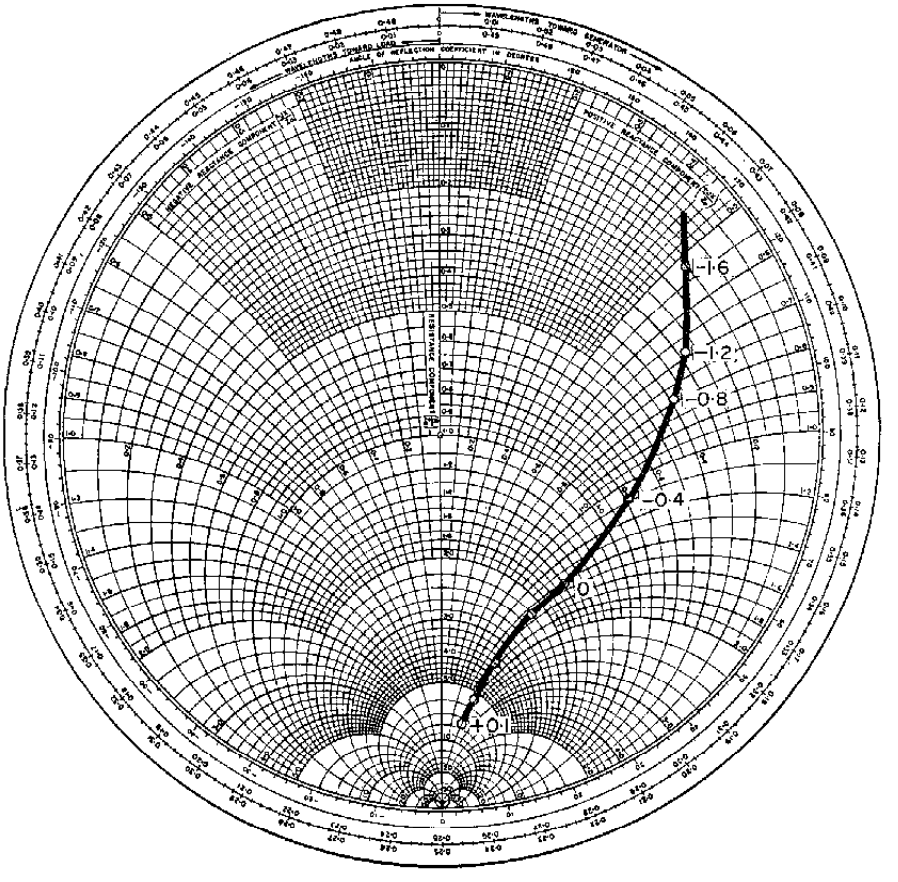


Figure 11—Smith chart plotting impedance change as a function of bias.

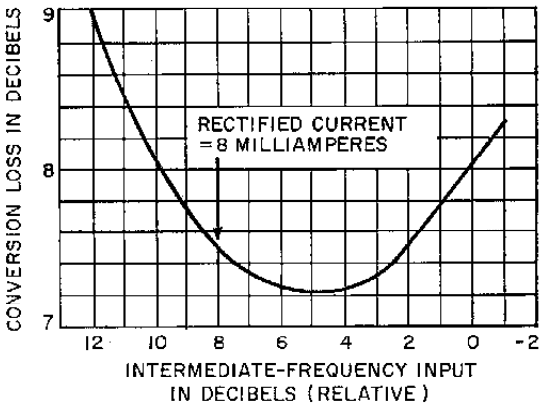


Figure 12—Conversion loss as a function of intermediate-frequency drive at 70 megacycles per second with 200 ohms of series resistance and a super-high-frequency output at +15 decibels referred to 1 milliwatt.

for work on whiskers, B. F. Armsby for making the tools, and D. R. Hill and R. C. Baron for painstaking measurements.

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J. B. Setchfield was born in Nottingham, England, on 9 February 1907. Starting as an amateur, he joined the radio industry in 1924 and until 1940 served with several firms in the development of broadcast receivers.

From 1940 to 1945, he was in the Royal Air Force, during the last half of which he was at the Telecommunication Research Establishment developing airborne radar equipment.

He joined Standard Telecommunication Laboratories late in 1945 and now heads a group studying the application of solid-state devices to microwave systems.

Vacuum Tubes for Submerged Repeaters

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The cost of an interruption in service and of replacing a submerged repeater in a multichannel telephone submarine-cable system places fantastic importance on the reliability of all parts of such repeaters and particularly on the vacuum tubes that are the essential active elements of these installations.

1. Properties of Submarine Repeater Tubes

1.1 Electrical

Unusual conditions must be anticipated in the design of vacuum tubes for long submarine-cable communication systems as they will be contained in amplifiers in sealed vessels at the bottom of the ocean accessible electrically only over the coaxial cable that connects the repeaters to each other and ultimately to the shore-based terminal equipment. Consequently, all power must be derived from a voltage drop across each repeater produced by current passing through the cable.

Calculations, as applied to optimized amplifiers, indicate need for high transconductance, low interelectrode capacitances, and low electrode voltages and currents.

Other things being equal, the higher the gain of the tube, the fewer will be the number of repeaters in the cable, provided that, at any point in the system, the signal-to-noise ratio does not fall below the specified limit. This calls for high transconductance. The lower the interelectrode capacitances, relative to high transconductance, the greater the transmitted bandwidth.

As all the amplifiers are operated in series from a voltage applied across the entire cable, low electrode voltages minimize this over-all voltage and low currents, particularly heater current, further reduce the operating power.

These basic properties must be associated with the two features, long life and utmost reliability, that put submarine repeater tubes in a class of their own.

1.2 Long Life

Long life implies maintenance of thermionic emission, constancy of electrical parameters, maintenance of insulation, and low evaporation rate from hot components.

1.3 Reliability

Reliability implies the possibility of an accident, a failure that is sudden and unexpected. Nevertheless a probability of failure can be assigned to a tube based on tests of a quantity of them.

2. Fundamentals of Design

2.1 General

The requisite properties of the tube are inextricably linked in a compromise, the success of which depends on the designer's skill in assessing the merits of conflicting factors. They will be treated generally here. Specific examples will appear in the later sections.

2.2 Electrical Characteristics and Dimensions

A high transconductance demands above all a close control of grid-to-cathode spacing. Small spacing presents insulation risks particularly in the presence of accidentally introduced conducting particles. If the spacing is less than the pitch of the winding of the grid wire non-linear effects result from the inhomogeneous electric field. Thin grid wires permit closer winding pitch but introduce mechanical weakness.

Low capacitances demand that only the active parts of relevant electrodes be exposed to one another. In practice, of course, connecting leads and essential screens make a substantial contribution to these capacitances.

Low electrode voltages demand closer electrode spacings to maintain the requisite electric field strengths, thus adding to capacitance.

Low heater currents are limited both by power

requirements and by the mechanical weakness of thin wire and evaporation if it is too thin.

Generally low-voltage operation increases the problem of maintaining uniformity among individual tubes as well as exposing the structure to greater risk of failure from accidentally introduced particulate contamination.

2.3 Components for Long Life

The outstanding requirement for long life is the maintenance of an adequate emission from the cathode.

Foremost, the alkaline earth carbonates must be satisfactory for this purpose by virtue of high purity and low evaporation rate. Secondly, the metal of the cathode core must be suitable. A vast amount of experience has been, and is still being, amassed on the constitution of the cathode core metal and its intentional and unintentional admixtures. Indeed, as the following history of submarine repeater tubes will reveal, certain aspects of the behaviour of cathodes were not clearly appreciated at the time and circumventing and limiting means were employed to avoid the deleterious phenomena observed.

For long life, the cathode must operate at the lowest possible temperature that will maintain stable equilibrium with the deactivating effect of residual gases in the tube. Evaporation is then least. In practice, with well-processed high-quality materials, the chosen optimum is 750 degrees centigrade.

Although not fundamentally an emissive property of the cathode, it is known that constituents of the cathode core metal can react, during operation, with the oxide coating to produce an interface layer having properties of an ohmic resistance. This causes degeneration in the cathode circuit and consequent loss of gain in the amplifier stage. Several new alloys have been devised since 1950 to avoid this effect.

A similar degeneration is brought about by the intrinsic resistance of the oxide layer. Control of this is partly one of correct coating and partly one of avoidance of gas contamination, which de-activates the layer and thus raises its resistance.

Aside from the deterioration problems of the cathode, the heater also has a wear-out function. Experience dictates the use of the lowest heater temperature that will maintain correct cathode operation.

A final point in protracted maintenance of constant tube characteristics is the possibility of relaxation of accidentally stressed components by the heat of operation producing an abnormal trend of characteristics during life.

2.4 Factors Affecting Reliability

The concept of reliability as the inverse of accident leads to the view that it can be obtained by perfection of all aspects of the tube and its preparation. Given the optimum design, it is necessary that the materials be as specified and that the execution of all the processes be likewise above criticism.

Translated into statistical terms, if the probability of failure of a tube be p then, in a system containing n repeaters each carrying redundancy in the form of twin parallel 3-tube amplifiers, the probability of breakdown of the system is $9\,np^2$. (A simple derivation of this expression is given in Section 9.) Thus, in a system containing 100 repeaters, the probability of the system failing is approximately equal to the probability of failure of an individual tube when the latter is 0·999 reliable over the acceptable operating life of the system.

It can be deduced that to attain a reliability at this level, the conventional system of statistical sampling becomes meaningless. The sample to be tested would be so large that mass production would be needed to handle it and the object would be defeated by the economics, at least, of controlling the operation at that level. Alternatively, accepting that, since accidents are random and their causes manifold, with a fairly large number of tubes the accident probability density will be uniform through

time, it might be concluded that the solution lies in this direction. Calculation again gives an impracticable answer. The time to life-test a manageable sample to give a reasonable certainty would exceed the acceptable life of the system (which is, say 20 years).

From these considerations one is driven inexorably to the conclusion that the desired reliability can be attained only by perfection in manufacture by all possible means and to seek, by virtue of the uniformity gained, to control the quality by supervision and elimination of disturbing influences. The practical realization of this is shown in Section 4 on manufacture.

3. Historical Survey

3.1 General Remarks

The physical form of the tubes concerned in this survey is shown in Figure 1. Table 1 gives details of their characteristics. The following notes on the individual types trace evolution to the present day and draw attention to the effect of fundamental discoveries on cathode behaviour and the advent of new constructional techniques.

3.2 *5A/153* Pentode

The *5A/153* pentode was a direct outcome of experience on coaxial repeater pentodes for land-based repeaters. At this time the basic known concepts for long life were low-current-density operation, low power dissipation, and low collector-electrode voltages. The phenomenon of interface growth was not recognized but selected batches of cathode nickel were known by experience to give consistent emission and small transconductance fall during running as well as low evaporation rates.

This type was used in 3-tube repeaters that served successfully for 10 years in 36-channel shallow-water cables. At the end of this time the repeaters were replaced by more-modern examples containing *5A/162* tubes to extend the capacity to 180 channels.

3.3 *5A/161* Pentode

As can be seen from the photograph the *5A/161* was generally similar to the *5A/153* type. The principal differences were small modifications to the grids to suit operation at a lower voltage and lower anode current. The heater was also altered to adapt it to the voltage needs of the repeater.

TABLE 1

Characteristics of Submarine-Repeater Tubes

	Type and Year Introduced						
	5A/153 1947	*5A/161* 1949	*5A/162* 1952	*5A/181* 1958	*5A/182* 1958	*LS882* 1958	*VX7158* 1960
Heater Voltage in Volts	10	25	5·5	10	10	10	7·4
Heater Current in Milliamperes	430	180	265	300	300	300	390
Anode Voltage in Volts	150	75	90	70	40	70	50
Screen Voltage in Volts	150	75	60	70	50	70	45
Anode Current in Milliamperes	9	2	6	10	3	10	15
Screen Current in Milliamperes	2·5	0·5	1·6	2·5	0·7	2·2	3
Transconductance in Milliamperes Per Volt	7·5	3·5	6	6·9	5·9	17	22
Control-Grid Voltage in Volts	−3	−2·5	−1·6	−3·5	−1·5	−1·7	−1·8
Anode Impedance in Kilohms	500	800	300	150	250	200	45
Input Capacitance in Picofarads	13·7	13·7	7·5	9·8	10·5	16	24·5
Output Capacitance in Picofarads	8	8	5	6·2	7·6	3·5	5
Anode-Grid Capacitance in Picofarads	0·007	0·007	0·020	0·020	0·025	0·020	0·025

The application of the repeaters was unusual. A 3-stage push-pull amplifier (6 tubes) was located at the continental-shelf edge at either end of a transatlantic telegraph cable. The greatly attenuated signal from the deep-sea portion of the cable was boosted and shaped before passing along the shallow-water section where ships produce electric interference. The resultant improvement in signal-to-noise ratio permitted the speed of the cable to be increased.

3.4 *5A/162* Pentode

The *5A/162* is equivalent to the British Post Office *6P12* tube [1]. Fundamentally, the electrical characteristics are adapted from the *6AM6* (*CV138*) miniature high-transconductance pentode to operate at the low anode and screen voltages indicated. Two noteworthy features of this tube made it long lived. A platinum cathode sleeve was used to avoid interface growth and very thorough attention was paid to elimination of gas generation [2], deleterious to cathode emission, by material purity and processing.

3.5 *5A/181* and *5A/182* Pentodes

These tubes are the respective equivalents of the British Post Office types *10P1* and *10P2* tubes [3]. They are respectively output- and input-stage versions of a common mechanical construction and are direct descendants of the *5A/162*. Further attention was applied to poisoning influences on the cathode, which is of the same tungsten-nickel alloy as a later modification of the *5A/162*. The heater-cathode insulation has been vastly improved by use of an intervening non-porous alumina sleeve to eliminate any danger of insulation breakdown inherent in operation with heater positive to cathode. This permits development of the anode voltage across the heater chain instead of in series as previously necessary.

3.6 *LS882* Pentode

In 1956 consideration was being given to the possibility of increasing the channel capacity

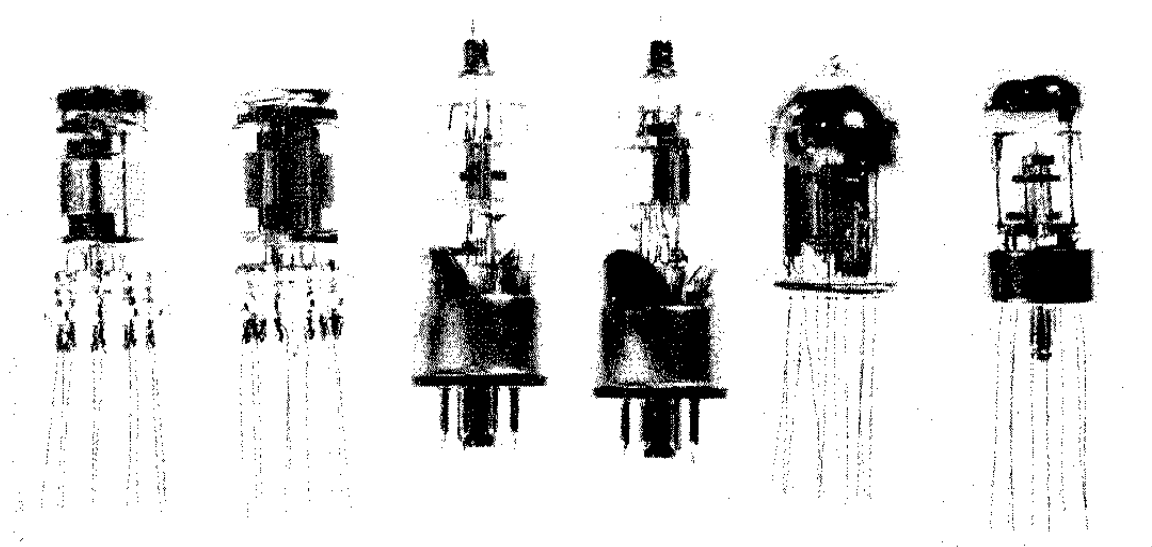


Figure 1—These submarine-repeater tubes are, from left to right, *5A/153*, *5A/161*, *5A/162*, *5A/181–5A/182*, *LS882*, and *VX7158*.

of submarine cables beyond that feasible with existing tubes. The characteristics of the *5A/162* and *5A/181* class of tube limit bandwidth to 0·5 megacycle per second for repeater spacing of 20 nautical miles. (A wider frequency band is possible by reducing the spacing since the cable attenuation is proportional to the square root of frequency.)

It was clearly recognized that no substantial improvement could be obtained with existing designs of tubes and that a parallel development in conventional repeater tubes must be exploited. The frame-grid construction [4, 5] offered, as it had for land-based repeaters, a large increase in transconductance with relatively little change in the important capacitances.

Calculations showed that, at the low working voltage, the spacing between control grid and screen grid would be quite small and the use of a conventional self-supporting screen grid would introduce electric field inhomogeneity and consequent lower efficiency. It was decided, therefore, to employ a frame grid for this grid also.

The resulting characteristics show the tube to be suitable for an amplifier having a bandwidth of 2 megacycles per second and providing 200 to 240 channels. This tube was the prototype of the *VX7158*.

3.7 *VX7158* Tetrode

The development exercise on the *LS882* proved the feasibility of the new construction. However, it appeared that the working voltage of the *LS882* would make the over-all voltage on long-haul systems excessive. Consideration was given to applying the principles of the *LS882* to a lower-voltage tube. A theoretical and practical investigation showed that the desired characteristics could be best obtained with a tetrode structure.

It may be remarked that in tubes designed for greater bandwidths, such as the *LS882* and the *VX7158*, the phase angle must be held to a low value since low-voltage operation gives a substantial transit angle. For this reason inductance of the leads must be minimized and a pressed glass base is indispensible.

4. Manufacture

4.1 General Remarks

It will be clear from the foregoing discussion that successful manufacture of submarine-repeater tubes calls for the highest standards in cleanliness of components and of fabrication. Coupled with this is the closest attention to dimensional control to ensure uniformity of the product.

The underlying principles involved in the design of the manufacturing section were: complete separation from other factory activities; restricted access by location and control of entry; ordered arrangement of processes; isolation of the assembly area, which has its own cleaned-air system; provision for close inspection, under engineering control, of all stages of manufacture; and adjacent testing areas, including life test, for rapid feedback of results.

4.2 Procedure

The layout of the section is shown in Figure 2. Entrance to the area is restricted to properly garbed personnel by a steward stationed in the store immediately adjacent to the double entry doors. The operation is arranged in a logical sequence so that there is an ordered flow of the components towards the assembly area, which is pressurized so that air flow into the rest of the system is counter to the component flow. (The latter is indicated by the thick line in the figure.)

Practically all the piece parts are fabricated from the raw material within the piece-part section on hand presses or winding machines as appropriate. This ensures that there is no danger of "cross-infection" from factory machinery used with other materials and processes.

The inherent relevant impurity content of

piece parts (mostly gas such as oxygen and carbon monoxide) must be relentlessly eliminated. The unwanted gas content is reduced to less than 2 percent of the original amount. This compares with 90 percent in ordinary receiving tubes.

The operators enter the assembly area through air locks in which they change their nylon coats for fresh more-thoroughly enveloping apparel including shoes and headgear.

Inspectors in the room check each assembler's work, providing immediate information on quality.

The subsequent pumping operation is lengthy and designed to expel the last dregs of gas from the tube.

4.3 Testing

All electrical tests are performed on high-precision equipment capable of measuring transconductance to 4 significant figures. A second mechanical inspection is made prior to life testing.

Life testing takes place on racks with closely controlled current supplies, where batches are run for a period of approximately a year. The tubes are tested at intervals during this time to gauge the trends of the principal parameters. Finally, on the basis of satisfactory trends, tubes are selected for repeater use. Tubes from the same batches are retained as shadow life tests to provide background information on the behaviour of their submerged counterparts. The final selection takes into account not only the electrical trends observed during the year but also the mechanical inspection results, which are repeated and counterchecked at the end of the period. The final tests are supplemented by X-ray stereo photographs to reveal hidden components, such as heaters and lead wires in the base and by special electrical tests, as for harmonic generation and noise, dictated by the specific circuit requirements.

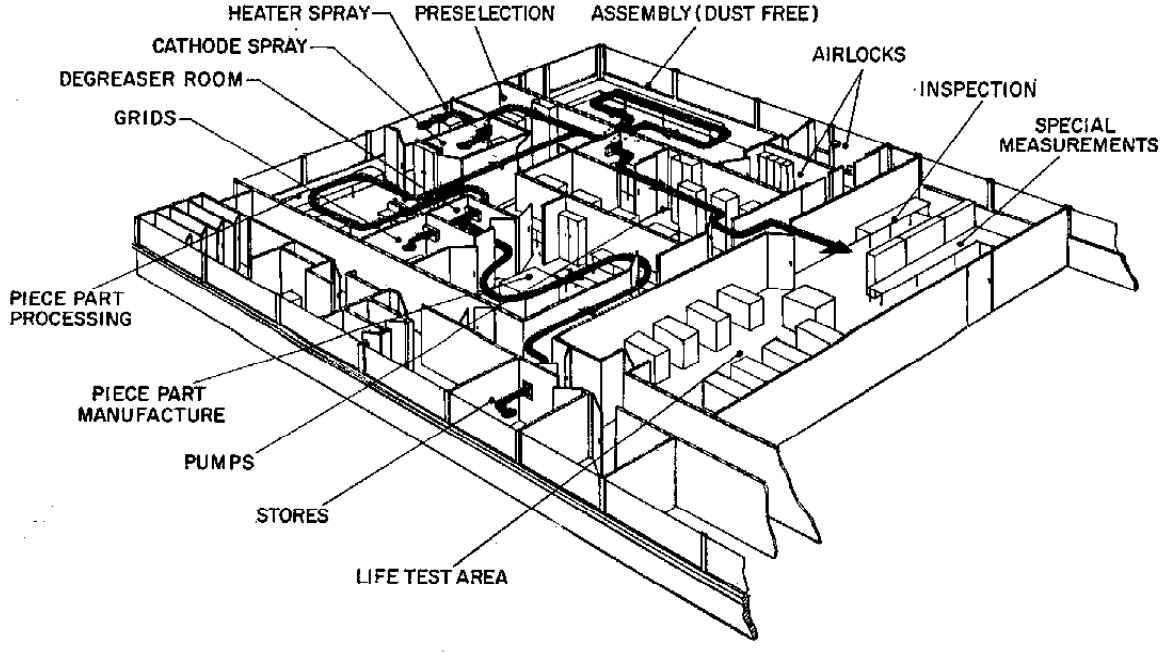


Figure 2—Layout of manufacturing area.

5. Performance

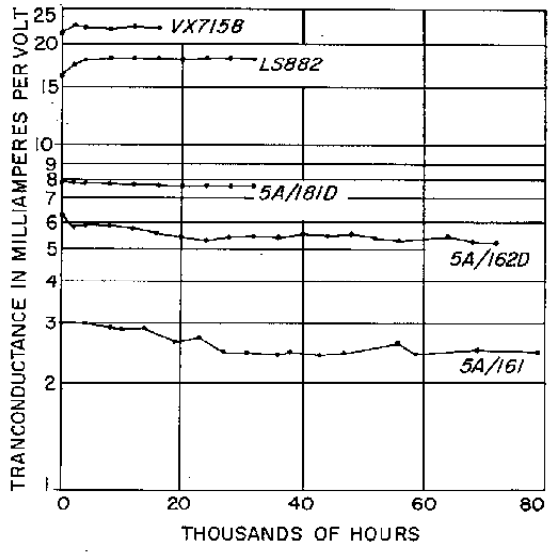
5.1 Uniformity of Product

In view of the need to maintain the tube parameters within close limits despite the complication introduced by low-voltage operation, it is worth demonstrating by a histogram what has been achieved. Figure 3 shows the control exerted over transconductance.

5.2 Life Performance

The shadow life tests on the various types continue to provide impressive evidence on the long-term stability of the tubes described, as Figure 4 illustrates. As it is difficult to present the very-small changes adequately it is worth remarking that a batch life test on *5A/181* tubes of early manufacture, which has now

Figure 4—Transconductance trends with life.



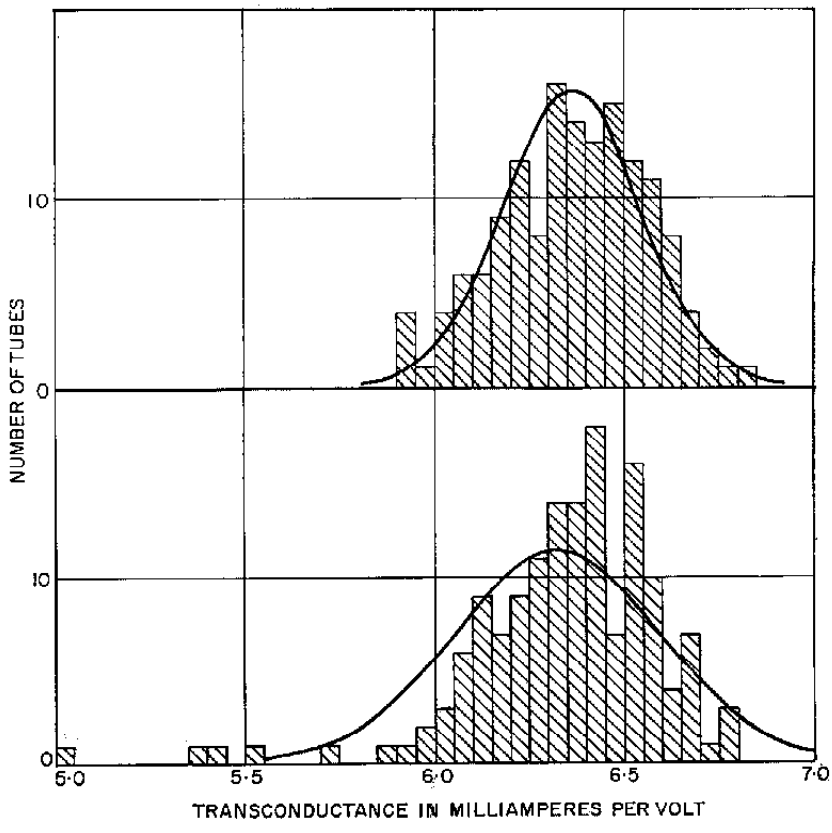


Figure 3—Histograms of transconductance of a group of *5A/182D* tubes. The lower curve is at the start of life test and the upper curve is after 1500 hours.

reached 4 years running time, shows a transconductance fall of 0·5 per cent per year over the past year. Additionally there is evidence that this already small rate of decline is diminishing with time.

6. Applications

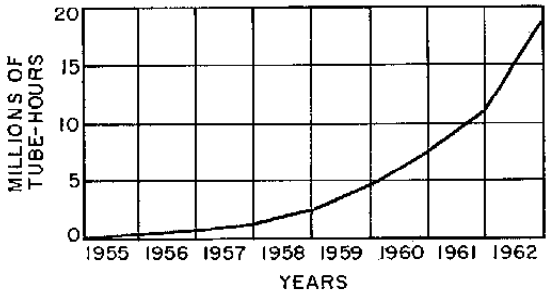
Examples of the tubes described in the foregoing are operating successfully in a large number of submarine-cable systems throughout the world. These include many installations around Europe. Half of the tubes installed in the repeaters of the United Kingdom-Canada cable laid in 1961 and the trans-Pacific cable now being laid between Vancouver, Canada, and Sydney, Australia, were manufactured in the plant surveyed above.

The chart in Figure 5 of tube-hours versus time continues to rise more steeply and it seems likely that the success of prior installations will encourage the planning and laying of wider-band systems with the newly developed tubes described above.

7. Acknowledgments

The writer is grateful for the help of all his colleagues in the submerged-repeater-tube section and especially to Mr. P. E. Douglas and Mr. N. B. Evershed. Thanks are also due to Mr. D. C. Rogers for some fruitful discussions during the course of preparation of this paper.

Figure 5—Cumulative tube-hours of tubes in submarine-repeater service from 1955 to 1962.



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3. M. F. Holmes and F. H. Reynolds, "Post Office Valves for Deep-Water Submarine Telephone Repeaters," *Proceedings of the Institution of Electrical Engineers*, volume 107, Part B, pages 165–171; March 1960.

4. E. J. Walsh, "Fine-Wire Type Vacuum Tube Grid," *Bell Laboratories Record*, volume 28, pages 165–167; April 1950.

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9. Appendix

Consider a system comprising n repeaters in series. Each repeater contains two 3-tube amplifiers connected in parallel.

Let the probability of failure of any valve be p. Thus the probability of no failure $= (1 - p)$. The probability of no failure in any of 3 tubes $= (1 - p)^3$.

Thus the probability of failure

$$\begin{aligned} &= 1 - (1 - p)^3 \\ &= p(3 - 3p + p^2) \\ &= S. \end{aligned}$$

With the parallel arrangement of amplifiers, the probability of failure of both channels

(which will fail the repeater) = S^2 and of no failure $1 - S^2$.

With n repeaters in series, the probability of the system not failing

$$= (1 - S^2)^n.$$

Thence the probability of breakdown of the system

$$= 1 - (1 - S^2)^n$$
$$= 1 - (1 - p^2(3 - 3p^2 + p^2)^2)^n$$
$$\approx 9np^2.$$

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He joined Standard Telephones and Cables in 1950 as production engineer on valves. Since 1955 he has been in charge of the development and production of submarine repeater valves

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Ceramic-Insulated Vacuum Tubes for Very-High-Frequency Industrial Heating

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The past decade has witnessed a remarkable expansion of the use of high-power tubes for industrial heating. Broadly speaking, this field can be divided into two distinct branches, induction heating and dielectric heating.

In induction heating the metal part to be heated is placed within or near a coil through which an alternating current is passed. A current induced in the metal produces heat as a function of the resistance offered to its flow.

In dielectric heating the material to be heated is placed between two metal plates with which it need have no physical contact. The two plates are connected to a radio-frequency voltage, and the material is heated by the current flowing through it as a function of its dielectric loss, which increases with frequency. Thus as frequency is increased, not only does the rate of heating increase, but materials having low dielectric losses can also be heated within reasonable times. The art of dielectric heating is still in the exciting stage where each new increase in power at a higher frequency heralds additional applications. Some present uses of dielectric heating are listed below.

(A) Pre-heating plastics and other materials to reduce production time and wear on tools.

(B) Welding of plastic materials, such as raincoats, inflatable toys, and upholstery.

(C) Packaging, as for shampoo packs.

(D) Wood-gluing.

(E) Unfreezing and melting such foods as chocolate, strawberries, and butter.

(F) Sterilizing of hospital packs.

(G) Disinfestation of pre-packed cake mixture, rice, flour, et cetera.

(H) Drying rolls and breakfast cereals.

(I) Drying textiles, veneers, paper products, bath cubes, foamed plastic and rubber, asbestos ceiling boards, and tobacco.

New uses for one of the tubes described in this article are as follows.

(A) Biscuit baking.

(B) Thawing bales of rubber frozen during transport through a very-cold climate.

(C) Thawing ice cream where a fault has occurred in processing.

(D) Thawing frozen fish.

(E) Sintering carborundum discs.

1. Tube Requirements

A new range of ceramic-insulated tubes have been designed for the highest frequencies now in use and to have adequate safety margins for operation at even higher frequencies for future applications. Other features are discussed below.

General mechanical robustness is essential because in the industrial heating field equipment will probably be used by unskilled operators and may be located in a machine shop, where vibration would be excessive for the normal radio types of tubes.

High safety factors against flashover across external terminals must be provided because a tube may be operated under very dirty and grimy conditions as in a wood shop or foundry and also in equipments where protection is not provided against voltage transients. The filament must be able to operate without mains regulation, as the filament voltage is not regulated in most industrial heating applications.

High efficiency is needed for economic reasons. In a 20-kilowatt-output radio-frequency generator, the cost of electric power represents about 75 per cent of the continuous running cost when depreciation and cost of tube replacements are taken into account.

Life must be not less than 5000 hours. In many cases tubes must be replaced by the manufacturer of the industrial heating equipment; a shut-down of a production line for tube replacement can be expensive.

The radio-frequency load may vary from full load to no load. This is reflected in the grid drive, where the grid dissipation under no-load conditions can be several times the dissipation under full load. The tube could fail if it is not designed to have an adequate grid safety factor.

The anode must be able to withstand momentary overloads.

2. New Range of Triodes

The range of tubes developed to meet these exacting requirements is given in Table 1.

To meet the high-frequency requirements these tubes have been designed with ceramic envelopes. Glass tubes have two main disadvantages, the dielectric loss in the glass at high frequencies and the overheating of the seals in handling the circulating radio-frequency currents, which can exceed 100 amperes. At these high frequencies the currents flow only through a very-thin skin of the conducting part, penetration generally being less than 0·001 inch (0·025 millimetre) at 100 megacycles per second. It is given by the expression

$$\mathrm{E} = k_1(R/\mu f)^{1/2} \qquad (1)$$

where E = skin depth

R = resistivity of the conducting material

μ = permeability of the conducting material

f = frequency in megacycles per second

k_1 = a constant.

The resistance of the conducting material is given by

$$R_{rf} = k_2(R/\mathrm{E}) = k_3(R\mu f)^{1/2}$$

where k_2 and k_3 are constants.

The radio-frequency circulating current is proportional to frequency, whence the radio-frequency losses P_{rf} are

$$P_{rf} = k_4 f^{3/2} R^{1/2} \mu^{1/2}$$

where k_4 is a constant.

In glass-to-metal seals, the metal is normally of a magnetic material, such as Kovar. As Kovar has high resistivity and permeability, the radio-frequency losses at the seals are therefore high, and at high frequencies cracking and/or glass suck-in near the seals can result.

With ceramic-to-metal seals this problem is minimized because the radio-frequency circulating currents at the seals flow through the metallizing and plating on the ceramic. The resistivity is low, and the permeability is unity.

Ceramics also have the additional advantages over glass of lower dielectric loss and higher safe operating temperature. The safe operating temperature of the seal may be between 220 and 250 degrees centigrade as against 180 degrees for Kovar-glass seals. Higher bake-out temperatures during evacuation increase both reliability and life.

Ceramic tubes will stand higher thermal and mechanical shocks than those with glass envelopes. They can be manufactured to more-exacting dimensional tolerances. Also, less chemical cleaning is required, since all brazes and final sealing-in can be carried out in a protective atmosphere.

TABLE 1
TUBE PERFORMANCE

Tube Type	Output Power in Kilowatts (Measured in Load)	Efficiency in Per Cent	Direct Anode Voltage	Maximum Frequency of Operation in Megacycles per Second
3JC/187E	5	80	6000	>220
3JC/203E	12	80	6000	220
3RC/223E	24	80	6000	>100
3ZC/262E	40	80	6000	100

TABLE 2

CHARACTERISTICS OF THE TUBES

	Length in Centimetres (Inches)	Amplification Factor μ	Mutual Conductance in Milliamperes per Volt
3JC/187E	21·8 (8·6)	12	22
3JC/203E	23·4 (9·2)	12	32
3RC/223E	26·2 (10·3)	12	60
3ZC/262E	36·5 (14·4)	12	60

TABLE 3

3JC/203E, TYPICAL MEASURED OSCILLATOR PERFORMANCE

Output Power, In Load	12 Kilowatts
Efficiency	80 Per Cent
Anode Direct Voltage	6 Kilovolts
Anode Direct Current	2·5 Amperes
Grid Direct Current	450 Milliamperes
Grid Bias Resistance	1500 Ohms
Peak Positive Grid Voltage	280 Volts

Tubes of the new range have smaller dimensions and lower inductances than previous designs, which contributes to efficiency at higher frequencies. Their tapered co-axial structures provide for easy insertion into high-frequency cavities.

The ceramic may be replaced by glass for induction-heating applications without other change of piece parts. Figure 1 shows the *3JC/187E* in both its glass and ceramic forms. The other three tubes are depicted in Figures 2 and 3.

3. Design of Metal-to-Ceramic Seals

The temperature coefficients of expansion of the metal and ceramic members of the seals must match as closely as possible, so that the seals can withstand repeated thermal cycling without fractures or vacuum leaks. Exact matching from ambient to brazing temperature is not possible with known materials, so careful attention must be given to the design of the seals. The high-alumina ceramic used has a tensile strength of 27 000 pounds per square inch (1898 kilogrammes per square centimetre) compared with a compressive strength of 10 times this value. The ceramic-to-metal seals have therefore been designed so that the ceramic is in compression after the brazing operation through the selection of a metal having a higher total thermal expansion than that of the ceramic and by brazing the metal components to the outside of the ceramic rings. The metal is a nickel-iron-

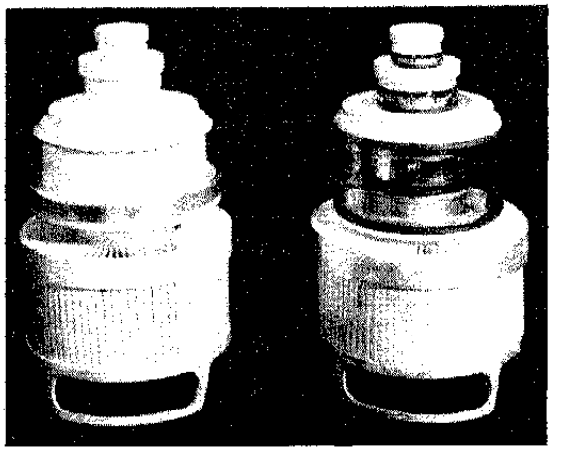


Figure 1—The ceramic and glass versions of the *3JC/187E* 5-kilowatt industrial triode.

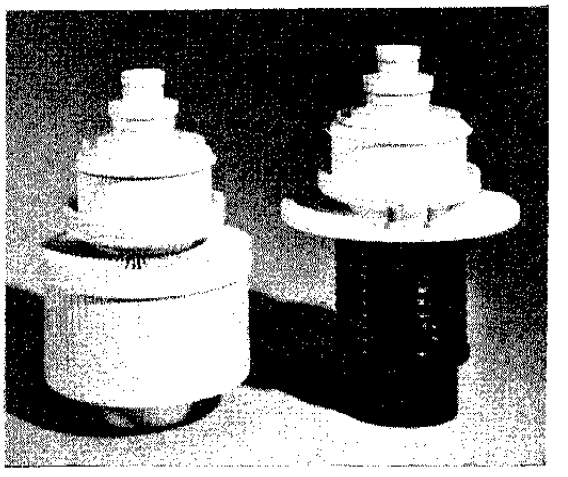


Figure 2—The *3JC/203E* 12-kilowatt and the *3RC/223E* 24-kilowatt industrial triodes.

cobalt alloy, and the ceramic is a 97·6-per-cent alumina.

4. Characteristics and Performance

4.1 Characteristics

To meet the important requirement of a high grid safety factor, the power dissipated in the grid must be kept low and the thermal emissivity of both the grid and anode surfaces should be as high as possible. To minimize grid power dissipation, the grid-to-cathode and grid-to-anode distances must be as short as mechanical stability and freedom from flash-over permit while a relatively coarse grid pitch, which in these tubes is about 1·4 times the grid-to-cathode distance, is used.

A more-detailed analysis of these design considerations is given in Section 12. The values of amplification factor and mutual conductance are given in Table 2.

The tubes are so similar in performance that only the *3JC/203E* will be examined in detail. Data for this tube are given in Table 3, and its characteristics in Figures 4 and 5.

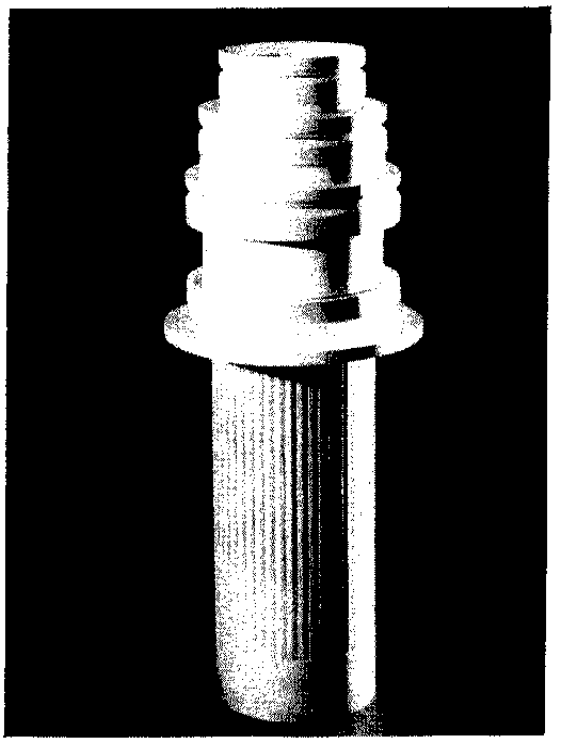


Figure 3—The *3ZC/262E* 40-kilowatt industrial triode.

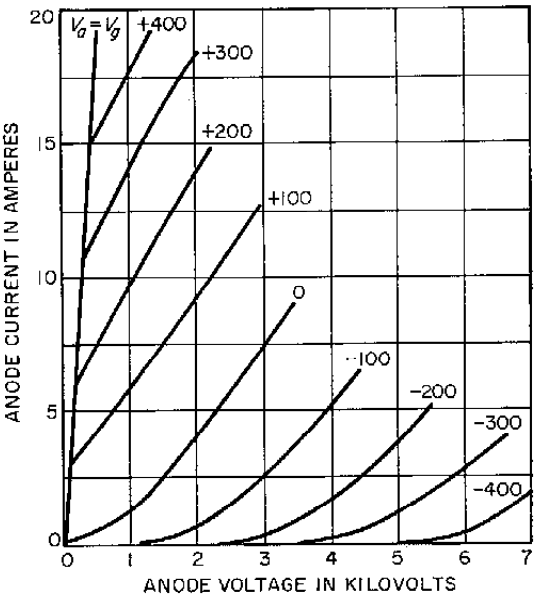


Figure 4—Anode current versus anode voltage for the indicated values of grid voltage for the *3JC/203E*.

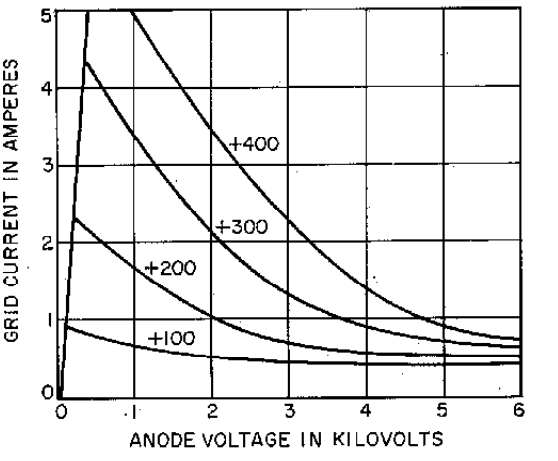


Figure 5—Grid current plotted against anode voltage for indicated grid voltages for the *3JC/203E*.

4.2 Efficiency

Using the symbols given in Section 10, the anode efficiency η is given by

$$\eta = \frac{(V - V_a)I_{1a}}{2VI_{da}}. \qquad (2)$$

For an anode-current flow angle of 140 degrees, which is representative of full-load operation, with a good compromise between adequate output power and efficiency [1], then $I_{1a}/I_{da} = 1{\cdot}77$, which substituted in (2), gives

$$\eta = 0{\cdot}885\,\frac{V - V_a}{V}$$

$$= 0{\cdot}885\left(1 - \frac{V_a}{V}\right). \qquad (3)$$

It can be seen from (3) that the maximum theoretical efficiency obtainable with a 140-degree flow angle is 88·5 per cent, when $V_a = 0$. In practice the anode "bottoming" voltage cannot be made as low as this because of the very-high grid dissipation that would result. However, the design principles that ensure a low value of grid dissipation with an adequate safety margin also allow a lower bottoming potential, and hence higher efficiency, than other available tubes of comparable power output. The *3JC/203E*, for example, can be operated with $V = 6$ kilovolts, $V_a = 570$ volts and $\phi = 140$ degrees, giving an anode efficiency of 80 per cent, with an adequate grid safety factor greater than 2:1. An even-greater efficiency could of course be obtained at the expense of power output and/or grid safety factor.

5. Filament Power and Emission

Among the earlier-listed requirements for industrial heating tubes is that the filament be able to operate without mains regulation and have a long emission life. These are, of course, conflicting requirements that must be weighed against each other in the design of the filaments.

5.1 Filament Temperature and Power

Thoriated-tungsten filaments in tubes for radio transmitting stations, where the filament voltage is normally controlled to better than ±1 per cent, are normally operated between 1950 and 2050 degrees kelvin. Emission increases with temperature, of course, for a given filament power, but a higher temperature shortens the emission life for a given wire diameter and carbide depth. To ensure adequate life within filament voltage variations of ±5 per cent these tubes were designed to operate at a low filament temperature at the nominal filament voltage. To ensure an adequate margin of emission under these conditions, it was calculated that the filament power must not be less than 38 watts per ampere of maximum useable emission. This adequate filament power also allows operation of the tubes with short-time variations of filament voltage as great as −10 per cent.

5.2 Carburization of Filaments

It is extremely difficult with high-frequency dielectric heating equipments, using voltage-step-up cavities to supply power to high-impedance loads, to prevent occasional circuit flashovers because of unfavourable atmospheric working conditions. In grounded-grid operation, external flashovers cause the anode voltage to appear momentarily on the grid, which can cause internal grid-to-filament flashovers, with consequent removal of the mono-molecular layer of thorium at the surface of the filament. Inability to replenish the thorium layer quickly can result in premature failure of the tube.

As is well known, a thoriated-tungsten filament is carburized to maintain at its surface a mono-molecular layer of thorium that, because of its low work function, provides the emission. The thorium results from the reduction of thoria in the following chemical equation:

$$2W_2C + ThO_2 \rightarrow 2CO + 4W + Th.$$

Several forms of tungsten carbide can exist, depending on the carburizing temperature and

pressure and on the nature of the carburizing gas. One of these forms of carbide reduces the thoria more readily than do the others. Hence, the carburizing conditions for the filaments of this range of tubes have been selected to provide this preferred carbide structure, so that the tubes can withstand the above-mentioned flashovers without loss of emission.

6. Mechanical Features

6.1 Envelope

A coaxial structure was chosen for its mechanical robustness and superior high-frequency performance over the more-conventional pin-type tubes. The coaxial structure leads to smaller dimensions and reduced inductances in the tube itself, thereby facilitating the use of these tubes in cavities for very-high-frequency operation. With exception of the *3ZC/262E*, all the tubes use the same stem parts, thus reducing tooling and jigging, multiplicity of parts, and costs. A cross-section of the *3JC/187E* showing details of the envelope structure is given in Figure 6.

The ceramic sub-assemblies, which are brazed in hydrogen at low pressure, are carefully jigged to ensure accurate concentricity and alignment of electrodes. No chemical cleaning is necessary after the brazing operations; thus, a major hazard of glass tubes, chemical residue leading to low emission, is eliminated.

The final seal-in of the tube is by argon-arc welding, thus avoiding any possibility of oxidization of the inside of the tube, which is a risk in glass-to-glass sealing.

If a dielectric-heating set employing a high-Q cavity is switched on without the load being connected, the peak radio-frequency voltage across the grid-to-anode ceramic can rise to several times the direct-current supply voltage. For this reason the metal parts at the grid and anode ceramic-to-metal seals have been carefully shaped to avoid field concentration and arcing across the anode-to-grid ceramic.

6.2 Filament

In all the tubes except the *3ZC/262E*, a mesh filament is used. It consists of a mesh of criss-crossed wires, wound on a cylindrical mandrel and welded together at the cross points. This type of filament resembles a uniformly emitting cylinder, thus enabling electrodes to be smaller than with more-conventional filaments, an important requirement for high-frequency operation. Another important feature is that the resistance to mechanical distortion is very great, which allows a smaller grid-cathode spacing. The filament is supported rigidly at one end, and at the other end is connected to a number of flexible tapes,

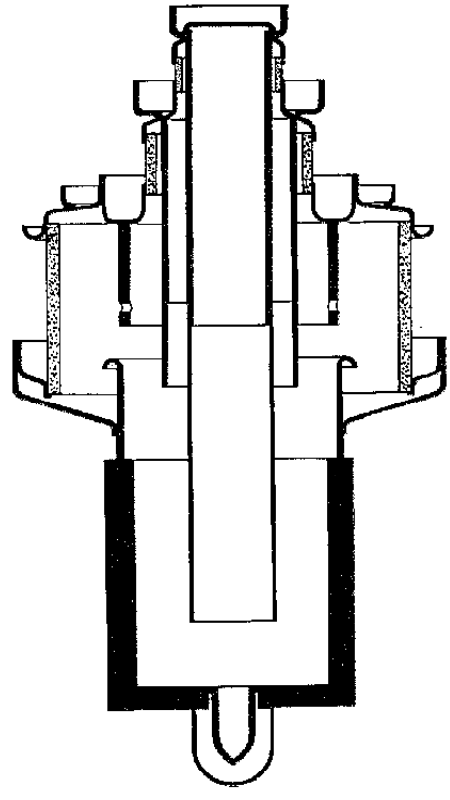


Figure 6—Cross section of the envelope structure of the *3JC/187E*.

as shown in Figure 7, which, in turn, are connected to the central tube that is used both for support and for conducting the filament current. The tapes allow expansion of the filament in the axial direction, thus minimizing mechanical strain.

The *3ZC/262E* has been designed with a long small-diameter anode to permit operation in a small boiler accommodated within a cavity. This shape does not readily lend itself to the mesh construction and so a more-conventional free-hanging filament is used with a slightly bigger grid-to-cathode spacing. Special design features allow for differential expansion between the individual filament strands and also prevent any barrelling of the filament in the radial direction. A special arrangement of the filament wires, in relation to the grid supports, allows an almost uniformly emitting cylinder to be obtained.

6.3 Grid

The grids of the mesh-filament tubes are made from expanded metal, specially processed to increase the thermal emissivity and to reduce grid electron emission to a very low value. This form of grid does not require grid supports; hence, it allows smaller electrode spacings.

The grid of the *3ZC/262E* is the conventional helically wound type, processed similarly to the expanded-metal grids. The grid and filament are mounted rigidly on a stem, but are fixed in a radial direction by a high-alumina bush that freely allows their differential axial expansion with respect to each other.

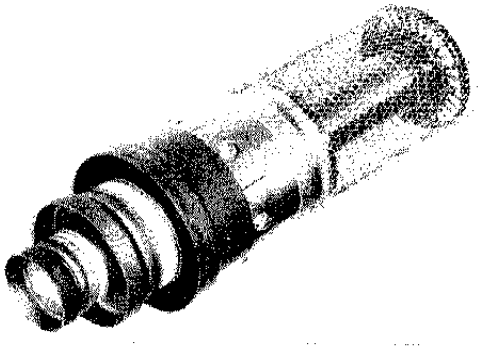


Figure 7—Grid-cathode assembly.

7. Cooling

Four different means of cooling are used.

(A) Air cooling (prefix J).

(B) Water cooling with normal water jacket (prefix Q).

(C) Water cooling with integral water jacket in the form of a coil hard soldered to the anode wall (prefix R).

(D) Vapour cooling (prefix Z).

All of the tubes, whatever the form of cooling, have thick-walled copper anodes to provide a large thermal overload capacity.

The air-cooled tubes have the fins brazed directly to the anode. This feature, coupled with the use of louvres on the fins to ensure turbulent flow, results in a high-efficiency radiator of minimum size. The *3JC/187E*, for example, requires only 50 cubic feet (1·42 cubic metres) per minute of cooling air at a water-gauge pressure of 0·4 inch (10·2 millimetres), for an anode dissipation of 3 kilowatts at an ambient temperature of 20 degrees centigrade.

Of the two forms of water cooling, the wrapped-coil version is finding increasing favour with manufacturers of industrial heating equipment. It is becoming the most favoured of all four cooling methods for medium-size equipments because of the very-small initial capital outlay for cooling apparatus, and also because of the small amount of cooling water required, an important feature where water is expensive. The required water flow is as low as that required for a vapour-cooled tube. The *3RC/223E*, for example, requires only 1 gallon (3·79 litres) per minute at a pressure of 7·5 pounds per square inch (0·53 kilogrammes per square centimetre) for 12 kilowatts of anode dissipa-

tion at a water inlet temperature of 20 degrees centigrade.

The advantages of vapour cooling are well known and will not be enumerated here.

The vapour-cooled versions of these tubes have vertical grooves machined into the anodes. This results in a large increase in the anode dissipation before heating occurs, and permits a much-greater anode dissipation than do the other water-cooled and the air-cooled versions. However the rated maximum anode dissipation of the vapour-cooled tubes has not been increased, and the advantage of vapour cooling is passed on to the customer in the form of an anode dissipation safety factor of over twice that of the normal safety factor for water- and air-cooled types.

8. Conclusions

The experience gained from the operation of a large number of glass tubes in a diverse range of industrial heating equipments, operating under many different adverse conditions, has been used in the design of a new range of improved ceramic-insulated tubes for this service.

Their high efficiency, generous safety factors, and high maximum frequency of operation provide a significant advance to the designer of very-high-frequency dielectric heating sets with output powers ranging from 5 to 40 kilowatts.

9. Acknowledgments

The author is greatly indebted to a large number of people who helped in the successful development of this range of tubes, especially to Messrs. H. M. Jenkins, G. H. G. Phipps, and A. R. Manners for their enthusiastic support and to Mr. D. C. Rogers for helpful criticism and encouragement.

10. Symbols

d = diameter of grid wire
I_a = peak instantaneous anode current per unit area
I_{da} = anode direct current per unit area
I_{dg} = grid direct current per unit area
I_g = Peak instantaneous grid current per unit area
I_k = Peak instantaneous total cathode current per unit area
I_{1a} = peak value of fundamental component of anode current per unit area
l_a = grid-to-anode distance measured to the centre of the wires
l_g = grid-to-cathode distance measured to the centre of the wires
p = pitch of grid wires
P_g = total grid dissipation per unit area of grid
R_b = grid-bias resistance
V = anode supply direct voltage
V_a = minimum instantaneous anode voltage relative to cathode
V_b = bias voltage
V_g = peak value of instantaneous grid voltage relative to cathode
V_s = peak value of grid voltage relative to bias voltage
η = efficiency
μ = triode amplification factor
ϕ = angle of grid-current flow in electrical degrees.

11. References

1. F. E. Terman, "Radio Engineering," McGraw-Hill Book Company, New York: 1932; pages 325 and following.

2. J. H. Fremlin, "Calculation of Triode Constants," *Electrical Communication*, volume 18, pages 33–49; July 1939.

3. J. H. Fremlin, R. N. Hall, and P. A. Shatford, "Triode Amplification Factors," *Electrical Communication*, volume 23, pages 426–435; December 1946.

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12. Appendix, Design Considerations

To establish design principles, expressions are required that relate tube electrode dimensions to the peak currents and voltages.

12.1 General Equations

An accurate expression [2] for the total space current is

$$I_b = \frac{2{\cdot}34 \times 10^{-6}\left(V_g + \frac{V_a}{\mu}\right)^{3/2}}{\left[l_g^{1/2} + \frac{(l_a + l_g)^{3/2}}{\mu}\right]^2}, \quad \text{amperes per square centimetre} \qquad (4)$$

where voltages are expressed in volts and spacings in centimetres. Symbols are as given in Section 10.

For the grid current, the approximate relationship for a parallel-wire grid is

$$I_g = I_a\left(\frac{l_g + l_a}{l_g}\right)^{3/2}\frac{d}{p-d}\left(\frac{V_g}{V_a}\right)^{1/2}. \qquad (5)$$

The grid flow angle ϕ is given by

$$\cos\frac{\phi}{2} = \frac{V_b}{V_s} = \frac{V_b}{V_b + V_g}. \qquad (6)$$

A commonly used approximate expression [3] for μ is

$$\mu = \frac{2\pi l_a}{p\log_e\left(2\sin\frac{\pi d}{2p}\right)}. \qquad (7)$$

Substituting (7) for μ in (4) and since in practice l_a is several times larger than l_g, then

$$I_b = \frac{2{\cdot}34 \times 10^{-6}\left(V_g + \frac{V_a K}{l_a}\right)^{3/2}}{[l_a^{1/2} + l_a^{1/2}K]^2}, \quad \text{amperes per square centimetre} \qquad (8)$$

where $K = \dfrac{p\log_e\left(2\sin\frac{\pi d}{2p}\right)}{2\pi}$.

12.2 Grid Dissipation

An expression [4] for the grid dissipation is

$$P_g = \frac{\phi}{4\pi} V_g I_g. \qquad (9)$$

One of the important tube requirements for industrial heating is a high grid safety factor, that is that P_g should be small.

From (9) it is evident that, to obtain this desirable feature, ϕ, V_g, and I_g should all be as small as possible.

The various electrode spacings and dimensions will now be examined to discover how the above requirements can best be met.

12.3 Grid-Anode Spacing

From (8) it is evident that for a given space current and current division l_a should be made small to keep V_g as low as possible. In other words the amplification factor should be made as small as possible consistent with an adequate safety factor as regards freedom from flashover. It might be argued that the resulting low amplification factor would necessitate a high drive power. This, however, is not true for the range of tubes described in this article because the drive power increases not only with a decrease in μ, but also with increases in both I_g and V_g, and the design features of these tubes ensure that V_g and I_g are small.

12.4 Grid-Cathode Spacing

Examination again of (8) shows that, for a given space current, l_g also should be made small to keep V_g as low as possible. In other words, the mutual conductance of the tube should be made as high as possible consistent with mechanical accuracy and stability of the filament and grid structures and supports.

It can be seen from (5) that for a given anode current, I_g also decreases as V_g is made smaller by achieving a low amplification factor and a high mutual conductance.

Further, from (6) it can be seen that a low value of V_g also results in a low angle of grid current flow ϕ.

Thus the above design considerations satisfy the requirement of low grid dissipation as shown in (9), that is, that V_g, I_g, and ϕ should all be small.

12.5 Grid Pitch

The larger the number of grid wires, the higher is the value of intercepted grid current for a given space current, resulting in a smaller anode current and, hence, power output. For a given amplification factor the grid safety factor is not affected because additional wires can dissipate additional power. The grid pitch should therefore be large, not only for the above reason, but also because this results in a smaller amplification factor, the benefits of which have been explained in 12.3 above. There is a limit, though, to increasing the grid pitch beyond a certain value because of the onset of "island" effects and "long tail" characteristics leading to low efficiencies. A grid pitch equal to 1·4 l_g is a good compromise [2].

13. Appendix, Grid Temperature and Emission

13.1 Runaway Conditions

The bias for an industrial heating oscillator is normally derived by the grid current I_{dg} flowing through a grid resistance R_g.

If in an oscillator an increase in the radio-frequency grid voltage $\int V_g$ is postulated, then resulting changes in the grid circuit will be:

(A) An increase in the normal grid current $\int I_{dg}$.

(B) An increase in the grid dissipation $\int P_g$.

(C) An increase in the grid emission current $\int I'_{dg}$.

The change in the bias voltage V_b is therefore given by

$$\int V_b = R_g[\int I_{dg} - \int I'_{dg}] \qquad (10)$$

and the tube will run away if

$$-R_g[\int I_{dg} - \int I'_{dg}] > \int V_g. \qquad (11)$$

If now a graph is plotted of $R_g[I_{dg} - I'_{dg}]$ against V_g with both ordinates to the same scale, then the tube will run away when the slope at that point is negative and greater than 45 degrees. It is interesting to note from this, that it is the slope of the grid emission curve that is important, not the absolute value of the grid emission. Since emission is related to temperature in such a way that the emission slope increases rapidly with temperature in accordance with the well-known emission law

$$I_g = A_0 T^2 \exp[-b_0/T] \qquad (12)$$

it is therefore very important that the grid temperature should be as low as possible.

13.2 Grid Temperature

Assuming that the power conducted along the grid and filament supports is small compared to the power radiated from the grid and filament to the anode, it can be shown that the grid temperature T_g is fairly accurately given by the expression

$$T_g^4 = \frac{P_g}{e_1 \sigma A_2} + \frac{P_f S}{\sigma A_2} + \frac{(P_g + P_f)(1 - e_2)S}{\sigma A_2 e_2} \qquad (13)$$

where A_2 = total surface area of grid

e_1 = total thermal emissivity of the grid

e_2 = total thermal emissivity of the inside surface of the anode

P_f = filament power

P_g = power received by the grid by electron bombardment
S = shadow area ratio of grid
σ = Stefan-Boltzmann constant.

Examination of (13) shows that the thermal emissivity of both the grid and anode surfaces should be as high as possible in order to keep the grid temperature low.

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He joined Standard Telephones and Cables, New Southgate, in 1952, working on the development of high-power transmitters. Mr. Behenna, in 1957, transferred to the valve division at Paignton, and is now in charge of the high-power vacuum-tube development laboratory.

Travelling-Wave Tubes for 6-Gigacycle-Per-Second Radio Links

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The recent development of the communication bands around 6 gigacycles per second has led to the design of two travelling-wave tubes for radio links. One of these, the *W5/1G*, operates at 3000 volts and 40 milliamperes for an operational output power of 5 to 8 watts. For 1800-channel systems of conventional design, the *W5/2G*, with an operating power output of 10 to 12 watts, is available. The required radio-frequency input power is about 1 milliwatt. Other important radio-frequency characteristics are noise factor, spurious noise, and both input and output matching.

A fairly high efficiency is necessary and is achieved by using a collector potential less than the helix potential. The reduced collector dissipation permits natural convection cooling. A periodic-permanent-magnet structure [1] requires no power for focusing the electron beam and is relatively small. The amplifiers comprise two parts: firstly, a mount with the waveguide feeds and magnets and, secondly, the travelling-wave tube. All parts subject to wear or failure are contained in the tube, which is easily replaced.

1. Design and Construction

The dimensions and other parameters of the tubes are given in Table 1 and tube and mount are shown in Figure 1. The helix is supported

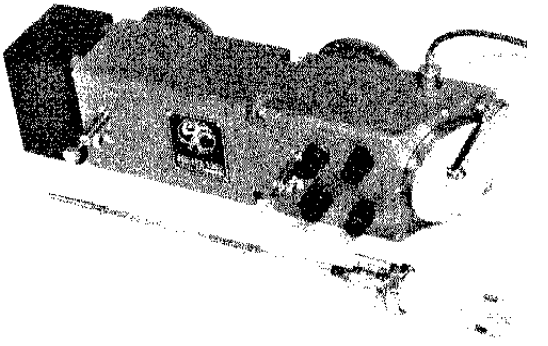


Figure 1—Travelling-wave tube *W5/1G* and mount.

TABLE 1
DIMENSIONS AND PARAMETERS OF TRAVELLING-WAVE TUBES

	W5/1G	*W5/2G*
Frequency Range in Gigacycles Per Second	5·85–7·05	5·85–6·45
Heater, Volts	6·3	6·3
Heater, Amperes	0·7	0·7
Grid-2 Volts, Approximate	2000	2300
Helix Volts, Approximate	2900	3400
Collector Volts	2000	2000
Collector Milliamperes	40	50
Helix Milliamperes, Approximately	0·5	0·5
Helix Inner Diameter, Inches (Millimetres)	0·088 (2·24)	0·088 (2·24)
Helix Wire Diameter, Inches (Millimetres)	0·012 (0·3)	0·012 (0·3)
Helix Pitch, Inches (Millimetres)	0·034 (0·86)	0·0345 (0·88)
Helix Length, Inches (Millimetres)	6·4 (163)	6·4 (163)
c/v	9·4	8·7
γa	1·62	1·5
F	0·39	0·39
Dielectric Loading Factor,* Per Cent	80	80
C	0·051	0·054
QC	0·2	0·2
Theoretical Synchronous Saturated Gain, Decibels	42	40
Helix Wire Loss, Decibels	6	6
Central Loss, Decibels	70	70
Cathode Diameter, Inches (Millimetres)	0·2 (5·1)	0·2 (5·1)

* Calculations based on Pierce [2] and Tien [3].

by 3 insulating rods with a carbon deposit to give a central attenuating region. A fusible link enables the helix to be outgassed by passing current through it on the pump. The helix pitch must be accurately maintained; the turn-to-turn spacing is checked with an image-splitting eye-piece [4] ensuring that errors are less than 1 per cent of the pitch. The electron gun is assembled on ceramic rods and is supported between the helix assembly and a metal ring at the base of the tube. The first grid of the electron gun is connected internally to the cathode. The beam current is controlled by the second grid. The third grid, also connected internally to the helix, has an external connection in the base ring. Ceramic lead-through insulators provide for connections to the electrodes and permit pumping through a relatively large copper tubulation in the metal base. Lead-throughs and tubulation are protected by potting in Silastomer after final processing of the tube. Individual components are heated in vacuum whenever possible before assembly to minimize the evolution of gas during life.

2. Mount

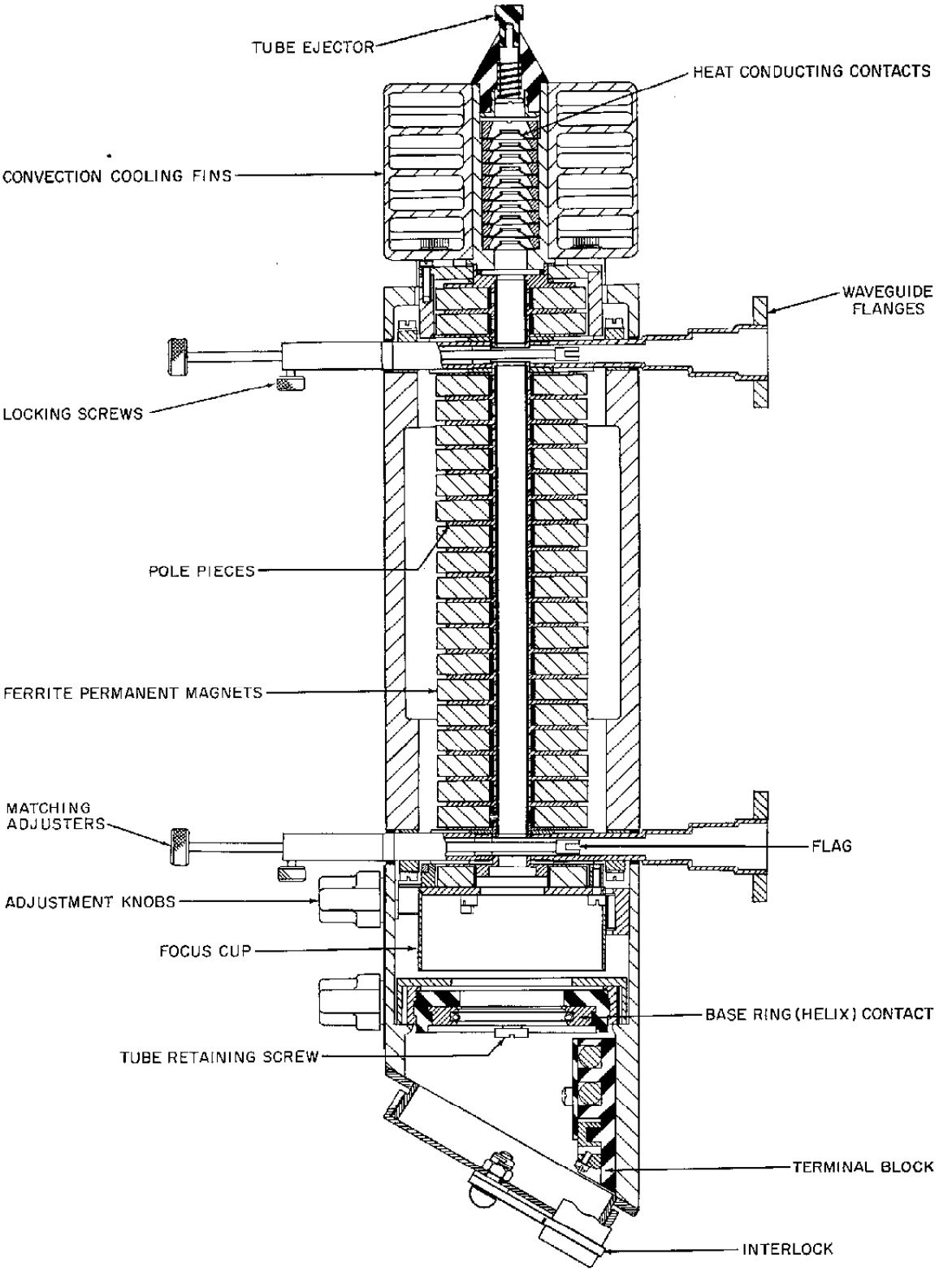
The mount of the *W5/1G* is shown diagrammatically in Figure 2.

Adjacent permanent magnets have opposite polarities, greatly reducing the leakage field and the total flux compared with that which would be obtained from a magnet providing a uniform field. The volume and weight are reduced accordingly. Short magnets are necessary to avoid defocusing due to resonant interaction with the electron beam and hence the pole-pieces need to have a small internal diameter to give the required field. This resonance will still occur at low voltages and leads to a rapid deterioration in the focusing at voltages appreciably below the normal operating voltage. Bar magnets, not shown in the diagram, provide the field across the waveguides and are mounted with the central magnet stack in an aluminium casting of rectangular external cross-section. The cooler, attached so as to conduct little heat back to the mount, has cooling fins of aluminium with a total surface area of approximately 120 square inches (774 square centimetres). Good thermal contact between the rigidly mounted cooler and the collector of the tube is provided by approximately 80 spring fingers pressing flexibly against the collector. Silver-bearing copper, 0·010 inch (0·25 millimetre) thick, is used since it does not set under stress at temperatures below 200 degrees centigrade. Since each finger gives two or three points of contact the total heat drop between the collector and the interior of the cooler is less than that obtained with a normal mechanical fit between the collector and a solid cooler. For 80 watts of dissipation, this temperature drop is about 10 degrees centigrade and the cooler itself runs about 70 to 80 degrees centigrade above ambient temperature. The fins of the cooler are staggered to increase the turbulence of the ascending air and side pieces are incorporated to give a chimney effect. The tube can be freed from the grip of the spring contacts by an ejector that applies pressure to its end.

The base ring of the tube is held at the other end of the mount in a ring whose position can be altered by two knobs to correct for any slight errors in alignment of the base. Two other knobs move the first pole-piece to give variable transverse magnetic fields in the region of the electron gun. Both adjustments are limited to prevent any misalignment that would damage the tube. Electrical adjustment of the current in special deflector coils is an alternative to movement of the pole-piece.

The waveguide in the mount is 1·375 by 0·25 inch (3·5 by 0·64 centimetre) with a binomial matching section to *WG137* dimensions. Some fine trimming is necessary to obtain the best match at both input and output either with an adjustable flag, as shown in Figure 2, or with

Figure 2—Structural details of the mount. ⟶


TUBE EJECTOR
HEAT CONDUCTING CONTACTS
CONVECTION COOLING FINS
WAVEGUIDE FLANGES
LOCKING SCREWS
POLE PIECES
FERRITE PERMANENT MAGNETS
MATCHING ADJUSTERS
FLAG
ADJUSTMENT KNOBS
FOCUS CUP
BASE RING (HELIX) CONTACT
TUBE RETAINING SCREW
TERMINAL BLOCK
INTERLOCK

four matching screws between the waveguide flange and the mount.

The flexible leads from the tube are connected to screwed terminals just inside the mount. Resistors are provided in the helix and grid-2 leads from these terminals to limit surges in the unlikely event of momentary voltage breakdown in the tube. The leads have polytetrafluoroethylene insulation and are taken, via a ferrite that absorbs any radio-frequency energy, through a cable whose outer braid is connected to the mount and should be earthed.

A lid, which screws to the mount, gives good radio-frequency screening. On removing the lid, a plug swinging on an attached arm is disconnected from its socket in the mount and removes all high voltages.

The higher-power tube *W5/2G* uses a slightly modified mount incorporating mechanical adjustment of the tube position at the collector end.

3. Electron Beam Focusing

The magnetic field distribution in the periodic structure is shown in Figure 3. Slight increases in period at the waveguides avoid the use of extremely narrow guides. A gradual increase in magnetic field over the last four periods before the output waveguide compensates for defocusing effects at maximum power output.

The values of the first two peaks of magnetic field were found experimentally to give best focusing. The magnetic field gradually falls in the region of the electron gun until it is zero at the cathode.

The ferrite permanent magnets used have an appreciable temperature coefficient but the helix current does not rise by more than 50 per cent over the temperature range from −10 to +60 degrees centigrade. The materials and the working point of the ferrite are such that exposure to temperatures of −60 degrees centigrade gives no irreversible change in the magnetic field.

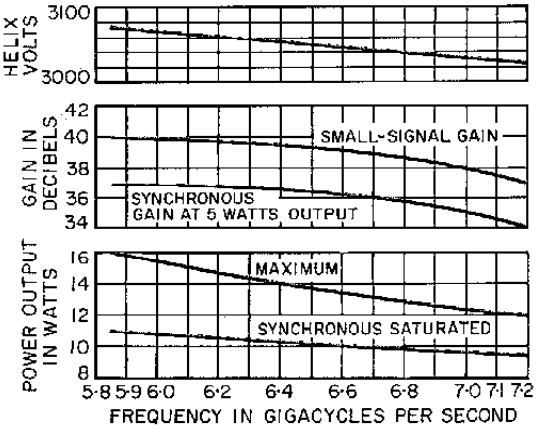


Figure 4—Gain, power output, and helix voltage for synchronous operation as a function of frequency for the *W5/1G*.

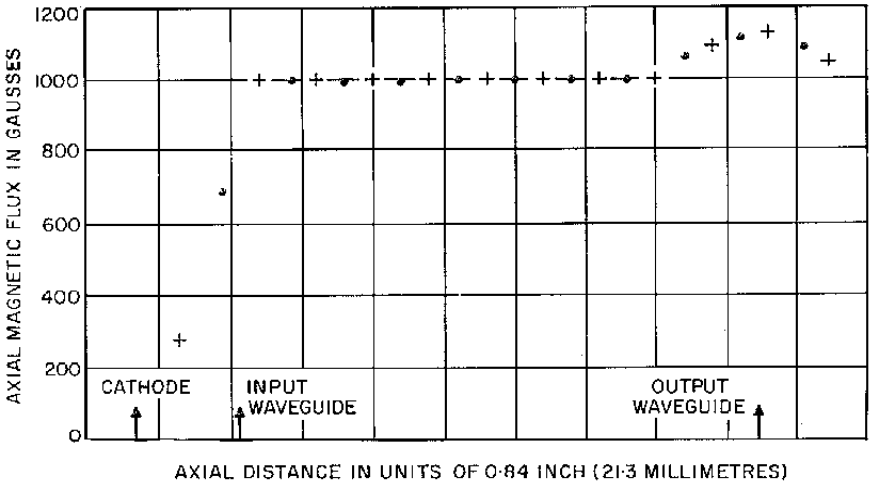


Figure 3—Magnetic field distribution in the periodic-permanent-magnet structure.

Periodic focusing is not effective at low voltages, which means that precautions are necessary on first switching on. If the beam current is held off by keeping the second grid at cathode potential until the desired helix voltage is set up, the current surge as V_{g2} rises to the normal operating level will be less than 10 milliamperes. If it is kept a constant fraction of the helix voltage, the current surge may be 15 milliamperes. No excessive voltage drop, due to power supply impedance, should occur with these surges.

Measurements of helix current at maximum power output on a considerable number of tubes and mounts gave a figure of less than 0·5 milliampere for half the results and 1 milliampere was exceeded in only 2 per cent of the cases. The normal operating maximum in field use is specified as 1·5 milliamperes initially and the supplies should be removed if 2·5 milliamperes is exceeded.

4. Radio-Frequency Performance

4.1 Gain and Power Output

Gain and power output are shown as functions of frequency for the *W5/1G* and *W5/2G* in Figures 4 and 5. Typical graphs of output power versus input power for three different helix voltages at a frequency of 6·4 gigacycles per second are shown in Figure 6 for the *W5/2G*. Higher helix voltages give greater saturated output but the tubes are designed

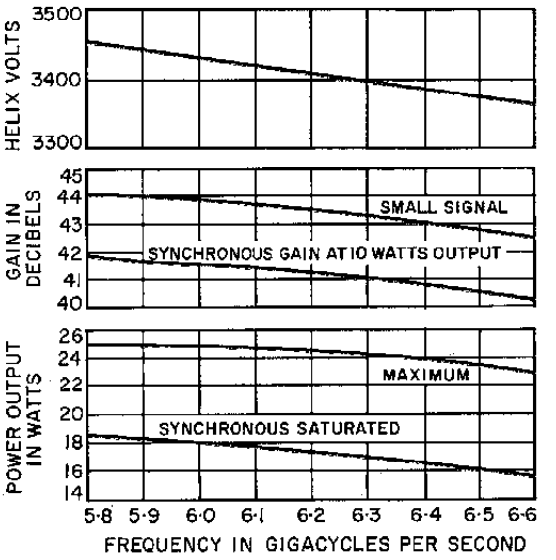


Figure 5—Gain, power output, and helix voltage for synchronous operation as a function of frequency for the *W5/2G*.

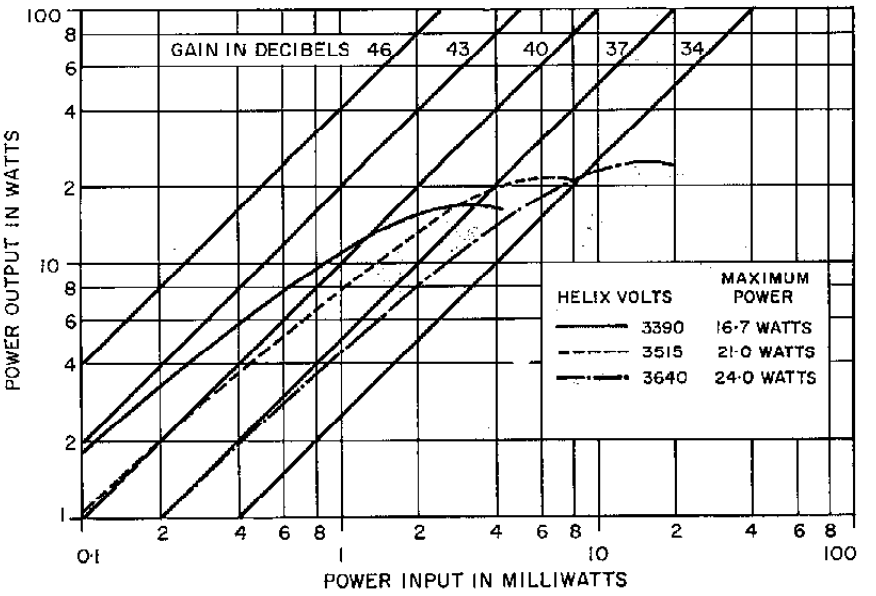


Figure 6—Output power plotted against input power for *W5/2G* at 6·5 gigacycles per second.

to maximize the power obtained at the helix voltage giving maximum low-level gain, so as to give less distortion from the conversion of amplitude modulation to phase modulation.

A slight fall in the saturated output power may be seen sometimes immediately after switching on the radio-frequency input. This lasts a minute or so and is due to parts of the tube increasing in temperature and in radio-frequency loss. For the *W5/1G* this fading gives a 12-per-cent power drop at a saturated output of 15 watts and 10-per-cent at 10 watts but at the same power level the *W5/2G* has a drop of only from 15·5 watts to 15 watts, perhaps due to the use of fused silica rods instead of steatite rods as in the *W5/1G*.

4.2 Signal–to–Noise Ratio

The noise factor of both tubes lies in the range of 26 to 28 decibels. Experiment has shown this to be independent of the exact magnetic field at the cathode over the range from 0 to 30 gausses, the discrepancy with earlier reports of high noise factor at zero cathode fields [5] being due perhaps to the way the field extends close to the cathode and controls the beam over almost all of its path.

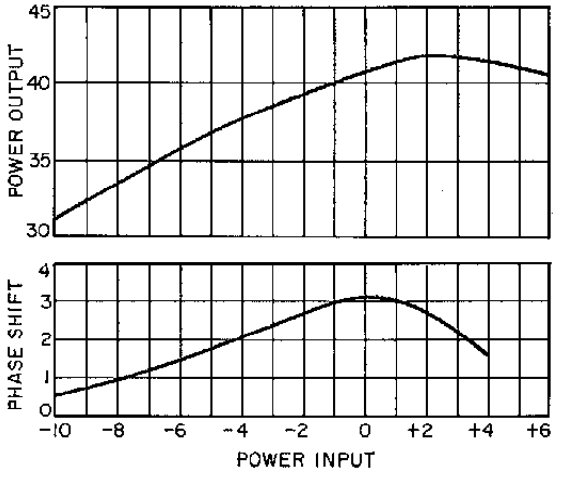


Figure 7—The effect of power input on power output, both in decibels referred to 1 milliwatt, and on phase shift in electrical degrees per decibel of input power are plotted above for the *W5/1G*.

The modulation noise due to ion-plasma oscillations has been studied and considerably reduced by attention to processing of the tubes. Sidebands between 20 kilocycles per second and 100 kilocycles per second wide at frequency distances around 3 megacycles per second and 10 megacycles per second from the carrier may initially have an amplitude of 10 to 20 decibels above the normal noise level but this falls rapidly during activation and testing to a level about 5 decibels above the shot noise. Improvement continues until no noise is detectable, although during the first 50 hours slight alterations in the operating conditions may make it perceptible again.

4.3 Leakage and Attenuation

Screening is good enough to ensure at least 65 decibels of attenuation of leakage between either of the waveguides and a radiating waveguide horn held in any position 2 inches (5 centimetres) or more from the mount.

Reverse attenuation, measured from tube output with the tube running at normal voltage and current, exceeds 70 decibels, effectively the same as the cold attenuation. Some reflected electrons from the collector decrease this value slightly with the helix about 200 volts below the normal operating value, corresponding to synchronism with a reflected electron beam of very low current.

4.4 Matching

For best results, matching adjustments should be made at the operating frequency when a new tube is inserted in a mount, giving a bandwidth of greater than 15 megacycles per second for a reflection coefficient of 5 per cent looking into the output of the tube, the limit being set by reflections at irregularities in the helix, which are amplified by the electron beam. At the operating level the gain is lower and the bandwidth improved. The bandwidth of the input match is considerably better, because of the lower gain before the attenuation region.

Alternatively, mounts can be pre-set for oper-

ation in any specified 600-megacycle-per-second range to give a voltage standing-wave ratio of less than 1·6 on all tubes without adjustment.

4.5 Phase Characteristics

The electrical length of the *W5/1G* is 30 wavelengths at 6 gigacycles per second and the *W5/2G* is 28 wavelengths long. Phase sensitivity is about −0·75 degree per volt change of helix potential and +0·125 degree per volt change of the second-grid potential at the normal operating power level. No measurable variation with collector voltage exists. The phase shift also varies with input power because of the slowing down of the electron beam as more power is taken from it. The differential of this quantity, expressed in degrees per decibel, is shown in Figure 7 for the *W5/1G* at synchronous voltage. At higher voltages, giving increased output, the maximum value can be 6 degrees per decibel.

4.6 Harmonic Output

Non-linearity near maximum power output produces many harmonics from travelling-wave tubes although this output may vary considerably since the impedance match at the harmonic frequencies will be generally bad and erratic. Second-harmonic power from the *W5/1G* at saturation output can be as much as 50 to 100 milliwatts.

5. Life

Cathode-current density is a most-important factor in determining life. Previous designs of medium-power travelling-wave tubes have given average lives of 7000 to 8000 hours at a cathode-current density of 400 milliamperes per square centimetre, and around 20 000 hours at 200 milliamperes per square centimetre. The *W5/1G* operates at this lower figure and the *W5/2G* at a value about 20 per cent higher. Hence a life of 15 000 to 20 000 hours is expected though, as this represents several years of running, it has not been verified yet.

Automatic control of the cathode current by the second-grid voltage is possible as with previously developed tubes. This control may not be necessary if maintenance visits are possible every few weeks since these will permit manual adjustments during the last part of the tube life.

The cathode temperature should be carefully controlled by regulating the heater voltage to within ±3 per cent of the recommended value of 6·3 volts, although short-term excursions as large as ±5 per cent are permissible. This establishes the optimum condition between any accelerated aging at high temperatures and irreversible poisoning at low temperatures.

6. Conclusions

Two travelling-wave tubes suitable for use in 6-gigacycle-per-second links have been developed. Although primarily developed for a superheterodyne type of microwave repeater, they can be used in other types as well [6]. Tubes giving similar performance in the adjacent higher frequency bands up to 8·5 gigacycles per second are in an advanced state of development.

7. Acknowledgments

This development could not have been successful without considerable and unstinted efforts from all of those concerned. Their numbers necessitate their anonymity, and the author's departure from this is solely an acknowledgment of responsibility for the shortcomings of this paper.

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In 1948, he joined Standard Telephones and Cables Limited and is engaged in the development and engineering of travelling-wave tubes.

Mr. Burke is a graduate member of the Institution of Electrical Engineers.

Effect on an Electron Beam of Variations in Periodic Permanent-Magnet Focusing Systems *

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The theory and design of a periodic permanent-magnet focusing system are based on the assumption that both the period and peak magnetic field are constant. This ideal, however, is never attained in practice. Non-uniformities in the period or magnetic field may be due to various tolerances or to the mechanical design of the circuit. On the whole, undesirable variations tend to increase the beam ripple which, in the case of the travelling-wave tube, may lead to loss of gain, increased frequency sensitivity, and high intercepted current.

In this paper, the effects of various practical non-uniformities in the period and magnetic field are examined by calculating the appropriate beam profiles.

The effect of small random misalignments of magnetic cells is also considered, and a simple expression is given for the total beam displacement from the intended circuit axis.

1. Beam Profiles for Non-Uniform Period or Magnetic Field

1.1 Outline of Analysis

The beam profiles were obtained by solving numerically Harker's equation [1] with a more-general expression for the axial magnetic field.

$$\frac{d^2\sigma}{dx^2} + 2\alpha B^2\sigma - \frac{2\alpha K}{\sigma^3} - \frac{\beta}{\sigma} = 0 \qquad (1)$$

where $\alpha = \dfrac{\eta L^2 \hat{B}_o^2}{64\pi^2 V}$

$$\beta = \frac{\rho_o L^2}{16\pi^2 \epsilon_o V}$$

$$K = \left(\frac{r_c}{r_o}\right)^4 \left(\frac{B_c}{\hat{B}_o}\right)^2$$

$$\sigma = r/r_o.$$

The suffix c indicates the values at the cathode plane, and o is the values at the plane of entry into the focusing circuit. V is the beam voltage, ρ is space-charge density, and r is the beam radius.

Equation (1) is derived by starting with three equations of motion in cylindrical co-ordinates. The radial force due to space charge is obtained from Gauss's theorem and this is reasonably correct provided that the envelope slope is less than, say, 0·15. Since the beam moves in a unipotential tunnel, the axial acceleration is negligible and the z equation can be ignored. With cylindrical symmetry in the magnetic field, the θ equation can be integrated, leading finally to (1).

Instead of commonly used

$$B = \cos x$$

the axial magnetic field is written here as

$$B = \bar{B}_n \cos\{l_n(x - X_n) + (n - 0{\cdot}5)\pi\} \qquad (2)$$

where $X_{n+1} = \dfrac{\pi}{l_n} + X_n$

$$X_o = -(\pi/2l_o)$$

$$l_n = L/L_n$$

$$\bar{B}_n = B_n/\hat{B}_o$$

and

L = optimized period of a uniform stack

L_n = period of the nth magnetic cell in a non-uniform stack

n = magnetic-cell reference number

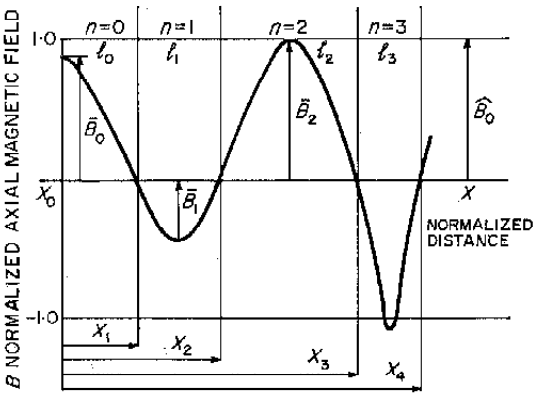
$\hat{B}_o$ = optimized peak magnetic field of a uniform stack

B_n = peak magnetic field of the nth magnetic cell in a non-uniform stack.

Equation (2) represents a cosine curve, which may have an arbitrary amplitude and/or period in any interval between two zero values, as shown in Figure 1.

* Presented in part at the International Congress on Microwave Tubes at Scheveningen, Holland, in 1962.

Figure 1—Cosine function of variable period and amplitude distribution of magnets.



The usual initial conditions were used, that is at $x = 0$, $\sigma = 1$ and $d\sigma/dx = 0$. Unless otherwise stated, α and β are the optimized values.

1.2 Random Variations in Period

Small random variations in the period of a focusing circuit are caused by the tolerances in magnets, pole-pieces, and circuit assembly. They will also occur when the magnetic-field distribution does not coincide with the cell geometry.

Variations of ± 1 per cent about the optimized period L were considered. Beam profiles were calculated for three different assemblies, each consisting of 15 magnets. The periods expressed in terms of l_n are given in Table 1, where l_n is the ratio of optimized period to actual period; and the maximum increase in beam radius is given in Table 2, where $\Delta\sigma$ is the maximum percentage increase in the normalized beam radius above unity.

It is interesting to note that for small values of α and β the increase in beam radius is of no practical importance.

1.3 Random Variations in Peak Magnetic Field

The field within a magnetic cell of an assembled periodic-permanent-magnet focusing stack is the result of the superposition of the fields in a cell alone, and of the stray fields of the neighbouring cells. Due to inhomogeneities in magnets and pole-pieces, which cause variations in stray fields and in the reluctance of the various paths, the peak magnetic field in an assembled stack is seldom absolutely uniform.

Again, as in the last section, three stacks were considered. As before, each stack had 15 magnets, but this time the random variation was obtained from actual measurement on experimental circuits. Table 3 gives the

TABLE 1
Variations in the Period

l_n	Assembly		
	1	2	3
l_0	1	1	0·9901
l_1	1·01	1·01	1·01
l_2	1	1·01	1·01
l_3	0·9901	1	0·9901
l_4	0·9901	1·01	1·01
l_5	1·01	0·9901	1
l_6	0·9901	1·01	1
l_7	1·01	0·9901	1
l_8	0·9901	0·9901	0·9901
l_9	1·01	1	0·9901
l_{10}	1·01	1·01	1
l_{11}	0·9901	0·9901	1
l_{12}	0·9901	1·01	1
l_{13}	0·9901	1·01	1
l_{14}	1	1·01	1

TABLE 2
Per Cent Increase in Normalized Beam Radius for Period Variation in Table 1

K	α	β	$\Delta\sigma$ in Per Cent		
			Assembly 1	Assembly 2	Assembly 3
0	0·1	0·0959	0·23	0·15	0·12
	0·3	0·2630	2·1	2·2	1·4
	0·5	0·3893	12·6	5·7	5·0
0·1	0·1	0·0749	0·33	0·35	0·22
	0·3	0·1857	2·3	2·9	1·2
	0·5	0·2129	5·1	4·6	5·1

TABLE 3

VARIATIONS IN THE PEAK MAGNETIC FIELD

Normalized Peak Magnetic Field $\bar{B}_n$	Assembly		
	1	2	3
$\bar{B}_0$	0·966	1·050	1·037
$\bar{B}_1$	0·915	0·973	0·954
$\bar{B}_2$	1·034	0·990	1·037
$\bar{B}_3$	0·983	0·994	1·037
$\bar{B}_4$	1·051	1·006	0·986
$\bar{B}_5$	1	0·99	1·012
$\bar{B}_6$	1·027	1	0·981
$\bar{B}_7$	1·017	1·017	0·996
$\bar{B}_8$	1·017	1·006	0·964
$\bar{B}_9$	1·003	1·006	1·020
$\bar{B}_{10}$	1·027	1·006	0·986
$\bar{B}_{11}$	1·017	1·006	1·008
$\bar{B}_{12}$	1·010	1·013	0·974
$\bar{B}_{13}$	1·003	1·006	1·06
$\bar{B}_{14}$	0·949	0·957	0·981

TABLE 4

PER CENT INCREASE IN NORMALIZED BEAM RADIUS FOR FIELD VARIATION IN TABLE 3

K	α	β	$\Delta\sigma$ in Per Cent		
			Assembly 1	Assembly 2	Assembly 3
0	0·1	0·0959	12·3	5·4	9·3
	0·3	0·2630	39·3	13·6	30·8
	0·5	0·3893	27·8	37·3	38·6
0·1	0·1	0·0749	13·1	6·2	8·7
	0·3	0·1857	85·3	13·2	42·8
	0·5	0·2129	27·0	11·5	44·2

normalized random values of the peak magnetic field, and Table 4 gives the corresponding increases in the beam radius.

For the comparison with a 1-per-cent variation in the period, beam profiles were also calculated for a 1-per-cent variation in the peak magnetic field. The values of $\bar{B}_n$ are now identical with l_n in Table 1, Assembly 1.

The relatively large percentage increases in the beam radius shown in Tables 4 and 5 are somewhat unexpected, as it is sometimes

TABLE 5

PER CENT INCREASE IN NORMALIZED BEAM RADIUS FOR A 1-PER-CENT PEAK MAGNETIC-FIELD VARIATION IN TABLE 1 ($l_n - \bar{B}_n$, ASSEMBLY 1)

K	α	β	$\Delta\sigma$ in Per Cent
0	0·1	0·0959	7·7
	0·3	0·2630	20·1
	0·5	0·3893	28·9
0·1	0·1	0·0749	5·4
	0·3	0·1857	12·3
	0·5	0·2129	9·2

believed that relatively large variations in peak magnetic field are quite tolerable.

It was thought that this discrepancy was caused by the operating α being always larger than its theoretical optimized value. For this reason some of the above computations were repeated with $\alpha = 2{\cdot}5\ \alpha_{opt}$; the lower curve in Figure 2 appears to confirm this. The same reasoning should apply to the effect of variations in the period.

1.4 SINGLE INCREASE IN PERIOD

The radio-frequency coupling to a travelling-wave tube can be obtained via a helical coupler, a radial or co-axial cavity, or by a waveguide transducer. Unlike a helical coupler, which may be small enough to be inserted inside the pole-pieces, a waveguide or cavity must be inserted between the magnets.

It is possible to retain the correct field across the waveguide, either by using additional bar magnets or utilizing the stray fields from the adjacent ring magnets.

The height of a standard waveguide section is usually much greater than half of the necessary period, and consequently, to avoid an increase in the period, the waveguide is stepped down. Very often drastic reduction in the waveguide height makes radio-frequency matching unsatisfactory; and then an increase in the period is unavoidable.

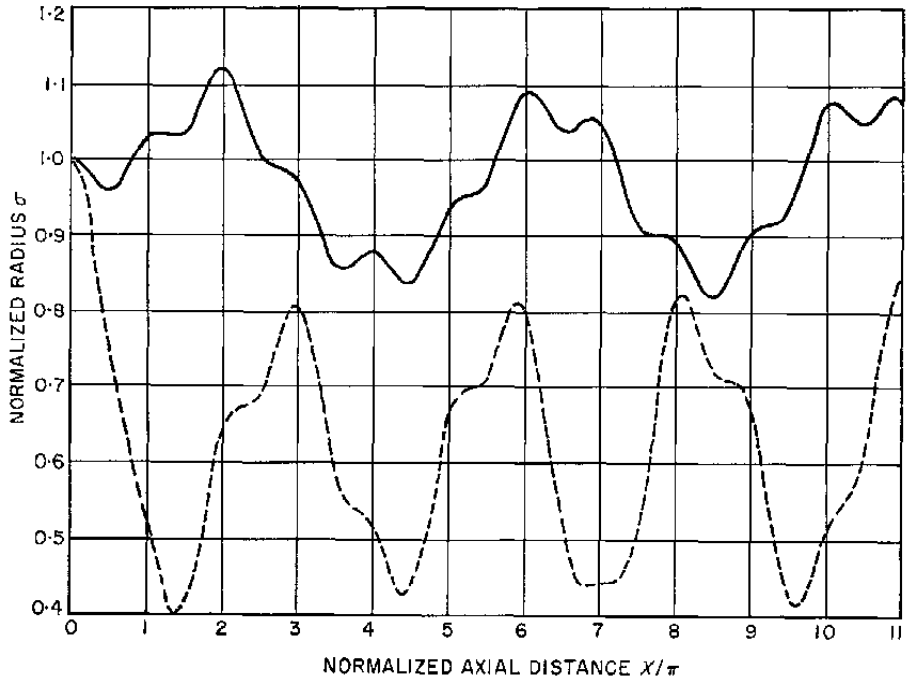


Figure 2—Beam profiles for random variations in the peak magnetic field of assembly 1 of Table 3 with $K = 0$ and $\beta = 0{\cdot}0959$. The solid line is for optimum $\alpha = 0{\cdot}1$ and the broken line is for $\alpha = 0{\cdot}25$.

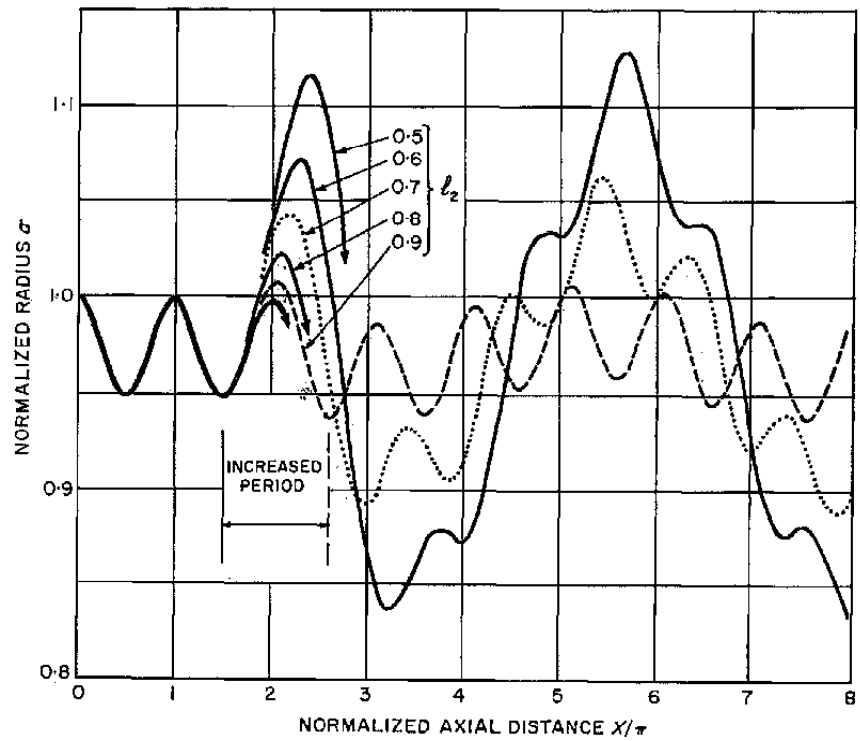


Figure 3—Beam profiles for a single increase in period for $K = 0$, $\alpha = 0{\cdot}1$, and $\beta = 0{\cdot}0959$.

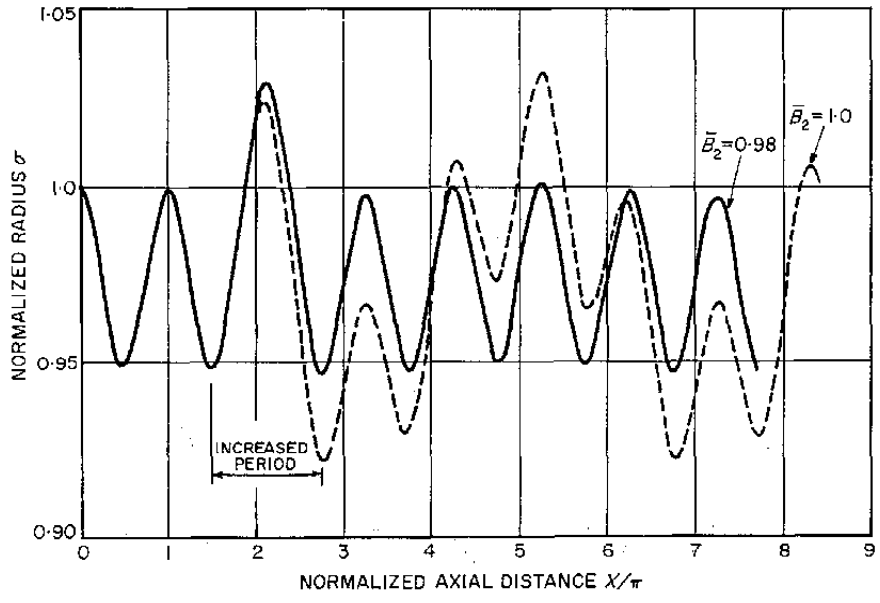


Figure 4—Beam profile before and after compensation for a single increase in period for $l_2 = 0{\cdot}8$, $K = 0{\cdot}1$, $\alpha = 0{\cdot}1$, and $\beta = 0{\cdot}0749$.

The effect of such a single increase in the period is shown in Figure 3. The usual sinusoidal beam profile is considerably distorted as l_2 becomes smaller, that is, the gap is increased. If, however, an increase in the period is accompanied by an appropriate reduction in magnetic field, the beam profile outside the waveguide remains undisturbed. Thus, for a particular set of parameters as in Figure 4, the correct field across the waveguide is $0{\cdot}98\ \hat{B}_o$.

According to Figure 3, doubling the period, that is, $l_2 = 0{\cdot}5$, causes only about 15-per-cent increase in the maximum beam radius in the middle of the gap. Although this would be acceptable in practice, experience shows that if the period increase exceeds about 20 per cent, focusing becomes too sensitive to the manufacturing tolerances of the valve.

Larger increases in the period can be avoided by inserting a pole-piece into the waveguide or cavity itself, as shown in Figure 5. The axial length of this pole-piece must be less than $0{\cdot}5\ \lambda_g$. This not only helps the matching, but

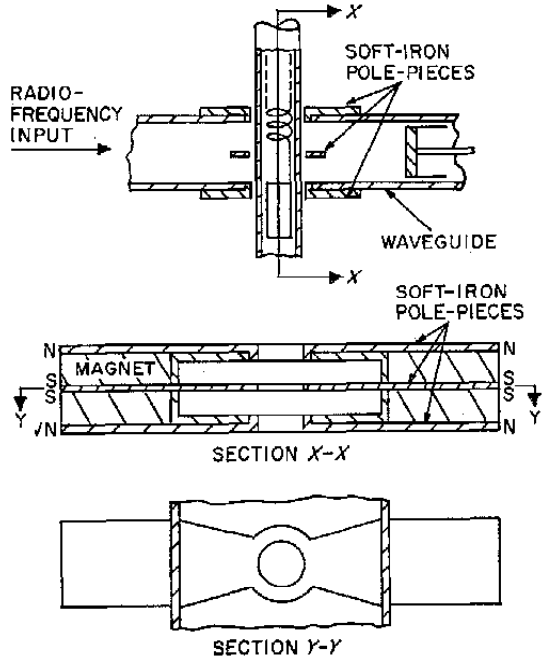


Figure 5—Waveguide-to-helix transducer with an internal pole-piece.

also eliminates the need for two pistons, one above the other. Matching characteristics of a waveguide transducer with and without an internal pole-piece are shown in Figure 6.

1.5 Step Increase in Peak Magnetic Field

The design of a periodic permanent-magnet focusing stack is always based on an unmodulated beam. This is satisfactory unless it is required to focus a power travelling-wave tube. Radio-frequency defocusing forces in the output section may no longer be negligible, and consequently the peak magnetic field in the output section must be increased. A typical increase is about 20 per cent.

Beam profiles in Figure 7 show that a sudden increase in peak field introduces a large beam ripple. A step increase by 20 per cent consisting of at least four steps compresses the beam by about 20 per cent without any apparent change in its profile. Increasing the number of steps beyond four makes no significant improvement. The profiles in Figure 8 correspond to nine steps, the steps increasing either linearly or cubically. There is no practical difference in the final profiles.

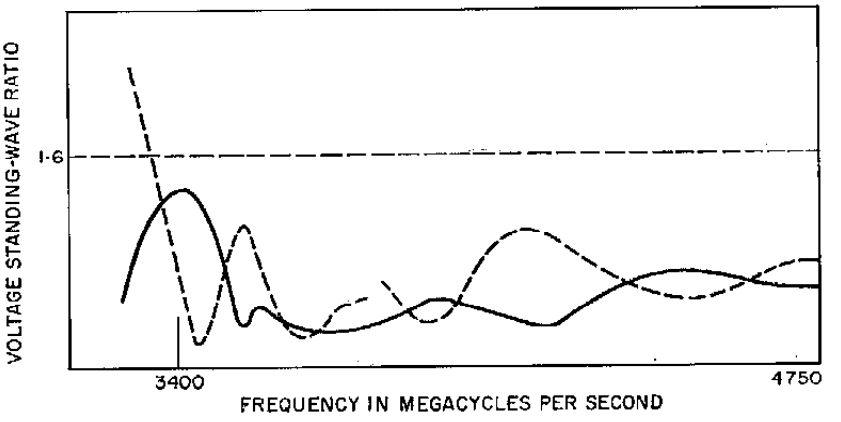


Figure 6—Voltage standing-wave ratio of a waveguide transducer with (broken line) and without (solid line) an internal pole-piece.

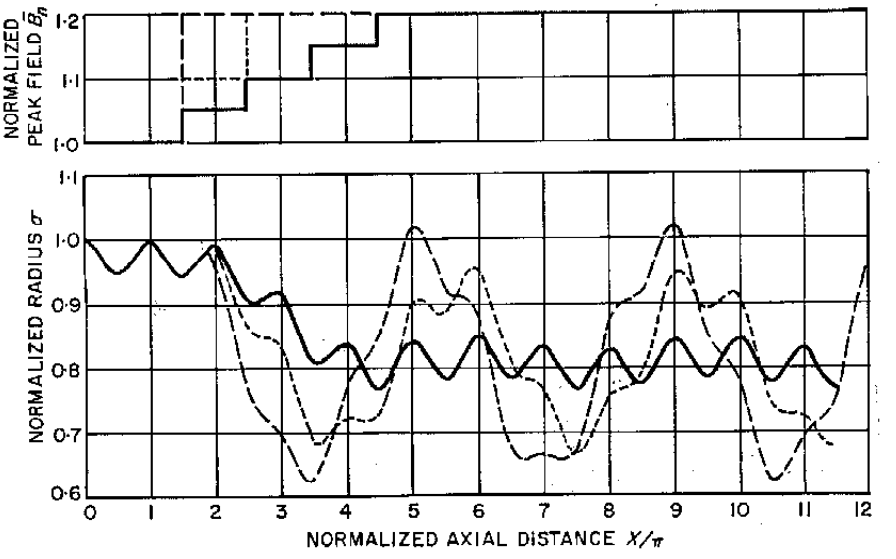


Figure 7—Beam profiles for a step increase in peak magnetic field showing the effect of the number of steps. $K = 0$, $\alpha = 0{\cdot}1$, and $\beta = 0{\cdot}0959$.

1.6 Step Increase in Period and Peak Magnetic Field

Some computations have been done also for a sudden step increase in period alone, and simultaneous increase in period and peak magnetic field. The beam profiles are shown in Figures 9 and 10 respectively. Step doubling of the period in Figure 9 approximately doubles the beam ripple without apparently increasing its maximum diameter.

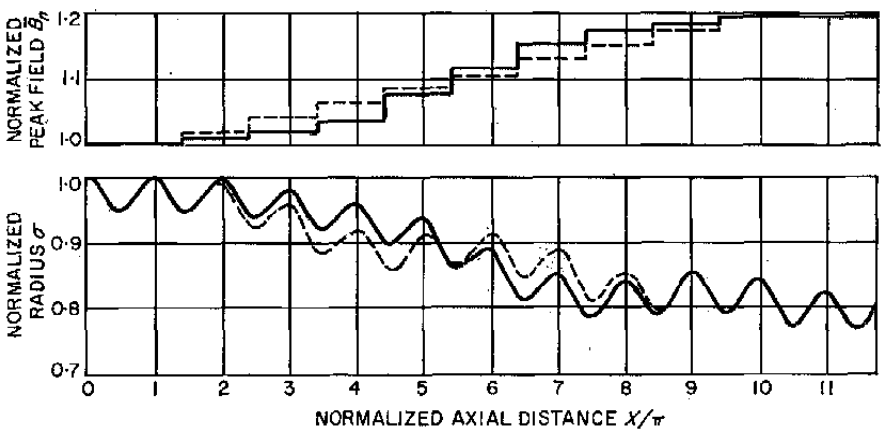


Figure 8—Beam profiles for a step increase in peak magnetic field showing effects of step amplitude. $K = 0$, $\alpha = 0{\cdot}1$, and $\beta = 0{\cdot}0959$.

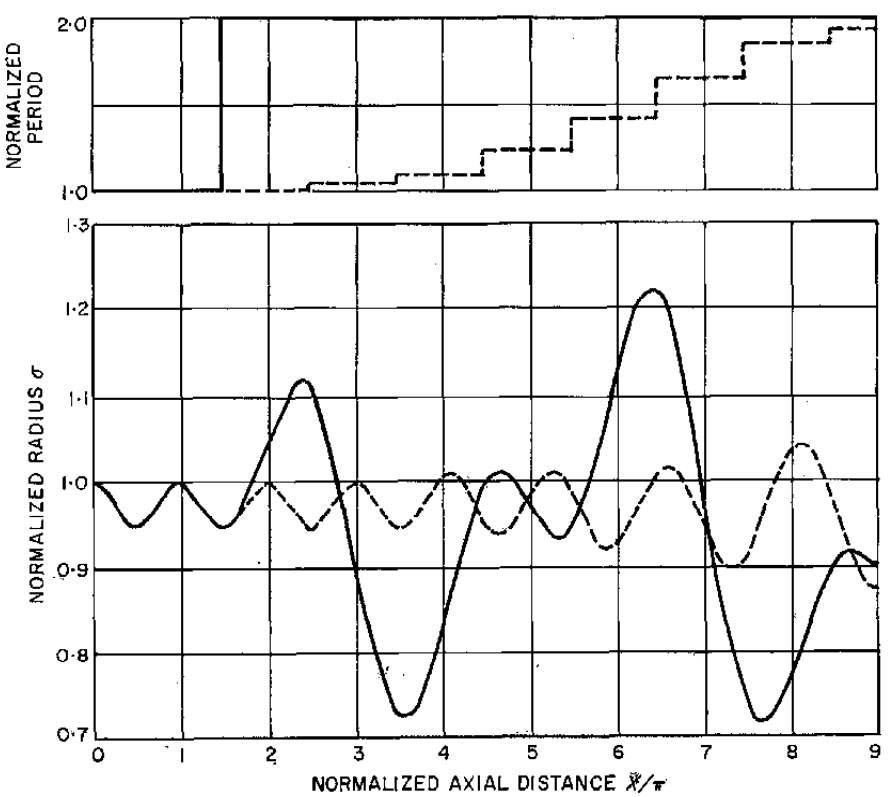


Figure 9—Beam profiles for a step increase in period. $K = 0$, $\alpha = 0{\cdot}1$, and $\beta = 0{\cdot}0959$.

2. Beam Displacement due to Misalignments of Magnetic Cells

Any crookedness of the magnetic field in a periodic permanent-magnet focusing system will deflect the beam from its desired path. We consider here the effect of each section between two pole-pieces, corresponding to a half period, having a magnetic axis parallel to the desired path of the beam, but displaced radially from it.

For the purpose of this analysis, assume that the beam is compressed close to its axis, that is, the beam radius to be smaller than the displacement of the magnetic cell axis.

We use slightly changed cylindrical co-ordinates d, θ, z with the initial entry point and the desired axis of the beam given by $d = 0$; the entry conditions into the nth section are $d = d_n$ and $\theta = \theta_n$. The coordinates of the displaced magnetic axis are D_n and Θ_n, Figure 11.

Considerable simplification in algebra is obtained, and no generality is lost, if θ_n is maintained at zero by rotating the reference co-ordinates at entry to each magnetic cell.

It is shown in the Appendix, Section 6, that if Ω is the angular displacement of the beam axis through one magnetic cell

$$d^2_{n+1} = d_n^2 + 4D_n^2 \sin^2 \Omega/2 + 4D_n d_n \sin \Omega/2 \sin (\Theta_n - \Omega/2). \quad (3)$$

If the different sections are randomly oriented the last term in (3) may be neglected because the average of $\sin (\Theta_n - \Omega/2)$ over all values of argument is zero. Assuming that initially the beam axis coincides with the circuit axis, $d_o = 0$ and (3) reduces to

$$d^2_{n+1} = 4 \sin^2 \Omega/2 \sum_1^n D_n^2 \quad (4)$$

or

$$d_{n+1} = 2n^{1/2} s |\sin (\Omega/2)| \quad (5)$$

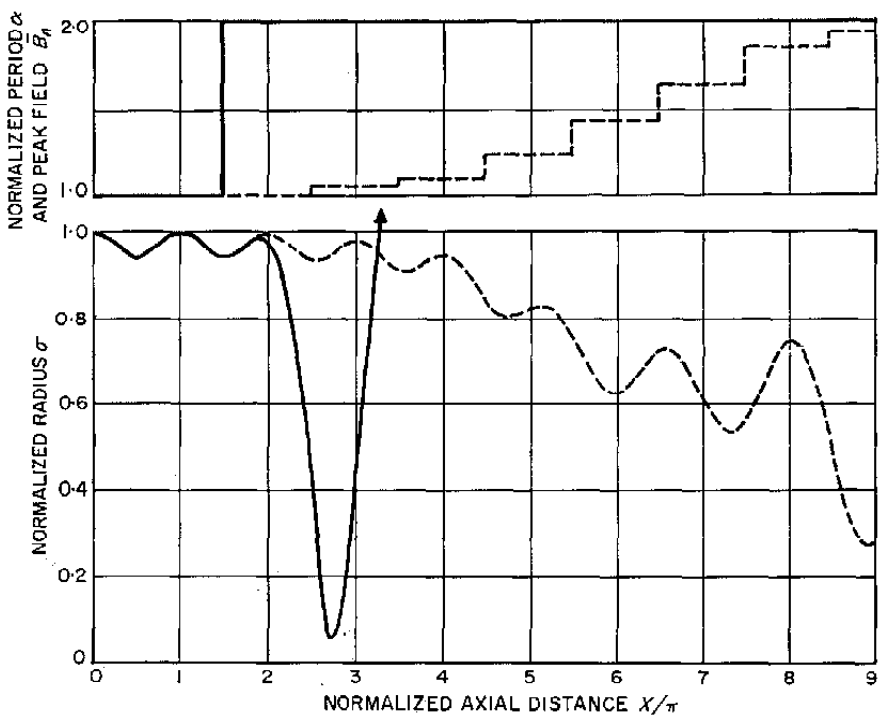


Figure 10—Beam profile for a step increase in period and peak magnetic field. $K = 0$, $\alpha = 0{\cdot}1$, and $\beta = 0{\cdot}959$.

where s is the standard deviation from the intended beam axis.

The angular displacement per magnetic cell Ω is found from Busch's theorem, that is,

$$\Omega = 2(2\alpha)^{\frac{1}{2}}\{(-1)^n - 0{\cdot}5\pi K^{\frac{1}{2}}\} \qquad (6)$$

thus finally, for a shielded cathode

$$d_{n+1} = 2n^{\frac{1}{2}}s|\sin(2\alpha)^{\frac{1}{2}}|. \qquad (7)$$

Can this analysis be applied to all forms of misalignment in periodic circuits? Although formally it does not cover the case where the field is tilted, rather than axially displaced, it would seem that a similar equation would be obtained with perhaps a different definition for the magnetic-axis displacement.

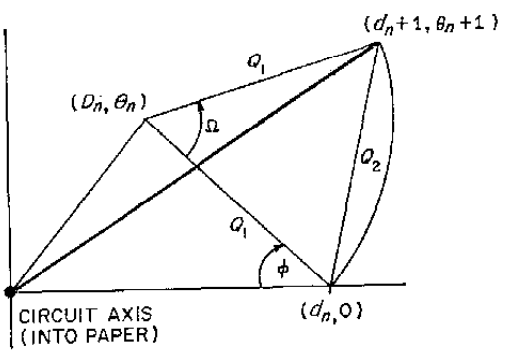
In cases where one or two gross irregularities occur the term $\sin(\Theta_n - \Omega/2)$ in (3) may have to be evaluated.

3. Conclusion

The results show that the undesirable effects of various non-uniformities in the period and magnetic field can be controlled, either by a suitable choice, or by an appropriate variation, of circuit parameters. If α and β are small, as they usually are, the discussed random variations, as well as waveguide gaps, have little effect.

The beam profile for a step increase in the period suggests that one could reduce the number of magnets in a stack by starting with a larger period at the input and tapering down to its optimized value towards the output end of the circuit.

Figure 11—Geometry for electron-beam displacement.



4. Acknowledgments

Thanks are due to Mr. D. Hunter and Mr. M. Lawn of Standard Telecommunication Laboratories computer department for advice on the computer programming.

The approach in Section 2, except for some small modifications, is due to Mr. P. F. C. Burke, and a good deal of this section and the whole of Section 6 have been taken, in changed notation, from his technical note [2].

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2. P. F. C. Burke, "Electron Beam Displacement Due to Misalignment in a Periodic Focusing System," Standard Telephones and Cables Internal Technical Note; 1962.

6. Appendix

With reference to Figure 11

$$d^2_{n+1} = d_n^2 + Q_2^2 - 2d_nQ_2\cos\{\phi + \tfrac{1}{2}(\pi - \Omega)\} \qquad (8)$$

$$\frac{D_n}{\sin\phi} = \frac{Q_1}{\sin\Theta_n} = \frac{d_n}{\sin(\phi + \Theta_n)} \qquad (9)$$

$$Q_2 = 2Q_1\sin\Omega/2. \qquad (10)$$

Using (9) to eliminate Q_1

$$Q_2 = 2D_n\frac{\sin\Theta_n}{\sin\phi}\sin\Omega/2 \qquad (11)$$

$$\tan\phi = \sin\Theta_n\{(d_n/D_n) - \cos\Theta\}^{-1}.$$

Substituting in (8)

$$d^2_{n+1}=d_n^2+4D_n^2\sin^2\Omega/2 \times\{1+(d_n/D_n)^2-2(d_n/D_n)\cos\Theta_n\} +4D_nd_n\sin\Theta_n\sin\Omega/2 \times\left\{\cos\Omega/2-\frac{(d_n/D_n)-\cos\Theta_n}{\sin\Theta_n}\sin\Omega/2\right\}$$

$$=d_n^2+4D_n^2\sin^2\Omega/2+4D_nd_n \times\sin\Omega/2\sin(\Theta_n-\Omega/2). \quad (12)$$

Borivoje Minakovic was born in Zagreb, Yugoslavia, on 16 September 1923. He obtained a B.Sc. (Eng.) degree from London University in 1955.

He worked from 1952 to 1955 for Ferranti Ltd.

Mr. Minakovic in 1955 joined Standard Telephones and Cables and is presently engaged in research and development of travelling-wave tubes.

United States Patents Issued to International Telephone and Telegraph System; May–July 1962

Between 1 May 1962 and 31 July 1962, the United States Patent Office issued 56 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

L. J. F. Andre and R. Van Den Bossche, Bell Telephone Manufacturing Company (Antwerp), Mechanism for Extracting a Document out of a Document Carrier, 3 035 726.

R. P. Arthur and N. J. Simkus, ITT Kellogg, Fuse Holder, 3 038 977.

R. W. Blanchard, ITT Laboratories, Transistor Phase-Shift Oscillators, 3 045 191.

F. H. Bray and D. G. Bryan, Standard Telephones and Cables (London), Electrical Calculating Circuits, 3 039 683.

A. E. Brewster, Standard Telecommunication Laboratories (London), Magnetic Information Storage Arrangements, 3 040 304.

A. E. Brewster, Standard Telephones and Cables (London), Magnetic Information Storage Devices, 3 045 216.

A. T. Brown III, Federal Telecommunication Laboratories, Automatic Gain-Control Circuit, 3 036 276.

F. Buchwald and H. Fliegner, Mix & Genest (Stuttgart), Arrangement to Separate Piled Flat Articles From Each Other, 3 035 695.

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J. E. Cox, ITT Kellogg, Two-party Line Discriminating Circuit for Telephone System, 3 046 343.

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B. M. Mindes, ITT Laboratories, Diversity Combining System, 3 045 114.

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R. K. Van Vechten, ITT Laboratories, Tape Transports, 3 035 748.

R. C. Welch, ITT Federal Laboratories, Oscillatory Apparatus, 3 031 878.

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E. P. G. Wright and J. Rice, Standard Telephones and Cables (London), Electrical Notations Converting Circuits, 3 039 689.

H. Zschekel, Mix & Genest Werke (Stuttgart), Magnetic Storage System, 3 045 213.

Production of Pure Semiconductor Material

3 044 967

H. F. Sterling and F. J. Raymond

This is a process of producing silicon by thermal decomposition of silane using a metal crucible having high thermal conductivity maintained at a temperature lower than the decomposition temperature of silane. A small

charge of silicon is placed in the crucible and heated to the silane decomposition temperature. A steam of pure silane is then directed to the surface of the silicon on the heated surface.

Transistors

3 040 219

W. Fulop

The emitter-base breakdown voltage is increased in a transistor having an emitter and a collector region of one conductivity type and separated by a base region of the opposite conductivity type by providing in the base region a narrow region adjacent to the emitter having a lower impurity concentration than the remainder of the region.

Collision Course Warning Device

3 040 314

H. R. Hesse

This is an automatic device for warning craft operators that they are on a collision course with another craft. The positions and bearings of all aircraft within the range of a plan-position-indicator radar relative to the craft carrying this equipment are recorded on a magnetic storage device. This recorded bearing information is compared with later instantaneous bearing information. An alarm is sounded when the comparison indicates that the craft are on a near-collision course.

Tape Transports

3 035 748

R. K. Van Vechten

This arrangement prevents damage to magnetic tape during drive of the tape in either of two directions. The drive is accomplished by forcing the tape into contact with capstans continuously rotating in opposite directions, by means of respective pinch rollers. The pinch rollers are selectively moved into operative position against the capstans. To prevent tape damage by wrapping around the pinch rollers when the tape is stopped and reversed, these rollers are made with grooves. Protective fingers extend from the mounting plates of the pinch rollers into these grooves. Thus they are not in contact with the tape in normal operation. The pinch roller mounts are also arranged to have a limited axial rotation to assure parallel contact with the capstans.

Facsimile Apparatus

3 044 768

K. G. Hoer

This facsimile system is provided with a magazine for holding a plurality of sheets for successive transfer to a scanning drum. The sheets are separated by rods and the separating members are moved in succession to transfer one sheet at a time to the facsimile scanning drum. Suitable linkages and escapement devices are provided to assure that only one sheet is transferred to the drum for scanning at any one time.

Principal ITT System Products

Telecommunication Equipment and Systems

Automatic telephone and telegraph central office switching systems
Private telephone and telegraph exchanges—PABX and PAX, electromechanical and electronic
Carrier systems: telephone, telegraph, power-line
Long-distance dialing and signaling equipment
Automatic message accounting and ticketing equipment
Switchboards: manual, central office, toll
Telephones: desk, wall, pay-station
Automatic answering and recording equipment
Microwave radio systems: line-of-sight, over-the-horizon
Radio multiplex equipment
Coaxial cable systems
Submarine cable systems, including repeaters
Data-transmission systems
Teleprinters and facsimile equipment

Military/Space Equipment and Systems

Aircraft weapon systems
Missile fuzing, launching, guidance, tracking, recording, and control systems
Missile-range control and instrumentation
Electronic countermeasures
Electronic navigation
Power systems: ground support, aircraft, spacecraft, missile
Radar
Simulators: missile, aircraft, radar
Ground and environmental test equipment
Programmers, automatic
Infrared detection and guidance equipment
Global and space communication, control, and data systems
Nuclear instrumentation
Antisubmarine warfare systems
System management: worldwide, local

Industrial/Commercial Equipment and Systems

Distance-measuring and bearing systems: Tacan, DME, Vortac, Loran
Instrument Landing Systems (ILS)
Air-traffic control systems
Direction finders: aircraft and marine
Ground and airborne communication
Data-link systems
Inverters: static, high-power
Power-supply systems
Altimeters
Flight systems
Railway and power control and signaling systems
Information-processing and document-handling systems
Analog-digital converters
Mail-handling systems
Pneumatic tube systems
Broadcast transmitters: AM, FM, TV
Studio equipment
Point-to-point radio communication
Marine radio
Mobile communication: air, ground, marine, portable
Closed-circuit television: industrial, aircraft, and nuclear radiation
Slow-scan television
Instruments: test, measuring
Oscilloscopes: large-screen, bar-graph
Vibration test equipment
Magnetic amplifiers and systems
Alarm and signaling systems
Telemetering
Intercommunication, paging, and public-address systems

Consumer Products

Television and radio receivers
High-fidelity phonographs and equipment
Tape recorders
Microphones and loudspeakers
Refrigerators, freezers
Air conditioners
Hearing aids
Incandescent lamps
Home intercommunication equipment
Electrical housewares

Cable and Wire Products

Multiconductor telephone cable
Telephone wire: bridle, distribution, drop
Switchboard and terminating cable
Telephone cords
Submarine cable
Coaxial cable, air and solid dielectric
Waveguides
Aircraft cable
Power cable
Domestic cord sets
Fuses and wiring devices
Wire, general-purpose

Components and Materials

Power rectifiers: selenium, silicon
Parametric amplifiers
Transistors
Diodes: tunnel, zener, parametric
Semiconductor materials: selenium, germanium, silicon
Capacitors: wet, dry, ceramic
Ferrites
Tubes: power, transmitting, traveling-wave, rectifier, receiving, thyratron
Picture tubes
Relays and switches: telephone, industrial
Magnetic counters
Resistors
Varistors
Fluorescent starters
Transformers
Quartz crystals
Crystal filters
Printed circuits
Hermetic seals
Magnetic cores

Trends in Component Developments
High-Reliability Testing and Assurance for Electronic Components
Microelectronic Devices
Silring Mounting of Silicon Power Rectifiers
Cause and Prevention of High Reverse Currents in Large-Area High-Voltage Diffused-Silicon Rectifiers
Defects in Vapour-Grown Silicon
Silicon Epitaxial Planar Transistors
Mixer Diodes for 6 Gigacycles Per Second
Vacuum Tubes for Submerged Repeaters
Ceramic-Insulated Vacuum Tubes for Very-High-Frequency Industrial Heating
Travelling-Wave Tubes for 6-Gigacycle-Per-Second Radio Links
Effect on an Electron Beam of Variations in Periodic Permanent-Magnet Focusing Systems

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This Issue in Brief

Analog Signal Multiplier—A cascade-bridge type of analog signal multiplier forms its product in stable parametric fashion, using an uncomplicated structure. The signal voltage required may be of the order of 0.1 volt, or less.

The use of voltage-adjustable capacitors, or parametric diodes, permits the desired product to be effected over a very-wide frequency range. Products of complicated time functions can be obtained within the present bandwidth capabilities of approximately 400 kilocycles per second in the developmental model. Extension to megacycle-per-second bandwidths is entirely feasible.

Products of various types of analog signal voltages are illustrated. A number of applications for the multiplier are also suggested, ranging from the practically instantaneous phase reversal of waveforms to the more-sophisticated application of cross correlation to improve signal-to-noise ratios.

Transmission-Loss Measurements and Diversity Tests Made on Two Knife-Edge Diffraction Paths in Europe—Results are given of radio propagation measurements made at 1965 megacycles per second on two knife-edge diffraction paths. The measured transmission loss indicated that the diffraction gain was not the same on the two paths. On each path, the measured loss was less than was estimated for normal beyond-the-horizon paths having the same angular distance, but was greater than the values calculated from knife-edge diffraction theory.

Diversity tests made with receiving antennas spaced 47 feet (14.3 meters) horizontally and 10 feet (3 meters) vertically showed that there was little correlation on either path between the carrier-level variations at the two diversity receiving antennas. A reason for this low correlation may be that in each case the knife-edge obstruction was not smooth but was a ridge varying in elevation across the path.

Marine Transmitter with Single-Sideband Facility—A marine transmitter for the 400-to-535-kilohertz and 1.6-to-25-megahertz ranges produces a peak power of 1200 watts. For telegraph operation, the carrier may be keyed or the carrier and one sideband of audio-frequency modulation may be keyed. For telephony, single sideband may be transmitted with either suppressed or full-amplitude carrier.

A linear power amplifier using four *4X250B* tubes in parallel is driven by a preamplifier, which is operated from a frequency-determining and modulating exciter. The transmitter uses 94 quartz crystals, without temperature control, to produce the multiplicity of channels in the various bands. A frequency-drift-cancelling circuit is used.

The carrier and single-sideband signal components are generated independently of each other and may be combined in any proportion as required.

A π-L network permits proper loading of the many types of antennas found on shipboard without modification of either the equipment or antennas.

Submarine-Coaxial-Cable Manufacture at Southampton Factory 2—The construction of the second submarine-cable factory at Southampton was initiated in 1961. It is used for the manufacture of a new design of armourless deep-sea cable developed by Bell Telephone Laboratories for their *SD* long-distance submarine-cable telephone systems. The new cable is of coaxial construction and has the strength member, consisting of 41 high-tensile-steel wires, inside the inner-conductor copper tube. These two components are produced from one machine in which a strander, a tube-forming mill, an argon-arc seam welder, and a tube-drawing mill are operated in tandem. A complex electronic controller maintains correctly the relative speeds of the four units of this machine during starting, running, and stopping. A current proportioner varies the arc current as a preset function of the line speed.

The core insulant is a high-molecular-weight low-density polythene, and after extrusion the external diameter is shaved at 20 degrees centigrade to within a tolerance of 0·1 per cent of the specified diameter of 1·000 inch (25·4 millimetres). The shaving operation also corrects for extruder failure to keep the inner conductor concentrically within the polythene insulant.

A single copper tape is applied longitudinally with overlapping edges over the shaved core to form the outer conductor. In these operations, the tubing mill is driven in tandem with the sheathing extruder. A sheath of high-molecular-weight high-density black polythene plastic compound completes the new deep-sea cable, for which the manufacturing unit is a 20-nautical-mile (37-kilometre) repeater-section length.

Telepuls and FMX Modular Electromechanical and Electronic Remote Control—A group of standard units permit the needs of various types of users of remote indication and control systems to be met without special designs. Telepuls 10 and 12 are pulse-code binary systems using electromechanical and electronic means, respectively, and are so fully compatible that their units may be intermixed for certain purposes.

A mark and space, each of 40 milliseconds duration, represent a binary 0 and a similar 160-millisecond pair indicate binary 1. By reference to two 60-millisecond pulses that are initiated by the start and the end of each received signal pulse, the shorter pulses cancel each other and do not affect a flip-flop circuit but the longer signals turn the flip-flop on and then off, thus selecting only the 1's for decoding the signals.

Memories supply both the addresses of the equipment to be queried or controlled and the proper form of message for the type of equipment. The usual mimic diagram shows the system conditions by lighted lamps. Analog information is converted into decimal and then binary code for transmission.

The *FMX1* system uses 15 tone frequencies for identifying equipment and is limited to party-line control of substations having only 1 or a few controlled units. Presence of a tone indicates "good" or "on" and its absence the alternate condition.

Digitel 108, A Digital Telemetering System—All data are transmitted in binary digits, suitable transducers being employed to convert analog measurements to this form. A basic group of 8 inputs has been selected and multiple groups may be used for larger installations. A telemetering precision within ±0.5 percent is obtained. Transmission is within a 120-hertz voice-frequency telegraph channel, which may be derived from a telephone channel.

Binary coding uses a frequency deviation of −30 hertz for 0 and +30 hertz for 1. Address and message codes are taken from memories and a parity pulse is added if necessary to make an even number of pulses for each complete telegram. Converters permit parallel inputs and outputs to be transmitted serially over the telegraph channel.

Plug-in printed circuits and transistors permit great flexibility in the design and assembly of systems. Convenient test points, built-in light indicators, reference oscillograms in the instruction manual, programs for routine tests, and rapid replacement of units minimize out-of-service time.

Transradar—A Narrow-Band Radar Relay Equipment Using Storage Capacitors—For air-traffic control, it is helpful to use several surveillance radars whose coverages overlap or supplement each other. For economy and flexibility, the megahertz video band from each radar may be greatly compressed to permit remote plan-position indicators to be operated over telephone channels.

Bandwidth compression is obtained by sequentially charging a bank of capacitors, each corresponding to a different increment of range from the radar antenna. A writing timer under

control of the radar trigger determines the sequence in which the capacitors are charged from the radar video signal and the charges are read at a slower rate. This derived narrow-band video signal, supplemented by trigger and azimuth information in digital form, is encoded. It is then transmitted for long distances, using vestigial-sideband modulation, over conventional radio or telephone lines. The transmitting unit may be installed at unattended radar stations, which may then be operated by remote control.

The received signal may operate a slow-scan plan-position indicator directly. However, if the conventional high-speed indicator is used, another bank of storage capacitors will be required at the receiver to permit bandwidth expansion corresponding to the compression in the transmitter. By time sharing, signals from more than one radar may be reproduced on a single indicator.

Optimization of Network Configurations Through Queueing Theory—A major problem in system communication is the optimum location of manned and/or machine information-processing stations. The problem is further complicated if the information is processed on a priority basis.

A queueing-theory model that simplifies the process of obtaining an optimum network configuration and a computer flow-chart of the model are presented. The output of this model is a generalized set of curves showing traffic density versus queue length and traffic density versus waiting time, both as functions of processing rates and priority.

A numerical example is included to illustrate the method by which the expected queue lengths and waiting times can be found for a station in the iterative process of determining an optimum network configuration for a semi-automatic communication system.

Traffic Simulation and Its Application in Telephony—Approximate computation of the traffic-handling capacity of telephone switching networks may be verified by measurements based on artificially generated traffic. Electronic computers may replace the traffic machine to generate artificial traffic and at the same time simulate the switching network to be tested.

In a simple example, the number of rejected or lost calls from an infinite number of inlets to 3 outlets for a traffic of 2 erlangs is to be found. A series of random numbers from 1 through 6 is produced by repeated casting of a 6-sided die. Numbers 1, 2, and 3 indicate release of any call that may be present on the correspondingly numbered outlet. Numbers 4 and 5 initiate a call on the lowest-numbered free outlet. No action is initiated by 6. Although the successive steps do not correspond to constant time intervals, either real or artificial, this method produces calls having exponentially distributed holding times. The test should encompass several thousand numbers in the random series to ensure a balance of statistical variations.

In a practical test, the number of outlets and of calls would be much larger than in the example above. The paper is concerned with the application of a computer to the essential elements of developing the required random numbers, their use as artificial traffic, and the simulation of the switching network in handling the traffic offered to it, to find the quantity of calls rejected because outlets were not immediately available to them. Lost calls were not stored for later attempts for completion.

Automatic Switching for Telex—The telex offices of the American Cable and Radio Corporation in San Francisco and in New York, with extension to Washington, District of Columbia, now employ *7E* rotary switching equipment to expand facilities and increase automation. Compatibility of operation must be maintained with the domestic telegraph and telephone telex networks and with corresponding overseas networks to provide subscriber-to-

subscriber service with minimum operator assistance.

A 3-digit prefix that identifies the geographical area in which the telex subscriber's number applies permits worldwide calling on an automatic basis. Operator assistance may be required on calls switched through the domestic telephone network for certain tone signals that are not automatically translated into teleprinter signals.

Each telex operator position can handle 2 calls and make 6 connections simultaneously. Overflow positions are provided for peak-traffic intervals. Billing is automatic.

The telex subscriber may receive traffic and also send messages at the conventional per-word rate.

Telex subscribers of the domestic telegraph and telephone systems may gain access to the overseas network of American Cable and Radio Corporation.

The *7E* teleprinter switching system uses single-motion nonhoming 7-brush rotary switches having 100 outlets. Switches are positioned under the direct control of a register by matching 12 phases and 2 voltages of a 450-hertz supply to provide 24 distinguishable positions. Wire-spring relays provide long life and fast operation with low actuating power. Each line circuit has 2 relays and 2 resistors that are adjusted initially to compensate for the line characteristics and to set the bias current of the repeater in the connecting circuit.

High-Power Precision-Stabilized Rectifiers—Three-phase full-wave rectifiers and filters are described for producing direct currents as large as 450 amperes at powers above 100 kilowatts. Regulation maintains the load current and ripple within 5 parts in 10^5 for an 8-hour period and within 5 parts in 10^6 for approximately 5 minutes.

The voltage drop across a precision resistor carrying the load current is compared with a reference voltage derived from 3 cascaded stages of zener diodes. Any difference voltage goes to a direct-current amplifier and the greatly amplified output is connected to the base of one or more control transistors that also carry the full load current. The voltage drop across the emitter and collector of this transistor stage controls a 3-phase motor-driven regulator transformer. An auxiliary chopper and amplifier are provided to compensate for drift in the direct-current amplifier.

Recent Achievements

Laval Factory Opens—Mr. Jacques Marette, Minister of the French Posts and Telecommunications, presided over the opening on 5 July 1963 of the new Sept-Fontaines plant in Laval, about 290 kilometers (180 miles) west of Paris. The 300 guests were welcomed by Dr. Maurice Deloraine, president, and Mr. Philippe Lizon, managing director, of Le Matériel Téléphonique.

Covering an area of 15 000 square meters (161 460 square feet), the plant will assemble Pentaconta crossbar telephone equipment, its capacity being about 250 000 lines per year. In addition, manufacture of capacitors will soon be initiated. The present staff of 650 will be increased to 1000. Figure 1 shows a partially assembled Pentaconta switch being inspected.

Le Matériel Téléphonique
France

Quasi-Electronic Telephone Exchange—Richard Stücklen, Minister of the German Posts, Telegraphs, and Telephones Administration, may be seen in Figure 2 as he opened a quasi-electronic telephone exchange in Stuttgart on 12 July 1963 by placing a call from a push-button subscriber's set.

The 2000-line Blumenstrasse office shown in Figure 3 is part of the Stuttgart local area. In addition to the standard German facilities of subscriber dialling of long-distance and international calls and of identified out-dialling and in-dialling with private automatic branch exchanges, the new office provides push-button calling and alternative routing within the local area.

This trial installation will provide practical experience with the functioning of the equipment and system as well as the acceptance by subscribers of the new features that it makes available to them.

The *HE-60* quasi-electronic switching system uses hermetically sealed dry-reed switches and electronic components exclusively, combining the speed of electronic action with the reliability of mechanical contacts. A multistage link arrangement uses 4-contact crosspoints with individual reed switches and coils for each contact.

The desired connection is established in one marker operation by the register. While the



Figure 1—Examining a Pentaconta switch at the opening of the new Laval factory are, from left to right, Dr. Maurice Deloraine, president of Le Matériel Téléphonique; Mr. Jacques Marette, minister of French Posts and Telecommunications; and Mr. Philippe Lizon, managing director of Le Matériel Téléphonique.



Figure 2—Richard Stücklen opening Blumenstrasse quasi-electronic office in Stuttgart using a push-button subscriber's set.

central translator is determining the optimum route, the path through the network is found with the help of a guide wire. Subscriber lines are in groups of 2000 that are served on a one-at-a-time basis by a central control. The common units are duplicated and normally operate alternately but one unit is capable of carrying the entire load in case of trouble. Punched-tape reports are made automatically of all faults. Standardized plug-in units are used.

Standard Elektric Lorenz
Germany

Non-Linear Speech Amplitude Quantizer—The standard telephone channel bandwidth from 300 to 3400 cycles per second passes more information than is needed for correct recognition of human speech. To investigate this field, a non-linear amplitude quantizer in which the number of steps and their grading could be adjusted was used to find the minimum number of levels necessary for accurate reception of English language speech.

With 4 amplitude levels for each half wave having relative magnitudes of 1, 4.6, 22, and 100, corresponding to 13.4-decibel steps, the received speech was above minimum acceptable telephone quality. Optimum performance was obtained with peak clipping of about 8 decibels at the highest level. Speech became slightly harsh from harmonics of the vowel sounds when the signal was 20 decibels above this optimum.

Although the output waveform was a very crude copy of the input analogue waveform, it contained the elements necessary for understandability and for conveying the natural variations in loudness of the original speech. The understandability was maintained over a 30-decibel variation in the mean level of speech input.

Standard Telecommunication Laboratories
United Kingdom

Accelerometer Calibrator—The equipment shown in Figure 4 is used for the comparative calibration of accelerometers. When used with a reference accelerometer calibrated to within

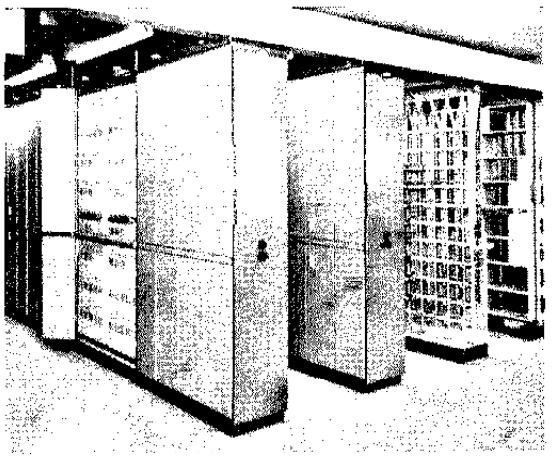


Figure 3—Quasi-electronic 2000-line exchange based on electronic switching of dry-reed contacts. It is part of the Stuttgart local-area telephone system.

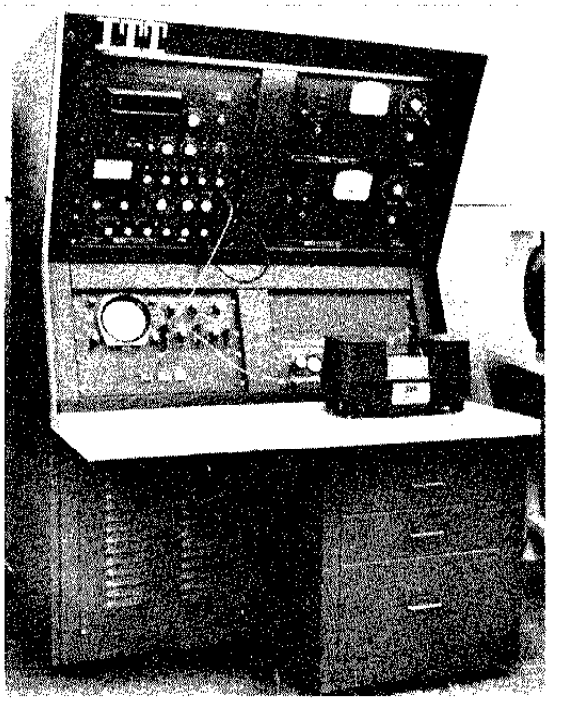


Figure 4—Equipment for comparative calibration of accelerometers.

±1 percent by the National Bureau of Standards, the system will have a total possible error of less than 2.4 percent and a probable error of only 1.2 percent.

A series of seminars on the techniques for calibrating vibration pickups conducted at our San Fernando, California, plant permitted about 50 specialists from the leading industrial companies active in this field to contribute to the design criteria for this apparatus.

ITT Industrial Products Division
United States of America

Pentomat Pentaconta Private Branch Exchanges—A new series of Pentaconta private automatic branch exchanges has been developed. Three sizes are available for 40 extensions and 5 trunks, 10 extensions and 3 trunks, and 6 extensions and 2 trunks, being designated Pentomat *40*, *10*, and *6*, respectively.

The entire equipment, including the mains-operated power supply, is housed in a metal cabinet. The Pentaconta relays are mounted on hinged frames for easy and rapid accessibility. Operation of this all-relay system is quiet and free of maintenance.

The Pentomat *40* is operated with an attendant's set; if traffic is heavy, the Pentomat *10* will be similarly equipped. The smallest unit does not normally need this facility. The attendant's set shown in Figure 5 is a compact desk unit with provision for both dial and push-button selection.

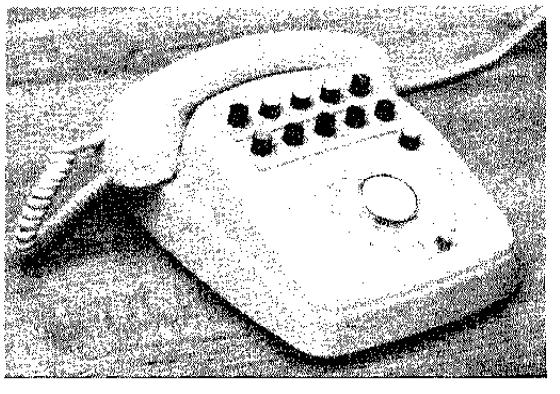


Figure 5—Attendant's set for Pentomat *10* private branch exchange.

Bell Telephone Manufacturing Company
Belgium

Transatlantic Telephone Cable—Now being laid over a distance of 3600 nautical miles (6667 kilometres), *TAT-3* will be the longest single submarine cable link in the world.

In Figure 6, *HMTS Alert* is taking on a supply of the new lightweight cable from the world's largest submarine-cable factory built by Standard Telephones and Cables at Southampton Docks in England. A joint project of the British Post Office and the American Telephone and Telegraph Company, this will be the first repeatered telephone cable to link directly the United Kingdom and the United States.

Standard Telephones and Cables
United Kingdom

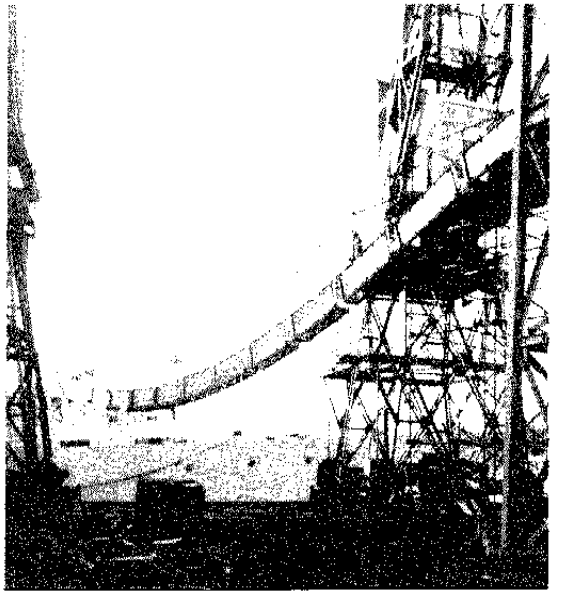


Figure 6—*HMTS Alert* taking on a cargo of the new lightweight coaxial cable from the largest submarine cable plant in the world at Southampton Docks, England.

Coherent Light Pierces Steel—Enormous energy may be concentrated into a very small area by focusing coherent light as shown in Figure 7.

A laser cooled by liquid nitrogen produces a coherent light that is focused by a photographic lens onto a steel plate. A single pulse lasting only 200 microseconds pierces a hole having a diameter of about 50 microns in the 1.5-millimeter (0.06-inch) plate.

Laboratoire Central de Télécommunications
France

Netherlands Automatic Telex Exchanges—The Leeuwarden telex exchange in The Netherlands was cut over on 18 April 1963. The network uses the indirect rotary system developed by Bell Telephone Manufacturing Company.

Separate telex exchanges have been built in Amsterdam, The Hague, Rotterdam, Bois-le-Duc, Groningen, Utrecht, and Arnhem. Additional exchanges are under construction for Venlo, Deventer, Hengelo, Eindhoven, and Haarlem.

There are more than 6000 private telex subscribers in the country. All national and 95 percent of the international connections are made automatically.

Nederlandsche Standard Electric Maatschappij
The Netherlands

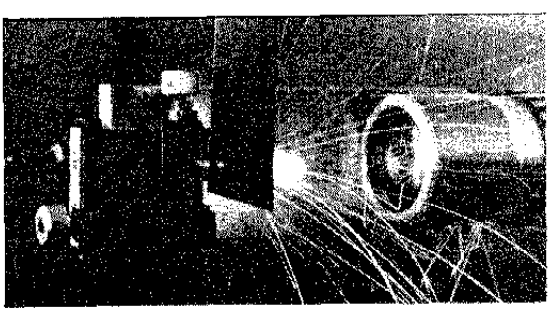


Figure 7—Coherent light from laser pierces hole in steel plate.

Digital Recording of Wind-Tunnel Tests—A characteristic of wind-tunnel testing is the large number of measurements that are taken in a short time.

These measurements may now be recorded by a page teleprinter that also punches a paper tape to be used as input to a Zebra computer for evaluation of the test results.

A second-generation design shown in Figure 8 makes increased use of electronic circuits for control of the teleprinter.

Standard Telephone & Radio
Switzerland

Poly Para Xylylene Capacitor—A very-thin film of poly para xylylene, having a permittivity of 1.7 and a breakdown strength of 2 megavolts per centimetre, is the dielectric of a new capacitor. The material may be deposited from the gas phase on to a suitable substrate such as aluminium foil, glass, et cetera, by passing para xylene vapour through a tube at 900 degrees centigrade and at a pressure of a few millimetres.

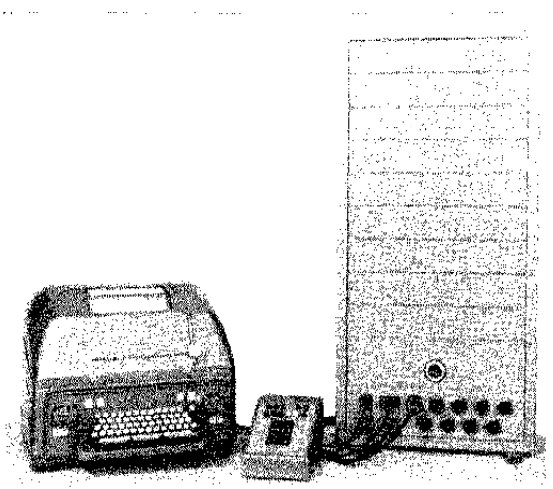


Figure 8—Combined page teleprinter and tape perforator with input equipment for recording wind-tunnel test data. The tape becomes input to a Zebra computer.

The poly para xylylene may exist in three forms, amorphous, α, and β, depending on the pressure, temperature of the gas, and temperature of the substrate. The amorphous form converts to α at about 50 to 140 degrees centigrade and then to the β form at 200 to 300 degrees centigrade. The β form thus produced is brittle in contrast to that obtained directly from the gas phase. The amorphous and α forms are both mouldable but change very quickly to the β form at moulding temperatures. Ultra-high pressures will reconvert β to α.

Typical capacitors having 1-square-centimetre electrodes, a calculated film thickness of 16 microns, and a weight of 0.01 gramme, have a capacitance of 130 picofarads, loss angle of 0.0012, and resistance of 1 megohm. After irradiation with 2-million-electron-volt electrons at doses up to 50 000 megarads, their electrical properties were changed to a capacitance of 122 picofarads, loss angle of 0.0026, and resistance of 500 kilohms.

Standard Telecommunication Laboratories
United Kingdom

New Factory in Eggenberg, Austria—On 21 June 1963, a new factory was inaugurated in Eggenberg, Austria. With a floor area of 3000 square meters (32 280 square feet), the plant will accommodate up to 350 workers. It is shown in Figure 9.

This factory will be used for assembling telephone bays, relay strips, Villaphone home telephones, and Dirigent intercommunication sets including loudspeaker and door-answering designs.

Standard Telephon und Telegraphen
Austria

Radio Altimeter—A new design of radio altimeter, coded *STR 52*, has simultaneous outputs of 0–2500 feet (0–762 metres) for terrain clearance in low-strike fighter operation and of 0–500 feet (0–152 metres) for use in automatic landing. The aerials may be as far as 50 feet (15 metres) from the equipment.

Standard Telephones and Cables
United Kingdom

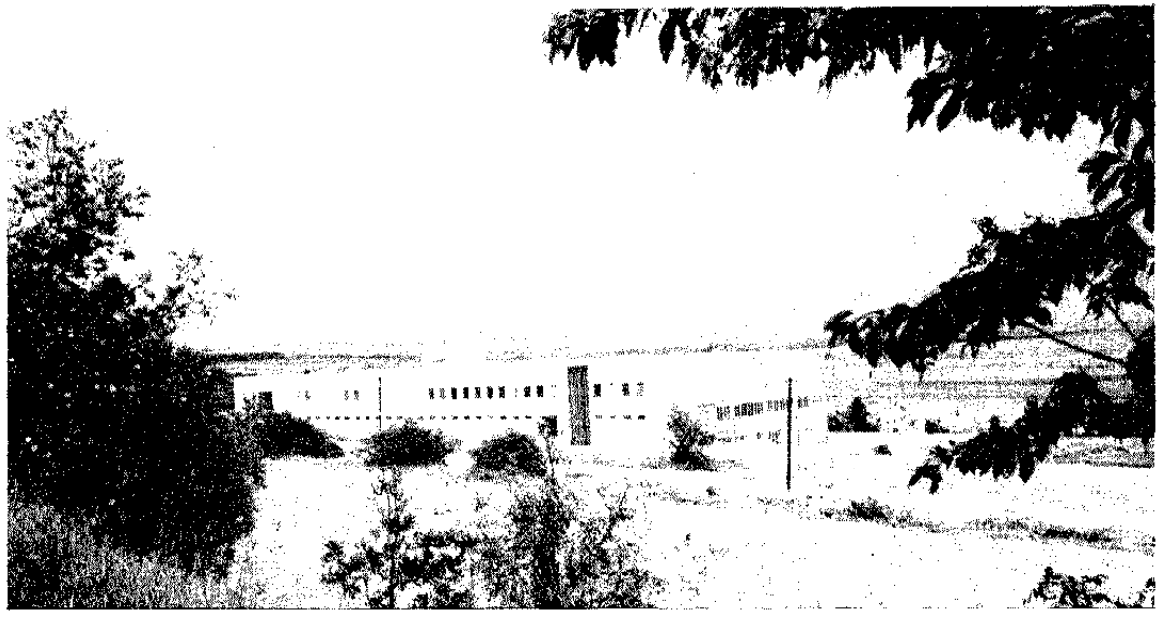


Figure 9—New telephone factory recently inaugurated in Eggenberg in the northern region of lower Austria.

Standard Elektrik Lorenz Installs 10 000th Crossbar Private Automatic Branch Exchange—In March 1963, Standard Elektrik Lorenz put in service its 10 000th crossbar private automatic branch exchange, which is in the new building of the central fire station of Ludwigshafen, Germany. This is the 11th installation of its kind for that city administration.

All private automatic branch exchanges having more than 10 lines are now supplied with crossbar switches by Standard Elektrik Lorenz, which has supplied crossbar equipment having a capacity of 650 000 lines to customers all over the world. Precious-metal contacts provide good noisefree connections; short travel of contacts permits high-speed, vibrationless, quiet operation; and the simple relay type of design ensures reliability, long life, and minimum maintenance.

Standard Elektrik Lorenz
Germany

Primrose Telephone Subscriber Set—Available in several colors to harmonize with any surroundings, the Primrose telephone subscriber set, with its illuminated dial, oblong shape, and depth of only 95 millimeters (3.7 inches), is particularly suited for bedrooms and hotels. It is shown in Figure 10.

The sound volume of the built-in ringer is adjustable. The lightweight handset is equipped with a high-quality microphone and receiver and can be furnished with a transmission equalizing circuit. The Primrose set has been approved by the Belgian telephone administration.

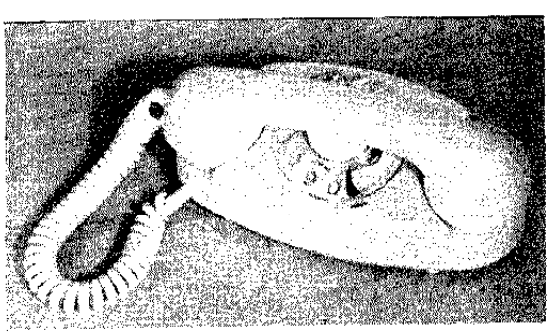


Figure 10—Primrose subscriber set with illuminated dial.

Bell Telephone Manufacturing Company
Belgium

Gallium-Arsenide Crystals and Lasers—Gallium-arsenide is a second-generation semiconductor suitable for solid-state electronic applications. Compared to silicon and germanium, its greater carrier mobility promises response at higher frequencies or operation at higher powers. Its wider energy gap should permit operation at increased temperatures and reduced leakage at reverse-biased junctions. Its recent use as a laser indicates new properties not exhibited by silicon and germanium.

For laser operation, gallium-arsenide must contain large and carefully controlled amounts of impurities such as selenium and silicon. By the beginning of 1963, single crystals having the required level of impurity and unusually high crystal perfection had been grown. Dislocation density was below 1000 per square centimetre and carrier mobility was particularly high, being above 1000 square centimetres per volt-second at 10^{19} carriers per cubic centimetre and 2500 square centimetres per volt-second at 10^{18} carriers per cubic centimetre. This material is particularly well suited for laser preparation because of its great homogeneity. A further finding was a method of growing from the melt p-type gallium-arsenide of the extremely high resistivity of 10^{8} to 10^{9} ohm-centimetres.

Standard Telephones and Cables
United Kingdom

Blind Telephone Operators—In Western Germany, about 2000 blind persons are working as telephone operators. Despite the complexity of the modern telephone exchange, they perform their duties effectively with the aid of finger guide bars, a few indications in Braille characters, and audible rather than visual signals. A typical switchboard is shown in Figure 11.

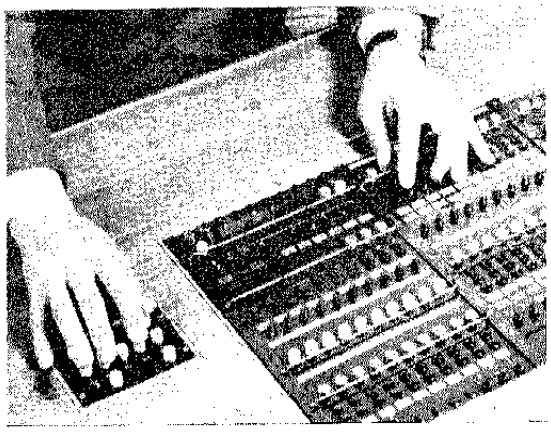


Figure 11—Telephone switchboard for the blind.

A report based on 30 years of experience in designing telephone equipment for the blind is now available. Requests for brochure 0130/364 should be addressed to Standard Elektrik Lorenz AG, 42 Hellmuth-Hirth-Strasse, Stuttgart Zuffenhausen, Germany.

Standard Elektrik Lorenz
Germany

Hospital Intercommunication Systems—The modernization of hospital intercommunication systems has greatly expanded the services they provide. The early push button to call an attendant has been replaced by a microphone-loudspeaker permitting the patient to converse with the nurse or with an operator during the night when calls are concentrated. The same loudspeaker may be used for music.

Recent improvements include the use of unbreakable housings for indicator lamps, improved acoustic properties of the microphone-loudspeaker units, printed wiring on individual boards for fewer interruptions in service, control switchboards using printed-circuit panels of washable materials, and standard equipment racks with plug-in drawers for transistor amplifiers and oscillators, groups of relays, and common circuits.

Bell Telephone Manufacturing Company
Belgium

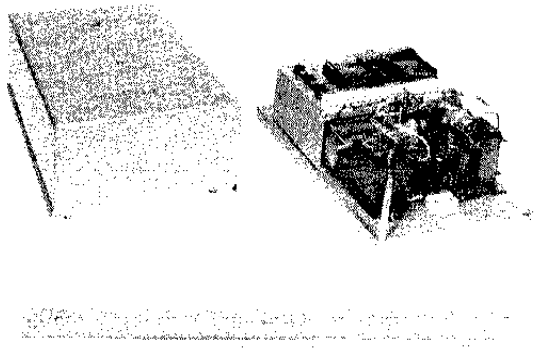


Figure 12—Magnetically stabilized rectifier-filter power supply for telephone subscriber sets.

Battery Substitute for Telephone Operation—Figure 12 shows a battery substitute operating from the 220-volt 50-cycle-per-second power mains that produces 25 volts at 0.7 ampere of direct current for speech and 75 alternating volts for ringing of telephone subscriber sets. Magnetic stabilization makes it insensitive to mains variations of ±10 percent and to load variations. Its dimensions are 375 by 240 by 142 millimeters (14.8 by 9.4 by 5.6 inches) and a transistor version reduces this volume to less than half.

Nederlandsche Standard Electric Maatschappij
The Netherlands

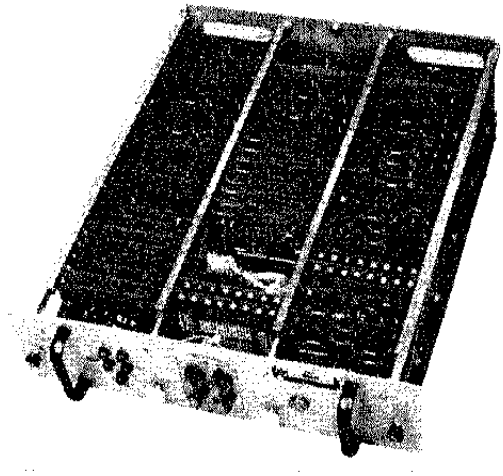


Figure 13—Line store for 256 characters operating at 10 000 bits per second.

Solid Tantalum Capacitors—The British Ministry of Aviation has issued its first such certificate for a design of sintered-anode tantalum capacitors. They have high capacitance per unit volume and may be used over a temperature range from − 55 to + 125 degrees centigrade. They are the only type of electrolytic capacitor capable of being hermetically sealed.

Standard Telephones and Cables
United Kingdom

Magnetic-Core Line Store—The line store shown in Figure 13 was developed for command and control systems. Up to 256 characters may be stored on magnetic cores at a rate of 10 000 bits per second with character-by-character output. It may be used as a random-input device and as a real-time store. Unusual features include visual operating and fault indicators, message sequence counting, and nondestructive readout. Internal circuits detect errors in data information and malfunctioning of the logic circuits.

ITT Kellogg Communications Systems
United States of America

Safe Telephone Set—A special telephone set has been developed for use in coal mines and other hazardous places where combustible gases can be ignited by electric sparks. The energy flow, either in normal or fault conditions, is held to so low a value that the opening and closing of electric circuits will not ignite explosive gases. No further safety provisions are required.

Operation in a noisy environment is aided by a shortened handset, which is equipped with a high-quality dynamic microphone and receiver, and a transistor amplifier that accentuates the high-frequency response. The set is shown in Figure 14.

Standard Elektrik Lorenz
Germany

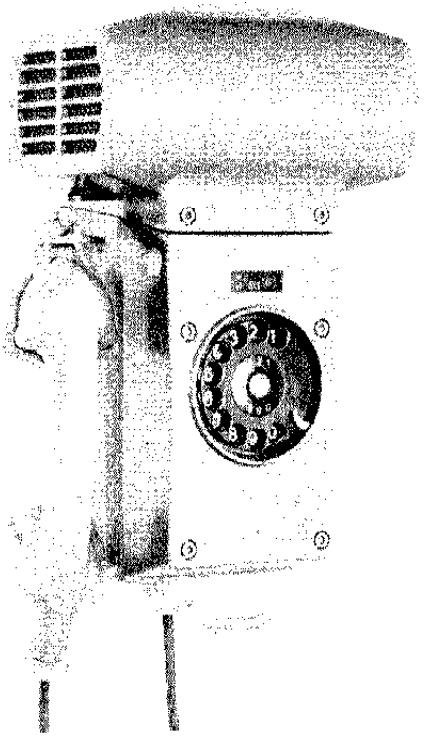


Figure 14—Telephone set for installation in mines and other places where explosive gases may be present.

Laser and Infrared Scope—The two boxes shown in Figure 15 house a gallium-arsenide laser and an image tube to convert the infrared output of the laser to visible radiation. This



Figure 15—Gallium-arsenide laser and image-tube scope to produce visible light from the infrared laser beam.

compact self-contained battery-operated equipment is suitable for laboratory experimentation.

The laser is of the cleaved gallium-arsenide type and has a very-low operating threshold of 4 amperes. It radiates an average output power of 90 milliwatts at a wavelength of 0.874 micron. The beam is pulsed at an adjustable rate between 100 and 1000 cycles per second. A liquid-nitrogen cooling head extends from the top of the box and a thermocouple is provided to monitor the temperature of the laser.

The scope is an infrared image tube that permits visual and photographic inspection of the laser beam. It can be equipped with either objective or eyepiece lenses or with both for the study of enlarged images.

ITT Industrial Laboratories
United States of America

Avalanche Rectifiers—Avalanche rectifiers are now available that are capable of absorbing reverse voltage transients. A unit rated for 1.25 amperes is able to dissipate reverse transient energy at a 4-kilowatt rate without damage. Previously, silicon rectifiers have required special networks or selenium rectifiers as surge absorbers.

Standard Telephones and Cables
United Kingdom

Transistors

Designated as "A Self-Instructional Programed Manual," this book was prepared under the editorship of I. L. Kosow by the training branch of Federal Electric Corporation, a subsidiary of International Telephone and Telegraph Corporation. The material is presented in the form of statements and questions, the answers to which follow immediately. Each answer is kept covered by the student until he has answered the associated question. If he is incorrect, the student should review enough to discover his error before going to the next question. The immediate testing of the understanding of the student is an important element in this new written teaching method of programed self-instruction.

The book is divided into the following 7 parts that are in turn subdivided into a suitable number of sets, each containing one major idea or circuit description.

1. Transistor Theory
2. Transistor Symbols
3. Rectification and Amplification
4. Oscillation
5. Modulation-Demodulation
6. Transistor Pulse and Switching Circuits
7. Logic and Gating Circuits

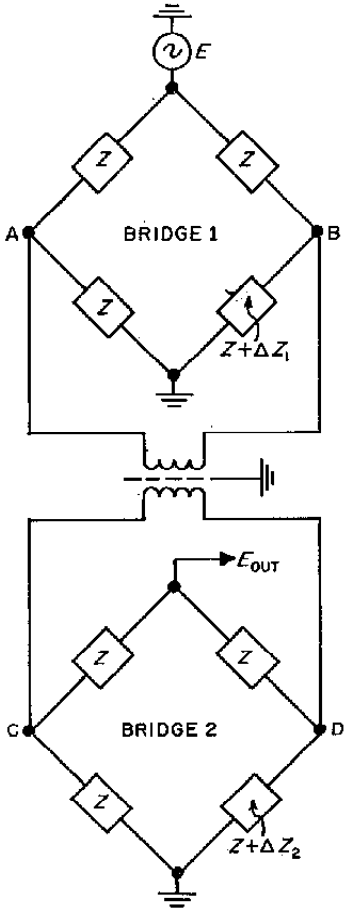
The page size is 9¼ by 6½ inches (23.5 by 16.5 centimeters) and the book contains 430 pages. It is published by Prentice-Hall, Englewood Cliffs, New Jersey, at $12.00 per copy.

Analog Signal Multiplier

O. S. MEIXELL
ITT Federal Laboratories, A Division of International Telephone and Telegraph Corporation; Nutley, New Jersey

1. Introduction

Communications and electronics are challenging fields for the application of a real-time analog signal multiplier. This is particularly true if the multiplier can furnish the true product of time-varying signals simultaneously with their variations. The multiplier described in this paper has been designed to achieve these objectives as closely as possible.



$$Z + \Delta Z_1 = Z(1 + \Delta Z_1/Z) = Z(1 + \mathcal{E}_1)$$
$$Z + \Delta Z_2 = Z(1 + \Delta Z_2/Z) = Z(1 + \mathcal{E}_2).$$

Thus, $\mathcal{E}_1$ and $\mathcal{E}_2$ are fractional unbalances of their respective bridges.

Figure 1—Basic multiplier. E is the alternating-current carrier excitation.

2. Basic Principles

The basic multiplier comprises two equiarm impedance bridges in cascade, with the unbalance or detector output of the first bridge applied as excitation for the second. This is illustrated in Figure 1.

The output E_{AB} of the first bridge is given closely by $k_1\mathcal{E}_1E$, where k_1 is a proportionality constant. Assuming an ideal unity-ratio transformer, this becomes the excitation for the second bridge, $E_{CD} = k_1\mathcal{E}_1E$. Correspondingly, the output of the second bridge is defined by $k_2\mathcal{E}_2E_{CD}$. Thus, $E_{\text{out}} = k_1k_2\mathcal{E}_1\mathcal{E}_2E$.

This relation shows that, for a given alternating-current voltage applied to the first bridge, the output of the second bridge is proportional to the product of the two fractional unbalances, $\mathcal{E}_1$ and $\mathcal{E}_2$. If these unbalances are caused by electrical or mechanical deviations, then the output of the second bridge is also proportional to the product of these deviations. For each bridge, the departure from linearity is only about half the fractional impedance unbalance.

A voltage-adjustable capacitor, or parametric diode, is an excellent means to implement the foregoing principles. The desired change of impedance ΔZ can be obtained merely by applying a signal voltage. This is accomplished over a very-wide frequency range at minimum cost of signal energy.

2.1 Fundamental Circuit

Two impedance bridges are used. They should be highly sensitive, with practically dissipationless elements that exhibit a conjugate-reactance relation at the bridge diagonal points. Figure 2 shows the fundamental circuit [1]. Obviously, there is no output voltage E_{out} when E_1 and E_2 are both zero. This is also true if either E_1 or E_2 is zero. However, the simultaneous presence of both voltages causes two incremental capacitances ΔC_1 and ΔC_2 to appear. Thus

$$C_v = C_{vo} + \Delta C_1$$
$$= C_{vo}(1 + \Delta C_1/C_{vo})$$
$$= C_{vo}(1 + \mathcal{E}_1)$$

for bridge 1, and

$$C_v = C_{vo} + \Delta C_2$$
$$= C_{vo}(1 + \Delta C_2/C_{vo})$$
$$= C_{vo}(1 + \mathcal{E}_2)$$

for bridge 2.

For peak analog voltages not exceeding about 10 percent of the direct-current voltage used to bias the voltage-adjustable capacitors, the fractional unbalances $\mathcal{E}_1$ and $\mathcal{E}_2$ vary closely with E_1 and E_2, respectively. As will be shown, E_{out} is therefore proportional to $Q^2E_1E_2E_c$, where E_1 and E_2 are the analog voltages, E_c is the carrier voltage, and $Q = \omega_o L/R_s$ with $\omega_o = 2\pi f_o$.

Thus, for a constant carrier voltage, the output is proportional to the product of E_1 and E_2.

In practice, a portion of the carrier voltage is applied, with proper phase, in series with E_{out}. This becomes an amplitude-modulated wave with a carrier frequency f_o and an envelope that is the product of E_1 and E_2. By demodulating the wave, the true instantaneous product of E_1 and E_2 is obtained. E_1 and E_2 may be any type of signal within the bandwidth capabilities of the two bridges. Thus, the products of sine waves, nonsinusoidal waves with or without direct-current components, square waves, pulses, noise, et cetera, can be formed.

2.2 Comparison with Other Multipliers

The product of two analog voltages is formed in essentially linear fashion. This is significant because the respective bridge unbalances caused by these voltages result from differential excursions over the voltage-adjustable capacitor's characteristic. In fact, the smaller the analog voltages are, the more linear is the formation of their product. As in other analog devices, there is a limit to the dynamic range that can be accommodated. The present model exhibits a 40-decibel range in its product. By refinement of the technique for obtaining quiescent (no-signal) bridge balance and by other means, this range can be extended materially. In contrast, the quarter-square technique [2] requires diodes or vacuum tubes that exhibit a square-law characteristic as closely as possible. Also, undistorted multiplication can take place only when the instantaneous sum or difference of the two analog signals is within the square-law region.

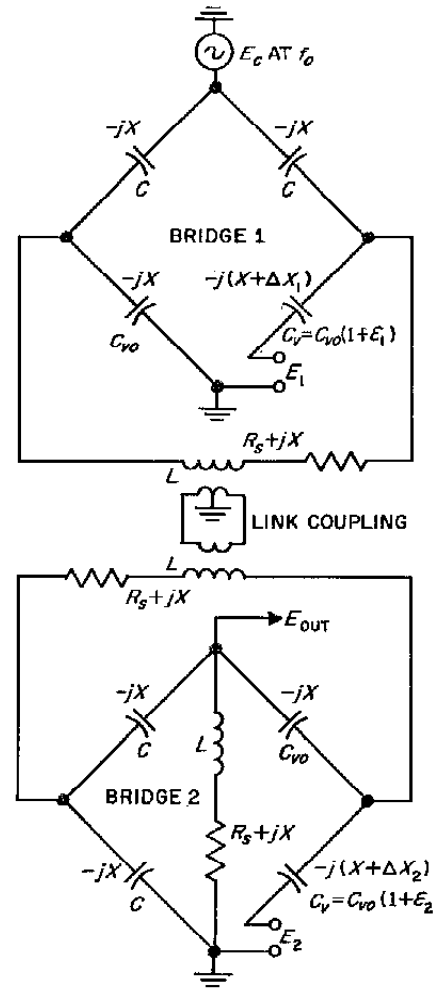


Figure 2—Multiplier with low-loss elements. C_v is a voltage-adjustable capacitor whose no-signal value $= C_{vo}$. C is the arm capacitance $= C_{vo}$. Therefore, both bridges are balanced when $E_1 = E_2 = 0$.

Multipliers using the Hall effect [3], [5], [7] require conversion into a high current for one analog voltage and a high magnetic field for the other. The latter conversion imposes a severe bandwidth limitation on the formation of the desired product.

Mechanical and electronic computers are available that perform multiplication and other mathematical operations accurately. The office calculating machine provides one-quadrant multiplication but not on a real-time or high-bandwidth basis. Electronic computers can perform four-quadrant multiplication. However, they are complex and expensive, and their signal bandwidth is limited because of the use of time-sharing and storage techniques, or for other reasons.

The analog signal multiplier presently does not have computer-like dynamic range and precision. However, it has unique advantages in its simplicity and ability to effect a four-quadrant low-noise product of complicated time functions over a large bandwidth.

2.3 Analytical Considerations

It might be very hard to develop and appraise a quantitative analysis of the multiplier output for generalized forms of E_1 and E_2. However, a set of conditions can be postulated that greatly simplifies the task and yet provides a representative picture of quantitative relations. Refer to the Appendix for an analysis based on the simplified assumption that the two analog voltages E_1 and E_2 are both sinusoidal, are at the same frequency, and differ only in amplitude and relative phase. Thus

$$e_1 = E_{1m} \sin \omega_s t$$

$$e_2 = E_{2m} \sin (\omega_s t + \theta_s)$$

where

$\omega_s = 2\pi f_s$

f_s = signal frequency

θ_s = phase angle between E_1 and E_2

When a carrier voltage $e_c = E_{cm} \sin \omega_c t$ is applied as excitation for the first bridge, the instantaneous output of the second bridge is described by

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} [2 \sin (\omega_c t) \cos \theta_s - \sin (\omega_c t - 2\omega_s t - \theta_s) - \sin (\omega_c t + 2\omega_s t + \theta_s)] \quad (1)$$

where k_o is a proportionality constant.

Two special conditions are instructive. The first is when E_1 and E_2 are in phase, and the second is when they are in quadrature.

For the in-phase condition $\theta_s = 0$, (1) becomes

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} [2 \sin \omega_c t - \sin (\omega_c t - 2\omega_s t) - \sin (\omega_c t + 2\omega_s t)]. \quad (2)$$

A 100-percent amplitude-modulated wave is described in (2). Its modulation envelope represents the instantaneous product of the two analog voltages, which can be restored by demodulation of the wave. Note that the frequency of this product is twice that of the analog voltages.

For the quadrature condition $\theta_s = -90$ degrees, (1) becomes

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} \times [- \sin (\omega_c t - 2\omega_s t + \pi/2) - \sin (\omega_c t + 2\omega_s t - \pi/2)]. \quad (3)$$

Equation (3) shows that only the sideband components of the wave exist. The envelope no longer describes the instantaneous product of the two analog voltages. To do so, a sufficiently strong fixed carrier component must be added to e_{out} before demodulation. If this fixed carrier component is in proper phase (90 degrees with respect to the impressed carrier), the envelope then becomes the true instantaneous product of E_1 and E_2 regardless of their relative phase.

The foregoing quantitative relations are illustrated in Figure 3 by waveforms derived from the multiplication of two sinusoidal analog voltages having the same amplitude and frequency but differing in relative phase (0, 90,

and 180 degrees). Row *A* shows the analog voltages and row *B* the corresponding outputs of the multiplier. These outputs are pictorial representations of (1). However, the output shown for the in-phase condition of e_1 and e_2 is described by (2) and for the 90-degree condition by (3).

If the multiplier outputs are rectified at this point, the resultant waveforms in row *C* would be false products except for the fortuitous case where e_1 and e_2 are in phase. To prevent this, a fixed carrier component must be added before demodulation. This results in the waveforms in row *D*. The carrier level shown is minimum; a higher level is advisable in practice, to improve rectification efficiency.

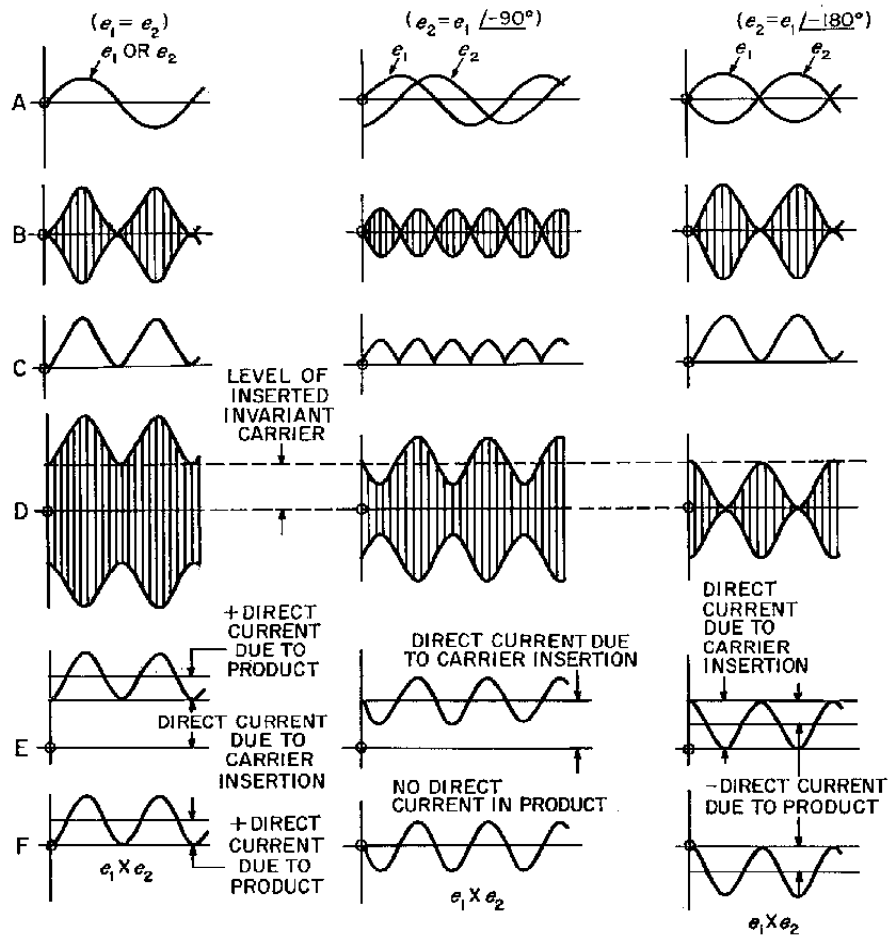


Figure 3—Analog signal multiplier waveforms.

A = analog voltages for 3 different phase relations.

B = products before insertion of invariant carrier.

C = demodulated false products.

D = products with inserted invariant carrier.

E = demodulated products.

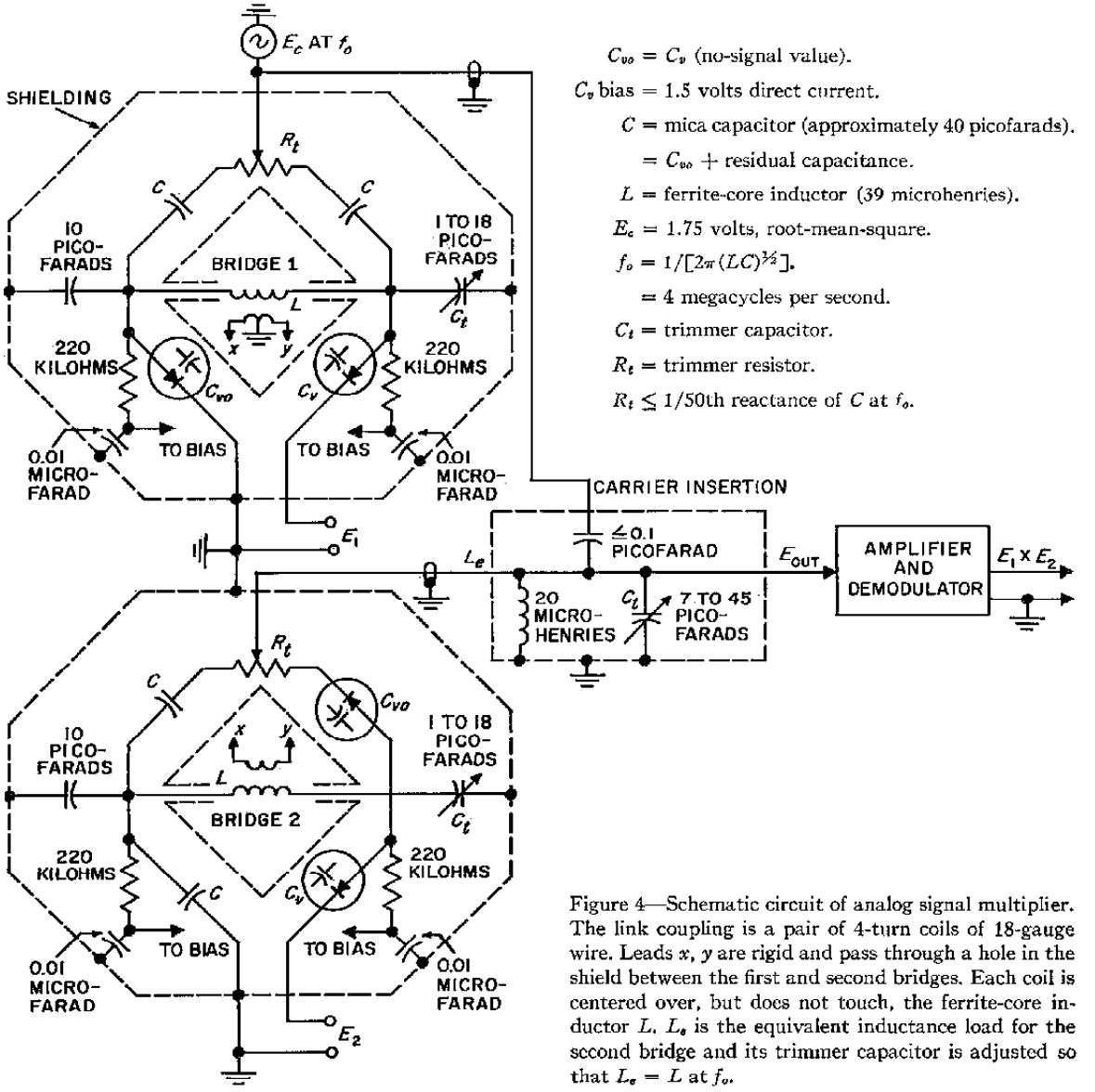
F = true products with direct-current voltage due to invariant carrier removed.

Row E shows that the demodulation now provides the product of the original analog voltages raised above a fixed direct-current level. Row F shows the true instantaneous products after the fixed level has been removed.

Note that row F represents the instantaneous power function when e_1 is proportional to the voltage applied to a load and e_2 is proportional to the current.

3. Design of Developmental Model

Figures 4 and 5 show, respectively, the schematic diagram and a photograph of the developmental model of the analog signal multiplier.



C_{vo} = C_v (no-signal value).

C_v bias = 1.5 volts direct current.

C = mica capacitor (approximately 40 picofarads).

= C_{vo} + residual capacitance.

L = ferrite-core inductor (39 microhenries).

E_c = 1.75 volts, root-mean-square.

$f_o = 1/[2\pi(LC)^{1/2}]$.

= 4 megacycles per second.

C_t = trimmer capacitor.

R_t = trimmer resistor.

$R_t \leq$ 1/50th reactance of C at f_o.

Figure 4—Schematic circuit of analog signal multiplier. The link coupling is a pair of 4-turn coils of 18-gauge wire. Leads x, y are rigid and pass through a hole in the shield between the first and second bridges. Each coil is centered over, but does not touch, the ferrite-core inductor L. L_e is the equivalent inductance load for the second bridge and its trimmer capacitor is adjusted so that $L_e = L$ at f_o.

It has four major connections. One is for application of the carrier voltage E_c to the first reactance bridge, located in the lower half of the structure. The analog voltage E_1 is also applied to this bridge. Through the action of the voltage-adjustable capacitor* in series with E_1, a suppressed-carrier modulation appears across the primary winding L of a special transformer which link-couples this output to the input of the second reactance bridge. The over-all coupling is made low and thus the inductance of the primary winding is virtually unchanged relative to its unloaded value. Furthermore, the carrier frequency f_o is such that the reactance of this high-Q winding is the conjugate of the reactance presented to it by the four capacitance arms of the bridge. Thus, under the quiescent and balanced condition $E_1 = 0$, if $+X$ is the reactance of L, each of the four capacitance arms has a reactance of $-X$. In simpler terms, this corresponds to a resonant condition of L and C at the carrier frequency.

The other analog voltage E_2 is correspondingly applied to the second of the reactance bridges, located in the upper half of the structure. Through its action on the associated Varicap, the suppressed-carrier modulated voltage coupled to the input of this bridge is again modulated to become E_{out}.

For sinusoidal analog voltages at the same frequency, instantaneous E_{out} appears as shown in row B of Figure 3. Adding the original invariant carrier voltage at the proper 90-degree phase results in the waveform of row D. This envelope displays the true instantaneous product of E_1 and E_2 for the conditions postulated.

* Varicap, type *PC-116-22*, a product of Pacific Semiconductors, Incorporated.

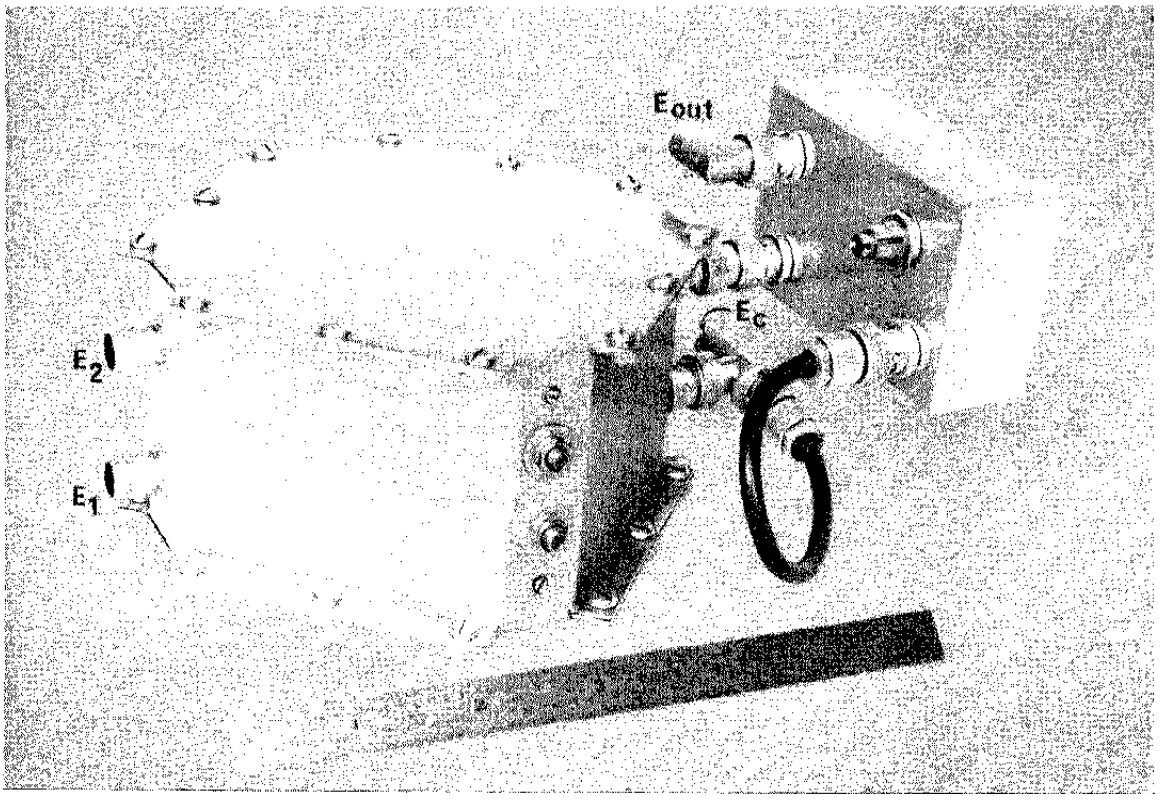


Figure 5—Analog signal multiplier. The smaller box contains the means for carrier insertion.

The method of coupling the output of the first bridge into the second is especially important. The over-all coupling must be strictly inductive and on a balanced-to-ground basis. There can be no capacitive or longitudinal coupling because a capacitance change caused by the analog signal of one bridge would also appear as a change in the other. The link coupling solves this problem since its impedance is so low that the induced voltage gradient across the winding is practically zero. When center tapped and grounded, as shown in Figure 4, the link-coupled windings constitute a Faraday shield that effectively prevents any capacitive coupling between the output of the first bridge and the input of the second.

An important design feature is the means used for dynamic compensation to improve linearity and stability of performance. Figure 4 shows an additional Varicap in the arm contiguous with that containing each signal-controlled Varicap. Note that in the first bridge half of the impressed carrier voltage E_c is impressed cophasally on each Varicap. Since these are identically poled, the variations in their capacitances caused by the carrier voltage are essentially equal and of like sign during its positive and negative excursions. Thus, these capacitance variations cause no unbalance of the bridge. Unbalance can occur only when E_1 is present. The second bridge works similarly and can only be unbalanced when E_2 is present.

Stability is also improved by the use of these additional Varicaps. Since they are all of the same type and characteristics, capacitance variations caused by temperature changes are of like sign and magnitude, and quiescent balance is maintained.

The source impedances of E_1 and E_2 should be low at f_o compared to the reactance of C at f_o and should offer a direct-current return path. If these requirements are not met, suitable shunt capacitance and resistance must be applied at the analog-signal input terminals. When a common analog source is used, as for obtaining the square of a signal, two aperiodic voltage dividers should be used between the source and the analog-signal input terminals for decoupling purposes.

The multiplier equivalent circuit in the Appendix shows that the permissible bandwidth of the analog signals is a function of the equivalent Q value of the inductors used in the link coupling and the output. It is also proportional to the carrier frequency. Since the intrinsic Q value of the ferrite-core inductors is high, suitable series or shunt resistance can be used to lower it, although at the expense of sensitivity. For the present model, the Q value of the inductive elements has been reduced to about 10 at the 4-megacycle-per-second carrier frequency. A product bandwidth has thus been observed from direct current to 400 kilocycles per second, which is 10 percent of the carrier frequency. It appears possible to modify the existing structure for operation with a 40-megacycle-per-second carrier and achieve a tenfold extension of the product bandwidth, to 4 megacycles per second.

Several design principles are useful regardless of the carrier frequency chosen. First, choose a voltage-adjustable capacitor whose Q value is high at the carrier frequency f_o. Use the lowest possible negative bias to obtain the largest usable quiescent capacitance C_{vo}, remembering that this bias should not be less than about half the carrier peak maximum $E_c/2$ plus the peak analog signal. The largest C_{vo} provides the highest possible sensitivity in the presence of unavoidable residual capacitances to ground. The arm capacitance C is the sum of the quiescent and residual arm capacitances. This establishes the inductance L in the output arm of the first bridge and the input arm of the second in accordance with $L = 1/\omega_o^2 C$.

Second, the mechanical layout of each of the constituent bridges should be symmetrical. All arms must be electrically isolated from each other by shielding. Also, the detector arm of the first bridge and the input arm of

the second should be isolated carefully from the bridge arms containing the fixed and voltage-adjustable capacitors. The second bridge must be isolated from the first except for the balanced link coupling which should be mechanically and electrically symmetrical.

Third, the output of the multiplier should be amplified 20 to 40 decibels before demodulation. This can be accomplished best by a band-pass amplifier with a center frequency at f_o and a bandwidth sufficient to accommodate the sidebands in the product of E_1 and E_2.

Most signal sources are single-ended. Also, a usable direct-current component may reside in the analog signal. The present model is adapted to these conditions. However, if balanced-to-ground or push-pull analog signals are used, distortion can be reduced by applying E_1 and E_2 suitably to both Varicaps in their respective bridges. Of course, if transformers are used for this purpose, any direct-current signal components are lost and the bandwidth is limited to that of the transformer.

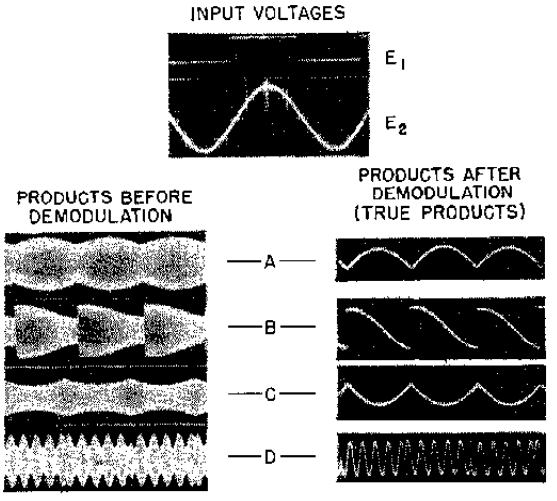


Figure 6—Products of rectangular and sine waves. E_1 and E_2 are the input signals at 300 cycles per second. Their amplitudes are 0.16 and 0.30 volt, respectively, peak to peak. The carrier is 1.75 volts (root-mean-square) at 4.1 megacycles per second. Oscilloscope sweep = 500 microseconds per centimeter.

A = E_1 and E_2 in phase.

B = E_2 advanced 90 degrees.

C = E_2 advanced 180 degrees.

D = E_2 increased from 300 to 3000 cycles per second.

4. Multiplier Performance

Figures 6 through 9 illustrate typical multiplier output waveforms for various analog signal voltages. Figure 6 shows the products of a rectangular wave and sinusoids. Note that discontinuities are cleanly formed and little distorted, confirming the high bandwidth capability and good dynamic range.

Figure 7 shows the products of sinusoidal analog voltages. The first group is a superimposed display of the products of two 300-cycle-per-second sine waves at relative phase displacements of 0, 90, and 180 degrees. Observe that these are replicas of the waveforms shown in row F of Figure 3. Further evidence of bandwidth and dynamic range capability is provided by the two remaining displays of Figure 7, where the frequency of the second sinusoid is increased 10 times and then 100 times above the first, with no change in amplitude.

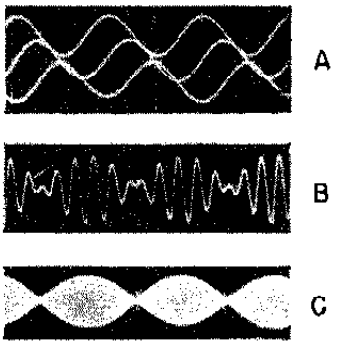


Figure 7—Products of sine waves. The amplitude of each voltage is 0.30 volt peak to peak. Oscilloscope sweep = 500 microseconds per centimeter. At A, the top curve is the product for E_1 and E_2 in phase. The middle and bottom curves of A are for E_2 shifted 90 and 180 degrees, respectively. E_1 and E_2 are each at 300 cycles per second and the displays are superimposed.

At B and C, are the products of E_1 at 300 cycles per second and E_2 at 3000 and 30 000 cycles per second, respectively.

The square of a triangular wave, shown in Figure 8, consists of contiguous parabolas. Also, the bottom waveform in this figure demonstrates the capability to multiply two signals each having a direct-current component (in this case, equal to half the peak-to-peak value of the analog).

Figure 9 demonstrates the capability to obtain the product of two identical noise functions; first as instantaneously cophasal and, second, in an instantaneous antiphasal relationship. Observe that the product for the cophasal condition contains a positive direct-current component, while the product for the antiphasal condition contains a negative one.

5. Multiplier Applications

One of the most important applications of an analog multiplier is as a correlator to improve the signal-to-noise ratio of various types of signals. For example, when a noisy signal, $f(t)$ + noise, is multiplied by a noise-free replica that periodically passes through time coincidence, sufficient integration can reveal the presence of the signal even when it is deeply immersed in noise. By using this cross-correlation technique [4], the analog signal multiplier has detected sinusoidal signals having a signal-to-noise ratio of 1/30. Of course, the bandwidth of the signal information that can be restored by this process suffers with increased integration time. The signal-to-noise ratio can be improved materially for noisy pulse-like signals when they

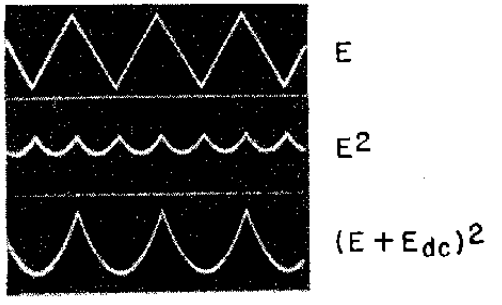


Figure 8—Square of a triangular wave. E = 920 cycles per second at 0.07 volt peak. E_{dc} = peak value of E.

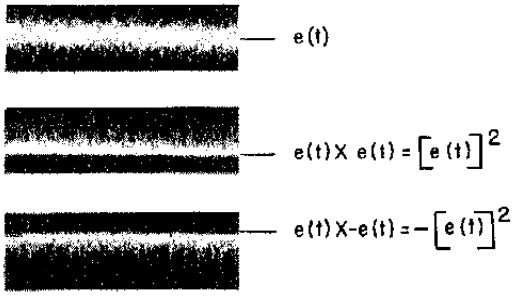


Figure 9—Products of white noise over a band from zero to 20 kilocycles per second.

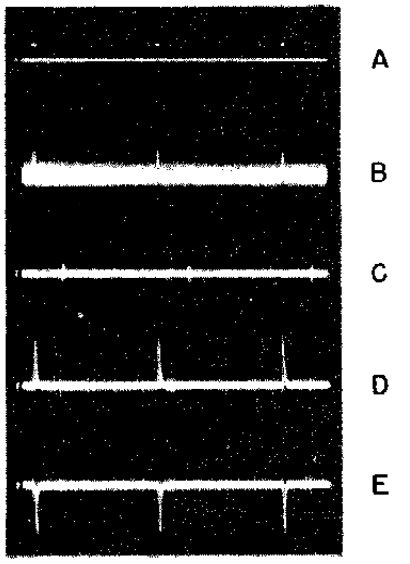


Figure 10—Products of noisy and noise-free pulses. Pulses are of 10-microsecond duration with a repetition frequency of 1000 per second. Oscilloscope sweep = 200 microseconds per centimeter. The light output of the oscilloscope was integrated over a 0.1-second exposure for each display on Pola Pan 200 (Type 42) film. The lens was adjusted to f5.6. Electrical time constant is about 22 microseconds.

A = signal pulses E_1 at 0.03 volt. (The correlating pulses E_2 at 0.3 volt are not shown.)

B = E_1 + 0.01 volt root-mean-square white noise from zero to 500 kilocycles per second.

C = E_1 + noise × E_2, not correlated.

D = E_1 + noise × E_2, correlated.

E = $-E_1$ + noise × E_2, correlated.

are multiplied by time-coincident noise-free pulses. This is illustrated in Figure 10. Here the signal-to-noise improvement is obtained with only a moderate amount of integration. Additional improvement should result from cross-correlating the product in a second multiplier.

An analog multiplier can be employed as a cross-correlator for signal recognition or analysis. The use for signal recognition contemplates the formation of the product of an incoming signal with voltages having various normalized trial waveforms. On time-sweeping with these waveforms and integrating the various products, the largest of the maxima identifies the waveform that is closest to the unknown. Williams and Denker [5] describe the application of a Hall-effect multiplier for analysis of a nonsinusoidal wave. It appears that the bridge-type multiplier can implement this analysis technique more readily and, certainly, over a greater bandwidth.

Each bridge of the analog signal multiplier constitutes a low-level linear modulator. The device (in toto) offers a means for a wide variety of modulations. Before delineating these in detail, it will be helpful to compare more-conventional modulation techniques with those used in the bridge-type analog multiplier.

It is common practice to obtain amplitude modulation of a carrier by applying the carrier and signal voltages as inputs to a circuit that has a nonlinear output-input characteristic. For various types of vacuum-tube, transistor, or diode modulators, this relation may be represented by the series expansion

$$e_o = k_1 e + k_2 e^2 + k_3 e^3 + \cdots + k_n e^n$$

where e_o is the output, $e = e_c + e_s$ (the carrier-plus-signal input), and k_1 through k_n are amplitude coefficients.

Let the first three terms suffice to describe a typical nonlinear characteristic. Substituting $e = e_c + e_s$, we obtain

$$e_o = k_1 e_c + k_1 e_s + k_2 e_c^2 + 2k_2 e_c e_s + k_2 e_s^2 + k_3 e_c^3 + 3k_3 e_c^2 e_s + 3k_3 e_c e_s^2 + k_3 e_s^3. \quad (4)$$

The terms $k_1 e_c + 2k_2 e_c e_s$ in (4) afford undistorted amplitude modulation for a modulation index of less than unity. The other terms are not wanted since they include modulation of the carrier by a square of the signal, modulated and unmodulated second and third harmonics of the carrier, et cetera. Normally, most of these are removed by the use of a band-pass filter that accommodates only the desired carrier frequency and sideband components.

Assuming $e_c = E_{cm} \sin \omega_c t$ and $e_s = E_{sm} \sin \omega_s t$, then

$$e_o = k_1 e_c + 2k_2 e_c e_s = k_1 E_{cm} \sin \omega_c t + 2k_2 E_{cm} \sin \omega_c t E_{sm} \sin \omega_s t. \quad (5)$$

This result can also be achieved more directly by a linear multiplication process such as occurs in either bridge of the analog signal multiplier. Within its allowable dynamic range, the output-input characteristic of each bridge is defined by $e_o = k e_c e_s$ where k = amplitude coefficient.

Assume a carrier $e_c = E_{cm} \sin \omega_c t$ is applied as excitation for the first bridge. Let $e_s = e_{dc} + E_{sm} \sin \omega_s t$ be applied to the analog signal port of this bridge (see Figure 4). Thus

$$e_o = kE_{cm} \sin \omega_c t (e_{dc} + E_{sm} \sin \omega_s t) = ke_{dc} E_{cm} \sin \omega_c t + kE_{cm} \sin \omega_c t E_{sm} \sin \omega_s t. \quad (6)$$

Except for the constant coefficients, (6) is the same as (5). Accordingly, undistorted modulation can be realized directly without the unwanted products that characterize usual practice. This can be accomplished at signal levels as low as a fraction of a volt. In fact, the lower the levels of carrier and signal, the more linear is the formation of the modulation. Any

small extraneous products are removed by the band-pass action of the bridge network.

Table 1 summarizes the modulation and related capabilities of the analog multiplier when energized by a carrier voltage E_c.

This versatile modulation capability permits additional uses. For example, a version of the multiplier can serve as an electronic phase shifter of the carrier. Two bridges, identical in form with the first bridge, would be excited in parallel; the first by E_c and the second by E_c shifted 90 degrees. If $E \sin \alpha$ is applied to the E_1 port and $E \cos \alpha$ to the E_2 port, the vector sum of the outputs of the two bridges, taken in series, is the original carrier with a phase shift of α. When $\alpha = \omega_s t$, it can be shown that the vector sum of the outputs consists solely of the lower-sideband modulation of E_c and the vector difference consists solely of the upper sideband. This single-sideband modulation technique has been anticipated [6] using conventional types of modulators but use of the bridge structures for this purpose deserves consideration because they constitute low-level linear modulators having a minimum of complexity.

Waveform *A* of Figure 6 obviously duplicates the full-wave rectification of a sinusoid by conventional means. The upper group of waveforms in Figure 7 suggests use of the multiplier either as a synchronous detector [7] or for the detector function in a synchrodyne type of receiver. The latter is treated in the literature under other designations such as homodyne and product-detector types of receivers.

6. Extension of Design

Size can be reduced advantageously in the present model and extension to higher carrier frequencies appears feasible, with further increase of allowable bandwidth in the product. For designs at carrier frequencies higher than 50 megacycles per second, the present Varicap is not suitable at the low bias employed. Other types of parametric diodes may have to be used singly or in parallel groups. The use of ultra-high-frequency and microwave carriers

TABLE 1

MULTIPLIER MODULATION CAPABILITIES

Modulation or Process	Signals to be Applied		Desired Modulation or Process Appearance
	at E_1 Port	at E_2 Port	
Modulation of E_c by E:			
Suppressed carrier	E	E_{dc}	E_{out} port with no inserted E_c, or by coupling to the input of the second bridge, which is not used.
Normal amplitude modulation	$E + E_{dc}$	E_{dc}	
Modulation of E_1 by E_2 or vice versa:			
Suppressed carrier	E_1 (or E_2)	E_2 (or E_1)	Output of amplifier-demodulator whose input is E_{out} + inserted E_c.
Amplitude modulation of E_1 by E_2	E_1	$E_2 + E_{dc}$	
Amplitude modulation of E_2 by E_1	$E_1 + E_{dc}$	E_2	
Commutation or phase reversal of:			
E_c (method 1)	$\pm E_{dc}$ steps	E_{dc}	Same as for modulation of E_c by E.
E_c (method 2)	E_{dc}	$\pm E_{dc}$ steps	E_{out} port with no inserted E_c.
E (method 1)	E	$\pm E_{dc}$ steps	Same as for modulation of E_1 by E_2.
E (method 2)	$\pm E_{dc}$ steps	E	
On-off switching of:			
E_c	Same as for commutation of E_c and E, respectively, but with $\pm E_{dc}$ step = 0 for off condition.		
E			

will no doubt require distributed-parameter bridge elements. Extension of designs into these areas will lead to many more-challenging applications.

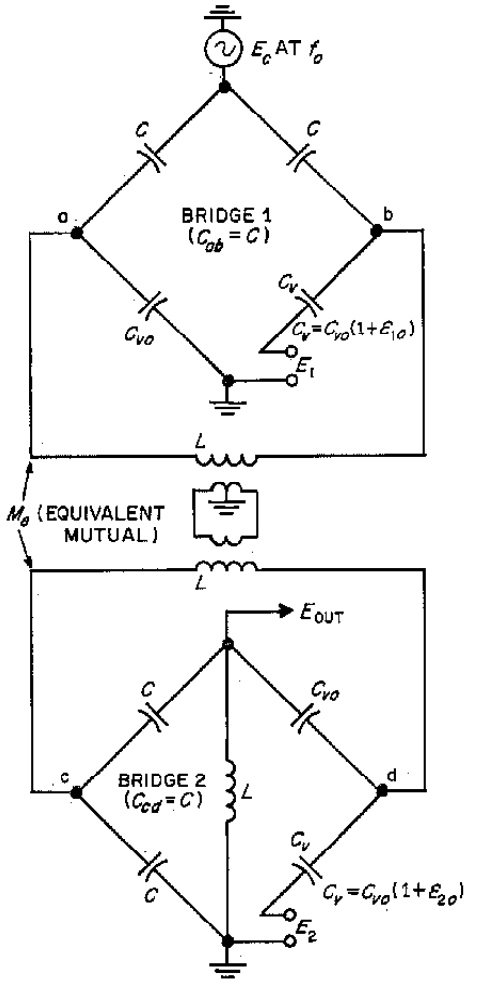


Figure 11—Multiplier equivalent circuit.

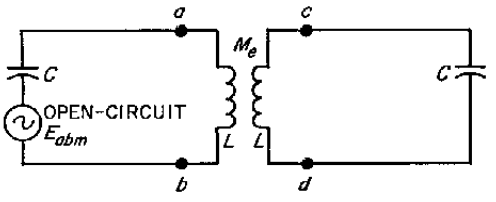


Figure 12—Thévenin's equivalent circuit.

7. Acknowledgements

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9. Appendix

The multiplier equivalent circuit is shown in Figure 11.

C_v = voltage-adjustable capacitance (no-signal value = C_{vo})

C = total arm capacitance
($C = C_{vo}$ + fixed residual capacitance)

L = inductance arm

Carrier frequency $(f_o) = 1/[2\pi(LC)^{1/2}]$.

Let $\mathcal{E}_1$ be fractional unbalance of first bridge, then

$$\mathcal{E}_1 = \frac{\Delta C_1}{C} = \frac{\Delta C_1}{C_{vo}} \frac{C_{vo}}{C} = \mathcal{E}_{1o} \frac{C_{vo}}{C}.$$

ΔC_1 is caused by E_1 and C_{vo}/C takes residual capacitance into account.

For $\mathcal{E}_1 \leq 0.05$

$$\text{open-circuit } E_{ab} = \frac{\mathcal{E}_1 E_c}{4} \text{ (very closely)}$$

$$\text{open-circuit } E_{ab} = \frac{\mathcal{E}_{1o}}{4} \frac{C_{vo}}{C} E_c$$

$$\text{open-circuit } E_{abm} = \frac{\mathcal{E}_{1o}}{4} \frac{C_{vo}}{C} E_{cm}$$

where subscript m = maximum value.

Characteristics of a Varicap are such that

$$\frac{\Delta C_1}{C_{vo}} = \mathcal{E}_{1o} = \frac{E_{1m}}{2E_{bo}} \text{ (closely for } E_{1m} \lesssim 0.1\, E_{bo})$$

where

E_{1m} = maximum value of E_1

E_{bo} = applied direct-current bias.

Accordingly,

$$\text{open-circuit } E_{abm} = \frac{1}{8} \frac{E_{1m}}{E_{bo}} \frac{C_{vo}}{C} E_{cm}$$

$$= K_1 E_{1m} E_{cm} \qquad (7)$$

where

$$K_1 = \frac{1}{8} \frac{C_{vo}}{CE_{bo}}.$$

This leads to the Thévenin's equivalent circuit shown in Figure 12. For critical coupling $(1/Q)$ or a coupling up to twice this value (or $2/Q$)

$$E_{cdm} \doteq \frac{Q}{2} E_{abm}$$

where Q is the Q value of the identical primary and secondary inductors L in the presence of coupling.

From (7)

$$\text{open-circuit } E_{abm} = K_1 E_{1m} E_{cm}$$

$$\therefore E_{cdm} = \frac{Q}{2} E_{abm} = K_1 \frac{Q}{2} E_{1m} E_{cm}. \qquad (8)$$

With further application of Thévenin's theorem, it can be shown readily that the output of the second bridge is

$$E_{\text{out } m} = j \frac{Q}{8} \frac{E_{2m}}{E_{bo}} \frac{C_{vo}}{C} E_{cdm} = jQK_2 E_{2m} E_{cdm} \qquad (9)$$

where

Q = Q value of load L on second bridge

$$K_2 = \frac{1}{8} \frac{C_{vo}}{CE_{bo}}.$$

When the bridges are identical, with the same Q values and with a common bias E_{bo} applied to all Varicaps, $K_1 = K_2 = K$. Substituting (8) in (9), we obtain

$$E_{\text{out } m} = jK^2 (Q^2/2) E_{1m} E_{2m} E_{cm}. \qquad (10)$$

Equation (10) shows that, for a constant E_c, the output of the second bridge is proportional to the product of E_1 and E_2.

Rewriting (10) in terms of instantaneous sinusoidal voltages

$$e_{\text{out}} = k_o e_1 e_2 e_c$$

where

$$k_o = jK^2(Q^2/2)$$
$$e_1 = E_{1m} \sin \omega_s t$$
$$e_2 = E_{2m} \sin (\omega_s t + \theta_s)$$
$$e_c = E_{cm} \sin \omega_o t$$
$$\omega_o = 2\pi f_o$$
f_o = carrier frequency.

Thus

$$e_{\text{out}} = k_o E_{1m} E_{2m} E_{cm} [\sin \omega_s t] \times [\sin (\omega_s t + \theta_s)] [\sin \omega_o t].$$

By straightforward trigonometric development

$$e_{out} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} [2 \sin (\omega_o t) \cos \theta_s - \sin (\omega_o t - 2\omega_s t - \theta_s) - \sin (\omega_o t + 2\omega_s t + \theta_s)]. \quad (11)$$

When e_1 and e_2 are in phase ($\theta_s = 0$), (11) becomes

$$e_{out} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} \times [2 \sin (\omega_o t) - \sin (\omega_o t - 2\omega_s t) - \sin (\omega_o t + 2\omega_s t)]. \quad (12)$$

Equation (12) is for a 100-percent-modulated carrier whose envelope is the true product of E_1 and E_2.

When e_1 and e_2 are in quadrature ($\theta_s = 90$ degrees), (11) becomes

$$e_{out} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} \times [- \sin (\omega_o t - 2\omega_s t - \pi/2) - \sin (\omega_o t + 2\omega_s t + \pi/2)]. \quad (13)$$

Only the upper and lower sidebands exist in this case. The envelope is not the true product. It becomes so upon insertion of a carrier component of proper phase and having sufficient magnitude.

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He spent five years as an engineer with the Western Electric Company. From 1936 to 1943, he was an instructor in communication and electronics at RCA Institutes in New York, becoming a department head in 1938. During 1943 and 1944, he was a senior instruction-book writer for the Hazeltine Electronics Corporation.

In 1944, he joined ITT Federal Laboratories and is presently technical assistant to the director of the Electronic Defense Laboratory.

Mr. Meixell has been granted four patents. He is a Senior Member of the Institute of Electrical and Electronics Engineers.

Transmission-Loss Measurements and Diversity Tests Made on Two Knife-Edge Diffraction Paths in Europe *

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ITT Communication Systems Inc., on loan from ITT Federal Laboratories, a Division of International Telephone and Telegraph Corporation; Nutley, New Jersey

1. Introduction

As part of a program of transmission-loss measurements on beyond-the-horizon paths in the Mediterranean area, tests were made on two knife-edge diffraction paths having a common receiving terminal near Genoa in northern Italy. The tests were made at a frequency of 1965 megacycles per second in August 1961.

The path-loss measurements indicated that the loss on one path was about 27 decibels less, and on the other 8 decibels less, than would be expected for normal scatter paths having the same angular distances.

Diversity tests, using receiving antennas spaced 47 feet (14.3 meters) horizontally and 10 feet (3 meters) vertically, showed that there was little correlation on either path in the received carrier-level variations at the two antennas.

From the transmission-loss measurements and diversity tests, it is estimated that high-quality 72-voice-channel systems could be established on these paths, using 10-kilowatt transmitters and 60-foot (18.3-meter) antennas in the 2000-megacycle-per-second band.

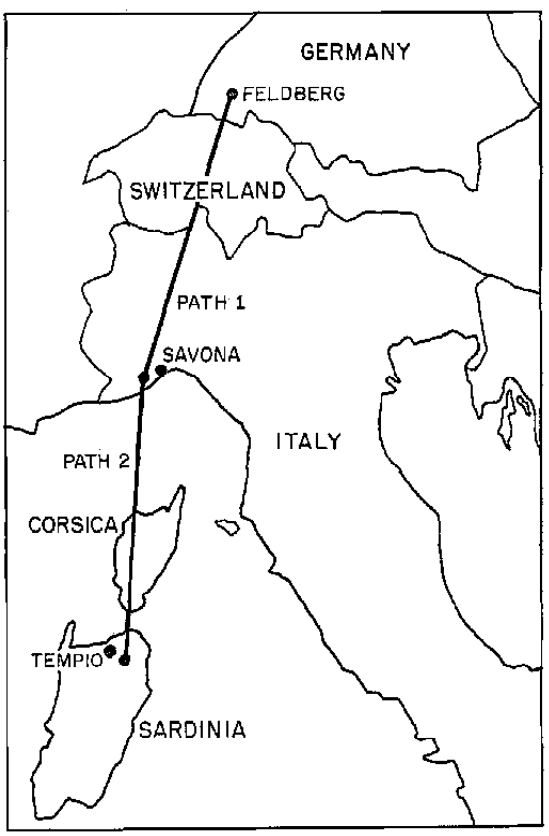


Figure 1—Transmission paths.

2. Transmission Paths

Path 1 extended from Feldberg, in southern Germany, across the Alps to a site near Savona in Italy; and path 2 from a point near Tempio, in northern Sardinia, across Corsica to the same receiving site. Figure 1 shows the location of the two paths.

The characteristics of the two transmission paths were as follows:

	Savona–Feldberg	*Savona–Tempio*
Height of transmitting terminal above sea level in feet (meters)	4900 (1493)	3970 (1210)
Height of receiving terminal above sea level in feet (meters)	3370 (1029)	3370 (1029)
Height of common obstacle above sea level in feet (meters)	13 200 (4030)	5400 (1650)
Length of path in statute miles (kilometers)	251 (404)	240 (385)
Distance from receiving terminal to common obstacle in statute miles (kilometers)	159 (256)	134 (215)
Distance from transmitting terminal to common obstacle in statute miles (kilometers)	92 (148)	106 (170)
Angular distance in milliradians (assuming $K = 4/3$)	53	28.2

* The tests described were part of a program of measurements made by ITT Communication Systems, Inc., under Air Force Contract AF 19(626)-4, "Eur-Med Tropo Scatter Communication System."

To allow for normal refraction, the effective radius of the earth is about one-third longer than the true radius, and K is defined as the ratio of the effective radius to the true radius. Figures 2 and 3 show the profiles of the two paths, assuming a 4/3 earth radius.

3. Equipment

3.1 Transmitting Installations

3.1.1 *Feldberg Site*

A 10-watt crystal-controlled transmitter operating on 1965 megacycles per second and a 28-foot (8.5-meter) Kennedy-type transportable antenna were installed at this terminal.

3.1.2 *Tempio Site*

Two 10-watt crystal-controlled transmitters, one operating on 1965 and the other on 1965.3 megacycles per second, and two 10-foot (3-meter) antennas were installed at this site. The centers of the antennas were spaced 55 feet (16.8 meters) horizontally and 15 feet (4.6 meters) vertically at right angles to the transmission path.

3.2 Receiving Installation

At the receiving site near Savona, two crystal-controlled narrow-band (500-cycle-per-second) path-loss-measuring receivers and two antennas, one 28 feet (8.5 meters) and the other 10 feet (3 meters) in diameter, were employed. The receiving antennas (see Figure 4) could be directed toward either Feldberg or Tempio and were spaced 47 feet (14.3 meters) horizontally and 10 feet (3 meters) vertically at right angles to the transmission path. The output of each receiver went to a separate recorder, scaled linearly in decibels. The time constants of the recorders could be varied from

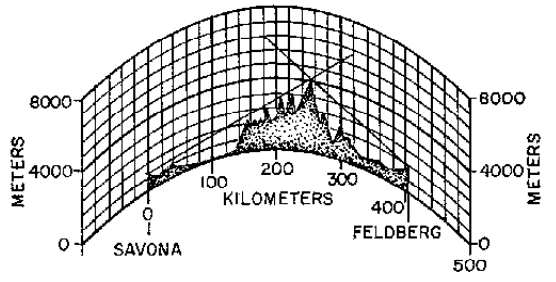


Figure 2—Profile of Savona-Feldberg path for $K = 4/3$.

TABLE I
Vertical Spacing Between Antenna Centers

Path	Calculated Antenna Spacing in Feet (Meters) for $K = 4/3$	
	Receiver	Transmitter
Feldberg–Savona	11.4 (3.5)	6.0 (1.8)
Tempio–Savona	27.9 (8.5)	15.4 (4.7)

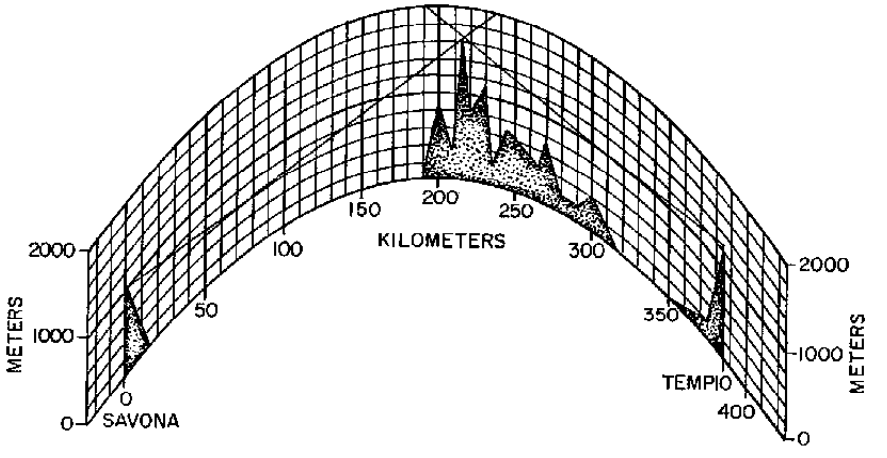


Figure 3 — Profile of Savona-Tempio path for $K = 4/3$.

3 to 50 seconds, and the speed of the recorder charts could be varied from 1 inch (25 millimeters) per minute to 1 inch per 15 minutes. Each receiver had a microammeter indicating the rectified receiver output; the over-all time constant of these meter indications was less than 1 second.

3.3 Antenna Spacing

In a space-diversity system, the antennas are spaced so that there is little correlation between the received level variations at the two antennas. On a normal scatter path, a horizontal antenna spacing of about 100 wavelengths is found to be adequate to reduce the correlation coefficient to a value of less than +0.4. On a knife-edge diffraction path, however, interference may exist between a direct ray and one or more rays reflected from points between the terminals and the knife-edge. In this case, vertical antenna spacing is required, and the correct distance between antennas may be calculated from the path profile. Assuming a single reflection at each side of the knife-edge, Table 1 shows the calculated vertical spacing between antenna centers necessary to ensure that when one antenna is at a minimum, the other is at or near a maximum of the interference pattern formed by the direct and indirect rays.



Figure 4—Receiving antennas at Savona site.

4. Transmission-Loss Measurements

Path-loss measurements were made at 1965 megacycles per second, first on the Feldberg-Savona path for about 600 hours, and then on the Tempio-Savona path for about 500 hours. Figures 5 and 6 show the distribution of the 15-minute medians on the two paths.

Table 2 shows the measured median loss, the calculated loss assuming knife-edge diffraction, and the estimated annual loss for paths having the same angular distances as those measured.

The calculated loss is based on the method shown in National Bureau of Standards Report 6767, Part 5, Sections 6–13, June 1961, and the estimated annual median loss is taken from an empirical curve of the average of a large number of transmission-loss measurements made on tropospheric-scatter paths.

The Feldberg-Savona path showed a higher diffraction gain than did the Tempio-Savona path. This may be due to the fact that the knife-edge on the Tempio-Savona path was the rounded side of a mountain on Corsica, whereas the knife-edge on the Feldberg-Savona path was a relatively sharp ridge in the Alps. The fading recorded on the two paths was very similar to that normally associated with tropospheric-scatter paths.

It has been estimated from the test results that the median loss for the worst hour of the worst month at 2600 megacycles per second for the Feldberg-Savona link would be 234.6 decibels. Using this value, the performance in

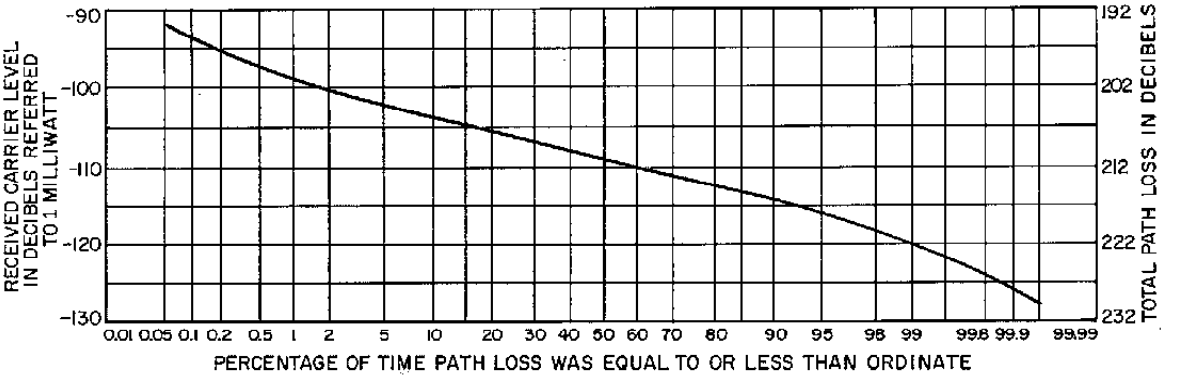


Figure 5—Path-loss results, Feldberg-Savona, for 2439 compilations of 15-minute medians.

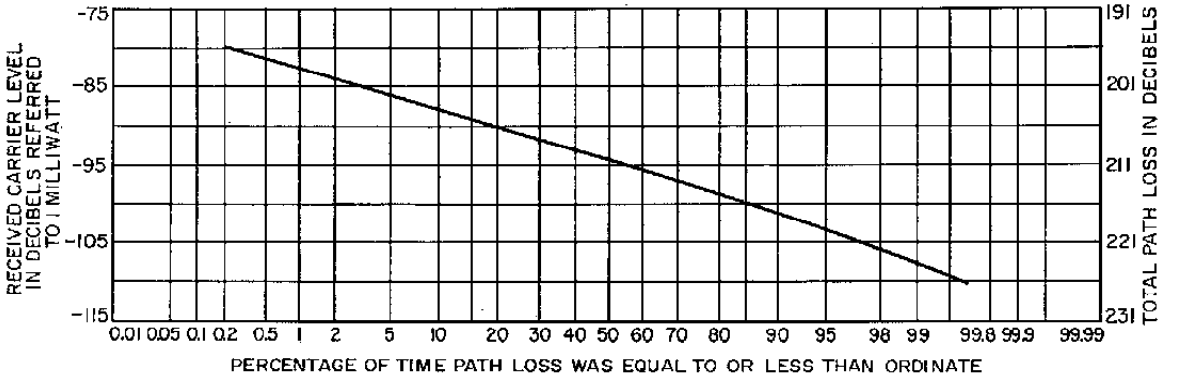


Figure 6—Path-loss results, Tempio-Savona, for 2029 compilations of 15-minute medians.

the worst channel of a 72-voice-channel system during the worst hour of propagation is outlined in Table 3.

It is re-emphasized that Table 3 represents the performance for the median of the worst hour of the year.

Since the measured diffraction gain on the Tempio-Savona path was only 8 decibels compared with 27 decibels on the Feldberg-Savona path, it is estimated that the seasonal change in loss would be about 4 decibels greater on the Tempio-Savona path.

Thus, using the same power and antenna size, the performance during the worst month on the Tempio-Savona path is likely to be about 4 decibels poorer than on the Feldberg-Savona path.

TABLE 2
MEDIAN LOSSES IN DECIBELS

Path	Measured Median Loss	Calculated Median Loss	Estimated Median Loss
Feldberg–Savona	210	198	237
Tempio–Savona	211	188	219

TABLE 3
WORST PERFORMANCE CHARACTERISTICS

Transmitter power in kilowatts	10
Antenna diameter in feet (meters)	60 (18.3)
Antenna gain in decibels	86
Transmission loss in decibels	234.6
Transmission-line loss in decibels	3
Received carrier level in decibels referred to 1 milliwatt	−81.6
Carrier-to-noise ratio in decibels*	25
Bandwidth improvement in decibels	27.4
Frequency-modulation improvement in decibels (100-kilocycle-per-second root-mean-square deviation per channel)	−8.1
Weighting, pre-emphasis, and 4-fold diversity improvements in decibels	13.5
Channel signal-to-noise ratio in decibels	57.8

* Assuming 2-decibel noise-figure receivers with parametric amplifiers.

5. Diversity Measurements

Dual space-diversity tests were made by recording simultaneously the received carrier level on two receivers at the Savona terminal, and also by noting the reading on the two receiver output meters.

5.1 FELDBERG-SAVONA PATH

For diversity tests on this path, a transmitter and a single antenna were used at Feldberg, and two receivers with spaced antennas were used at Savona. The received carrier level was recorded on each receiver using a recorder with a time constant of about 3 seconds and a chart speed of 1 inch (25 millimeters) per minute.

5.1.1 *Diversity Test Results*

The results obtained are given in Figures 7 through 9, which show sample values taken from the recorded carrier levels. These were obtained at different times during a 4-day period in August 1961. The difference in median received level on the two antennas is a function of the difference in gain of the antennas and the difference in loss of the two antenna transmission lines; 10 divisions on the recorded scales represent a 5-decibel difference in receiver level.

5.1.2 *Correlation Coefficient*

The calculated correlation coefficients between the recorded carrier levels in decibels as shown in Figures 7 through 9 are given in Table 4.

5.2 TEMPIO-SAVONA PATH

For diversity tests on this path, two transmitters and two spaced antennas were available at Tempio, and two receivers with spaced antennas were used at Savona. The Tempio transmitters operated at slightly different frequencies so that the carriers could be separated at the Savona receivers. During diversity

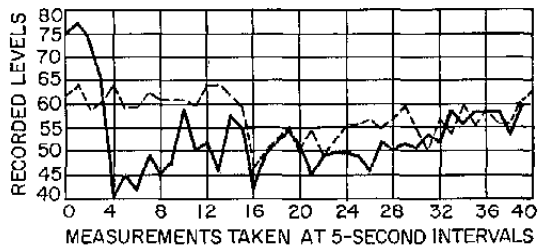


Figure 7—Recorded levels of measurements taken at 5-second intervals, Feldberg-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was +0.22.

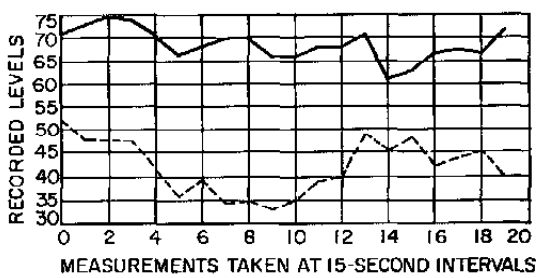


Figure 8—Recorded levels of measurements taken at 15-second intervals, Feldberg-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was +0.26.

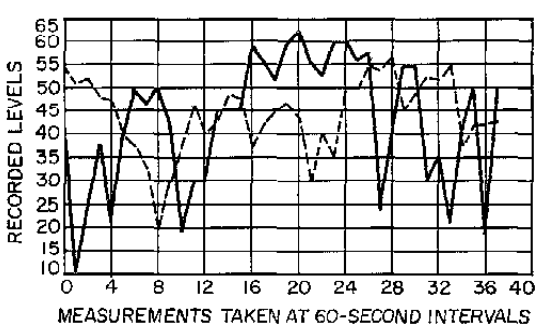


Figure 9—Recorded levels of measurements taken at 60-second intervals, Feldberg-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was −0.26.

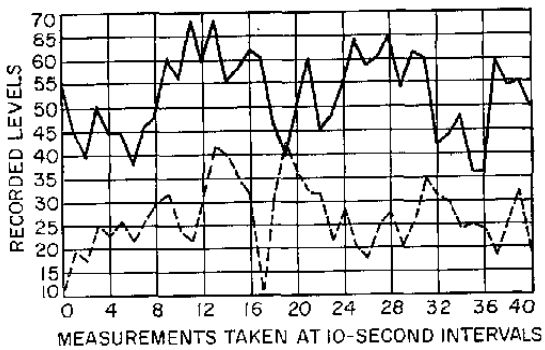


Figure 10—Recorded levels of measurements taken at 10-second intervals for one transmitting and two receiving antennas, Tempio-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was +0.05.

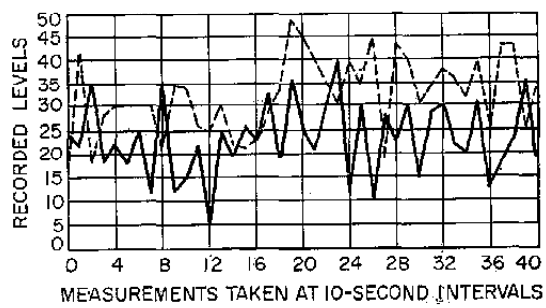


Figure 11—Recorded levels of measurements taken at 10-second intervals for two transmitting and two receiving antennas, Tempio-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was −0.08.

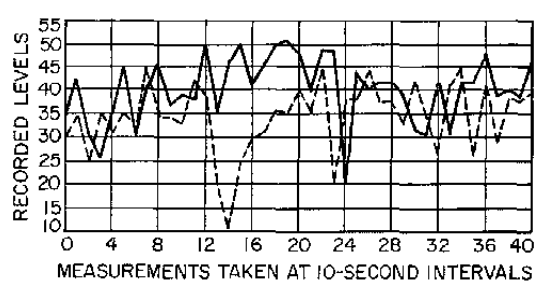


Figure 12—Recorded levels of measurements taken at 10-second intervals for two transmitting and one receiving antennas, Tempio-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was −0.15.

TABLE 4
CALCULATED CORRELATION COEFFICIENTS, FELDBERG–SAVONA

Time Interval in Seconds	Correlation Coefficient
5	+0.22
15	+0.26
60	−0.26

tests on this path, the following antenna combinations were used.

(A) One transmitting and two receiving.
(B) Two transmitting and two receiving.
(C) Two transmitting and one receiving.

5.2.1 *Diversity Test Results*

Figures 10 through 12, respectively, show samples of the results obtained with each of the 3 combinations of antennas.

5.2.2 *Correlation Coefficient*

The calculated correlation coefficients between the recorded carrier levels in decibels for the various antenna combinations are given in Table 5.

TABLE 5
CALCULATED CORRELATION COEFFICIENTS, TEMPIO-SAVONA

Antenna Combination	Time Interval in Seconds	Correlation Coefficient
1 transmitting and 2 receiving	10	+0.05
	60	−0.27
	900	+0.28
2 transmitting and 2 receiving	10	−0.08
	15	−0.14
	60	+0.10
2 transmitting and 1 receiving	10	−0.15
	15	−0.0046
	60	−0.16

6. Conclusion

Path-loss measurements and space-diversity tests were made at 1965 megacycles per second on two knife-edge diffraction paths in Europe. On each path, the measured loss was less than was estimated for normal beyond-the-horizon paths having the same angular distance, but was greater than the values calculated from knife-edge diffraction theory.

Diversity tests were made with antennas spaced 47 feet (14.3 meters) horizontally and 10 feet (3 meters) vertically. Correlation coefficients calculated from the recorded carrier level variations in decibels were in all cases less than +0.30.

A suggested cause of this low correlation is that in each case the obstruction was not a smooth horizontal knife-edge, but a ridge varying in elevation across the path. Such an obstruction would result in multipath transmission with an expected low correlation of amplitude variations on spaced receiving antennas and with the fading characteristics associated with multipath transmission. The recorded fading was similar to that observed on normal scatter paths.

It is estimated that high-quality 72-voice-channel systems could be established over these paths using 10-kilowatt transmitters and 60-foot (18.3-meter) antennas in the 2000-megacycle-per-second band. A Feldberg-Savona multichannel system would be of particular interest since it would form a direct communication link between Italy and Germany.

R. E. Gray was born in Wynford, England, on 15 February 1902. He received a First Class Diploma, equivalent to a bachelor's degree, from Faraday House, London, in 1924. From 1924 to 1927, he was with Standard Telephones and Cables, London, and from 1927 through 1939 was with Laboratoire Central de Télécommunications, Paris. During the

second World War, he was on loan to the Royal Aircraft Establishment and to the Telecommunication Research Establishment in England.

He participated in the first microwave communication link between France and England, in the introduction of the ground-controlled-approach aircraft landing system in England, and in the original commercial transatlantic radiotelephone service. He was also active in the installation of the first Spain-Argentina radiotelephone link, as well as the multichannel radio link joining Scotland and Ireland.

In 1945, he joined ITT Federal Laboratories, where he is presently a senior scientist. He continues his radio wave propagation studies with particular attention to forward-scatter systems.

He was on loan to ITT Communication Systems for the survey discussed in this issue.

Eaglesfield Receives Institution of Electrical Engineers Premium

Charles Cecil Eaglesfield has received the Premium of the Institution of Electrical Engineers for his paper on "Optical Pipelines—A Tentative Assessment," which was published in the Proceedings of that society, volume 109, part 13, pages 26–32; January 1962.

Mr. Eaglesfield joined Standard Telephones and Cables in 1947 and was transferred to Standard Telecommunication Laboratories ten years later. He has been associated with the application of waveguides to long-distance communication and is presently concerned with the feasibility of communication at optical wavelengths.

Marine Transmitter with Single-Sideband Facility

F. HEDSTRÖM
Standard Radio & Telefon AB; Bromma, Sweden

1. Introduction

Efforts have been concentrated in recent years on decreasing the emitted bandwidth of high-frequency radio transmitters to make best use of the very-crowded bands. Some of the methods used to reach this goal have been careful pulse shaping in radiotelegraphy and advanced single-sideband techniques in radiotelephony. Both techniques require the use of linear power amplifiers, since the amplification that follows modulation or keying must be linear all the way to the antenna. The input to the linear amplifier must therefore have the correct frequency and the required kind and degree of modulation; frequency multiplication may not be used after modulation.

In marine radio communication, telephony is increasing in importance. Distances are often very long and transmission quality must be good to allow interconnection with the public telephone network. Since directive antennas are not practicable on shipboard, the high efficiency of single-sideband operation is a great advantage.

A modern marine transmitter must provide for telegraphy of both unmodulated (*A1*) and

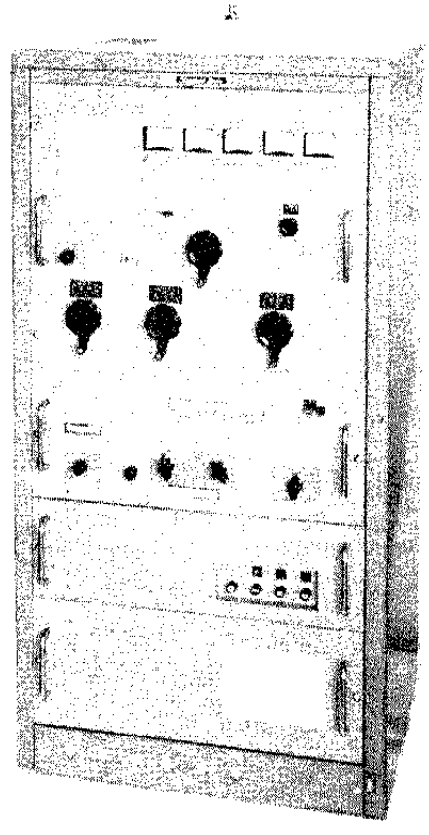


Figure 1—*ST 1200* transmitter. From top to bottom are shown the linear power amplifier, exciter unit, control unit, power supply, and fuses and connection unit.

audio-frequency modulated (*A2*) types, telephony with single-sideband and reduced carrier (*A3A*), and, to comply with international regulations, telephony with full carrier (*A3H*).

The lowest limit of the transmitter's output power (250 watts for some countries) is also based on full-carrier double-sideband operation. This corresponds to a peak effective power of 1 kilowatt from a linear power amplifier. Thus the power output of the *ST 1200* transmitter was designed to be 1.2 kilowatts peak effective power to include a margin for aging of tubes and other losses.

The main parts of the transmitter, which is shown in Figure 1, are:

(A) A linear power amplifier covering the frequency ranges from 400 to 535 kilohertz and from 1.6 to 25.5 megahertz. The linear power amplifier includes a matching network for the antenna impedances specified by different national rules to permit proper loading of the various types of shipboard antennas.

(B) A power supply with control devices for the linear power amplifier.

(C) An exciter unit that produces a signal of the required frequency and modulation at a power level suitable for the input of the linear power amplifier.

2. Linear Power Amplifier

2.1 Design Problems

To assure linear operation and also to reduce the load on the penultimate stage, the output tubes should give reasonable efficiency even when not driven into the grid-current region. The efficiency of such tubes is about 64 percent when driven to the grid-current point. Thus a peak effective power of 1.2 kilowatts gives an anode dissipation of 675 watts. The coupling network between the output stage and the antenna introduces losses of about 15 percent, so the output power from the tubes must be increased by about this amount. The efficiency of a linear stage decreases with decreasing output power. When the output power is zero, all the power delivered to the anodes is dissipated as heat. If this minimum level of plate input is lowered, the linearity of the stage deteriorates. Thus, as a function of linearity, this minimum level determines the required anode dissipation. As a rule of thumb, the obtainable peak effective power for modern tubes not driven into the grid-current region is about 1.2 times the allowable anode dissipation. Accordingly, the required anode dissipation in this case is about 1 kilowatt. The factor of 1.2 applies to tubes with a relatively low anode voltage of about 2000 volts; the factor increases for tubes with higher anode voltage.

Eimac tube type *4X250B* was chosen for the transmitter. To get the required power output, four of these tubes were connected in parallel. A single tube, such as Eimac type *4CX1000A*, could be employed but the type *4X250B* is more widely available. Also its anode voltage is only 2000 volts compared with 3000 volts for the *4CX1000A*. At the highest operating frequency of about 25 megahertz, 4 high-slope tubes can be operated in parallel by careful design of the input and output circuits.

Figure 2—Typical antenna coupling network for medium-frequency range from 415 to 535 kilohertz.

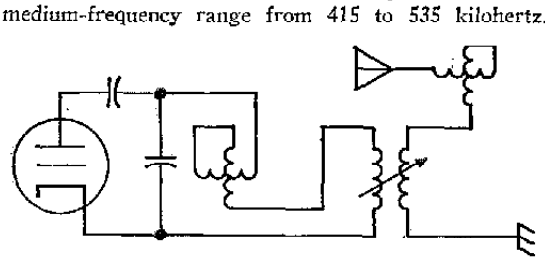
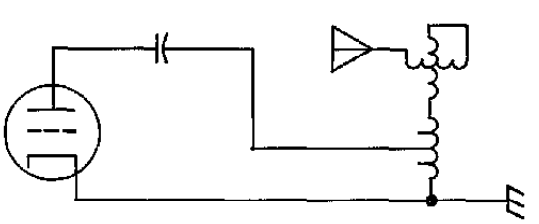


Figure 3—Simplified antenna coupling network for medium-frequency range.



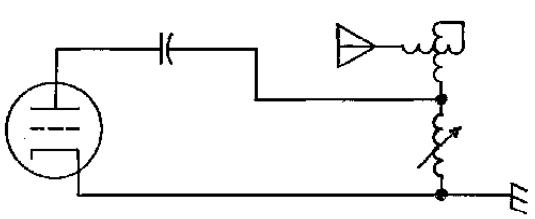
Another problem was the design of the matching network between the output stage and the antenna. Two different types of networks were required, one for the 400-to-535-kilohertz range and one for the 1.6-to-25.5-megahertz range. In the medium-frequency range, the Q and thus the selectivity of the antenna is always very high. The required selectivity for suppressing harmonics can easily be obtained in the antenna circuit without the need for extra tuned circuits or filters. The high Q of the antenna causes the antenna current and voltage to be high. Because of this, antenna circuit components tend to be rather bulky and expensive.

2.2 Antenna Coupling Networks

A commonly used medium-frequency antenna coupling network is shown in Figure 2. The anode and antenna circuits are separately tuned while the coupling between them is adjustable. The circulating currents are high in the two tuned circuits, especially in the antenna circuit, which requires careful design of components.

The coupling circuit used for the transmitter is shown in Figure 3. The single tuning element is an adjustable inductor in series with the antenna to form a tuned circuit with the antenna capacitance. Taps on the inductor are used to match the antenna resistance to the optimum loading resistance of the output stage. This circuit works only with an antenna of capacitance reactance but this is always the case with antennas in the medium-frequency range aboard ship. Under special conditions, a capacitor can be connected in series with the antenna to shorten it electrically. Figure 4 shows the practical circuit; to avoid taps on the inductor, a second adjustable small inductor is connected in series with the antenna tuning inductor and forms a continuously adjustable tap.

Figure 4—Practical antenna coupling network for medium-frequency range.



For the 1.6-to-25.5-megahertz range, another type of coupling network is required. Antennas in this frequency range cannot be depended on to suppress harmonics and the selectivity of the coupling network must be high. Thus three different requirements are imposed on the network:

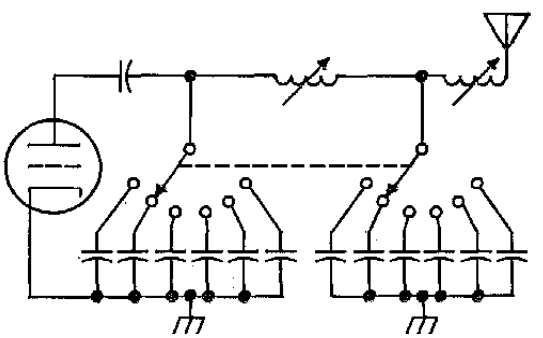
(A) Correcting the phase angle of the load (antenna) to zero.

(B) Transforming the antenna resistance to the correct value for optimal loading of the output stage.

(C) Providing the required selectivity to get sufficient suppression of harmonics.

To satisfy the first two requirements, at least two adjustable controls must be introduced in the network. Many different types of networks can be used and the most common one is the π filter. However, the harmonic suppression obtained with a simple π filter is limited and two components, one a capacitor, must be adjustable. Adjustable air dielectric capacitors are very bulky at this output power and vacuum capacitors are expensive. The problem was solved eventually by using a π-L filter as

Figure 5—π-L antenna coupling network used from 1.6 to 25 megahertz.



shown in Figure 5. The two adjustable components are now inductors, allowing the use of small and inexpensive fixed capacitors. It was possible to design the inductors to be usable over the entire frequency range. The fixed capacitors have to be switched in 6 bands.

In the coupling network for the medium-frequency range, it was explained that a small adjustable inductor was used in series with the antenna inductance. To minimize the number of components in the complete network, this same inductor is used as the loading coil in the π-L filter. Figure 6 shows the complete tuning and coupling network for all ranges.

The amount of amplification required from the linear power amplifier depends on the input power level and the input impedance. An input power level of 0.1 watt was chosen as a compromise among such considerations as the stability of the linear power amplifier, the amount of power available from inexpensive receiving-type tubes, and the number of amplifier stages required. For standardization, an input impedance of 50 ohms was chosen.

2.3 Preamplifier

The preamplifier is shown in Figure 7, which is a simplified schematic diagram of the entire linear power amplifier. Two preamplifier stages are employed, one a grounded-grid and the other a conventional grounded-cathode stage. A grounded-grid circuit is used in the first stage for impedance matching and stability without neutralization. The couplings between the first and second stages and to the grids of the output stage are tuned circuits. Because of the rather high impedances in the grid circuits of the penultimate and output stages, both must be neutralized. Neutralization holds for the complete frequency range and does not require adjustment during normal operation of the transmitter. Both tuned circuits are provided with continuously adjustable inductors. In combination with the switched fixed capacitors of the tuning network, they cover the entire medium-high and high-frequency range with a single coil per circuit. The coils for both circuits are ganged. In the medium-frequency range, broad-band circuits are used and no tuning is required.

2.4 Tuning Indicators

The tuning of a linear power amplifier must be carried out with precision or the linearity will be adversely affected. To simplify the tuning process, the preamplifier and the output stage are provided with phase discriminators

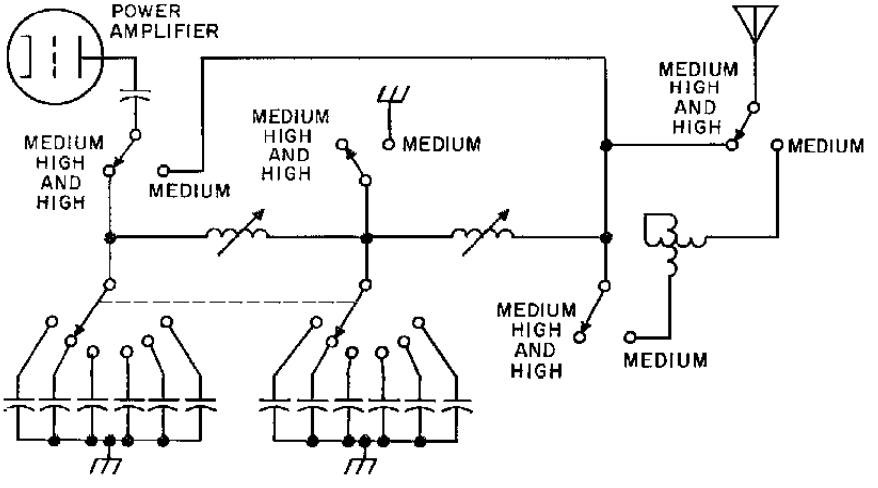


Figure 6—Complete antenna coupling network.

that indicate the resonant condition of the tuned circuits. The output stage is also provided with an amplitude-comparison circuit (loading discriminator) and all the discriminators are provided with zero-center meters.

3. Exciter Unit

The exciter unit provides the linear power amplifier with the proper input signal, which must have the required frequency plus the appropriate kind and degree of modulation. Therefore the exciter must contain two different circuits, a frequency-generating circuit and a modulating circuit.

3.1 Frequency-Generating Circuit

The number of frequencies required from a marine transmitter has increased very rapidly during the past few years and their harmonic relation, which was once a rule, does not fully hold today. This means that the number of crystals for a modern marine telegraph and telephone transmitter is very high; the *ST 1200* transmitter is equipped with 94 replaceable and switchable crystals. For the medium-high-frequency band, 1.6 to 3.8 megahertz, the crystal frequency is the same as the output frequency. For the high-frequency band, 4 to 26 megahertz, the crystal frequencies range from 2 to 3 megahertz and frequency multiplication is used to get the required output frequencies. Crystal ovens are not needed for the frequency stabilities required by international regulations.

When higher frequency stabilities are required for certain types of transmission, other methods must be employed. The *ST 1200* transmitter obtains the required output frequency by mixing frequencies from three different sources:

(A) A crystal oscillator in the 2-to-3-megahertz range.

(B) A crystal oscillator adjustable from 41.2 to 66.2 megahertz in 1-megahertz steps.

(C) A 9-megahertz amplitude-modulated frequency derived from a 1-megahertz crystal oscillator.

The simplified block diagram of Figure 8 illustrates the frequency-generating circuit for the

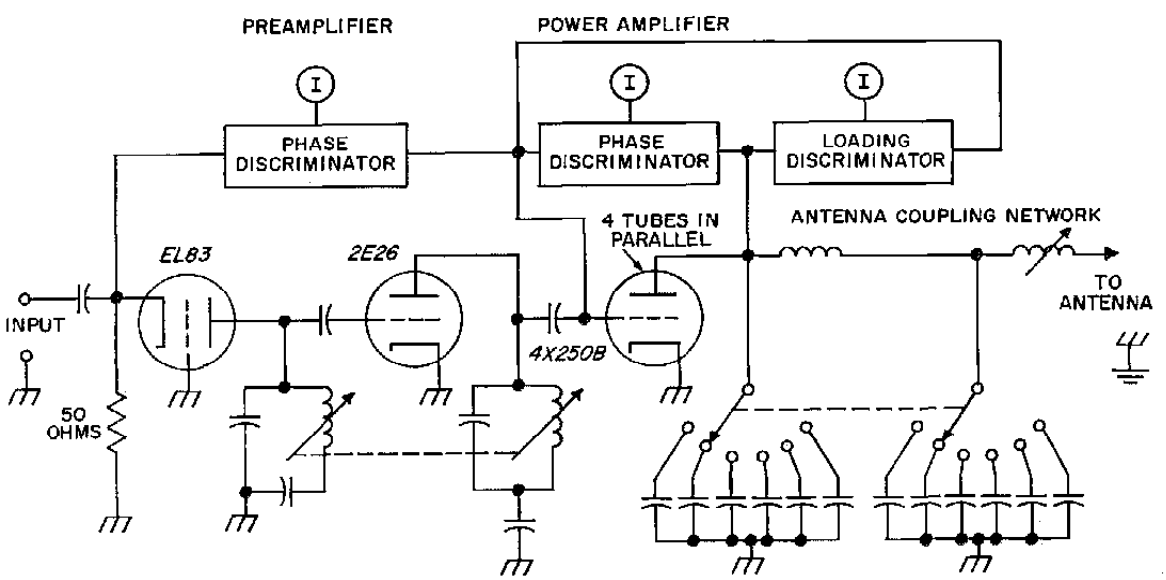


Figure 7—Diagram of the linear power amplifier.

medium-high and high frequencies. The 2-to-3-megahertz crystal oscillator is called the channel oscillator and the 41.2-to-66.2-megahertz oscillator is called the band oscillator. The reasons for these designations are that the channel oscillator generates the frequency required within a band, whereas the band oscillator determines the required frequency band.

The output from the channel oscillator is subtracted in mixer *M1* from the output of a correction oscillator that has been tripled to 34.2 megahertz. The difference frequency of approximately 31 to 32 megahertz passes through a band-pass filter to mixer *M2*, where it is added to the 9-megahertz amplitude-modulated frequency. The resulting 40-to-41-megahertz frequency and a frequency from the band oscillator are impressed on mixer *M3*. The frequencies from the band oscillator occur only at 1-megahertz intervals from 41.2 to 66.2 megahertz. The exact frequency from the band oscillator depends on the required frequency band, and the output from mixer *M3* is the required output frequency with the modulation contained in the 9-megahertz signal supplied to mixer *M2*.

If, for instance, the required frequency is 12 440 kilohertz, then the crystal frequency for the channel oscillator is 2440 kilohertz. The output frequency from mixer *M1* is 31 760 kilohertz, and the output frequency from mixer *M2* is 40 760 kilohertz, which means that the output frequency from the band oscillator must be 53.2 megahertz.

Frequency errors caused by the band oscillator are corrected as follows. The band-oscillator output is also supplied to mixer *M4*, which mixes it with sharp pulses having a repetition frequency of 1 megahertz controlled by a 1-megahertz crystal oscillator. Mixer *M4* develops a frequency spectrum containing mixer products between the band-oscillator frequency and the 1-megahertz harmonics. The 34.2-megahertz mixed product is supplied through a band-pass filter to a phase discriminator, which is also connected to the correction-oscillator circuit. The phase discriminator controls the frequency of the correction oscillator by means of a reactance element. Thus the output frequency of the tripler following the correction oscillator is exactly the same as the output frequency from mixer *M4* selected by

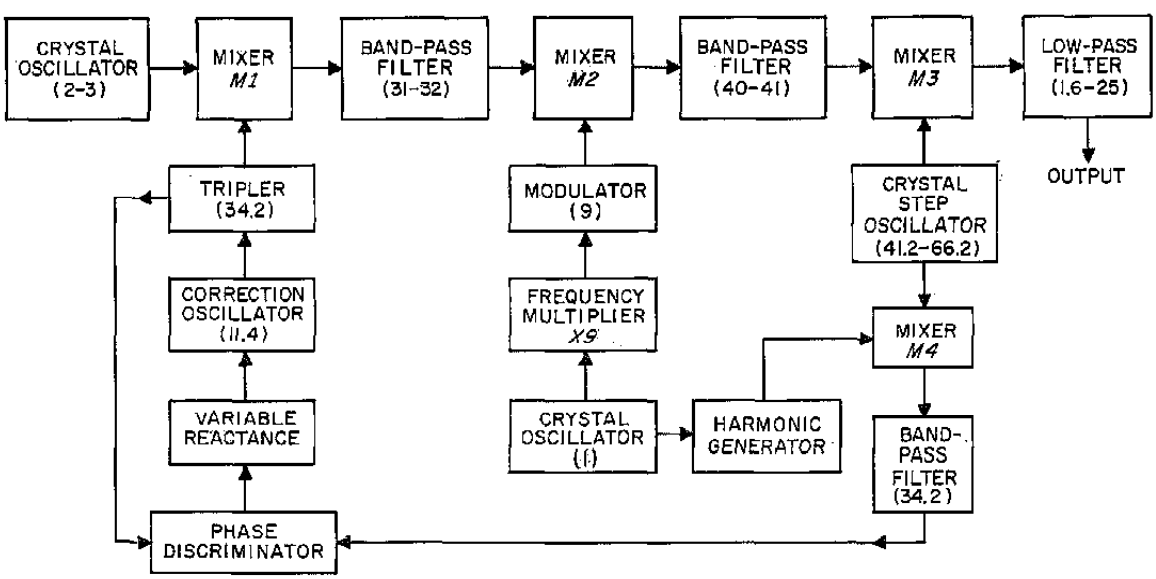


Figure 8—Diagram of frequency-generating circuit used on medium-high-frequency and high-frequency bands. The frequencies within parentheses are in megahertz.

the 34.2-megahertz band-pass filter. If the band-oscillator frequency varies by ΔF, the frequency from mixer *M4* also changes by this amount causing the frequency of the oscillator and tripler to change by the same amount. Since the latter frequency is also supplied to mixer *M1*, the frequencies in the chain from mixer *M1* to mixer *M3* will change. In this way ΔF is cancelled, and the only frequency errors are the sum of those in the highly stable channel oscillator and 1-megahertz oscillator. The correction oscillator with its associated tripler should be considered only as a filter and is necessary to minimize the spurious signals supplied to mixer *M1*.

An amplifier is used to raise the output level from mixer *M3*. This amplifier uses two wideband stages, with a gain control in the first stage. The nominal output is 0.1 watt peak effective power into 50 ohms.

3.2 Modulating Circuit

All modulation of the transmitter, including *A1* and *A2* keying, is introduced into mixer *M2* of the frequency-generating circuit. The 9-megahertz signal supplied to this mixer must contain all the necessary modulation.

Figure 8 shows that the 1-megahertz crystal oscillator supplies a signal to a frequency multiplier with an output frequency of 9 megahertz. The 9-megahertz signal is supplied via two separate amplifiers to two modulators, one for *A1* modulation and the other for *A3* modulation, as shown in Figure 9. The *A1* modulator supplies one of the inputs to a summation amplifier and the *A3* modulator supplies the other via the single-sideband filter. The output of the summation amplifier is supplied to mixer *M2* in the frequency-generating circuit.

The modulating signals, from either a microphone or a telegraph key, are supplied to the modulators via a switching unit, controlled by push buttons and rotary switches on the front panel. For *A1* emission, the telegraph key interrupts a direct-current source to produce the pulses representing telegraph signals. The direct-current pulses are routed through the switching unit to the *A1* modulator. A low-pass filter shapes the pulses to minimize the emitted bandwidth. Since the transmitter is

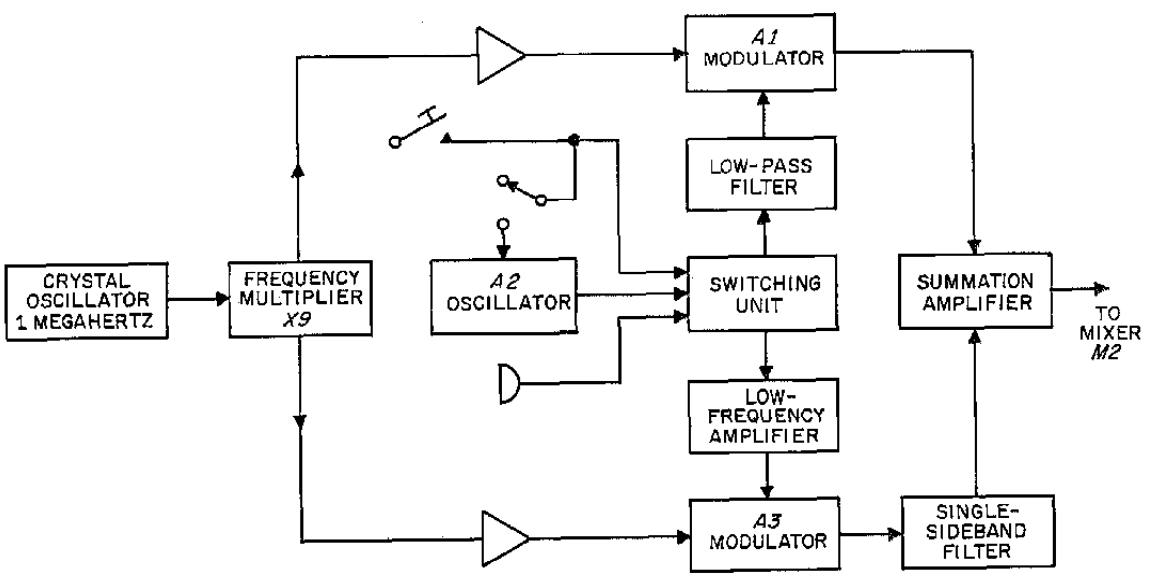


Figure 9—Diagram of modulating circuit.

linear from this point on, the pulse shape undergoes no further change. To suppress keying noises, the telegraph key operates a keying unit that provides the output amplifier with a blocking voltage.

For *A2H* modulation, in addition to the *A1* operation described above, the telegraph key also operates an *A2* oscillator whose output goes via the switching unit and the low-frequency amplifier to the *A3* modulator, which emits the sideband. Accordingly, both the *A1* and *A3* modulators are used during *A2H* operation; the *A1* modulator supplies the keyed carrier and the *A3* modulator supplies the keyed sideband.

For voice operation, *A3A* and *A3H*, the *A3* modulator also supplies the sideband and the *A1* modulator supplies the carrier. Adjustment of the *A1* modulator controls the carrier level. A switch, usually placed in the microphone handle, turns the transmitter on and off in the same way that the telegraph key does. The low-frequency amplifier that increases the output signals from the *A2* oscillator is also used to amplify the voice signals.

4. Power Supply and Cooling System

Separate power supplies are used for the exciter unit and the linear power amplifier. The latter supply provides 2000 volts for the output stage. This power supply is a 3-phase full-wave rectifier with the high-voltage transformer designed for either delta or star connection of the primary, to allow the transmitter to be connected to a source of either 220 or 380 volts. The rectifier is a compact unit provided with silicon elements.

The screen-grid and control-grid voltages for the output stage are developed by electronically regulated power supplies.

The transmitter is forced-air cooled by two blowers. One blower cools the output tubes and the tuning coils, especially the antenna tuning coil for the medium-frequency band. The second blower cools the silicon rectifiers and transformers. Its air stream is also ducted to the control and exciter units. Both blowers are provided with replaceable front-panel filters. The cooling system makes it possible to operate the transmitter in ambient temperatures of +55 degrees centigrade (+131 degrees fahrenheit).

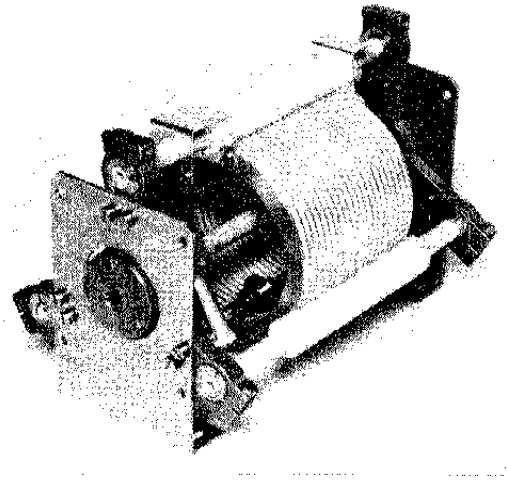


Figure 10—Inductor used in the π-L filter of the antenna-coupling network.

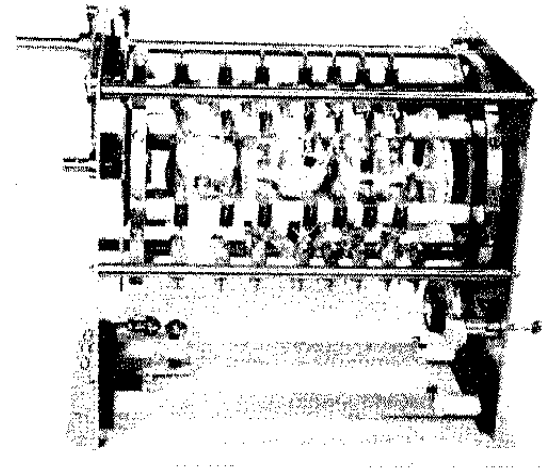


Figure 11—Main band switch.

TABLE I

ST 1200 TRANSMITTER PERFORMANCE

Characteristic	Specification
Frequency coverage	Medium-frequency: 400 to 535 kilohertz. Medium-high frequency: 1600 to 3800 kilohertz. High frequency: 4, 6, 8, 12, 16, 22, and 25 megahertz, within marine mobile bands.
Number of frequency channels	Medium-frequency: 8 telegraph channels. Medium-high frequency: 25 telephone channels. High frequency: 3 telegraph channels in each band; 7 telephone channels in each band except 3 channels in 6-megahertz band. All channels are crystal controlled.
Frequency tolerance attainable	Short term: ±40 hertz. Long term: ±100 hertz.
Types of emission	Medium-frequency: *A1, A2H.* Medium-high frequency: *A1, A2H, A3H, A3A.* High frequency: *A1, A3H, A3A.* *A1:* Telegraph-keyed carrier. *A2H:* Telegraph-keyed audio-frequency-modulated carrier with carrier at full amplitude. *A3A:* Telephony with single sideband and reduced carrier. *A3H:* Telephony with single sideband and full carrier.
Harmonic suppression	Greater than 50 decibels.
Suppression of lower sideband	Greater than 46 decibels.
Suppression of carrier (*A3A*)	Adjustable from 16 to 40 decibels below the sideband power. Normally set for 26 decibels.
Keying speed	50 bauds (approximately 60 words per minute).
Audio-frequency band	350-2700 hertz within 6 decibels.
Audio input	Dynamic or carbon microphone, 40 or 200 ohms, respectively.
Power into antenna	Medium-frequency: *A1:* 250 watts carrier. *A2:* 100 watts carrier, 50 watts sideband. Medium-high and high frequency: *A1:* 1200 watts carrier. *A3H:* 600 watts carrier and 600 watts sideband for 100-percent modulation. *A3A:* 1200 watts sideband.
Antenna impedance	Tunable to all normal antennas on shipboard.
Power supply	3-phase, 220 or 380 volts, 50 to 60 hertz.
Power consumption	3 kilowatts maximum, 0.9 power factor.

5. Special Components

The components used in the antenna-coupling network in the output stage of the linear power amplifier were designed especially for this transmitter. Figure 10 shows one of the continuously adjustable inductors used in the π-L filter. The winding is held by three ceramic rods. Grooved teflon pads are used to fix the turns and the spacing between them. To reduce the weight of the moving parts, the coil is fixed and the shaft with the contact arm and wheel moves. To avoid deterioration of the contact surfaces, rhodium plating is used. The

contact wheel has a graphite bushing bearing to avoid the need for lubrication.

A special main band switch was designed for the fixed capacitors in the output stage. Figure 11 shows its general design. The insulation is ceramic. The contact pressure is relieved before the switch is rotated, making rotation easy despite high contact pressures. The fixed tuning capacitors are mounted inside the switch turret, making very-short connections possible. The mechanism for turning the switch employs a Maltese cross to ensure very-positive positioning. The switch contacts can also be moved along the ceramic rods, permitting numerous switching combinations.

6. Conclusion

The performance specifications of the transmitter are shown in Table 1. Measurements are for typical production units. A number of transmitters have been installed on ships and confirm the known advantages of single-sideband operation over conventional *A3*. Installation of the transmitter on ships has proved to be easy, particularly since the antenna coupling network makes matching to the different antennas very simple.

Fred Hedström was born at Lysekil, Sweden, in 1914. In 1934, he graduated from the Technical Institute of Gothenburg as an engineer.

From 1939 to 1953, he was engaged in the development of receivers and amplifiers. In 1953, he joined Standard Radio & Telefon in Bromma, and now is in charge of the development of communication radio equipment.

From 1950 to 1953, Mr. Hedström lectured on radio engineering at the Norwegian Technological Institute in Oslo, and, from 1957 to 1959, he taught transmitter engineering at the Technical University in Stockholm.

Spicer Honoured

L. R. Spicer was appointed a Member of the Order of the British Empire by Her Majesty the Queen in the June Honours.

This recognition was made for his work on submarine cables for the Admiralty. Mr. Spicer is a cable development engineer for Standard Telephones and Cables at the Newport plant.

Submarine-Coaxial-Cable Manufacture at Southampton Factory 2

ERIC BAGULEY
Standard Telephones and Cables Limited; London, England

1. Introduction

An account has been given of the problems met with and the solutions achieved in setting up facilities for the large-scale production of submarine coaxial cables at Southampton [1] starting from a vacant building site in 1955 and completing the first 5000 miles of cable in 1962.

The most-significant project contributing to the rapid increase in demand for these repeatered submarine telephone cables was the 1956 transatlantic cable [2] for which flexible one-way submerged repeaters were developed by Bell Telephone Laboratories in the United States. In 1960 Bell Telephone Laboratories announced to cablemakers at a symposium in Boston, Massachusetts, their new *SD* 128-channel submarine-telephone-cable system based on two-way working with rigidly housed submerged repeaters. A new design of armourless submarine cable was specified.

The cable, shown in Figure 1, consists of a central strength member of 41 high-tensile-steel wires enclosed within a seam-welded copper tube that is the inner conductor of the coaxial core. The outer conductor is a second copper tube insulated from the inner conductor by a cylinder of polythene 1·000 inch (25·4 millimetres) in external diameter. Over the outer conductor is an extended sheath of black polythene compound 1·25 inches (32 millimetres) in diameter to afford protection to the outer conductor during handling in the factory and on board ship during laying. Thus, the cable design has only 5 components; a steel-wire strand, 2 tubed copper tapes, and 2 polythene extrusions. For these simple components precise control of dimension and of the properties of the materials used in their fabrication was specified.

After due consideration of these proposals it was decided to build a second submarine cable factory at Southampton. A vacant site was available next to the first factory, and building operations commenced with pile driving in January 1961. Production of the new design of cable was under way by the middle of 1962, and the first shipment was made on 3 January 1963.

2. Production Engineering

There are 5 principal processes in the manufacture of the cable, namely: centre-conductor assembly, core extrusion, core shaving, outer conductor combined with sheathing, and transmission testing. Repeater couplings are then fitted to the cable ends. Of these only the core extrusion process lacked novelty. Processes new to submarine-cable engineering, such as argon-arc seam welding and a new method of core shaving, had been tried out at a pilot plant by Bell Telephone Laboratories, and recommendations had been made as to how to implement them in commercial production. From the information received a factory layout was planned for a cable output rate of 80 to 100 nautical miles (150 to 185 kilometres) per week and consisted of 4 centre-conductor machines, 2 core extruders, 4 core-shaving machines, 2 sheathing extruders, a testing enclosure, and a jointing room for the repeater couplings.

For the *SD* system the repeater-section cable

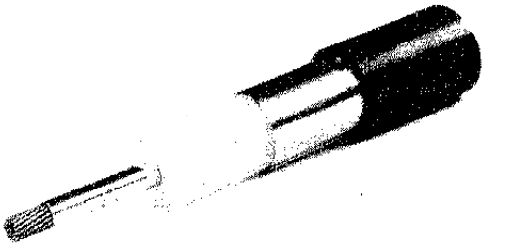


Figure 1—The new type *SD* armourless 1·000-inch (25·4-millimetre) coaxial submarine telephone cable comprising a central strength member, welded-copper-tube inner conductor, polythene dielectric, single-copper-tape outer conductor, and black polythene-compound outer sheath.

length is approximately 20 nautical miles (37 kilometres), and it was specified that this length should obtain at all stages of production, that is, 20-mile lengths of centre conductor, 20-mile lengths of core, and 20-mile lengths of completed cable, preferably without any joints in the central strength member. This manufacturing unit has to be taken into account in designing the various machines, in planning storage space for work in progress between machines, and for incoming materials. To ensure continuity of production at the stated rate it was decided to have storage space for not less than 3 weeks of stock of all materials and all work in progress. For some of the processes this means that storage facilities occupy more space than the production machines.

On completion of manufacturing, testing, and terminating, the 20-nautical-mile repeater-section lengths of cable have to be stored until required for loading into a cable ship. This is a major planning problem in a submarine-cable factory. To deal with the increasing demand for cables, new cable ships are being built capable of carrying up to 3000 nautical miles (5560 kilometres) of cable. The factory must, therefore, be able to store sufficient cable to enable ships to be loaded at a rate not less than 500 nautical miles (930 kilometres) per week.

A new method of storing completed cable was requested by the customer. Each 20-mile repeater-section length was required to be stored separately from all other lengths and to be accessible for ship loading in any sequence best suited to the transmission requirements of the system. Storage on drums was unacceptable because of the weight of cable involved, approximately 46 tons (47 tonnes). Instead, transportable cylindrical steel tanks were designed into which the 20-mile length of cable may be coiled down horizontally in the traditional submarine-cable manner. A 75-ton (76-tonne) travelling gantry crane presides over a storage yard in which up to 200 of the cable tanks may be moved around and stacked two high.

Part of the runway covered by the crane is inside the factory building, and within this enclosure the completed cable sections are coiled down into the tanks. When filled with cable the tanks are lifted and then lowered into fixed test tanks supplied with circulating water at 10 degrees centigrade, the temperature specified for the cable during transmission

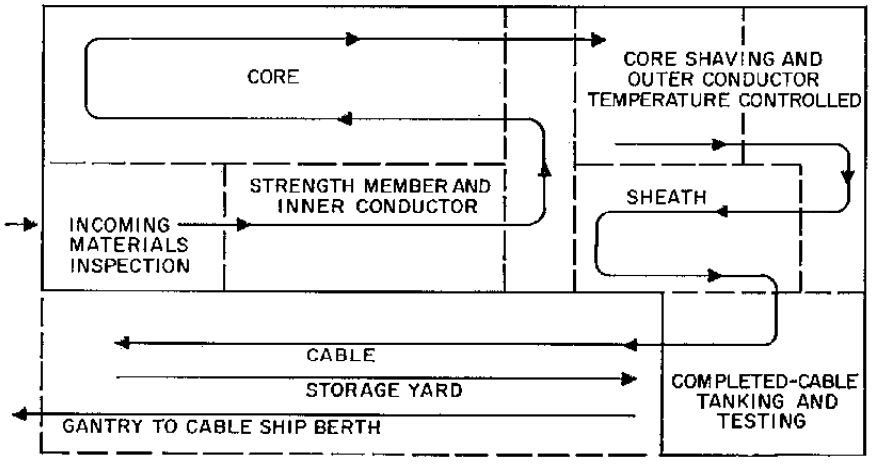


Figure 2—The layout of factory 2 in Southampton indicating diagrammatically the product flow line from raw-material intake to finished-cable delivery to ship.

tests. After testing, the cable tanks are taken out of the water tanks and placed near to the terminating room into which both ends of the cable are passed for the fitting of repeater couplings. The product line plan adopted for the new factory is shown diagrammatically in Figure 2.

3. Strength Member and Inner Conductor

There are 4 principal sequential operations to produce a copper tube tightly fitting round a steel-wire strand.

3.1 Steel-Wire Stranding

A conventional tubular stranding machine was selected as best suited to deal with 40 bobbins of steel wire, each containing 21 nautical miles (39 kilometres) of wire. In this type of strander, extensively used for stranding conductors of power cables, the bobbins do not rotate in a flat plane about the axis of the strand but are mounted in line in horizontal yokes that float inside a long tubular frame. This tube rotates about the axis of the strand, and the wires from the bobbins pass through guides mounted inside the tube. All 40 of them, plus the centre wire, emerge in a circular array from the output end of the tube. Here the wires pass through a closing die that brings them closely together twisted by the rotation of the tube, all in the same direction, with the specified length of lay of 6 inches (152 millimetres). The wires are of 5 different gauges to give a better space factor (ratio of steel cross-section to total cross-section) than the more-usual single-gauge wire strands. Five of these long machines are shown side by side in the background of Figure 3.

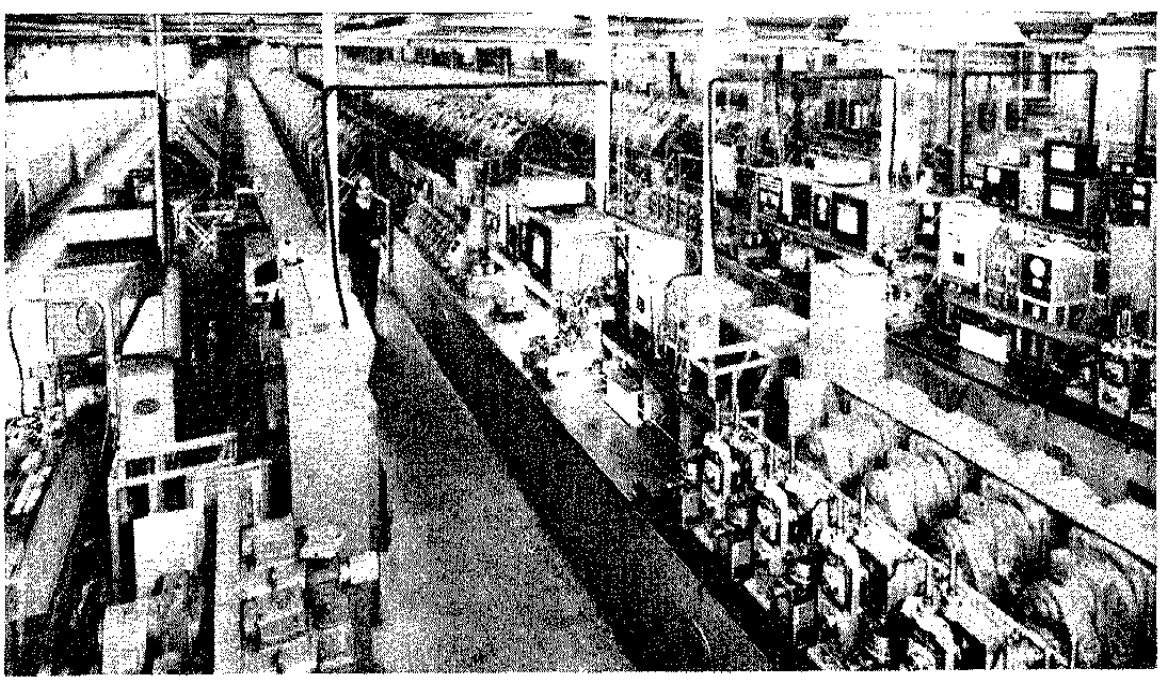


Figure 3—The inner-conductor lines. In the background are the 40-bobbin tubular stranding machines for the steel-wire strength member. In the right foreground is a tape-tubing mill, the argon-arc-welding unit, and the tube-drawing mill with completed inner-conductor assembly leaving the machine at the lower right.

3.2 Copper Tape Tubing

Copper tape for the inner conductor is supplied to the line immediately following completion of the steel-wire strand. The tape comes up from the floor below that on which the machines are mounted and passes along in a horizontal plane underneath the strand into a pair of slitting rolls that cut a narrow strip of copper off each edge. Both the tape and the strand then pass into a 6-stage tubing mill, which rolls the copper into a tube 0·50 inch (12·7 millimetres) in diameter enclosing the steel strand and with the butting edges of the tape along the top of the tube.

It will have been noted that the stranding machine does not have to stop to load steel wire during a 20-nautical-mile (37-kilometre) run. The same facility was required of the tubing unit, but this cannot be loaded with 20 miles of copper tape. The tape, being 1·65 inches (42 millimetres) wide, is not readily handled in pads weighing more than about 200 pounds (90 kilogrammes). This means that a new pad of tape has to be welded into the unit at intervals of about 2 hours. To avoid having to stop the machine when this occurs, sufficient copper tape for 15 minutes running is stored in a mechanical accumulator sited between the tape supply and the tubing mill.

3.3 Continuous Seam Welding

In Figure 4 the copper tube is shown on the left leaving the last stage of the tubing mill. It moves to the right into the welding enclosure where the butting edges of the tape pass under a direct-current arc torch with a non-consumable electrode operating in a piped supply of argon. The electrode is mounted vertically, and the top of it may be seen projecting above the enclosure. An indirect view of the arc is displayed in the dark-glass periscope unit to the right of the enclosure.

Basically the unit is a proprietary argon-arc welder, but a great deal of development effort was expended locally to obtain sufficiently precise control of the process to make possible non-stop runs of 20 nautical miles (37 kilometres) lasting 5 to 6 days. Brief reference may be made to a few of the difficulties encountered and overcome.

The copper tape is only 0·023 inch (0·6 millimetre) thick, and a sound seam weld will be obtained only when the arc provides the right amount of heat in the right place, directly over the butting edges of the tape. A round tube was found to have a tendency to rotate to and fro about its axis whilst passing through the machine, and at times the tape edges were not under the electrode. This was corrected by making the tube slightly oval. With the tube travelling at constant speed past the electrode there is an optimum value for the arc current; too little gives imperfect welding, and too much burns holes in the copper and may damage the steel strand inside the tube. The machine, however, does not run at constant speed all the time; it has to start and stop, and on occasions it has to run at less than full speed. Long studies were made of arc current as a function of tube speed, and of other variables such as electrode composition, electrode diameter, tip profile, and length of arc. The results of these studies were built into automatic arc-current–tube-speed controllers with facilities also for marginal manual adjustment by the machine driver. Improvements were found necessary to ensure more-complete shrouding in argon of the arc and of the hot copper on both sides of the tape [3].

After being welded the tube passes through a weld integrity checker shown on the extreme right of Figure 4. In this device two similar coils are coaxially mounted round the tube a few inches apart. They form part of a balanced high-frequency network whose output signal is displayed on a cathode-ray oscilloscope. When the portion of copper tube within one coil is in the same physical condition as the portion within the other coil the network is balanced and the oscilloscope trace is a flat ellipse; but when a faulty weld or defect in the copper

tape moves into one coil the network balance is upset and an alarm is sounded. There is a second welding unit about 6 feet (2 metres) downstream from the checker at which repairs may be effected. Welding may continue at this torch for a short time should adjustments to the main torch be necessary.

3.4 Tube Drawing

From the second welding torch the copper tube passes into a tube-drawing mill, one of which may be seen in the lower right of Figure 3. Here the tube is reduced in diameter from 0·5 inch (12·7 millimetres) to 0·33 inch (8·4 millimetres). At this new diameter the copper is drawn tightly on to the steel-wire strand inside it to form a compact centre-conductor assembly which is taken up in 20-nautical-mile (37-kilometre) lengths on to large steel drums as shown in Figure 5.

The external diameter and the effective thickness of the copper after draw-down are required to be held within very close limits of stated nominal values to ensure that the copper is firmly keyed into the steel-wire strand throughout all lengths. During the drawing-down process the linear speed at which the tube is moving through the machine increases, and the factor principally controlling the

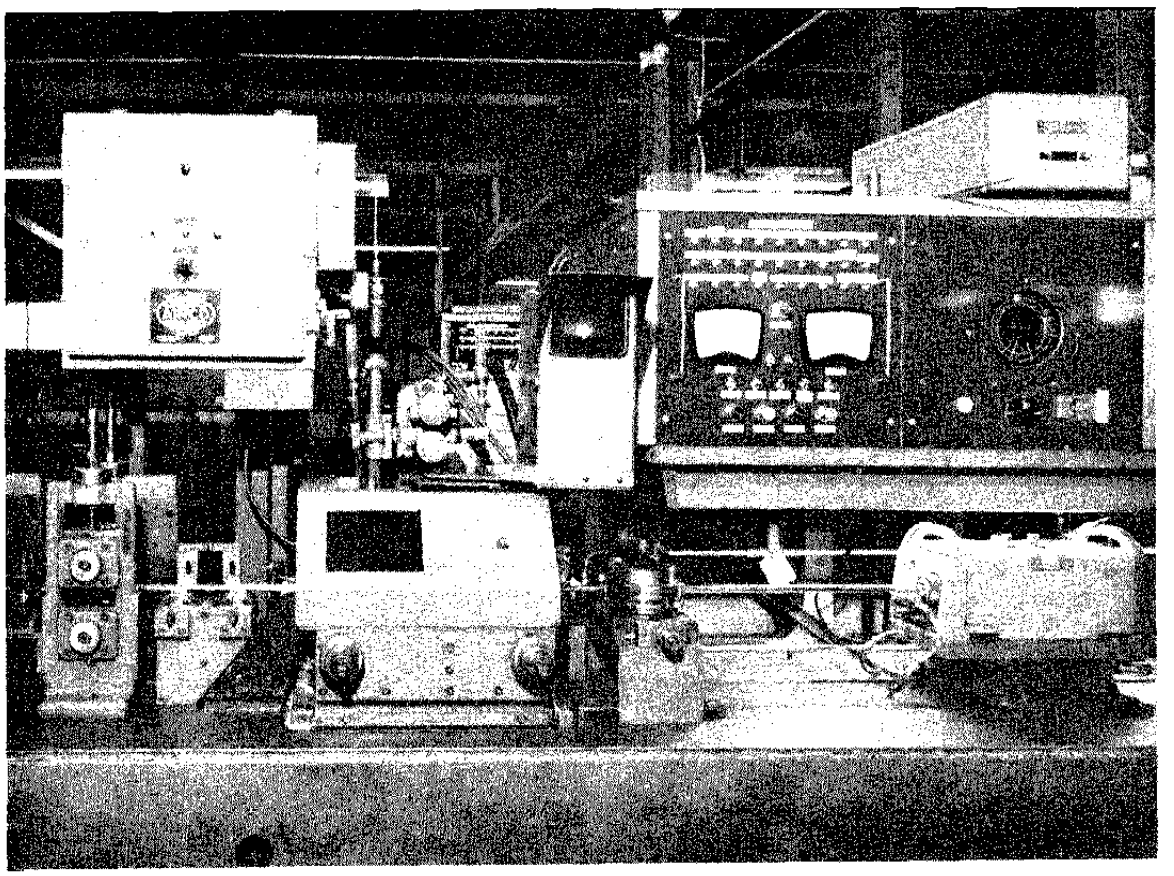


Figure 4—Inner-conductor argon-arc seam-welding unit with tube moving from the last stage of the tubing mill on the left into the welding box. Argon passes to the arc region through the electrode holder, which projects out of the top of the box. The arc is displayed in the viewer to the right. After seam welding the copper tube passes through the weld integrity checker at the extreme right.

dimensions obtained is the ratio of the speed of the finished conductor to the speed of the copper tape entering the welding unit. These two speeds are monitored by a ratiometer device, which displays the ratio on a dial and also provides a command signal to the complex control gear for the various electric motors driving the stranding machine, the 2 tubing mills, and the haul-off caterpillars. Ratio control during starting and stopping proved to be a difficult problem. The copper tube between the welding unit and the haul-off caterpillar is the mechanical link between the high-speed group of units and the low-speed group. It is a weak link, and unless the speed ratio is correctly held during speed changes the tube may either buckle or fracture.

In the tube-drawing mill it is necessary to lubricate the outside of the tube with oil, and then it is equally necessary to remove all traces of it after the tube leaves the final sizing die. This is effected in a trichloroethylene degreasing unit, which first sprays the tube with trichloroethylene and then dries it by air blast.

4. Core

Completed drum lengths of inner conductor are lifted by a travelling gantry crane from the take-up stands, Figure 5, and when re-

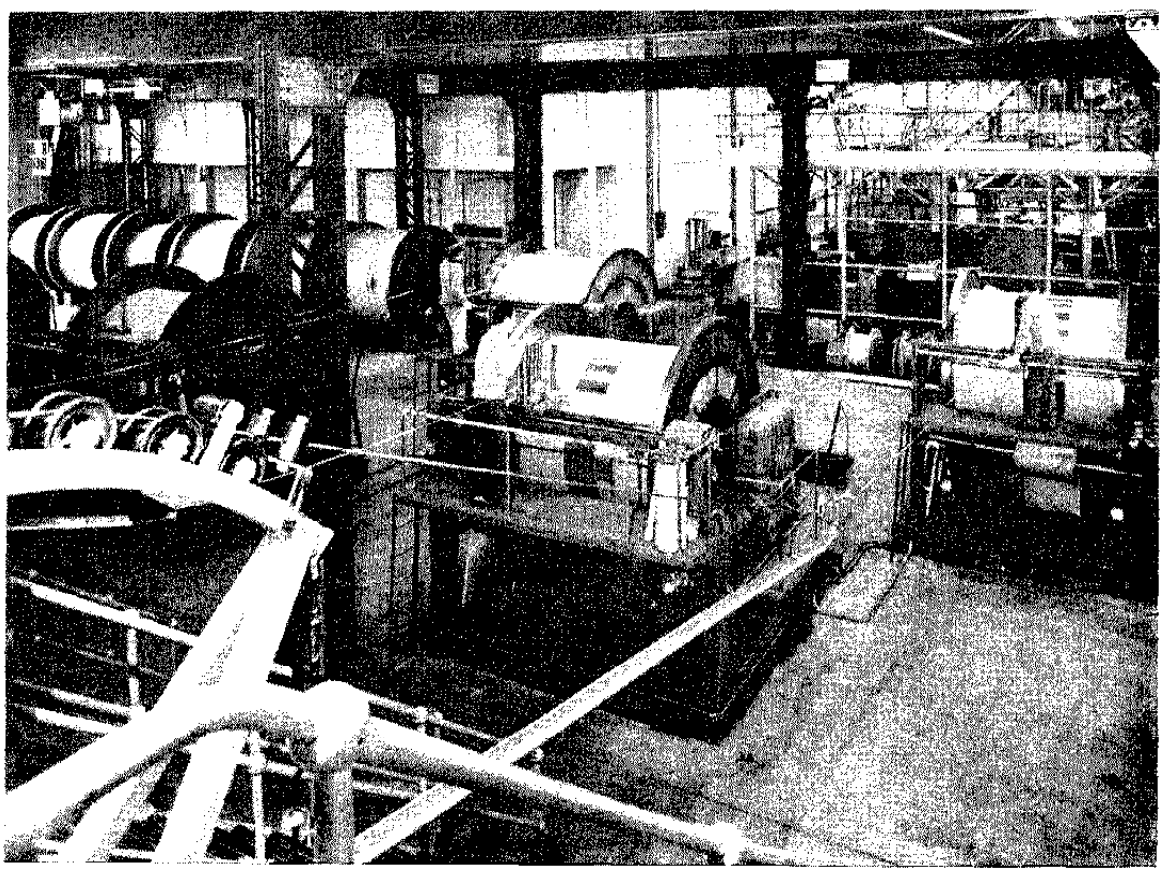


Figure 5—Completed inner-conductor assembly being taken up in 20-nautical-mile (37-kilometre) lengths on to large steel drums. Filled drums awaiting loading into extruder lines are stored in the left background. Drums are moved to left and to right on a 25-ton (25·4 tonne) travelling gantry crane.

quired are placed in the extruder supply stands at the left-hand end of the crane runway. Storage space for completed drum lengths forming the work-in-progress stock is shown in the left background of Figure 5.

The 2 core extruders are 6-inch-long (152-millimetre) screw machines designed to deal effectively with high-molecular-weight (melt index about 0·1) insulating-grade polythene. In common with all the machines in the factory, tractive effort is applied to the product by double-track caterpillar haulers equipped for servo-controlled line tension. (There is one on the right in Figure 8.) These replace the various types of single- and double-wheel capstans more-commonly used in cable making as the product is not subjected to bending and straightening stresses when passing through a caterpillar machine.

The first caterpillar in the extruder line hauls the inner conductor from its drum and pushes it into the extruder head at a linear speed controlled by a diameter measurement on the polythene-insulated core as it leaves the head. There are graded cooling troughs totalling 400 feet (122 metres) in length from the hot end, then a turn-round sheave followed by 400 feet of cold trough back towards the extruder. At the end of the first section there is a concentricity head, Figure 6, comprising two pairs of

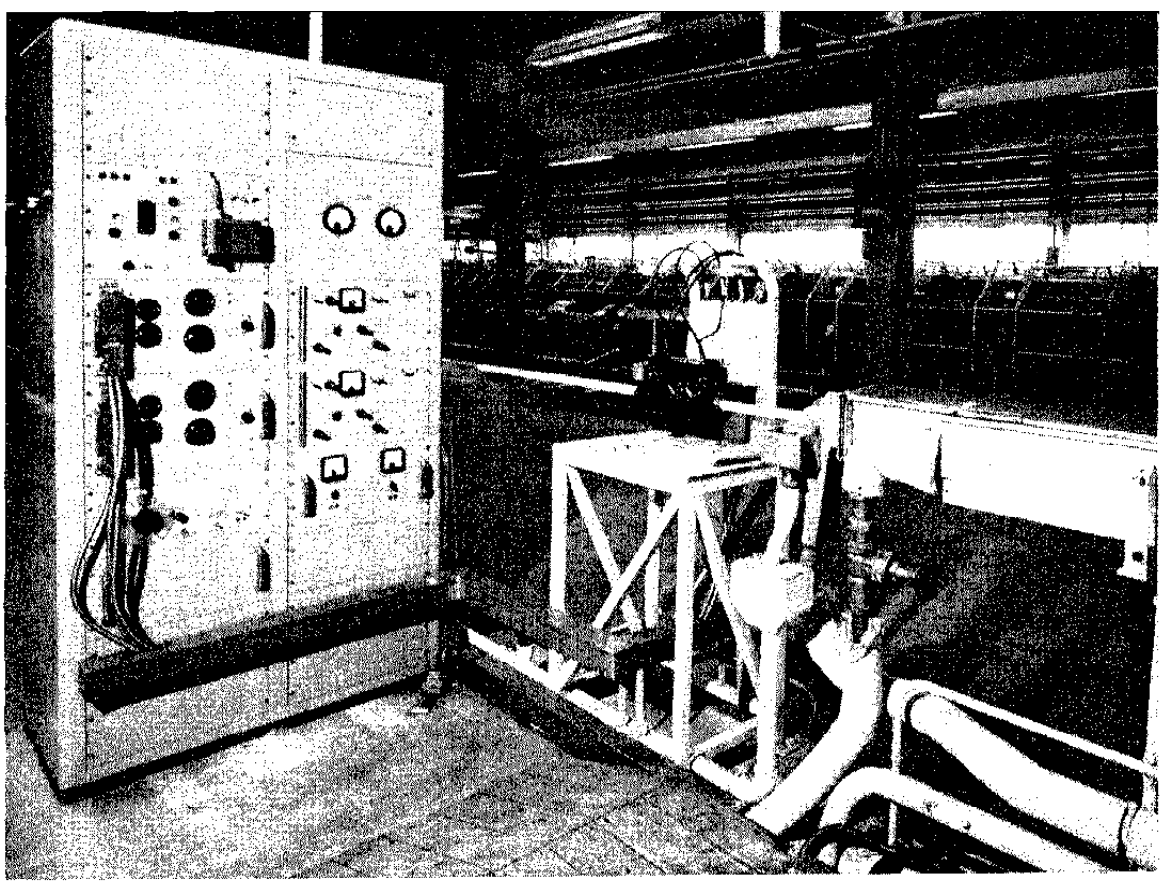


Figure 6—The eccentricity-monitoring head in the core-extruder line. The box containing two pairs of capacitance electrodes rides on the newly formed 1·000-inch-diameter (25·4-millimetre) core when it passes from one section of cooling trough into another section. Two racks of equipment at left integrate the signals from the electrodes and operate recorders (not shown) at the extruder console.

capacitance-sensing electrodes bearing on the core and aligned at right angles to each other. The capacitance measured between each electrode and the inner conductor of the core is a function of the distance between them, and by comparing the four readings the position of the inner conductor with respect to the axis of the polythene covering may be determined. When first setting up the core extruder, adjustments are made to the point and die to give an indicated conductor eccentricity not greater than 0·015 inch (0·4 millimetre) in the 1·05-inch (27-millimetre) core. Thereafter the capacitance readings are charted to provide quality assurance data.

5. Shaved Core

It was specified that the diameter of the core as presented to the outer-conductor tubing mill should be $1{\cdot}000 \pm 0{\cdot}001$ inch ($25{\cdot}4 \pm 0{\cdot}025$ millimetres) at 20 degrees centigrade, and that the eccentricity (displacement from core axis) of the inner conductor should not exceed 0·005 inch (0·13 millimetre). Such limits are not easily met by core extruders, so a new method of core shaving developed from a process first applied by Standard Telefon og Kabelfabrik of Oslo, Norway, was introduced.

Because of the comparatively large coefficient of expansion of polythene, variations of core diameter due to ambient temperature could be

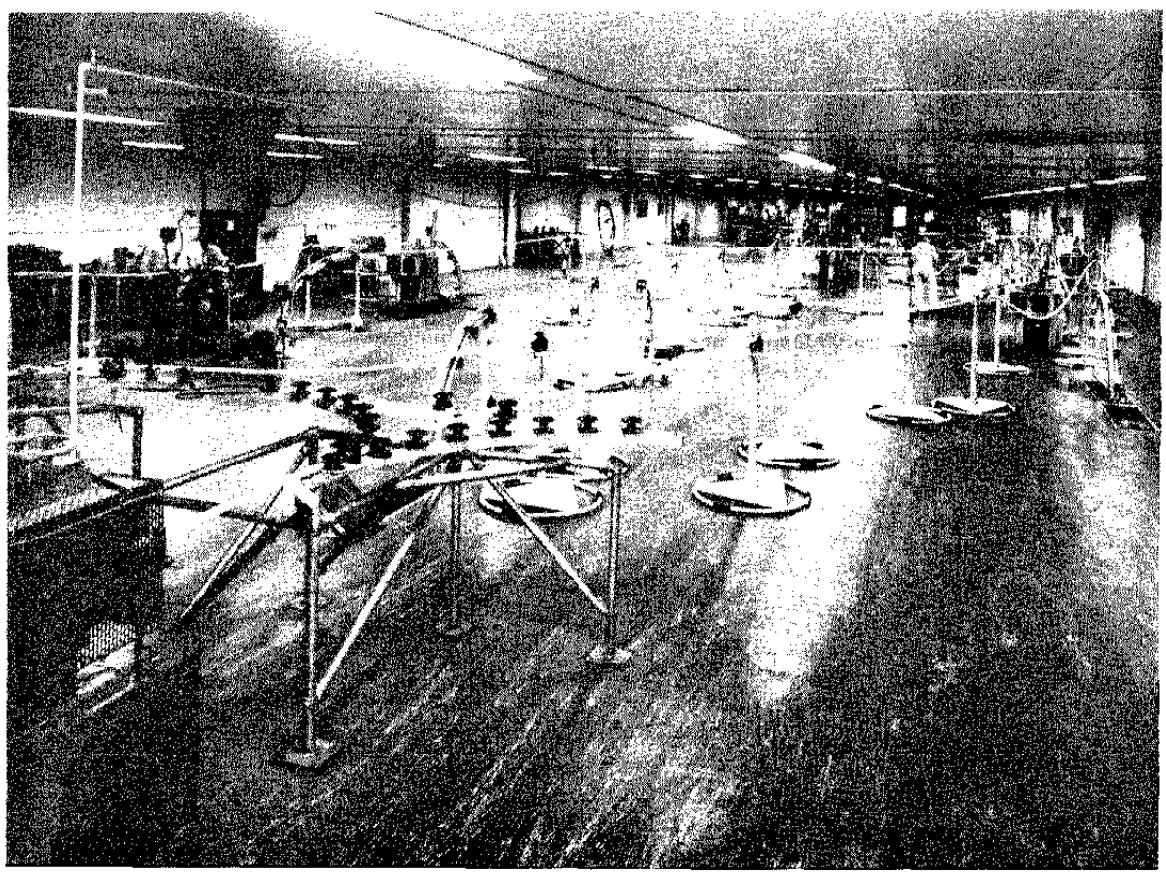


Figure 7—The mezzanine floor within the temperature-controlled core-shaving room. Newly completed cores from the extruder lines are entering the room from the left and go down through the floor into storage creels on the ground floor of the room. Also shown are core-jointing operations.

greater than 0·1 per cent. Thus the first requirement for the shaving process is a controlled working temperature, and to achieve this a large core-shaving room was built equipped with an air-conditioning plant capable of holding the temperature within it to 20 ± 1 degrees centigrade. A view of the mezzanine floor within this enclosure is given in Figure 7.

Cores from the extruders enter at this end of the room and pass down through slots in the floor to coil down in open creels on the floor below, which is within the temperature-controlled area. Each creel will hold a 20-nautical-mile (37-kilometre) length, and there is a storage accommodation for more than 300 nautical miles (560 kilometres). Each completed length may therefore be held here until it has become acclimatized to 20 degrees centigrade. The core may be run from one creel to another by way of the mezzanine floor, as shown, for quality inspection, cutting, jointing, and repairing of extrusion faults, as may be necessary.

When a length of core has passed process inspection and is acclimatized, it returns to the mezzanine floor and passes through one of the

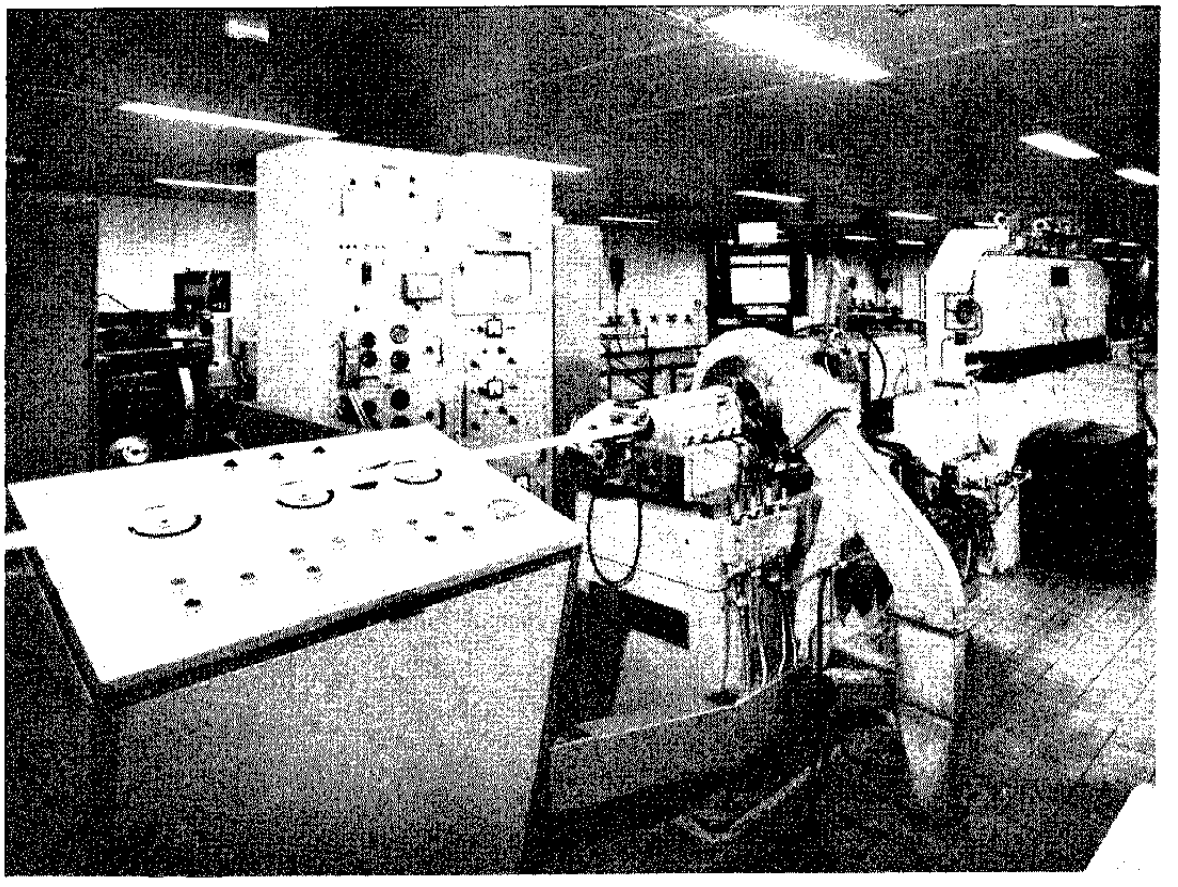


Figure 8—One of the core-shaving machines. At its extreme right is a caterpillar hauler for bringing the core to the shaving head within the circular blower casing where the polythene shavings are exhausted into the ducting going down through the floor. Up against this is the eccentricity head which, with the two racks of equipment behind the line, servo-controls the cutter axis about the conductor axis and records conductor displacement in two planes. Shaved-core-diameter record is behind the cutter head.

4 shaving machines at the balcony end of the floor in the centre of the air-conditioned room. The design of the shaving machine shown in Figure 8 stems from earlier development work at the pilot plant of Bell Telephone Laboratories and it has been further developed locally for commercial production use.

A high-speed 3-blade rotary cutter skims the polythene diameter to $1{\cdot}000 \pm 0{\cdot}001$ inch ($25{\cdot}4 \pm 0{\cdot}025$ millimetres) and a complex servo-controlled mechanism homes the axis of the cutter head around the axis of the inner conductor within the core rather than around the axis of the polythene covering, if the two differ. This takes care of the reduction of the conductor eccentricity to less than 0·005 inch (0·13 millimetre) after shaving.

The command signal for this control is obtained from a capacitance concentricity head (centre of Figure 8) of a type similar to that on the extruder lines. Both the shaved-core diameter and the eccentricity readings are recorded throughout the operation, as may be noted in Figure 8. The charts, as with other processes, form part of the quality assessment made prior to clearance for subsequent processing.

Each shaving head produces a large volume of

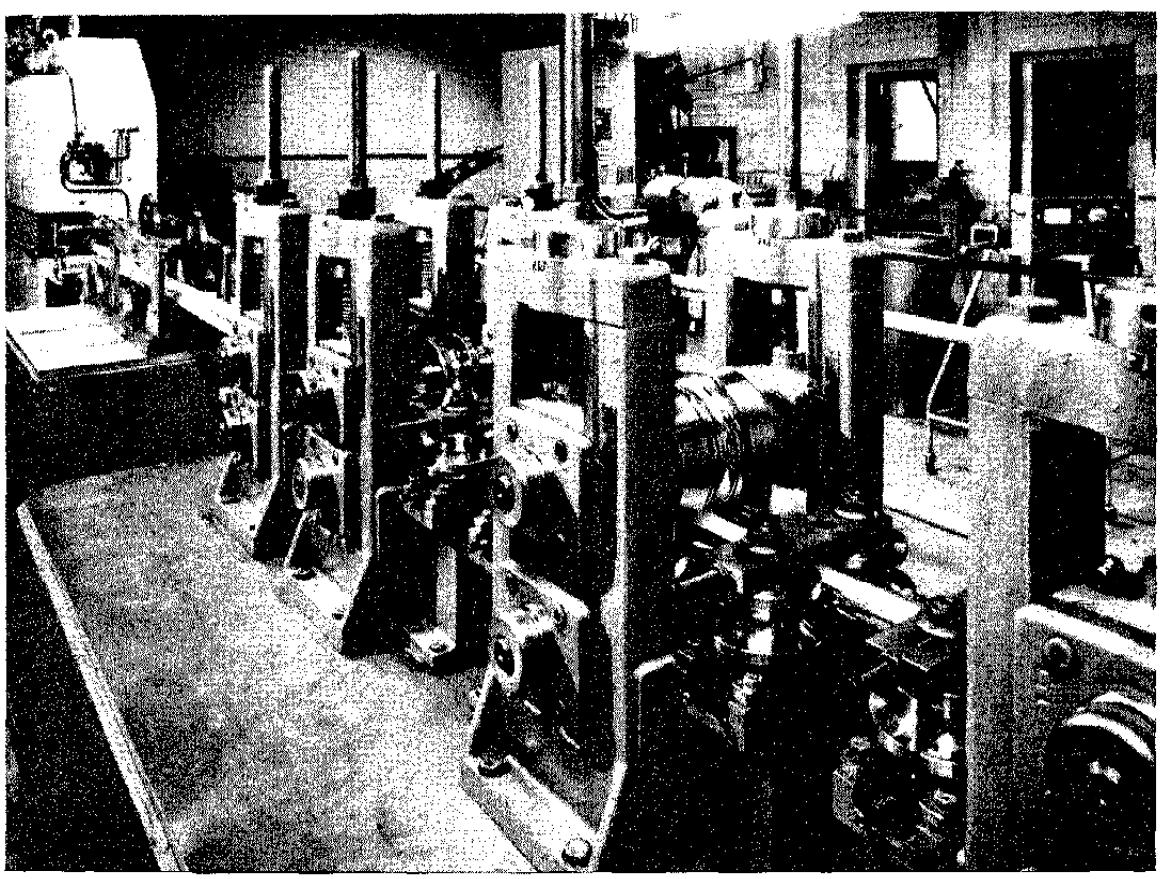


Figure 9—Tubing mill applying outer-conductor copper tape to shaved core coming into picture from caterpillar hauler at top left. Copper tape comes up from floor below where the core leaves the hauler. A ratiometer device monitors the relative speeds of the core and the copper tape as they enter the mill. By holding the ratio to nearly unity excessive stretching of the copper tape during tubing is avoided.

filamentary polythene shavings, which are sucked away from the core and blown down the trunking to the right of Figure 8 into a hopper room adjoining. There were problems here in keeping the polythene shavings moving along the ducting without stoppages caused by "bird nesting," and without acquiring excessive electrostatic charges. If the shavings are not removed, clogging may occur at the cutters, with damage to the core. Another difficulty proved to be maintenance of shaved-core diameter during starting and stopping of the machines; the cutters tended to dig into the polythene at very-low speeds. This is not a serious problem providing that not more than a few inches of core are undercut. A repair to the polythene may then be made in a core-jointing mould.

The shaved cores are coiled down into a second set of storage creels occupying the output end of the shaving room, where facilities for core examination, patching, and jointing are again provided.

6. Outer Conductor and Sheath

The most-efficient outer conductor for a coaxial cable is a copper tube and in the new *SD* submarine cable a very-close approximation to this is achieved by folding a single copper tape longitudinally around the shaved polythene core. The edges of the tape are overlapped but not joined together. A view of the tubing mill is given in Figure 9. Here the shaved core enters the mill from the caterpillar pusher (top left). The single copper tape comes up from the floor below and moves into the forming rolls with the core. All this equipment is inside the temperature-controlled shaving room so as to ensure that a core of constant diameter is presented to the copper tape.

When in position round the core the copper tape is not self-supporting, and in consequence the outer plastic sheath of the cable has to be applied, immediately after tubing, by a 6-inch (152-millimetre) extruder driven in tandem with the tubing mill. The completed cable leaving the extruder head is shown in Figure 10. Development problems here were concerned with keeping the copper tape in intimate contact with the polythene core, and free from ripples or waviness whilst passing from the tubing mill to the extruder. The copper tape supplied must accurately conform to specified dimensions and be free from edge bow and surface irregularities. A copper-tape accumulator similar to those on the inner-conductor lines is installed between the pad stand and the tubing mill so that the sheathing extruder need not be stopped whilst new pads of copper tape are loaded into the line. Unlike most cable-making processes, this combined operation is more-successfully carried out at high speed. A 20-nautical-mile (37-kilometre) length can be dealt with in less than 48 hours non-stop running time.

After passing through the extruder cooling troughs, the completed cable is coiled down into one of the repeater-section storage tanks, is tested, and terminated as already described.

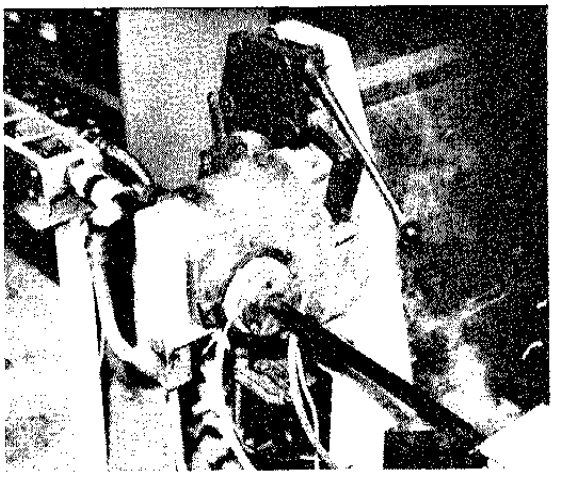


Figure 10—Completed sheathed cable leaving extruder head. This extruder is driven in tandem with the tubing mill in Figure 9. The copper-wrapped core comes from the mill in the enclosed channel way (top left), which is maintained under vacuum to remove air from underneath the copper tape to prevent extrusion blows in the hot sheath.

7. Conclusion

A brief account has been given of some of the development problems associated with the commissioning of the second submarine-cable plant at Southampton between 1961 and 1962. On 3 January 1963, the cable ship *H.M.T.S. Alert* left Southampton with 570 nautical miles (1056 kilometres) of the new *SD*-type cable on board. This now forms part of the submarine telephone cable between Port Everglades, Florida, and Kingston, Jamaica; it was completed on 10 February 1963 and was the first *SD* system to be installed.

8. Acknowledgements

The author acknowledges that his primary role has been to report on the achievements of the Submarine Cable Division of Standard Telephones and Cables in Southampton under the management of Mr. G. K. Duddridge. Also to be acknowledged is design data made available by the Bell Telephone Laboratories Ocean Cable Group under Mr. A. W. Lebert.

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Eric Baguley was born in Lancashire, England, in 1905. He graduated in electrical engineering from Manchester University in 1924.

On graduation, he joined Western Electric Company in London and was soon assigned to the engineering of communication cables. Since then, he has been concerned with all types of communication cables.

From 1927 to 1932, he was with ITT Laboratories at Hendon. He was then transferred to Standard Telephones and Cables and in 1940 was placed in charge of the telephone cable development laboratories at North Woolwich. In 1955, he was transferred to the newly formed submarine cable division at Southampton, where he is now chief cable engineer.

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Telepuls and FMX Modular Electromechanical and Electronic Remote Control

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1. Introduction

As customer requirements in Germany for remote control and indication systems were found to be not too divergent, systems were developed using standard components that could be combined to meet the most varied needs of such organizations as power companies, railroads, the post office, and gas, water, and pipeline operators.

Three systems were developed. Two of them operate with pulse-code transmission in binary form and are compatible with each other. The system called Telepuls 10 is electromechanical and the second system, Telepuls 12, is fully electronic. Using semiconductors throughout, Telepuls 12 operates at 1200 bauds compared with 25 bauds for Telepuls 10. Any combination of the standard units of the two systems may be used as shown in Table 1. This facilitates a progressive conversion from electromechanical to electronic operation. Furthermore, fully electronic substations are able to interwork with electromechanical central stations, which often are preferred for their particular methods of display.

The third system employs frequency multiplexing, which is advantageous if a number of substations along a party line require only one, or a few, indications each. This frequency-multiplex system, *FMX1*, may also be used for the transmission of pulse signals at speeds below 25 bauds.

2. Telepuls 10

In the design of the new Telepuls 10 system [1], experiments made with previous systems and an extended analysis of customer requirements were evaluated. The system must permit point-to-point, parallel, and star operation of substations, and transmission of digital and analog values must be possible over the indication path when it is not engaged with the transmission of indication signals. The code used must have a high degree of error detection but must avoid excessive redundancy. The elements for detecting changes in the positions of the indicating contacts in the substations must be very reliable. Operation must be ensured over a wide range of voltage fluctuations and with distorted pulses. The equipment must be divided into a minimum number of units that permit any combination within a system as well as interworking with the Telepuls 12 with only simple wiring changes.

TABLE 1
POSSIBLE COMBINATIONS OF TELEPULS 10 AND TELEPULS 12

Control Path			
Central Station	Converter	Converter	Substation
electromechanical	electromechanical	electromechanical	electromechanical
electromechanical	electromechanical	electronic	electronic
electronic	electronic	electronic	electromechanical
electronic	electronic	electronic	electronic
Indication Path			
electromechanical	electromechanical	electromechanical	electromechanical
electromechanical	electromechanical	electronic	electronic
electromechanical	electronic	electronic	electronic
electronic	electronic	electronic	electronic

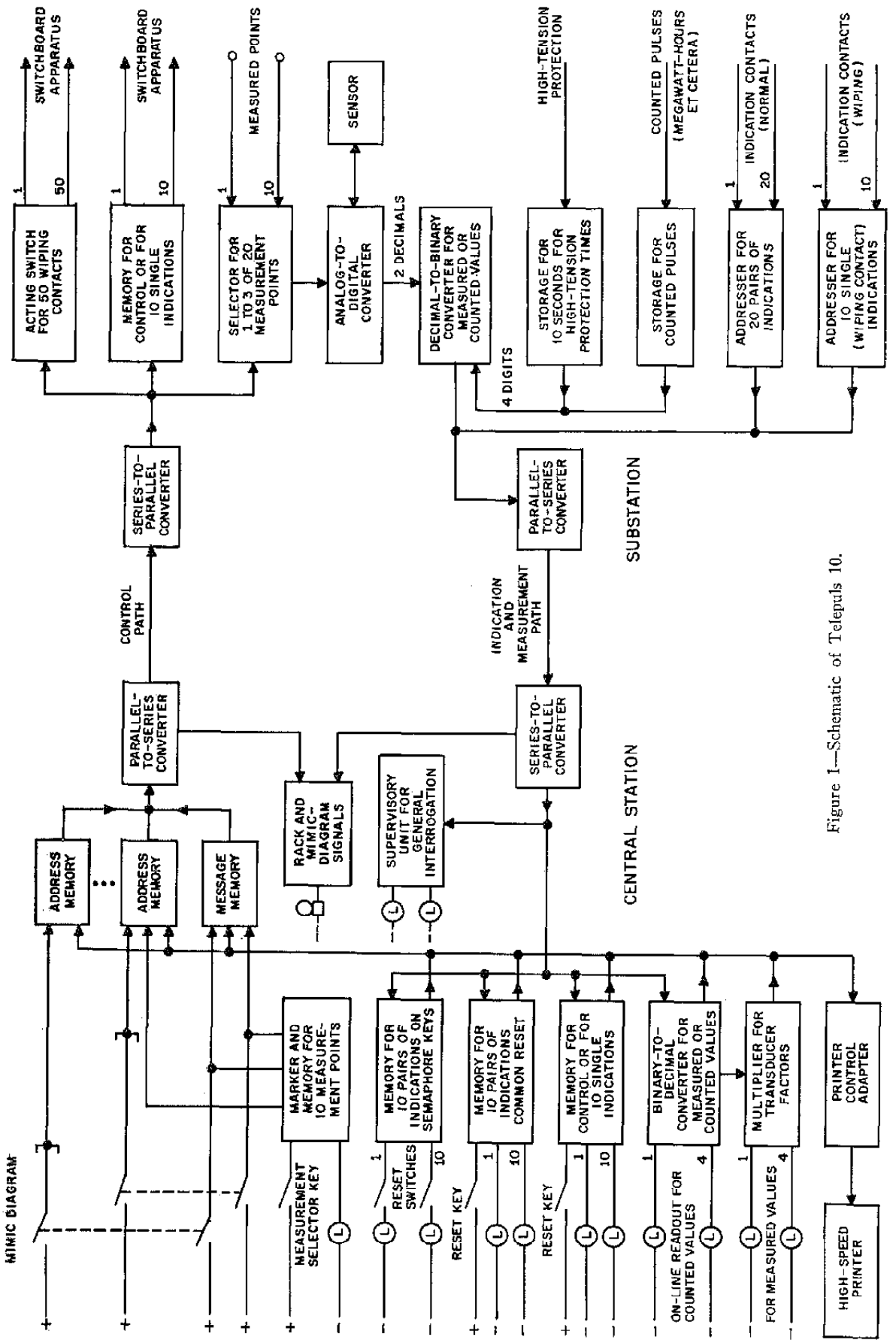


Figure 1—Schematic of Telepuls 10.

Figure 1, a schematic of the Telepuls 10 system, shows separate control and indication transmission paths.

In all pulse trains, the binary 0 is represented by a short mark and a short space each of 40 milliseconds. The binary 1 is a long mark and a long space, each of 160 milliseconds. A time detector at the receiver guarantees even greater safety against transmission errors than is achieved by the conventional method of a backward retransmission and a subsequent forward acknowledgment signal [5].

Each pulse train contains an address in the form of a 2-out-of-10 code, thus permitting 45 different combinations, and a message consisting of one or several groups of 10 pulses and spaces arranged in a code suitable for the type of information to be transmitted.

Control orders are given by operating a key or semaphore key, indications associated with the keys being shown by lighted or unlighted lamps. The address and message, the latter in a 1-out-of-10 code, are passed to a parallel-to-series converter, which transmits the two parts sequentially to the series-to-parallel converter of the receiver. Figure 2 shows an example of such a pulse telegram. After checking the pulse train for transmission errors, the decoded information is used to actuate the control equipment of the address unit.

Indications of contact position changes are reported by the control equipment to an addresser. Here a parallel-to-series converter forms the address and message, the latter containing in an n-out-of-10 code all indications concerned with this address. For each contact in a group of 10, the report is either a 1 for being closed or a 0 for being open. In the receiver, the message is converted into parallel output and, after a code check, is compared with the existing indication display associated with the address. Any change of information is then normally displayed by a flashing light.

In addition to control orders and indications, decimal counts and measured analog values may be transmitted from a substation to the central station. For transmission of counted values, a pulse counter in the substation stores the pulses, for example, those generated by a kilowatt-hour meter, and adds them up to a decimal figure. To handle measured analog values, the analog quantity is converted to a direct voltage between 0 and 10 volts, which is transformed by a special analog-to-digital converter into a decimal figure. For the decimal figures obtained from decimal counters and from analog values, each decimal digit is represented by marking 1 of 10 terminals. This marking is then changed by a converter into a 4-bit binary code. Thus two decimal digits are transmitted by the standard group of 10 information pulses, leaving 2 pulses for other purposes. Decimal figures with more than 2 digits are transmitted by a series of several 10-bit groups. In the central station, the receiver performs the reverse code conversion, resulting in a decimal output. It then marks 1 of 10 terminals for each decimal digit, which may be displayed by an on-line readout, a printer, or a tape perforator.

Auxiliary facilities are provided to check the proper operation of the system and to signal

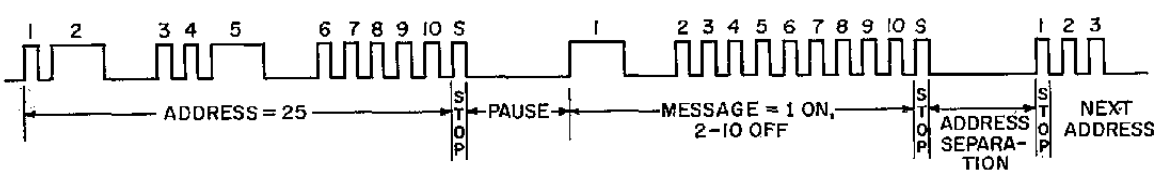


Figure 2—Example of a pulse-train. Each pulse and its following space of equal length represents a character. The short pulses and spaces are 40 milliseconds each and signify 0. The longer pulses and spaces are of 160 milliseconds and signify 1. The address-separation space is at least 320 milliseconds long.

such faults as blown fuses, transmission interruptions, and code failures. The transmission speed of 25 bauds permits operation on ordinary signalling and telegraph channels with a high safety margin; the average time for transmission of an address and message is about 3 seconds.

The mechanical design complies with German Industrial Standards *DIN 47432* and also permits installation of the units in racks conforming with International Telephone and Telegraph System practices. Figure 3 shows typical units for full supervision of an unattended 20-megavolt-ampere transformer and distribution station [4].

3. Telepuls 12

As mentioned previously, the electronic system, Telepuls 12 [2], is compatible with the electromechanical system, Telepuls 10. Accordingly, they use the same coding and have corresponding subunits.

Figure 4 is a schematic diagram of Telepuls 12; its similarity to Figure 1 is obvious. It is significant that in most cases there is a direct substitution of an electronic unit for a corresponding electromechanical unit. However, some of the units of Telepuls 10 are retained in Telepuls 12, notably the substation memory for the controls made by contacts and in the

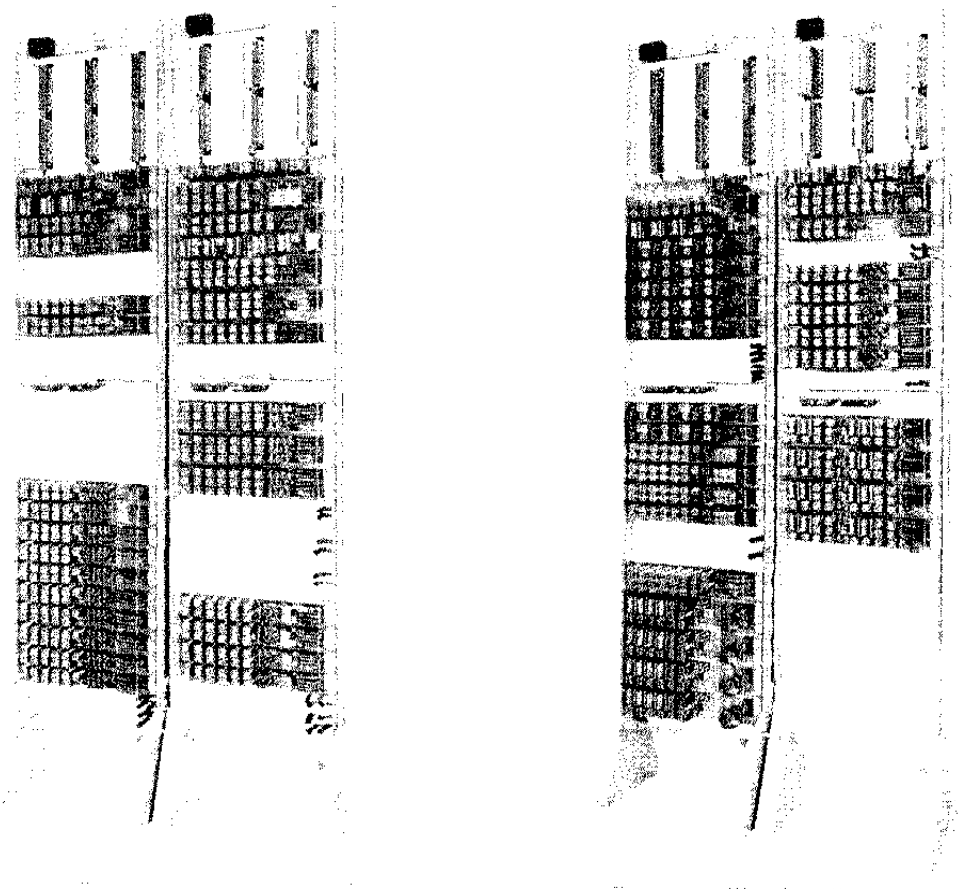


Figure 3—Racks for central and substation installations for Telepuls 10.

central station the memories for lamp indications on a board (flashing light, continuous light, acoustic signals, et cetera). It is not always economical to use transistors, which are most effective if indications are to be given by a change-over from one lamp to another.

In Telepuls 12, the magnetic counters of the series-to-parallel and the parallel-to-series converters used in the Telepuls 10 equipment are replaced by conventional shift registers. While Telepuls 10 uses relay flip-flops to generate pulses and relays with time delay to check the

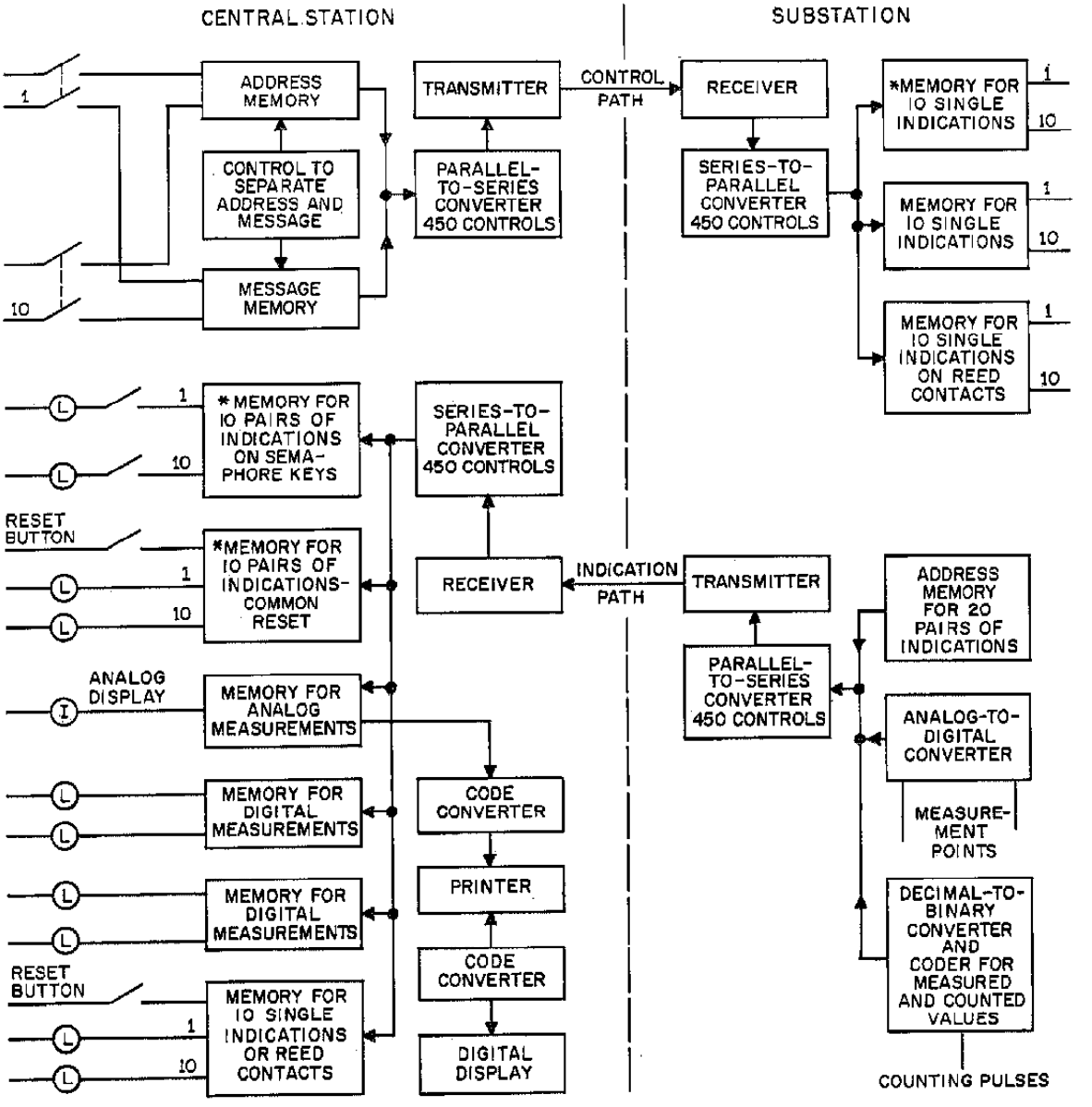


Figure 4—Schematic of Telepuls 12. All the units are electronic except those marked by an asterisk, which are electromechanical units of the Telepuls 10 System.

long and short marks, Telepuls 12 uses single-shot multivibrators, with coupled units in the master clock and single units for checking the marks.

Figure 5 shows how the pulse lengths are distinguished. At the input is shown a binary 0 of 40-millisecond mark and space followed by a binary 1 of 160-millisecond mark and space. The capacitor and rectifier in the left-hand branch permit the positive-going wave front at the start of a pulse to actuate the single-shot multivibrator to produce a 60-millisecond pulse, which is compared with the original signal by the AND gate.

In the right-hand branch, the input pulses are inverted and what was originally the negative-going wave front at the end of the pulse actuates a similar multivibrator, the output of which is also compared with its input by an AND gate. So long as there is an uninterrupted sequence of 40-millisecond marks and spaces, there will be no signal to operate the flip-flop but 160-millisecond marks and spaces will turn the flip-flop to mark and then to space. Thus the binary 1 is distinguished from the 0.

Using these electronic means, the transmission speed is limited only by the characteristics of the connecting telephone channel. At 1200 bauds, which is one of the speeds recommended by the Comité Consultatif International Télégraphique et Téléphonique for data transmission, it takes less than 0.1 second to transmit a pulse telegram.

The mechanical design complies with International Telephone and Telegraph System practices and is also compatible with German Standards *DIN 47432*. Figure 6 shows the units of a central station and of a substation, which have been installed together in one rack

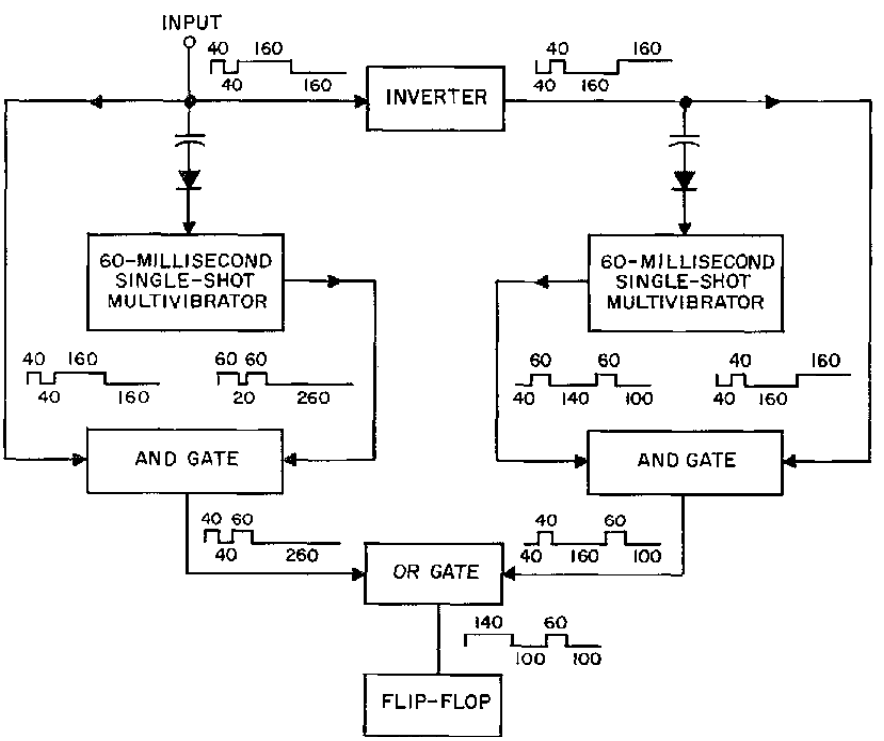


Figure 5—Distinguishing between 0 and 1 pulse lengths in Telepuls 12. The mark and space durations are given in milliseconds.

for testing purposes. The upper part comprises the equipment for the central station. These units are combined with Telepuls 10 panels, which provide for the controls and indications.

4. FMX1

As a reasonable relationship must exist between the cost of an entire installation to be supervised and that of the supervisory equipment, pulse-code transmission is uneconomical for very-small stations. Meshed power networks, for example, have a large number of stations that are required to transmit only 1 indication to a central station. For such installations, the *FMX1* system [3], which uses frequency separation of signals, was developed.

On a party line, each substation employs 1 or a few audio frequencies out of a total of 15 to identify the station and particular equipment being reported on. Reception of each tone at the control station indicates the state "good" or "on" and absence of the signal designates a state of "bad" or "off." Such indications are frequently stored in subcentral stations for common transmission by a pulse-code system to the dispatching center.

The *FMX1* system uses transistors exclusively. The signalling frequencies follow a geometric progression being separated by the factor of 1.224. This permits the use of very-simple filters and renders transmission distortion unimportant. Furthermore, the filter loss is almost the inverse to the cable loss for that frequency, thus permitting any required distribution along the party line. The generators and receivers have shunt-type impedance characteristics, that is, they have high input or output impedances outside their pass bands. Thus, no restrictions need be placed on their distribution along the line. Power consumption, especially of the generators, has to be very low. It was further demanded that it should be possible to use different supply voltages (6, 12, 24, or 60 volts) with a wide tolerance of ± 15 percent. The system uses neither contacts exposed to the atmosphere nor components subject to wear.

Figure 7 shows the four subunits. Only the

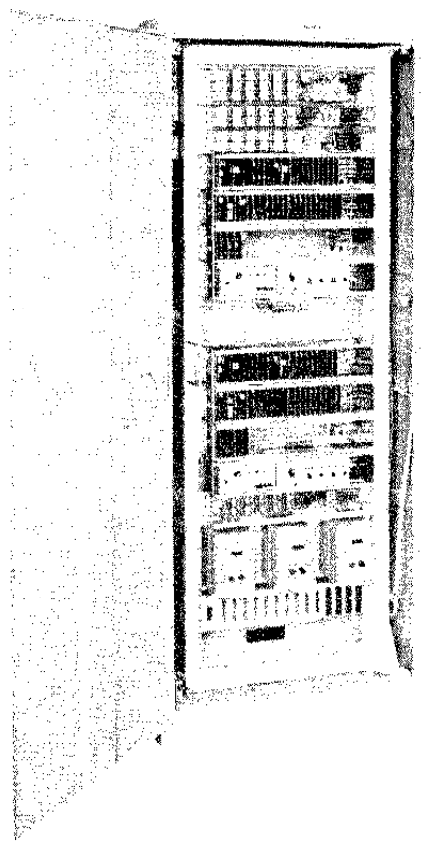


Figure 6—Telepuls 12 bay housing both the central and the substation equipments for testing purposes.

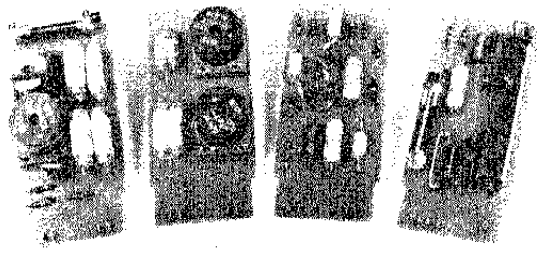


Figure 7—Subunits of *FMX1* system.

generator (left) and the band-pass filter (second from left) differ for the various frequencies. The amplifier and the alternating-to-direct-current converter (right) are identical for all 15 frequencies. Panels and housings, especially of the generators in the substations, are available in various designs to permit interworking of the receivers with the Telepuls 10 or Telepuls 12 components.

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Konrad Bartels was born in Bischofsburg, Germany, on 12 October 1919. He studied at the Technische Hochschule in Danzig.

After the second world war, he joined Standard Elektrik Lorenz in Pforzheim, where he is now chief of the laboratories for auxiliary equipment for radio links and remote-control equipment.

Mr. Bartels represents Standard Elektrik Lorenz in the study group on remote control of the Nachrichtentechnische Gesellschaft.

Digitel 108, A Digital Telemetering System

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1. Introduction

In telemetering, there are substantial advantages in converting the original analog values of the quantities under consideration immediately into digital or numerical values for transmission with modulation methods that provide protection against noise, that are not affected by distortion due to variations in line characteristics, that permit simple error-correction codes to be used, and in which several channels may be multiplexed for prompt transfer of information.

2. Digitel 108 System

A telemetering system must be able to handle all kinds of physical quantities, such as, current, voltage, power, pressure, flow, temperature, resistance, displacement, level, counting, et cetera. Transducers convert these quantities into proportional values of direct current that then actuate analog-to-digital converters.

The transducer may be a simple directly coupled adjustable resistor, thermocouple, differential transformer, or a more-complicated electrodynamic system.

For convenience, 8 inputs have been selected as a basic group. This is a good average of the needs met in practice; for larger installations multiple groups may be employed.

A precision within ± 0.5 percent has been adopted as being adequate for almost all industrial applications. Frequently, the telemetering system includes commercial recording equipment that has an accuracy within only 1.5 percent.

To make maximum use of high-cost transmission lines, time multiplexing is employed with several channels of coded information being sent in rapid predetermined sequence.

Existing frequency-modulated telegraph channels conforming with the recommendations of the Comité Consultatif International Télégraphique et Téléphonique may be used. These 120-hertz channels, which may be part of a 4000-hertz telephone channel, are sufficiently wide to accommodate several telemetering signals simultaneously.

Binary coding has been selected as having the highest information capacity for a given number of pulses. For error detection, the address, message, and such other pulses as may be required by the transmission system are treated as a complete telegram or parity block that must contain an even number of pulses, an extra parity pulse being added at the transmitter if the count would otherwise be odd. The loss of a pulse or the appearance of a parasitic pulse in the received block will cause its rejection.

The transmitted measurements are usually indicated on moving-coil instruments of various types. The trend is to present them in digital form by illuminated figures or to print them, particularly if there are large numbers of measurements and if it is desirable to maintain written records to provide for later analysis of operations. Ultimately, it should be possible to record the measurements on punched cards, on tape, and in some other similar form suited to data processing. If permanent records are not made, it is possible to store on a loop of magnetic tape sufficient transmitted information to permit an analysis to be made of any important occurrences.

3. Description of System

The Digitel 108 telemetering system is divided into plug-in units corresponding to functional elements.

3.1 Transmitter

As shown in Figure 1, the elements to be measured *X1, X2, ..., X8* are permanently connected to appropriate transducers that indicate their states by proportional voltages *V1, V2, ..., V8*.

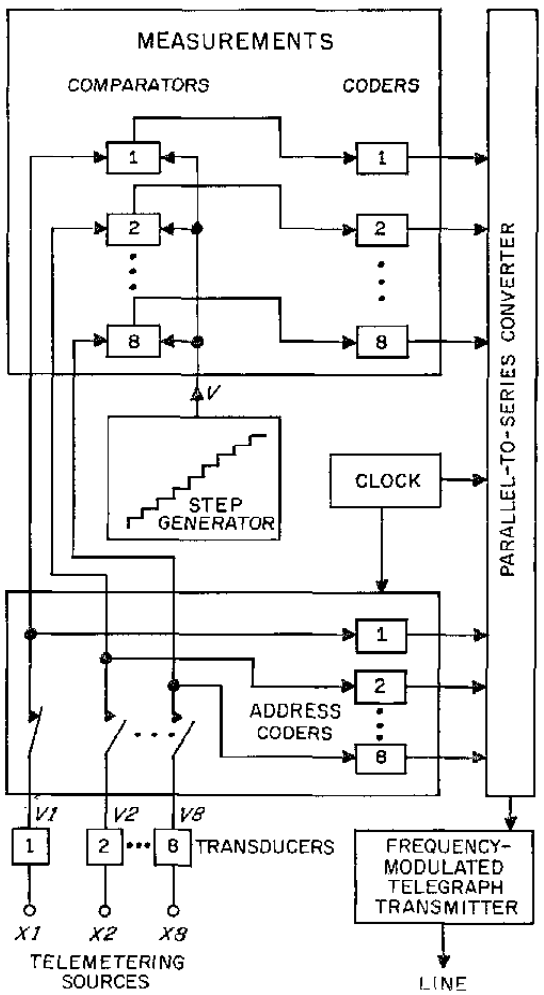


Figure 1—Transmitter. Under control of the clock, each of 8 units to be reported on is examined sequentially and its address and condition are transmitted as a frequency-modulated binary-code telegram to the receiver over a 120-hertz telegraph channel.

The address coder electronically switches each channel voltage $V1 \ldots V8$ sequentially to the measurement coder. As each channel is so connected, the appropriate address set up on bistable flip-flop circuits in binary code is connected to the parallel-to-series converter so that the address and message may be transmitted over the telegraph channel as a single telegram.

The measurement coder includes 8 comparators to which the proportional voltages of the channels are switched by the address coder. Each voltage is compared with V, a stepped voltage wave produced in the step generator. The step matching the proportional voltage to the comparator causes the appropriate output set up in binary code to be connected to the telegraph channel. The step generator is in principle a 7-bit counter controlling transistor switches that connect 7 resistors in a voltage-divider arrangement.

The parallel-to-series converter accepts both address and message and converts the binary signals into serial form for delivery to the telegraph transmitter. All the timing for the binary pulses and the switching sequence is provided by the clock, which is a crystal-controlled oscillator and frequency divider. Figure 2 shows the content of a telegram. A start pulse is followed by a group code to identify the group of channels if more than 8 inputs are involved. Following is the address code, the measurement code, and the parity pulse if needed. A synchronization pulse is then transmitted until the next start pulse. The space pulses are part of the synchronizing system.

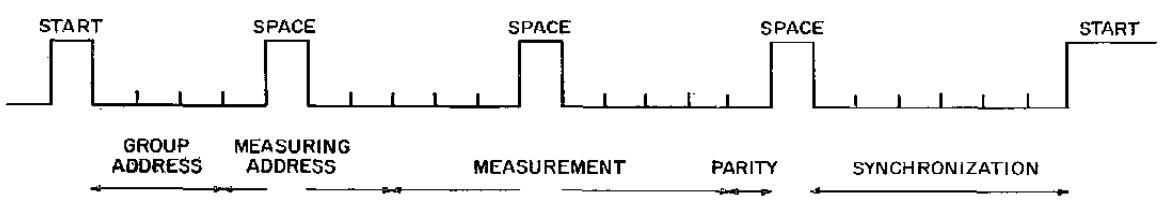


Figure 2—Typical binary-code pulse telegram.

The start and spacing pulses are transmitted at a voice frequency f_n, the binary 0 at f_n-30 hertz, and binary 1 at f_n+30 hertz.

3.2 Receiver

After demodulation, as shown in Figure 3, in the telegraph channel receiver, the pulses go to the distributor, where serial-to-parallel conversion takes place. The message and address are written in the bistable memories of their respective decoders, the measurement code being stored temporarily in a buffer memory while parity and synchronization are being verified.

The distributor, under control of the clock and synchronizer, connects the proper address switch in the address coder to the display converter and initiates the transmission of the corresponding measurement from its decoder to the display converter. Each display decoder receives its message in the original sequence established at the transmitter.

Each display converter can be equipped as an analog device giving a proportional output current corresponding to the encoded message value; producing an indication on a milliammeter or recorder. It may, optionally, contain an amplifier to operate relays of a binary-to-decimal converter to control luminous display units or a printing register, or both.

4. Construction and Maintenance

The use of plug-in printed circuits and transistors provides great flexibility in design of systems and in maintenance. Routine testing is possible with automatic program control and indicator lamps. Test points are provided for oscilloscope comparisons with reference pulse trains reproduced in an instruction manual. Thus, effective and rapid testing together with ready replacement of plug-in units minimizes out-of-service time.

Figure 4 shows a plug-in unit having 20 test points along the side and a switch for checking by the numbered light indicators on the panel the main functions and calibration of the unit.

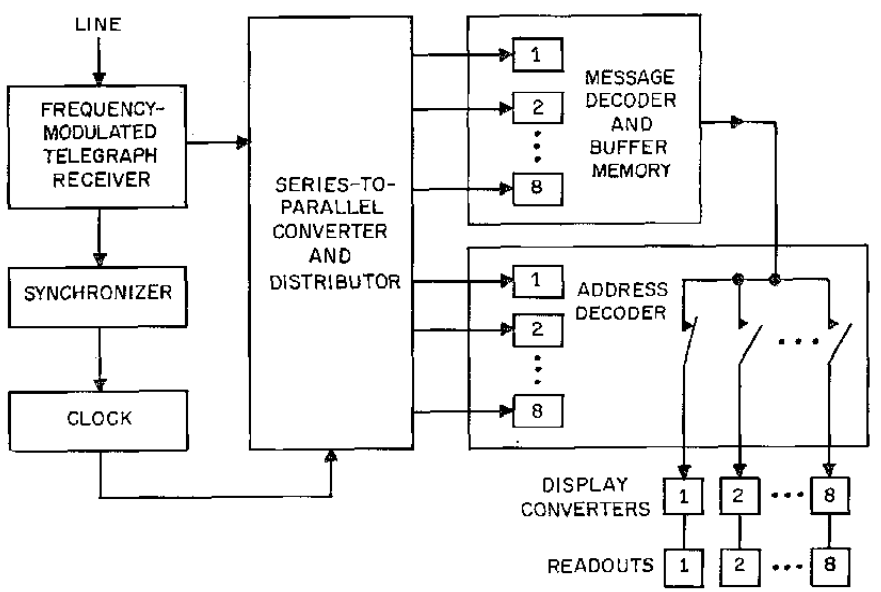


Figure 3—Receiver that reverses the processes accomplished in the transmitter.

Some typical printed-circuit boards are shown in Figure 6.

5. Combined Telemetering and Signalling

In many applications, remote signalling may be required in addition to remote measuring. Pipelines, for instance, may require supervision of many pumping stations and valves to control the flow. For such cases, the 7 information bits corresponding to the message in each telegram can be considered on a binary basis to indicate on-and-off conditions. Thus a group of 56 such reports can be accommodated by a full group of 8 channels. Using the full capacity of the group code, 448 on-and-off conditions can be transmitted. At the receiver, the signals are recorded on relays.

6. Applications

As examples, two applications are outlined, one being for a purely scientific purpose and the other is industrial.

6.1 Radio Astronomy

Radiation from astronomical bodies is not limited to light and heat but may be found in the radio spectrum. The radio telescope has consequently become an important astronomical tool and the Belgian Royal Observatory in Uccle has installed such equipment.

Signals are received from our sun on paraboloidal antennas tuned for operation at 153, 408, 600, and 900 megahertz. A system of 48 paraboloidal antennas in a cross arrangement operates as an interferometer at 408 megahertz.

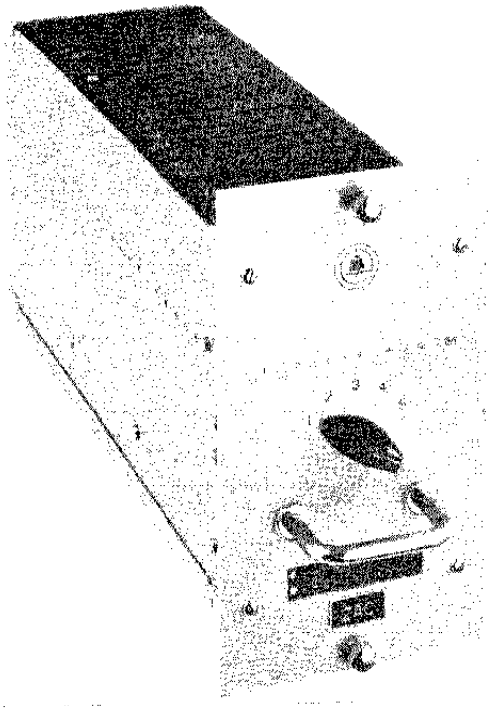


Figure 4—Typical plug-in unit with numbered lights on the panel and a row of test points along the side to facilitate testing and calibration.

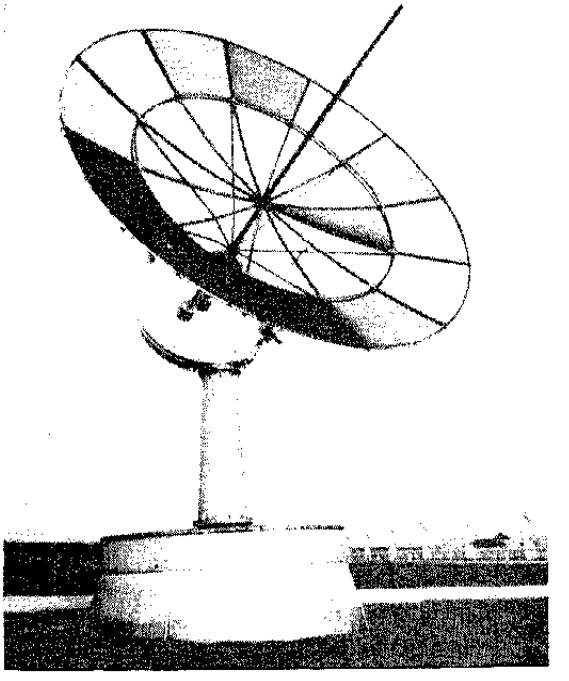


Figure 5—Paraboloidal antenna and in background part of 48-antenna array used for radio astronomy at the Humain field station of the Belgian Royal Observatory.

Another antenna working at 27 kilohertz is used for the study of signals originating in thunderstorms.

The information obtained from these antennas goes to a central control room to operate potentiometer-type recorders. The slide of an auxiliary potentiometer is coupled to certain of the recorders to provide sources of potential for telemetering transmissions.

Telemetering reports on 12 measurement systems are transmitted from the station at Humain to the observatory at Uccle over a 120-kilometer (75-mile) 2-wire telephone line. A 120-hertz band at a nominal frequency of 2460 hertz is used, the lower frequencies of the transmission path being retained for a telephone circuit with its attendant signalling frequencies.

At the receiver, the 12 telemetered reports are recorded on a 12-pen recorder. A punched-tape perforator will be added to permit simple transfer of the material to a computer.

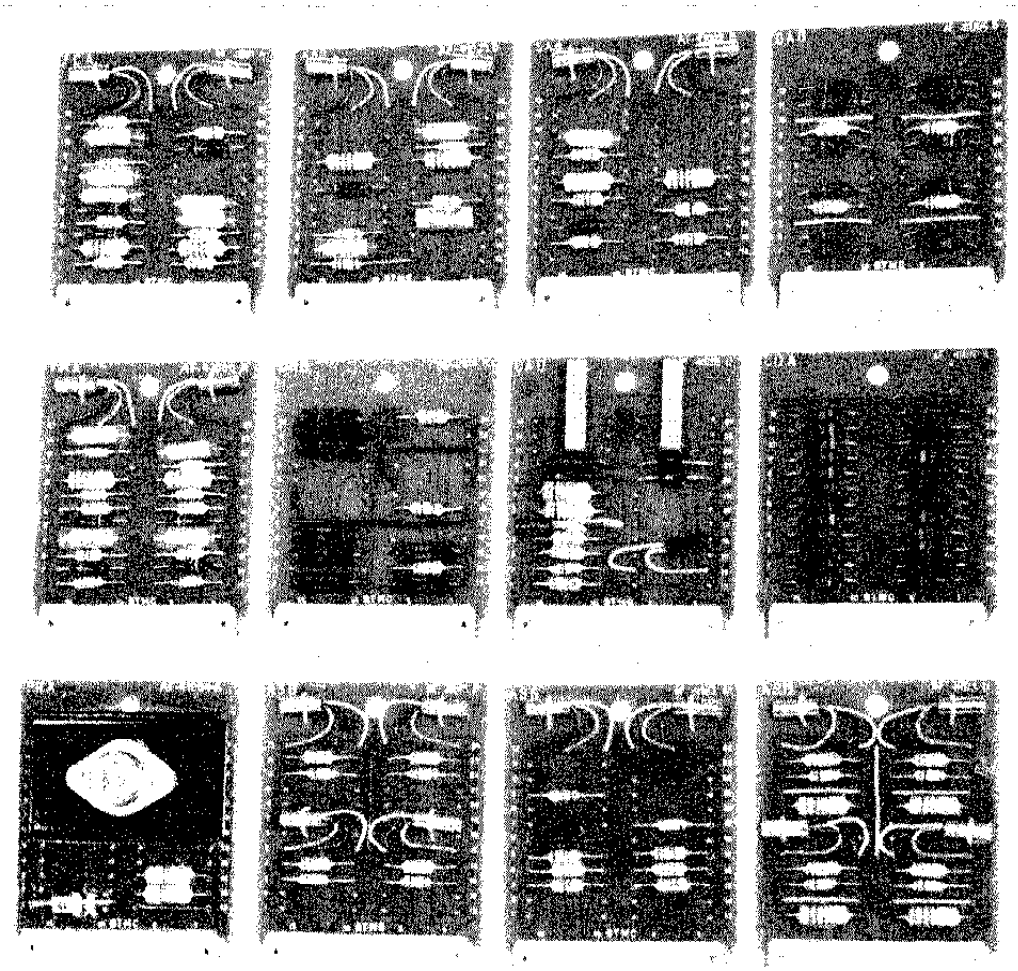


Figure 6—Assortment of printed-circuit panels.

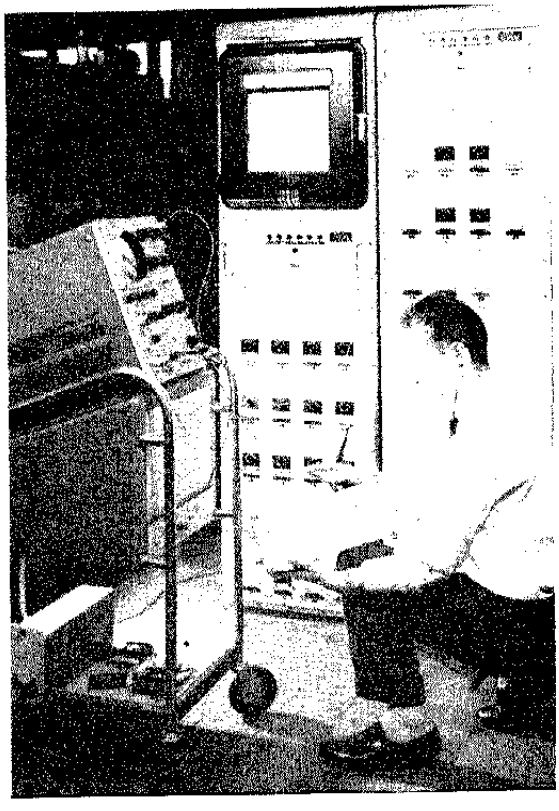


Figure 7—Transmitter and receiver installed at the Uccle observatory in Belgium.

6.2 Water System

In Holland, a water substation at Osdorp must keep the central station in North Amsterdam informed about two flow values derived from differential pressure-measuring systems with potentiometer indication, two pressures from manometers also having potentiometer outputs, input current to pump motors using transducers to convert alternating current into proportional direct currents between 0 and 5 milliamperes, and two integrated flows recorded on counters. An electromechanical transducer converts the count values to proportional pulse signals that are stored in bistable counting circuits for transmission.

Transmission is over a pair of telephone wires in a 120-hertz band at 2700 hertz. The transmission speed is 50 bauds. Remote control is obtained over the same channel but at a lower frequency.

At the receiver, the information is displayed on recorders for the flows and pressures, on milliammeters for the electric currents, and on electromechanical counters for the integrated flows.

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After working at Electrorail in Brussels, Mr. Mattlet joined Bell Telephone Manufacturing Company in Antwerp in 1958, and is now head of the instrumentation and control laboratory.

Transradar–A Narrow-Band Radar Relay Equipment Using Storage Capacitors

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1. Introduction

The effective coverage area of airspace surveillance radar equipment is limited, especially by shadowing effects such as from the horizon, mountains, and clouds. With increasing aircraft speed and traffic density, the time a target appears on a single radar screen is no longer sufficient for evaluation. Effective airspace surveillance requires the evaluation of indications obtained from a number of distant radar stations used in a meshed radar network. The information obtained from each radar station must be relayed to neighboring radar stations, evaluation centers, or other interested agencies.

Depending on the distances involved and the operational situation, one of the methods outlined in Figure 1 may be employed for transmission of the radar information.

Direct connection by means of several parallel cables is standard practice for short distances up to 5 kilometers (3.2 miles).

For longer distances up to 50 kilometers (32 miles), the signals must be suitably mixed to permit their transmission without complications. Such broad-band transmission, shown in *A* of Figure 1, is simple, but costly and inflexible [1, 2].

For distances greater than 50 kilometers (32 miles), narrow-band transmission as shown in *B* of Figure 1 is substantially superior.

The technique of automatic data transmission shown in *C* of Figure 1 requires much larger

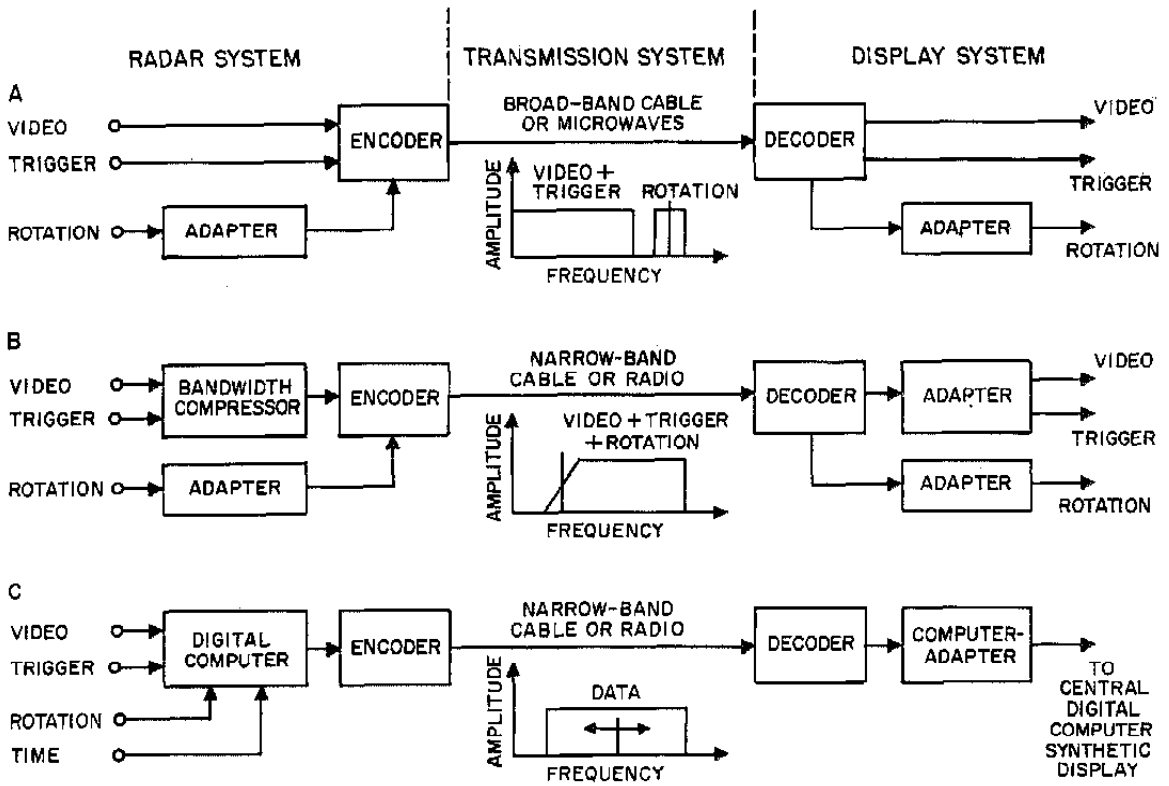


Figure 1—Relaying of radar indications over transmission systems of differing bandwidths. *A* = broad-band. *B* = narrow-band using bandwidth compression. *C* = narrow-band using digital computers.

equipment investments and is still experimental [3, 4]. Thus far, only semiautomatic data systems are being used in practical operations. These systems do not use the principal advantage of meshed radar networks, which is the correlation of overlapping information *before* detection. However, this lack of correlation is avoided by the narrow-band technique discussed below.

One of the decisive requirements for a practical meshed radar network is that the radar information must be compatible with the available transmission channels. These are:

(A) Telephone channels (bandwidth 3 kilohertz).

(B) Broadcast channels (bandwidth 8 to 15 kilohertz) or with similar data lines.

(C) Higher-order channels in carrier telephone systems (bandwidth 48 to 240 kilohertz).

(D) Television channels (5000 kilohertz).

An increasing volume of information can be transmitted with increasing channel width, but the transmission network becomes more inflexible, costly, sensitive to noise, and susceptible to failures. Satisfactory operation of a meshed radar surveillance system therefore depends primarily on the conversion of radar video, trigger, and rotation signals into a narrow-band signal. Since the signal-to-noise ratio is the cardinal criterion of radar technology, it must not be substantially deteriorated by bandwidth compression. Suitable means have to be used to ensure that a minimum of the reflected target energy is lost [5–9].

A bandwidth-compression method meeting these requirements is possible because of the redundancies contained in the radar signal [10]. For ground surveillance radar, to which this paper is limited, the redundancies consist of the following:

(A) Multiple scanning during one revolution of the antenna, with each target appearing in several (for example $n = 10$) successive pulse periods.

(B) Safety factor against overrange (for example $K = 2$), which is the ratio between the unambiguous and nominal maximum ranges, considering variations in target areas.

(C) Dead time when the operative range is shorter than the maximum range.

(D) For remote operation, possible reconsideration of the compromises among pulse length, relative resolution, and evaluation accuracy. For example, transmission accuracy can be adapted to possible location errors, depending on application and evaluating method, or adjusted to equal resolution in radial and tangential directions at half the maximum range [11].

(E) Low target density. Usually about 10 percent of the roughly 200 000 picture elements on the radar screen are activated at once. Of those activated, only a small number are associated with true flying objects while the rest are caused by ground clutter, clouds, and noise [4].

Elimination of redundancies A through D permits a reduction of the transmission bandwidth in most cases by 2 to 3 orders of magnitude, leading to the transmission of a corresponding time-extended radar signal. The equipments required for this bandwidth compression are relatively simple, inexpensive, and rugged, and their development is sufficiently advanced to use them for practical operation [12–15].

Elimination of redundancy E is more difficult. It normally requires extensive use of electronic computers [3, 4]. The transmitted signal containing the target coordinates is then a pulse train approximately 30 bits long per target. Substantially higher equipment investments are needed for such a digital data system, and stringent requirements would have to be imposed on the characteristics of the radar signal and the suppression of noise in it, since only a very-limited number of targets can be trans-

mitted over narrow-band channels (50 targets per second over a telephone line). Such a definitive limitation, however, is highly questionable, since radar surveillance attains its greatest importance during periods of heavy disturbances produced by bad weather and jamming. In practice it is advisable to introduce a margin of 1 to 2 orders of magnitude above the theoretical limitation. This results in approximately the same bandwidth for digital transmission as for narrow-band picture transmission.

Simple picture transmission with bandwidth compression, therefore, can be used both as a first stage in establishing an integrated radar system and later for interference-insensitive parallel operation. This type of transmission is well suited to digital data systems for monitoring, emergencies, and as a service channel. It will also be able to hold its place in all future integrated surveillance systems [16, 17].

2. Principle of Bandwidth Compression

Bandwidth compression requires the provision of a storage system between the output of the radar receiver and the input of the transmission path to the slave display unit. This storage unit can operate on the cathode-ray-tube picture produced at the radar or on the video output signal from the radar.

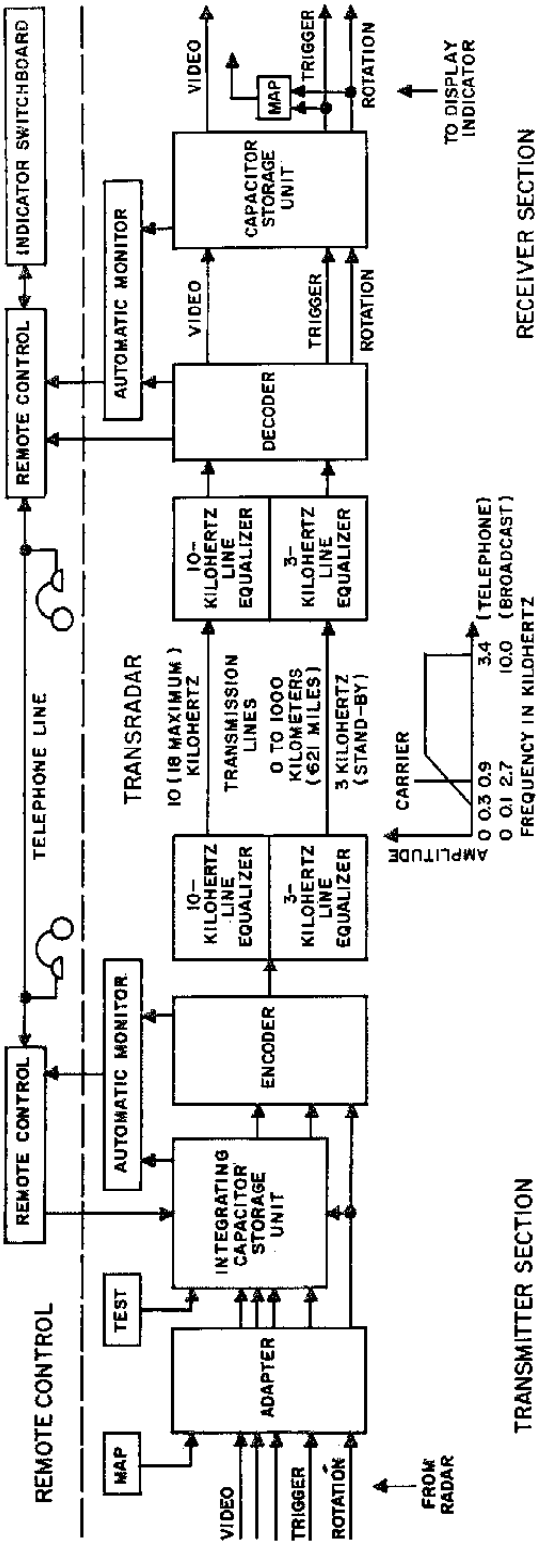
A picture storage unit has the duty of storing the complete radar signal received during one or several antenna revolutions. It therefore allows free choice of the scanning mode, which may be a conventional television scan [14, 15, 18], and under certain circumstances permits some use of the low target density for further bandwidth compression and track formation. Currently, the only available storage device for this purpose is the signal storage tube.

A video-signal storage unit, on the other hand, stores the information for only a small sector of antenna rotation. After integration and evaluation, the information is erased and replaced by that of the following sector [11–13, 17]. For airspace surveillance where the targets move rapidly, practically a full pattern of information must be transmitted during each antenna revolution. For this purpose, a rugged video-signal storage device for bandwidth compression at the origin of the transmission line, combined with a simple picture storage unit such as a storage tube or scan converter in the slave display unit for track formation, was found to be a most-suitable solution operationally and economically.

A thorough investigation of the principles that can be used for the realization of a video-signal storage device led to the choice of a bank of discrete capacitors, their quantity (250, 500, or 1000) depending on the required range resolution. This system uses pulse-operated gate circuits instead of tolerance-sensitive scanning by electron beams, and direct-current connections instead of coupling by secondary emission as is conventional for storage tubes.

The functions of this integrating storage device are thus based partly on analog and partly on digital techniques. Analog processing of the video signal achieves economy, excellent linearity of the integration characteristic, and lenient timing requirements. Performing all scanning operations with digital circuits ensures maximum reliability, high accuracy, and easy maintenance. Scanning can be started and stopped as required, using radar systems with nonconstant pulse frequency, and the scanning speed may be changed instantaneously with change of range or transmission bandwidth.

The behavior of a building block in a radar system with respect to natural and man-made interference is important from the viewpoint of information theory. The best arrangement for discerning a periodic pulse sequence in statistically distributed noise is a filter whose response-frequency curve provides periodically alternating pass and stop bands distributed so that all harmonics of the pulse frequency are within the narrow pass bands while the intervening noise power is suppressed by the stop bands. However, since the useful signal in a radar system is not absolutely periodic but



has a finite number of pulses per target, the pulse harmonics form sidebands that require wider pass bands. The practical realization of such periodic or cam filters presents certain difficulties as in obtaining very-steep filter characteristics and in the alignment procedure.

Much easier, however, is the realization of the equivalent arrangement in the time domain, this filter being a storage integrator or combining filter [5–9]. The selectivity then stems from linear addition of the coherent signals and increases with the number of identical summed pulse periods. Here also, the interference suppression is limited by the finite number of pulses per target [8, 9].

In this respect, the capacitor store shows its superiority. First, the integration characteristic of the storage elements is very linear. Second, digital switching by the pulse distributor readily accommodates aperiodic radar signals such as are encountered in phase modulation of the transmitted pulses (jitter) without loss of accuracy. To accomplish this, the digital pulse distributor is designed to stop when arriving at the last storage capacitor and to remain in start position on the first storage capacitor until the next triggering pulse arrives, thus providing dead time. The digital feature of this start-stop operation requires practically no transient time. The accuracy of distribution, and therefore both the effectiveness of interference suppression and the accuracy of distance measurement in the narrow-band signal, is independent of the duration of the variable dead time.

The writing and scanning circuits of the capacitor storage unit use only transistors as active elements. It is unsurpassed for ruggedness and reliability. The principle of operation practically excludes distortions in geometry as

Figure 2—Diagram of transradar system, with remote control, inserted between radar and display systems.

well as voltage and temperature influences. Adjustment to the operating data of individual radar sets is possible without using special measuring instruments.

3. Transradar Design

A bandwidth compression system has been developed for long-range narrow-band transmission of radar signals and operates on the principles outlined above. The equipment requires the same connections to the radar system as a normal display unit, and delivers the transmitted signals in the form required by the respective remote indicators. Thus it may be considered as an extension of the connecting cable between the radar receiver and the indicators.

Figure 2 is a block diagram of the equipment, with a remote-control facility, incorporated in a radar system.

The signals arriving from the radar equipment (video, trigger, rotation, and sometimes map video) are passed through adapter networks where they are matched to the operating requirements of the bandwidth-compression device. Free selection is possible of either normal, moving-target indication, or logarithmic receiver video signals, and another smaller-amplitude signal produced in synchronism (map; identification, friend or foe; et cetera) may then be superposed on the selected video signals. The rotation signals are converted into angular pulses, used in the bandwidth-compression equipment for controlling the narrow integration sector, and added within the encoder to the narrow-band transradar video signal to be transmitted. A number of different adapter networks are available to permit connection of the equipment to different types of radar sets.

Bandwidth compression is performed by the integrating capacitor storage unit, whose functions are illustrated in Figure 3. This video-signal storage device comprises a bank of 500 capacitors with electronic write and read

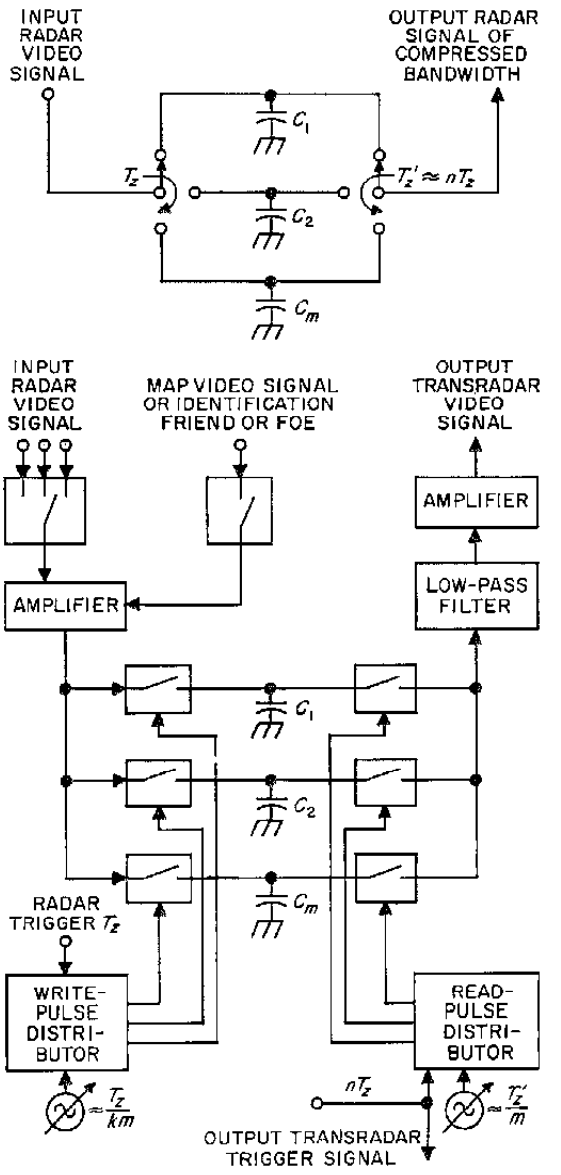


Figure 3—Diagram of the integrating capacitor storage unit. At top is the functional arrangement for storing the radar signals sequentially in capacitors, which discharge at a much slower scanning rate and thereby compress the frequency bandwidth. Below is the same functional arrangement using pulse-operated electronic gates to switch the storage capacitors. m = number of range elements. n = number of hits per target. k = unambiguous range divided by range in use.

gates, one capacitor being provided for each distance element (C_1 through C_m, $m = 500$).

A writing timer synchronized by the radar trigger distributes the video signals of consecutive scan periods n to the m storage capacitors arranged as a radial line. The capacitor line is traversed once during each pulse period. The writing time may be selected as required for range resolution in the storage unit.

The read-timing depends on the integrated angular sector derived from the radar antenna pattern plus the bandwidth of the transmission path (range resolution determined by transient time of the transmission line). The charges stored and integrated in the capacitors are removed via read gates and delivered to the transmission device via a low-pass filter.

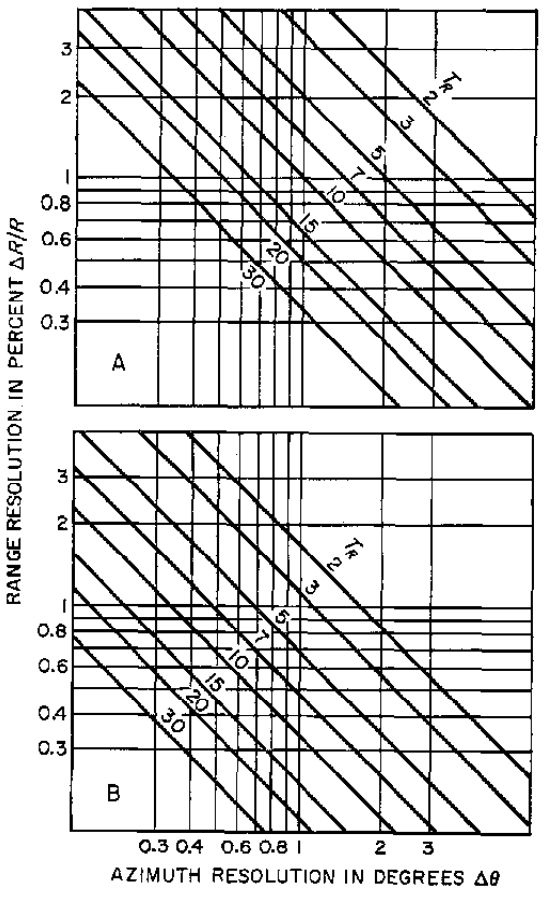


Figure 4—Resolution obtainable over 3-kilohertz telephone line *A* and 10-kilohertz line *B* for indicated values of antenna-rotation time in seconds.

The narrow-band transradar video signal, the read trigger, and the rotation pulses are first added to form a composite signal at the input of the carrier terminal equipment. This signal is then translated to the transmission frequency range by vestigial-sideband signal techniques, using double conversion and a Nyquist filter. The relative width of the vestigial sideband corresponds approximately to well-established television criteria, while the modulation depth can be held substantially lower since transradar signals are relatively insensitive to transmission interference. The carrier position at the lower end of the pass band permits both good and bad lines to be used, which is an important advantage in meshed networks made up of different line sections. The time characteristics are similar to that of a television signal with positive modulation, having large carrier amplitude for bright picture points and blanking of the carrier during synchronizing pulses. The system can work with two parallel carrier terminals to allow normal operation over a broadcast line and emergency operation over a telephone line.

The equalizers for the frequency delay characteristic are accommodated in the receiving terminal. The amplitude characteristic normally does not require equalization. The decoder is also located at the receiving terminal with the remote display unit. Here the composite signal is reconverted into video-, trigger-, and rotation-output signals in three separate channels for display.

No abnormal requirements are imposed on the transmission lines. However, the system design makes them the controlling factors for the quality of the transmitted picture.

The resolution attainable in practical operation over the most-commonly used transmission lines is shown in the diagrams of Figure 4, using the actual signal carrier frequencies and average equalization. The integrated angular

sector, or scanning frequency, whose magnitude influences the angular resolution, is determined from the angular pulses derived from the antenna rotation signals. The attainable range resolution thereafter results mostly from the bandwidth (transient time) of the transmission line. By varying the scanning frequency, the relationship between angular and range resolution obtainable for a given transient time can be adjusted (see Figure 4).

The mean duration of one antenna revolution of practical radar sets is proportional to the square of maximum range. Therefore, the absolute resolution is almost independent of the type of radar set used when, as is normal in radar networks, the full range is transmitted. The absolute resolution primarily depends on the type of transmission line, while the transmission distance depends on the quality of the transmission line. Normal transmission networks may span distances of 1000 kilometers (621 miles) or more. Along the transmission line, the signal may branch off by means of standard terminal equipments to a number of remote evaluation centers.

Figure 5 shows the time diagrams for operation with various displays. Line 1 shows a radar video signal displayed on the original plan-position indicator in synchronism with the rotating directional antenna. Lines 2 and 3 show, respectively, the bandwidth-compressed video signal at the output of the integrating capacitor storage unit and the line signal. After demodulation at the receiving terminal, various forms of display are possible, depending on the type of indicator (lines 4 through 6 of Figure 5).

Video and trigger signals can be displayed directly if a special slow-scan indicator is available (line 4). As this is seldom the case, a capacitor storage device, similar to the one at the transmitting end but with inverse operating mode, is usually incorporated in the receiving terminal. The incoming video signal is written into the storage unit at slow speed and subsequently read out at high speed one or several times. This method permits the use of normal fast-scan indicators (line 5). Of special interest is the extremely long dead time that occurs when the storage is scanned only once (line 6). This dead time permits inserting additional information such as direction-finder bearings, marking symbols, or other radar signals, in indicators with interscan facilities.

Various connecting facilities are available to

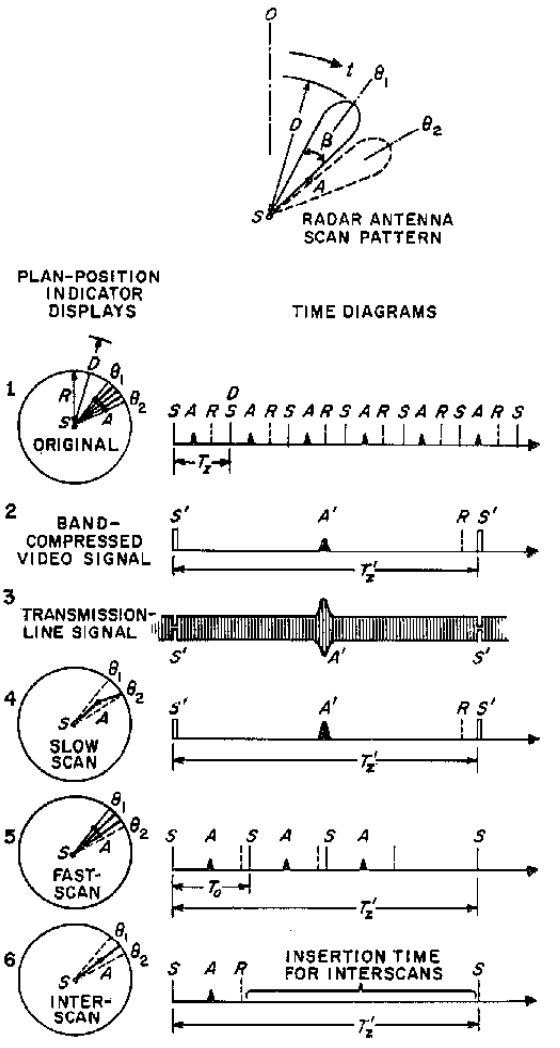


Figure 5—Plan-position indicator displays and corresponding time diagrams of signals. A = target returns. D = unambiguous range corresponding to pulse frequency. R = range in use. S = triggered pulse.

correlate the transmitted rotation signals to the deflection of the indicator systems. These facilities may be synchros with normal gear ratios or rotating shafts or they may be sine-cosine voltages varying with the azimuth. An automatic facility ensures fast reorientation to north after line interruptions or other disturbances of synchronous operation. The reacquired rotational information may also be used for synchronizing video mapping equipments or other evaluation aids.

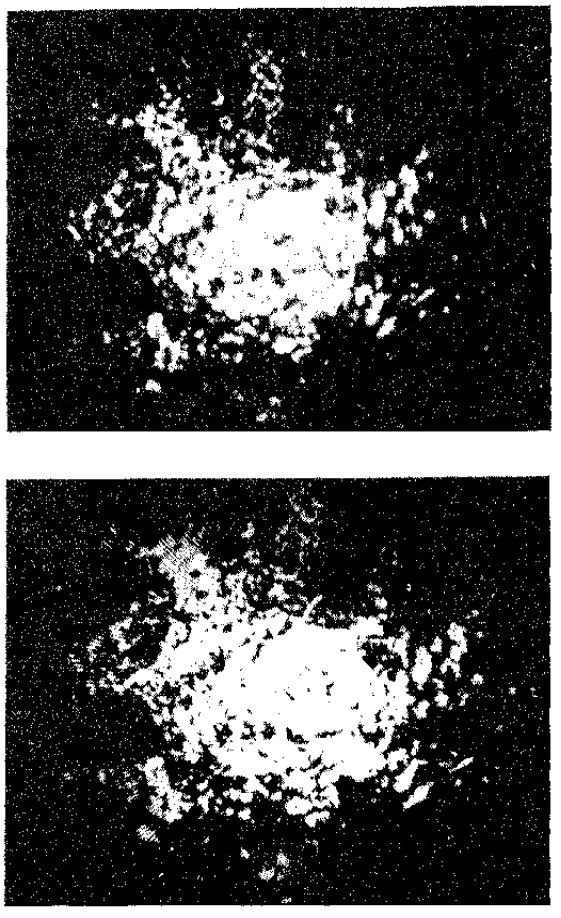


Figure 6—At top is an original radar picture, while below the same picture is shown after transmission over more than 600 kilometers (373 miles) of telephone lines. Each exposure was for exactly 1 antenna rotation. The pictures were taken on different days, which accounts for minor differences.

Figure 6 shows two photographs, one taken from an original indicator before bandwidth compression and the other from a slow-scan indicator after transmission over a normal telephone line of approximately 600 kilometers (372 miles). Both photographs were exposed for exactly one antenna revolution. Since the two pictures were taken on different days, because of the distance between the two locations, they do not correspond in all details. It is also very difficult to reproduce radar displays photographically.

Figure 7 shows the transradar transmitting and receiving equipments, which are very similar in design. They are subdivided into plug-in units, which simplify maintenance and rearrangement for special operational requirements. Transistors are the only active elements used. The equipment has been in operation for some time with very good evaluation results, and is now in production.

4. Applications

Narrow-band radar relay systems have the following important applications:

(A) Integrated radar network for airspace surveillance, permitting detection behind shadow areas, with great interscan time-sharing capacity and flexibility in emergencies.

(B) Relaying radar pictures from unattended or mobile gap-filled radar stations, with insertion into the long-range surveillance picture.

(C) Centralized air traffic control with common use of civil and military radar equipments (transmission before evaluation).

(D) Radar connection to positions not having their own radar sets.

(E) Target acquisition and coordination of short-range systems with the aid of the general display.

(F) Making radar surveillance immune to interference, through dense meshing of the radar sets working in space, time, and frequency

diversity, by combining the output signals of several transradar systems in time multiplex.

(G) Combining raw radar information and synthetic information, since the extremely long dead time permits simultaneous display of both types of information on one display unit.

(H) Centralized weather service.

(I) Documentation of the air-traffic situation in a realistically reproduceable form, by using standard audio magnetic tape recording.

(J) Training, by using selected portions of magnetic tape records.

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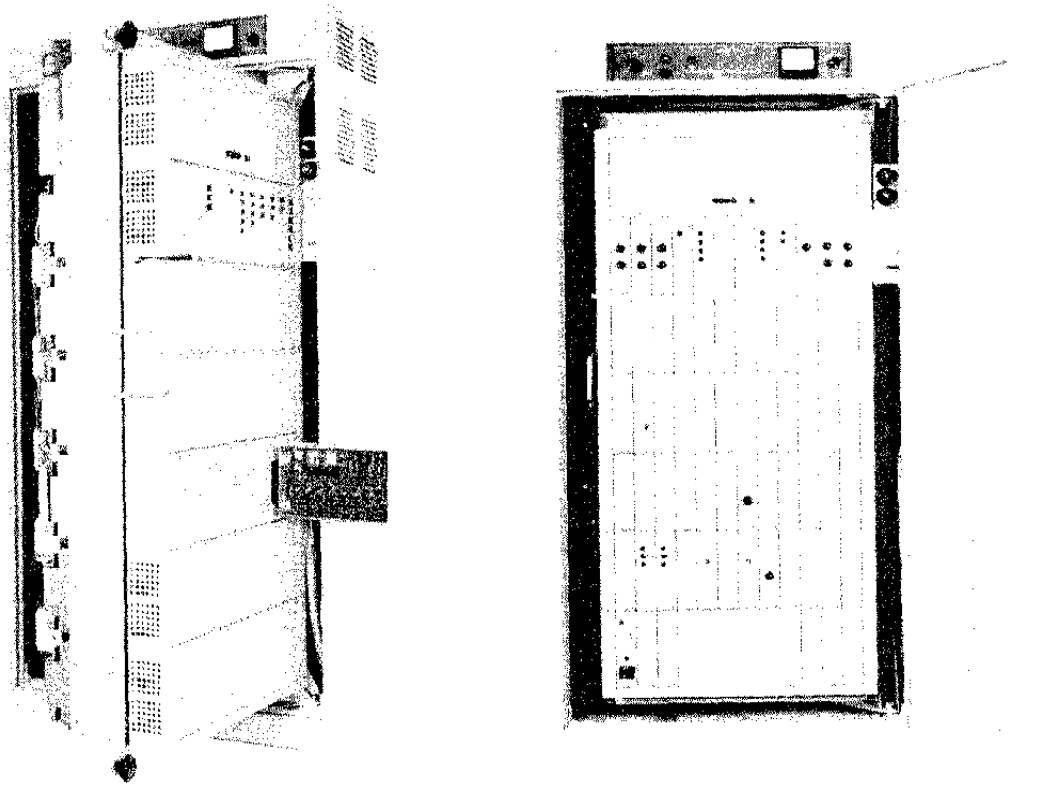


Figure 7—Transradar equipment. At left is the transmitter section for installation at the radar site. At right is the receiver section for installation at the remote indicator site. Dimensions of each unit are 1554 by 800 by 420 millimeters (61¼ by 31½ by 16½ inches).

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In September 1950, he joined Standard Elektrik Lorenz, where he was engaged in the development of the first television receivers, transmitters, and transmission lines. In 1953, he became head of the television laboratory, which included responsibility for the development of transradar equipment.

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Optimization of Network Configurations Through Queueing Theory

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1. Introduction

This paper describes a generalized queueing theory model that can be used to perform traffic studies for the optimization of network configurations with respect to time and equipment costs. The model considers parameters such as time of input arrivals, priorities, variable processing rates, and variable traffic densities, and may be used for both steady-state and dynamic conditions. The model demonstrates a way to reduce most multichannel problems to single-channel problems by analyzing a *type* of station rather than a *particular* station. That is, a multichannel processing station is considered as a single-channel processing station with a modified processing rate.

The queueing theory model is applicable only to networks having the property that from one point to another there is not more than one possible path. It follows that for a given configuration, the traffics corresponding to various junctions are known, whereas for a network not possessing the above property, most of those traffics would be additional unknowns.

A processing station is defined as a service unit that operates on inputs requiring service. It is considered in this paper as servicing only a single input at any specific point in time. If a processing station has parallel service units, it may be considered to be a single station with a processing rate equal to the sum of the individual processing rates of the parallel service units multiplied by an interaction constant. This constant corrects for the effect of interaction between the parallel service units. The interaction varies for each problem encountered and the constant, determined empirically, must be given as an input to the queueing model. Interaction may cause the processing rate of the station to be more or less than the sum of the individual processing rates. Where there is no interaction, the station processing rate is equal to the sum of the individual processing rates of the parallel service units

$$\text{(interaction constant} = 1\text{)}.$$

The expected queue length at a specific processing station is the expected total number of inputs requiring service within a particular time increment and awaiting processing because of the unavailability of a station.

The queueing model assumes a Poisson input rate and exponential holding time. The holding-time distribution is defined as that distribution generated by determining the probabilities that service is continuing on the input being processed at discrete points in time. Accordingly, holding time is the amount of time that an input spends at the station.

The queueing theory model develops the following outputs for a processing station:

(A) A generalized set of curves showing traffic density versus queue length as a function of processing rate and priority.

(B) A generalized set of curves showing traffic density versus waiting time as a function of processing rate and priority.

The model is time-independent; therefore, in analyzing a particular processing station, variable time increments Δt may be considered. From a generalized set of curves, it is possible to construct histograms for any particular station in a configuration. These histograms show queue length versus time of day and waiting time versus time of day, respectively. It is possible to determine the expected waiting time of an input if it arrives at a particular time of day and the expected number of inputs that must be processed before the input under consideration is processed. Both histograms are functions of priority.

An example is given of how this generalized set of curves may be used in determining the optimum configuration of a communication

network, subject to equipment and cost constraints.

2. Queueing Theory Model

2.1 General

For the application of queueing theory, the characteristics that must be specified include:

(A) Input distribution—the frequency or time distribution of occurrence of inputs to each station in the system.

(B) Holding time distribution—the time or distribution of times required by the station to perform its functions.

(C) Queue discipline—a set of rules indicating what the system will do about an input if it arrives when the station is still serving a previous input; that is, whether a waiting line forms or the input is lost.

The model has been developed under the following assumptions (explained in Section 2.1.2):

(A) The input distribution to any type of station is Poisson.

(B) The holding or service time distribution is exponential.

2.1.1 *Notation*

The notation that will be used throughout this paper is:

λ = total number of inputs introduced into the system during the time interval under consideration.

P_i = percentage of the total inputs having the ith precedence introduced into the station during the time interval under consideration.

$\lambda_i = P_i\lambda$ = number of inputs having the ith precedence introduced into the station during the time interval under consideration.

$\lambda'_k = \sum_{i=1}^{k} \lambda_i$ = number of inputs having precedence greater than or equal to k introduced during the time interval under consideration.

N = total number of precedences.

μ = number of inputs that the station can process one at a time during the time interval under consideration.

$\rho'_k = \lambda'_k/\mu$ = traffic intensity or utilization factor for a service facility.

Lq'_k = queue length composed of inputs introduced during the time interval under consideration and having precedence greater than or equal to k.

Lq_k = queue length composed of inputs of the kth precedence introduced during the time interval considered.

W_k = mean waiting time for an input having kth precedence introduced during the time interval under consideration.

It can be seen that if $\rho_k = (\lambda_k/\mu) > 1$, the queue must grow longer without limit. The probability of any specified queue length must then tend toward zero as time increases, as must the probability of any specified finite waiting time. However, it is clear that for finite intervals of time, ρ may be greater than 1, although the queue length need not increase without limit over a longer period of time. Where $\rho_k < 1$, the condition is termed steady state. Where $\rho_k > 1$, it is termed dynamic or nonsteady state.

2.1.2 *Assumptions*

The model is developed using the following assumptions:

(A) Poisson input distribution.
(B) Exponential holding-time distribution.
(C) Priority queue discipline.

(A) According to Saaty [1], a Poisson input exists when:

(1) The total number of arrivals during any given time interval is independent of the number of arrivals before the beginning of the interval.

(2) For any interval $(t,\ t + dt)$, the probability that exactly one arrival will occur is $\lambda dt + f(dt^2)$, where λ is a constant. The probability that more than one arrival will occur is of the order of magnitude of dt^2 and may be neglected. The input rate λ may, of course, vary for Δt time increments chosen at different times.

(B) Similarly, an exponential holding-time distribution exists when the following condition holds:

If a station is occupied at time t, the probability that it will become free during the following time increment is μdt, where μ is a constant. It follows that the frequency function of the service times is $\mu e^{-\mu t}$ while the mean duration of service is $1/\mu$, since the expected value of t is

$$E(t) = \mu \int_o^\infty t e^{-\mu t}\, dt = \frac{1}{\mu}. \qquad (1)$$

(C) Inputs of different types enter a station with assigned priorities for service. Whenever the station is free to accept an input, it selects the one of highest priority within each precedence class on a first-come first-served basis. However, if an input of higher priority enters the system while one of lower priority is in service, the service is not pre-empted; that is, the item being processed is not sent back to the waiting line. The model requires that provisions be made for storing queued items on a higher-to-lower-priority basis. It is understood that items of highest priority are also queued as required. Therefore, the expected length of the line is

$$Lq = \sum_{i=1}^{N} \lambda_i W_i. \qquad (2)$$

2.2 Queue Lengths

2.2.1 *Steady-State Condition* $\rho < 1$

It has been shown [2] that the expected number of units in the queue is given by

$$Lq = \frac{\rho}{1-\rho} - \rho = \frac{\rho^2}{1-\rho}. \qquad (3)$$

This presupposes that there are times when the channel is not occupied because of the nature of the Poisson distribution.

2.2.2 *Nonsteady-State Condition* $\rho > 1$

Clarke [3, 4] details a proof for the nonsteady-state condition when he derives the following expression.

$$Lq = (\rho - 1)\mu\tau + \frac{1}{\rho^m(\rho - 1)} + m + 0(\tau) \qquad (4)$$

where m is the queue length at the beginning of the time interval under consideration, τ is the length of the time interval, and $0(\tau) \to 0$ as $\tau \to \infty$. Since τ is considered large when it is much larger than $1/\mu$ (τ being considered a unit of time length, $\tau = 1$) $0(\tau)$ may be considered 0.

Therefore

$$Lq = (\rho - 1)\mu + \frac{1}{\rho^m(\rho - 1)} + m. \qquad (5)$$

However, since we are not considering the previous time interval in the generation of a model to derive a generalized set of curves, the leftover queue length can be set equal to 0, resulting in

$$Lq = (\rho - 1)\mu + \frac{1}{\rho - 1}. \qquad (6)$$

The above analysis assumes that the system has settled down. The time required for this is called the relaxation time. For the system under consideration (single-channel Poisson input and exponential holding-time distribution) this is approximated by

Relaxation time

$$= \frac{1}{(\mu^{1/2} - \lambda^{1/2})^2} = \frac{(\mu^{1/2} + \lambda^{1/2})^2}{(\mu - \lambda)^2}. \qquad (7)$$

For $\rho = 1$

$$Lq = 2\left(\frac{\mu\tau}{\pi}\right)^{1/2} + 0(\tau) \qquad (8)$$

where $0(\tau)$ is a bounded function for which $0(1)$ can be considered 0.

Therefore, for $\rho = 1$

$$Lq = 2\left(\frac{\mu}{\pi}\right)^{1/2}. \tag{9}$$

Considering priority, modify the preceding theory in the following manner.

Let $\lambda'_k = \sum_{i=1}^{k} \lambda_i$ = the number of units having priority greater than or equal to k introduced during the time interval considered.

$\rho'_k = \dfrac{\lambda'_k}{\mu}$ = the traffic intensity or utilization factor.

Lq'_k = the queue length composed of inputs, having priority greater than or equal to k, introduced during a specific time interval.

Lq_k = the queue length composed of inputs of the kth priority introduced during the time interval considered.

2.2.3 *Summary*

Since queue length is independent of priority within any group of inputs to a processing station, we conclude

$$\text{For } \rho'_k < 1,\ Lq'_k = \frac{(\rho'_k)^2}{1-\rho'_k},\quad k = 1, 2, \cdots, N. \tag{10}$$

$$\text{For } \rho'_k = 1,\ Lq'_k = 2\left(\frac{\mu}{\pi}\right)^{1/2},\quad k = 1, 2, \cdots, N. \tag{11}$$

$$\text{For } \rho'_k > 1,\ Lq'_k = (\rho'_k - 1)\mu + \frac{1}{\rho'_k - 1},\ k = 1, 2, \cdots, N. \tag{12}$$

Once Lq'_k has been computed for all k values, then Lq_k may be simply calculated by

$$Lq_k = Lq'_k - Lq'_{k-1}. \tag{13}$$

2.3 Waiting Times

Kesten and Runnenburg [5] show that for the steady-state condition $\rho_k < 1$, assuming an arbitrary holding time and Poisson input distribution, the expected waiting time for kth priority is given by

$$W_k = \frac{\sum_{i=1}^{N} \lambda_i \mu_i^{(2)}}{2\left(1 - \sum_{i=1}^{k-1} \lambda_i \mu_i^{(1)}\right)\left(1 - \sum_{i=1}^{k} \lambda_i \mu_i^{(1)}\right)},\quad k = 1, 2, \cdots, N \tag{14}$$

where

$$\mu_i^{(p)} = \int t^p \, dF_i(t) \tag{15}$$

and where $F_i(t)$ is the cumulative holding-time distribution for the ith priority units in a single-channel queueing operation.

In the saturation (nonsteady-state) condition $\sum_{i=1}^{N} \lambda_i \mu_i^{(1)} \geq 1$, we can find an integer s $(0 \leq s \leq N)$ such that

$$\sum_{i=1}^{s} \lambda_i \mu_i^{(1)} < 1 \quad \text{and} \quad \sum_{i=1}^{s+1} \lambda_i \mu_i^{(1)} \geq 1. \tag{16}$$

Testing whether precedence k is less than s

$$W_k = \frac{\sum_{i=1}^{s} \lambda_i \mu_i^{(2)} + \left(\frac{\mu^{(2)}_{s+1}}{\mu^{(1)}_{s+1}}\right)\left(1 - \sum_{i=1}^{s} \lambda_i \mu_i^{(1)}\right)}{2\left(1 - \sum_{i=1}^{k-1} \lambda_i \mu_i^{(1)}\right)\left(1 - \sum_{i=1}^{k} \lambda_i \mu_i^{(1)}\right)},\quad k = 1, 2, \cdots, (s-1). \tag{17}$$

For $k \geq s$, we modify (2) to derive the expected waiting time.

$$Lq = \sum_{i=1}^{N} \lambda_i W_i.$$

Therefore

$$Lq'_k = \sum_{i=1}^{k} \lambda_i W_i \tag{18}$$

and

$$W_k = \frac{Lq'_k - \sum_{i=1}^{k-1} \lambda_i W_i}{\lambda_k}. \tag{19}$$

An approximation that can be used instead of (19) is

$$W_k = \frac{(\rho'_k - 1) + W_{k-1}\left(1 + \frac{\lambda_k}{\lambda_{k-1}}\right)}{2}. \quad (20)$$

2.3.1 *Summary—Waiting Times*

2.3.1.1 *Steady-State Condition*

Using a Poisson input distribution and exponential holding-time distribution, (14) becomes

$$W_k = \frac{\rho'^N}{\mu(1 - \rho'_{k-1})(1 - \rho'_k)},$$
$$\rho'_k < 1, \rho'_o = 0. \quad (21)$$

2.3.1.2 *Nonsteady-State Condition*

Finding the first and second moment of the exponential holding-time distribution and substituting these expressions in (17) we obtain

$$W_k = \frac{1}{\mu(1 - \rho'_{k-1})(1 - \rho'_k)},$$
$$k < s, \rho'_o = 0. \quad (22)$$

When $k \geq s$

$$W_k = \frac{Lq'_k - \sum_{i=1}^{k-1} \lambda_i W_i}{\lambda_k} \quad (23)$$

or

$$W_k = \frac{(\rho'_k - 1) + W_{k-1}\left(1 + \frac{\lambda_k}{\lambda_{k-1}}\right)}{2},$$
$$\rho'_o = 0. \quad (24)$$

3. Generalized Application Procedure

Figure 1 gives a generalized procedure to determine an optimum configuration for any given network.

(**A**) Through a traffic analysis, determine the set or sets of message priority percentages to be considered.

(**B**) Using the queueing model, develop a set of generalized curves for the set of percentages desired.

(**C**) Considering operation functions and other appropriate factors, determine the processing rates for each station as a function of the length of the chosen time intervals.

(**D**) Hypothesize a station configuration for the specific application.

(**E**) Through a traffic analysis, determine

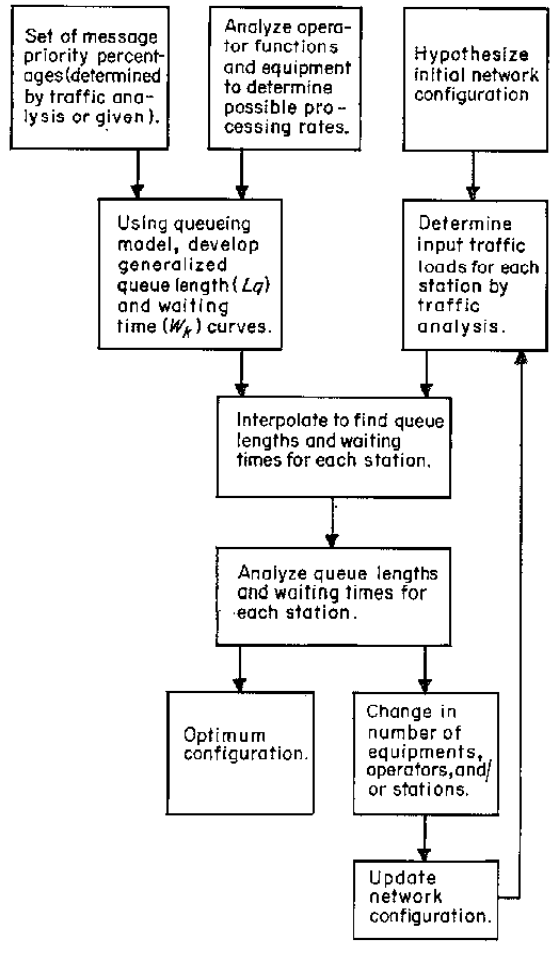


Figure 1—Procedure to determine optimum configuration within equipment and cost constraints.

representative message loads for each station in the proposed configuration as a function of time of day.

(F) Using the generalized set of curves, with the appropriate set of percentages for each station in the configuration, determine the expected queue lengths and waiting times.

(G) Analyze the expected queue lengths and waiting times for acceptability. If they are acceptable, determine whether the message loads are low using the present configuration. If so, modify the configuration to use less equipment; otherwise the chosen configuration is satisfactory. If the expected queue lengths and waiting times are not acceptable, modify the station configuration to use more equipment.

(H) If the station configuration is changed, return to (D) and proceed accordingly.

The described queueing-theory model may be used to test various network configurations in the iterative process of determining an optimum configuration, within time and equipment constraints, or it may be used as a model to determine expected queue lengths and expected waiting times of messages at any desired station(s).

Figures 2 and 3 are computer flow charts of the development of the generalized set of curves for queue lengths and waiting times.

4. Illustration of the Model

4.1 Application

The model was applied to the evaluation of a semiautomatic communication network designed for an Air Force base. Figure 4 shows a simplified configuration. The designed network consisted of a centrally located switching center at the base, several message centers, and the modified office typewriters (hereafter

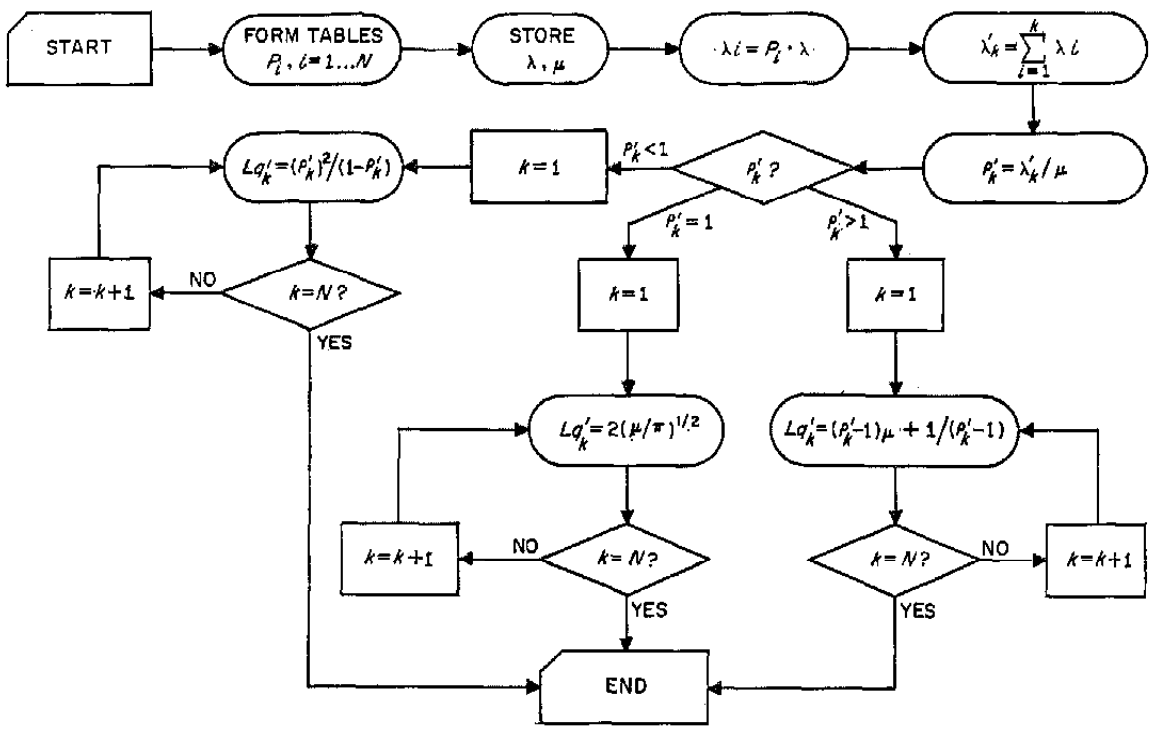


Figure 2—Program to develop generalized set of curves for queue lengths.

called typewriters) on which outgoing messages are prepared for automatic transmission to using offices.

All messages enter the base network at the switching center where they are automatically stored on tape. The information on the tape is then electrically transmitted to the proper message center. From here messages are hand-carried to the final destination. They are transported either by scheduled courier or are immediately hand-carried, depending on priority, location of the message center, and time of day.

Outgoing messages are originated at an office where, after proper coordination and authorization, each message is hand-carried to the typewriter servicing that office. This produces a paper-tape record that is to be transmitted and a hard copy of the message. The typewriter may be located in the office where the message originates, at the message center, or at some other location. The message, in paper-tape form, is then hand-carried to the message center associated with the originating office, where an operator electrically transmits the taped information to the switching center. The message is then automatically transmitted from the switching center to outside destinations over a communication network.

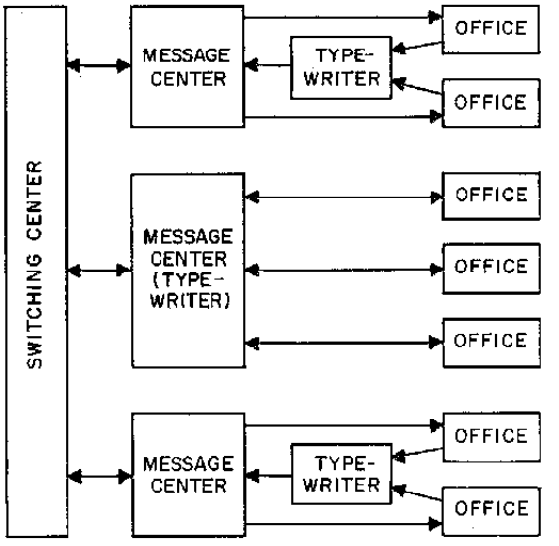


Figure 4—Simplified communication network.

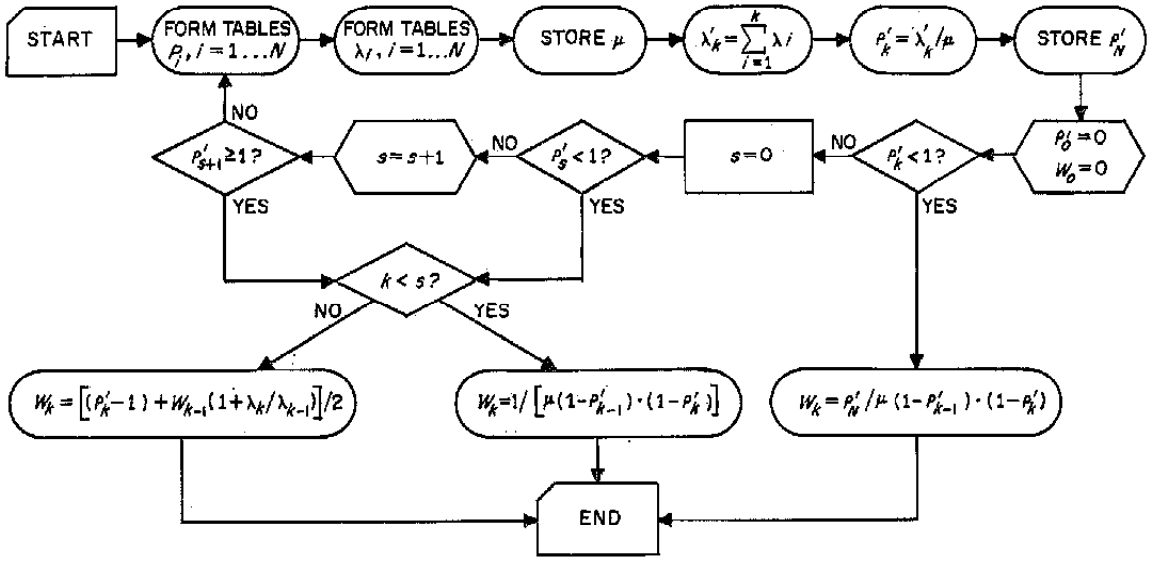


Figure 3—Program to develop generalized set of curves for waiting times.

4.2 Numerical Example

Studies of typing-speed requirements and other operator functions determined the processing rate for each typewriter to be 12 messages per hour. Where there are two typewriters, the processing rate is 24 messages per hour; that is, the interaction constant is equal to 1. At the message center, the processing rate was 30 incoming and 20 outgoing messages per hour. (The discrepancy in these processing rates is caused by different operator tasks in processing incoming and outgoing messages.) The switching center may process 80 incoming and 80 outgoing messages per hour.

Messages had one of four precedences: deferred, routine, priority, and operational immediate (in order of increasing precedence). A traffic survey was conducted at the base and the mean distribution of message traffic was found to be 40 percent deferred, 32 percent routine, 25 percent priority and 3 percent operational immediate. These percentages (variables in the described model and not necessarily constant over time) were obtained by averaging the traffic over all stations at all times of day. A generalized set of curves was developed for the expected queue lengths and waiting times at each type of station in the configuration (see Figures 2 and 3) based on the established percentages. Note that an exact set of percentages may be used for each station in the configuration, if desired, by developing different sets of generalized curves and using those appropriate to the particular station. Development of these curves for a set of priority percentages requires little computation and they may be obtained with a very limited amount of computer time. Generation of the curves shown in Figures 5 through 12 and similar curves for message centers and switching center requires less than 1 minute of *IBM* 7090 time.

Figures 5 through 8 and 9 through 12 are the generalized sets of curves showing expected queue lengths and waiting times, respectively, at a typewriter station for the percentages considered.

To illustrate the model, let us determine the expected queue lengths and waiting times for two typewriters serving a message center.

Figure 5—Typewriter queue length for operational-immediate traffic (P_1 = 3 percent of total).

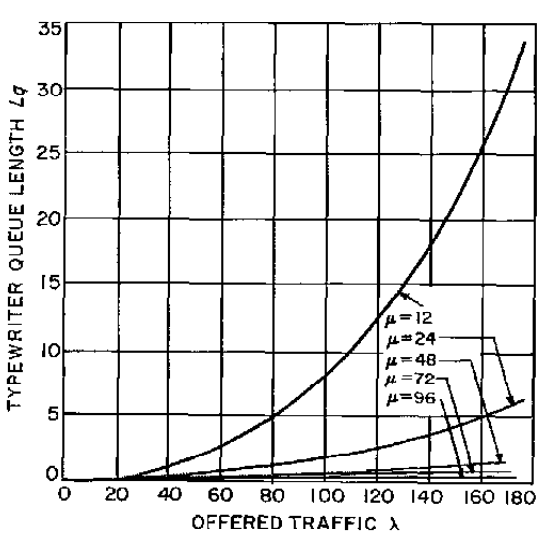


Figure 6—Typewriter queue length for priority traffic (P_2 = 25 percent of total).

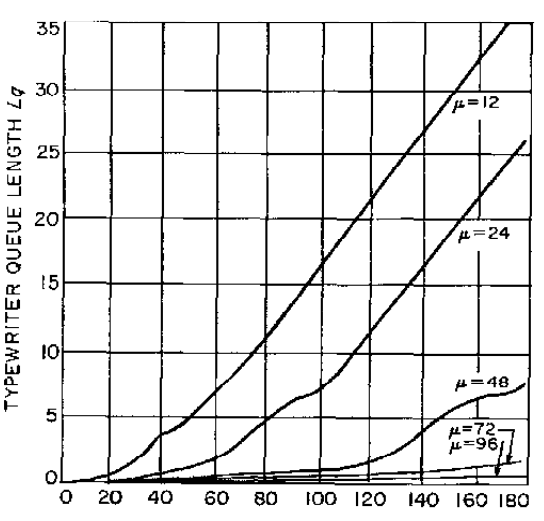


Table 1 lists their message loads as a function of time of day.

The processing rate used for a typewriter station is 12 messages per hour as determined by analysis of operation functions and other appropriate considerations. Using Figures 5 through 12, we interpolate to determine the readings in Tables 2 and 3.

The interpolation from Figures 5 through 12 is performed as follows:

TABLE 1

MESSAGE LOADS FOR TYPEWRITERS 1 AND 2

Time	Typewriter 1	Typewriter 2
0000–0100	0	0
⋮	⋮	⋮
0700–0800	0	0
0800–0900	4	1
0900–1000	4	1
1000–1100	8	2
1100–1200	8	3
1200–1300	2	1
1300–1400	15	10
1400–1500	19	7
1500–1600	11	1
1600–1700	7	1
1700–1800	0	0
⋮	⋮	⋮
2300–2400	0	0

Lq_4, for time 1400–1500 of Table 2, was determined by adding Lq_4 for time 1300–1400 to the message load input of time 1400–1500 and looking up this quantity in Figure 8. That is, finding $5 + 19 = 24$ along the horizontal axis of Figure 8 gives the queue length of 8 messages during this time interval. This queue length is added in turn to 11, the original message load during time 1500–1600, to determine the queue length of a fourth-precedence message during time 1500–1600, et cetera. Waiting times are determined the same way. For example, the mean waiting time for a third-precedence message during time 1500–1600 was found by adding the queue length for the third-precedence messages in time 1400–1500 to the original message load in time 1500–1600 and finding this quantity in Figure 11. That is, $4 + 11 = 15$ which from Figure 11 gives a mean waiting time for a third-precedence message of 0.46 hour.

If delays of 0.92 hour for a fourth-precedence message, 0.61 hour for a third-precedence message, 0.18 hour for a second-precedence message, and 0.09 hour for a first-precedence message are acceptable, then the typewriters are left unchanged; if not, then possibly some

Figure 7—Typewriter queue length for routine traffic (P_3 = 32 percent of total).

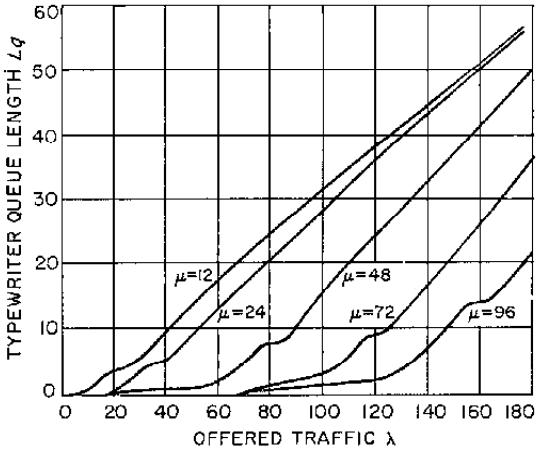
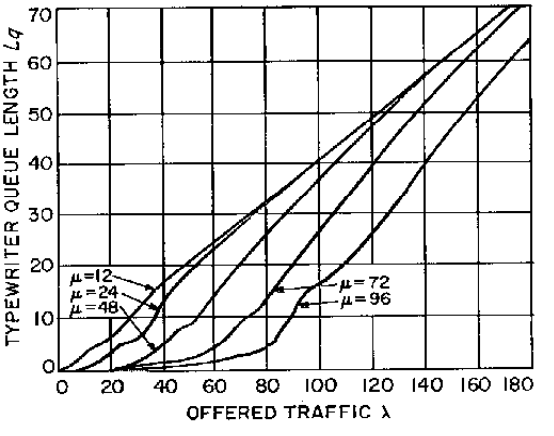


Figure 8—Typewriter queue length for deferred traffic (P_4 = 40 percent of total).



offices whose message traffic is included in the determination of the loads shown in Table 1 may be serviced by another typewriter.

This rearrangement would be made by detaching an office, on the basis of geographical location and its associated message load distribution, from typewriter 1 for service by another one associated with the same or with another message center. The message load distribution of the detached office is then subtracted from the load distribution of typewriter 1 in Table 1, taking into account the delivery time from the office to the typewriter. The message load distribution associated with

TABLE 2

EXPECTED QUEUE LENGTHS AND WAITING TIMES FOR TYPEWRITER 1

Time	Message Load	Lq_1†	Lq_2	Lq_3	Lq_4	W_1‡	W_2	W_3	W_4
0000–0100	0	0	0	0	0	0	0	0	0
⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
0700–0800	0	0	0	0	0	0	0	0	0
0800–0900	4	0	0	0	0	0.02	0.03	0.14	0.16
0900–1000	4	0	0	0	0	0.02	0.03	0.14	0.16
1000–1100	8	0	0	0	2	0.05	0.07	0.25	0.28
1100–1200	8	0	0	0	3	0.05	0.07	0.25	0.38
1200–1300	2	0	0	0	1	0.01	0.01	0.08	0.18
1300–1400	15	0	0	2	5	0.09	0.13	0.46	0.72
1400–1500	19	0	1	4	8	0.09	0.18	0.61	0.92
1500–1600	11	0	0	2	6	0.07	0.12	0.46	0.78
1600–1700	7	0	0	1	4	0.04	0.05	0.30	0.58
1700–1800	0	0	0	0	0	0	0	0.02	0.16
1800–1900	0	0	0	0	0	0	0	0	0
⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
2300–2400	0	0	0	0	0	0	0	0	0

† Lq_i is defined as the expected queue length for a message of ith precedence entering the station during the specified time interval.

‡ W_i is defined as the expected waiting time for a message of ith precedence entering the station during the specified time interval under consideration and is measured in hours.

Figure 9—Typewriter waiting time in hours for operational-immediate traffic (P_1 = 3 percent of total).

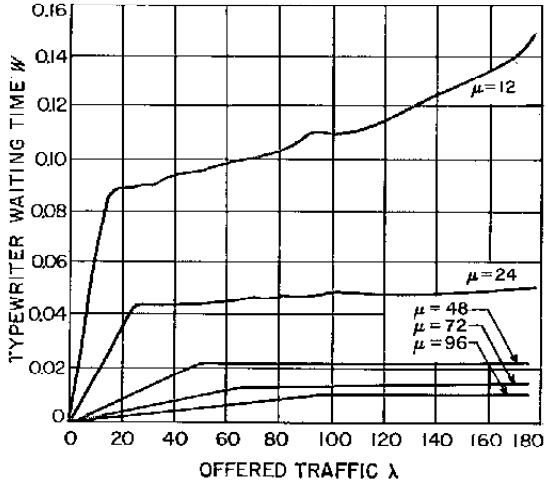
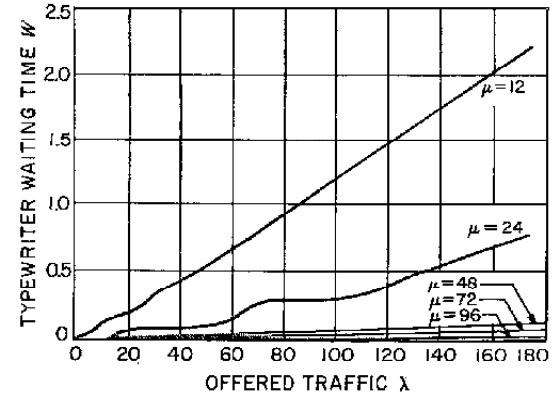


Figure 10—Typewriter waiting time in hours for priority traffic (P_2 = 25 percent of total).



the detached office is then added to the existing load distribution of the selected typewriter (again considering time displacement as shown in the following paragraph).

Assume that typewriters 1 and 2 are associated with the same message center and that a delay of more than 0.92 hour could be tolerated for a fourth-precedence message. Also assume that geographic considerations make it feasible to group the offices serviced by both typewriters and process the messages at typewriter 3. Assume further that typewriter 3 is to be located near the original ones to eliminate the requirement for adjusting the message traffic distributions. Therefore, the message load of typewriter 3 (refer to Table 4) for time 1300–1400, for example, is $15 + 10 = 25$.

Since we are using a single typewriter, the processing rate is still considered to be 12 messages per hour. Taking care that a Poisson input distribution still holds, the expected queue lengths and waiting times for this

TABLE 3

EXPECTED QUEUE LENGTHS AND WAITING TIMES FOR TYPEWRITER 2

Time	Message Load	L_{q1}	L_{q2}	L_{q3}	L_{q4}	W_1	W_2	W_3	W_4
0000–0100	0	0	0	0	0	0	0	0	0
⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
0700–0800	0	0	0	0	0	0	0	0	0
0800–0900	1	0	0	0	0	0	0	0.02	0.06
0900–1000	1	0	0	0	0	0	0	0.02	0.06
1000–1100	2	0	0	0	0	0.01	0.01	0.07	0.08
1100–1200	3	0	0	0	1	0.01	0.02	0.10	0.12
1200–1300	1	0	0	0	0	0	0	0.02	0.08
1300–1400	10	0	0	1	3	0.07	0.10	0.32	0.35
1400–1500	7	0	0	0	3	0.05	0.06	0.26	0.35
1500–1600	1	0	0	0	1	0	0	0.02	0.16
1600–1700	0	0	0	0	0	0	0	0	0
⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
2300–2400	0	0	0	0	0	0	0	0	0

Figure 11—Typewriter waiting time in hours for routine traffic (P_3 = 32 percent of total).

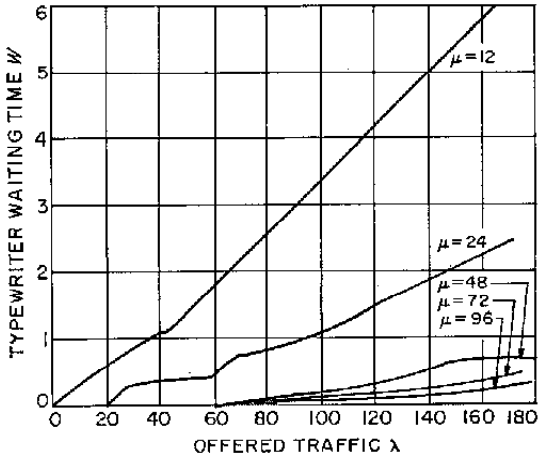
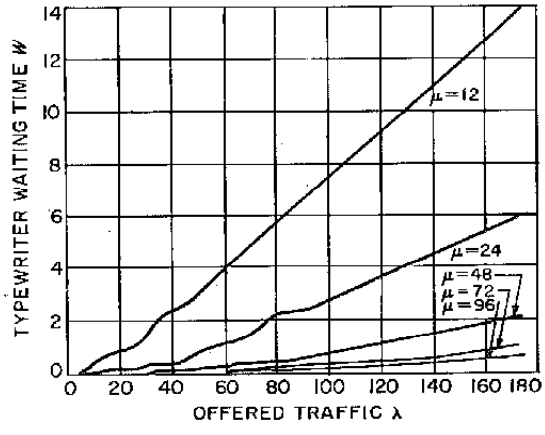


Figure 12—Typewriter waiting time in hours for deferred traffic (P_4 = 40 percent of total).



station are given in Table 5 (using Figures 5 through 12).

If the resulting mean waiting times of 2.00 hours for a fourth-precedence message, 0.85 hour for a third-precedence message, 0.26 hour for a second-precedence message, and 0.09 hour for a first-precedence message (during time 1400–1500) are acceptable, the configuration is left unchanged; if not, then the configuration should be modified to change the message load distribution of typewriter 3. This is done by having some of the offices, whose messages are processed by this typewriter, transfer their messages to another typewriter, if one exists, associated with the same message center or a typewriter associated with another message center.

However, if we must choose between the queue lengths and waiting times shown in Table 5 and those shown in Tables 2 and 3, then we must weigh the value of a maximum mean waiting time of 2.00 hours and the cost of a single typewriter against a maximum mean waiting time of 0.92 hour and the cost of two typewriters.

TABLE 4
MESSAGE LOAD FOR TYPEWRITER 3

Time	Typewriter 3	=	Typewriter 1	+	Typewriter 2
0000–0100	0	=	0	+	0
⋮	⋮		⋮		⋮
0700–0800	0	=	0	+	0
0800–0900	5	=	4	+	1
0900–1000	5	=	4	+	1
1000–1100	10	=	8	+	2
1100–1200	11	=	8	+	3
1200–1300	3	=	2	+	1
1300–1400	25	=	15	+	10
1400–1500	26	=	19	+	7
1500–1600	12	=	11	+	1
1600–1700	8	=	7	+	1
1700–1800	0	=	0	+	0
⋮	⋮		⋮		⋮
2300–2400	0	=	0	+	0

5. Acknowledgements

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6. References

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TABLE 5
EXPECTED QUEUE LENGTHS AND WAITING TIMES FOR TYPEWRITER 3

Time	Message Load	Lq_1	Lq_2	Lq_3	Lq_4	W_1	W_2	W_3	W_4
0000–0100	0	0	0	0	0	0	0	0	0
⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
0700–0800	0	0	0	0	0	0	0	0	0
0800–0900	5	0	0	0	1	0.03	0.04	0.17	0.20
0900–1000	5	0	0	0	1	0.03	0.04	0.17	0.22
1000–1100	10	0	0	1	4	0.07	0.10	0.30	0.37
1100–1200	11	0	0	1	4	0.07	0.12	0.37	0.68
1200–1300	3	0	0	0	1	0.02	0.03	0.14	0.16
1300–1400	25	0	1	4	9	0.09	0.24	0.72	1.04
1400–1500	26	0	1	5	14	0.09	0.26	0.85	2.00
1500–1600	12	0	0	3	9	0.07	0.14	0.52	1.04
1600–1700	8	0	0	1	5	0.05	0.08	0.36	0.75
1700–1800	0	0	0	0	1	0	0	0.03	0.20
1800–1900	0	0	0	0	0	0	0	0	0
⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
2300–2400	0	0	0	0	0	0	0	0	0

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Traffic Simulation and Its Application in Telephony

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1. Introduction

Theoretical computations of the traffic-handling capacity of telephone switching networks produce only approximate results, particularly for arrangements used in practical applications. The practical usefulness of the approximation equations must be verified by measurements performed with real or artificial telephone traffic.

Since measurements with real telephone traffic entail numerous difficulties, the idea of using so-called traffic machines [1–3] for the generation of artificial traffic was conceived several decades ago [4].

Lately, electronic computers frequently replace the traffic machine. Controlled by a program, the computer simulates the switching network to be tested as well as the artificial traffic flowing through it. This procedure does not require equipment investments, but merely writing a program. This program is known as a simulation program while the traffic trials are known as traffic tests or throwdowns.

Besides giving a general survey, this paper primarily describes the method of generating artificial telephone traffic and presents some examples. It also presents some known [5–7] and some new solutions to the problem based on accepted principles. Finally, some test results obtained with the aid of our *ER56* electronic computer are compared with theoretical results.

2. Simple Traffic Test

The principles underlying the generation of artificial telephone traffic, and consequently the investigation of switching networks, are explained with the aid of a simple example.

The number of calls rejected because no outlet is free is to be determined for a full-availability switching network with an unlimited number of inlets and three outlets (each inlet always having access to each free outlet), and with offered traffic $A = 2$ erlangs.

Although the beginnings and ends of the calls are random, probability theory establishes a certain order that permits mathematical definition of the telephone traffic. To simulate randomness, a random generator is required that will determine both the beginnings and ends of calls in the defined order.

For simple traffic tests, this can be achieved by repeatedly casting a six-sided die. We obtain a series of random numbers made up of 1, 2, 3, 4, 5, and 6, as for example 3 2 6 2 5 6 2 1 4 5 1 3 4 5 4 3 1 5 3 3 1 4 5 4 6 2 2 1 6 6. The random changing of the states of occupancy of the three outlets is now determined through the following use of these random numbers:

1, 2, and 3—releasing numbers—as soon as one of these numbers comes up, the corresponding outlet is released if occupied. If it is not occupied, the number has no effect.

4 and 5—call numbers—when one of these numbers appears, a call is originated and occupies a free outlet. If no outlet is free, a lost call is registered.

6—unused random number—this number has no function, and it causes no change in the state of occupancy.

This random distribution (refer to (6) with $R_0 = 2$) simulates offered traffic $A = 2$ erlangs. Figure 1 shows the changing states of occupancy caused by the above series of random numbers. The first few events are explained in Table 1. It is assumed that outlets

TABLE 1
CHANGING STATES OF OCCUPANCY

Number Cast	Effect
3	Call on outlet 3 ends.
2	Random number without effect because outlet 2 is free.
6	Unused random number.
2	Random number without effect because outlet 2 is free.
5	Originated call finds outlet 2 accessible.

1 and 3 are occupied at the beginning of the experiment.

Although the individual steps do not correspond to actual constant time intervals, it is evident from Figure 1 that the call duration varies considerably. One call may, for example, be longer than two other calls. In our experiment we offered $c_A = 10$ calls, and $c_v = 2$ of them, or 20 percent, were rejected.

To determine the loss

$$B = c_v/c_A \qquad (1)$$

through such a traffic test with a sufficient degree of reliability, the test duration should be several thousand times longer than in this simple example to ensure a balance of statistical variations.

3. Generation of Random Numbers

A very-long series of random numbers is required to generate enough artificial telephone traffic for a meaningful result. This series, however, must be made up from many more random numbers than 1 to 6 because there are switching networks that have more than 100 outlets, and higher offered traffic A that often have decimal positions. The numbers 1, 2, 3, . . . , do not suffice and must be supplemented by decimal numbers. The probability of any number in the series coming up next must be equal for all numbers and may not in any way depend on the preceding number (equal distribution and irregularity).

Some of the methods for generating a series of random numbers are particularly suitable for electronic computers. The mathematical instructions underlying these methods permit generating numbers x_i, starting with an arbitrary number x_0, according to the equation

$$x_i = f(x_{i-r}, \quad x_{i-r+1}, \cdots, x_{i-1})$$

or, for example, simply

$$x_i = g(x_{i-1}).$$

By omitting a certain number of decimal positions, the numbers z_i are derived from x_i. They belong to the quantity of numbers Z and form a series

$$Z_0, Z_1, Z_2, \cdots$$

If the mathematical equation for the generation of x_i and the modification method to z_i are suitably chosen, the series will have the desired properties although it is based on a formation law. Such a series of pseudo-random numbers has proved equal in many applications to a series of genuine random numbers. They have been used advantageously for traffic tests because they can easily and quickly be generated by electronic computers and are always reproduceable without having to be stored.

A description of the methods for generating pseudo-random numbers is beyond the scope of this paper. Some papers [8–10] containing detailed information are listed in Section 7, References.

4. Artificial Telephone Traffic

4.1 General

Switching networks in telephone systems are operated either on the busy-signal or the delay principle. Combining these two principles

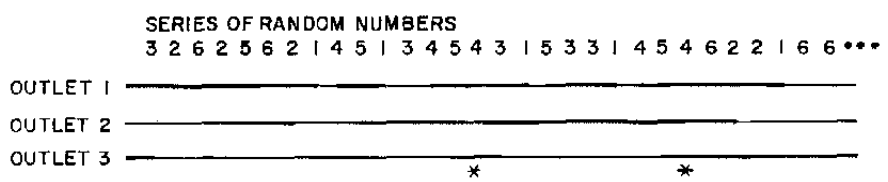


Figure 1—States of occupancy of 3 outlets.
* = lost calls.

is also possible and will be used whenever the number of waiting stores is insufficient or calls are terminated because of extended waiting times. In this paper, however, only the busy-signal principle is discussed, although simulation of a delay system or of a combination system is also possible.

Artificial telephone traffic must have the essential characteristics of real telephone traffic if it is to provide results reflecting practical operation. The tests simulate traffic that corresponds to the actual traffic during an arbitrarily determined number of main traffic hours (statistical equilibrium). This assumption underlies all conventional theoretical loss calculations.

When simulating systems that operate on the busy-signal principle, the duration of the different occupancy states of the switching network is insignificant. Merely the changes, that is, the probabilities of origination and termination of calls [7] are important. Actual timing is not simulated, not even in accelerated form.

The probability of a call originating is primarily dependent on the following:

(A) Number of sources or subscribers.

(B) Traffic offered and its distribution among the subscribers.

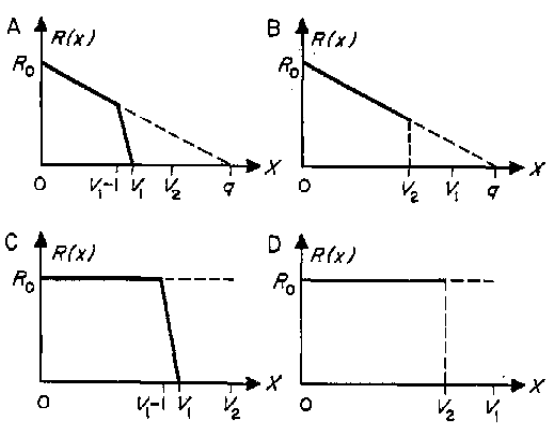


Figure 2—Possibilities of function R (x).

(C) Type and capacity of the networks between subscribers and the switching network being investigated.

(D) Treatment of rejected calls, which either are to be registered as lost or are routed via another group of overflow trunks to their destinations.

Assuming an equal distribution of the offered traffic among all subscribers q, the probability of a call originating is directly dependent on the number of simultaneously busy subscribers x, since a busy subscriber definitely does not originate a call. It thus must be proportional to the function

$$R(x) = R_0\left(1 - \frac{x}{q}\right). \qquad (2)$$

The factor R_0, which depends on the previously defined parameters, is explained as follows.

The number of sources q and the maximum possible number of simultaneous calls x influence the characteristics of the curve $R(x)$, which is defined for positive integral values of x only. On the one hand, the number of sources q may be finite or considered infinite; on the other hand, the variable x may be limited either through the number of inlets of switching network v_1, or of outlets v_2, or of links v_z. These four possibilities are indicated in Figure 2, the discrete points being connected to form curves to simplify understanding.

The probability of a call terminating depends solely on the mean call duration.

The treatment of rejected calls (lost or overflow calls) is important only when considered with a finite number of sources. It then becomes significant whether a source is immediately released after an unsuccessful attempt to call or whether the call-originating probability is reduced for the intended duration of the call. In the following derivation of the equations, the mean duration t_v of a rejected call is, for the time being, considered in a general way. It may later be selected as desired.

The equations required for the traffic tests are derived for condition A in Figure 2. Equations for the other three conditions are computed from this one. Condition A has little practical significance while the other conditions are of importance in practice. They are described in Sections 4.3 through 4.5.

4.2 Derivation of Condition A

The probability of a call terminating within a sufficiently short time Δt is

$$w_e = \Delta t / t_m \tag{3}$$

with t_m being the mean call duration.

With q as the equally loaded traffic sources, the probability of a call originating within Δt is

$$w_a(x) = (q - x)\frac{\Delta t}{p_m} \tag{4}$$

if x sources are busy. In this expression, p_m denotes the mean time interval between successive calls (time of inactivity of sources).

As a source can only be either active or inactive, the following condition is met.

$$c_y t_m + c_v t_v + c_A p_m = q \tag{5}$$

where

- c_A = mean number of calls offered per busy hour.
- c_y = mean number of calls handled per busy hour.
- c_v = mean number of rejected calls per busy hour.
- t_v = mean duration of rejected calls.
  - ≈ 0 if call is to be considered lost.
  - $= t_m$ if call finds an overflow path.

The equation for the mean time intervals between calls is obtained from (5).

$$p_m = \frac{1}{c_A}(q - c_y t_m - c_v t_v) \tag{6}$$

or, with the offered traffic $A = c_A t_m$ and the handling capacity $y = c_y t_m$, then

$$p_m = \frac{t_m}{A}(q - y - c_v t_v).$$

When examining condition A, it must be realized that the probability of a call originating is zero if all inlets v_1 of the switching network are busy. Assume the probability for this to be $w(v_1)$. The mean time interval between calls p_m in (4) is the time during which free sources may be expected to originate calls. The mean time interval as expressed in (6) contains passive times for condition A because an inactive source is not able to reach the switching network if all inlets v_1 are occupied. Only the mean active time interval $p_{m,a}$ however should be entered in (4).

We obtain

$$\begin{aligned} p_{m,a} &= p_m[1 - w(v_1)] \\ &= (q - y - c_v t_v)\frac{t_m}{A}[1 - w(v_1)] \end{aligned}$$

and therewith the probability of a call originating in condition A is

$$w_a(x) = \frac{q - x}{q - y - c_v t_v} \cdot \frac{A}{1 - w(v_1)} \cdot \frac{\Delta t}{t_m}, \tag{7}$$

for $x < v_1$

$w_a(v_1) = 0$.

In the simulation, the probabilities for the origination and termination of calls now have to comply with (3) and (7). For this purpose, a series of pseudo-random numbers is used that has the properties described in Section 3. The numbers are subdivided into releasing numbers $\mathfrak{L}$, calling numbers $\mathfrak{R}$, and not-used random numbers $\mathfrak{N}$, with the following agreement:

$$\begin{aligned} \mathfrak{L} &= [0, Z_1), \quad \text{with } 0 < Z_1 \\ \mathfrak{R} &= [Z_1, Z_2), \quad \text{with } Z_1 < Z_2 \\ \mathfrak{N} &= [Z_2, N), \quad \text{with } Z_2 < N. \end{aligned}$$

The mode of representation $[a, b)$ denotes that this quantity includes all numbers z_i with $a \leq z_i < b$.

The limits Z_1 and Z_2 must be so chosen that the probability of a releasing or calling number

appearing next in the series corresponds to (3) and (7). One of these two limits may be set arbitrarily.

To simplify writing the simulation program for the computer, it is advisable to set

$$Z_1 = v_2 \tag{8}$$

with v_2 denoting the number of switching network outlets. In this case, any individual outlet j that is occupied is always released if a random number z_i, with

$$j \leq z_i < j + 1$$

appears in the series. If the numbering of the outlets starts with 0, for example, the random number z_i will directly indicate by the number preceding its decimal point the outlet to be released. During the simulation process, the releasing number probability w_1 corresponds to the call-terminating probability w_e and thus is proportional to 1.

$$w_1 \sim 1. \tag{9}$$

The probability $w_a(x)$ of a call originating is linearly dependent on the state of occupancy x, according to (7). The upper limit of the calling number quantity, therefore, must also vary with x. Let us assume it is $Z_2'(x) < Z_2$. The calling number probability $w_r(x)$ during the simulation corresponds to the call-originating probability $w_a(x)$ and is proportional to $R(x)$ in (2). Thus

$$w_r(x) \sim R(x) = R_0\left(1 - \frac{x}{q}\right) = Z_2'(x) - Z_1. \tag{10}$$

To align the proportionality factors of (9) and (10), the quotient of calling-number and releasing-number probability is set equal to the quotient of call-originating and call-terminating probability

$$\frac{w_r(x)}{w_1} = \frac{w_a(x)}{w_e}.$$

From (3), (7), (9), and (10) we thus obtain

$$R(x) = R_0\left(1 - \frac{x}{q}\right) = \frac{q - x}{q - y - c_v t_v} \cdot \frac{A}{1 - w(v_1)}, \quad \text{for } x < v_1 \tag{11}$$

$$R(v_1) = 0$$

and for the unknown $R_0 = R(0) = Z_2 - Z_1$

$$R_0 = \frac{Aq}{[q - (y + c_v t_v)][1 - w(v_1)]}. \tag{12}$$

The limits for the quantities of random numbers are thus

$$\begin{aligned} Z_1 &= v_2 \\ Z_2 &= v_2 + R_0. \end{aligned} \tag{13}$$

The upper limit of the quantity of calling numbers, which depends on the state of occupancy, is, according to (10)

$$Z_2'(x) = v_2 + R_0\left(1 - \frac{x}{q}\right). \tag{14}$$

Practical applications of (12) are discussed in Sections 4.3 through 4.5.

With the offered traffic distributed equally to all sources, each successfully established call must reduce the upper limit by R_0/q and each release must increase it by the same amount. If no source is active, $Z_2'(0) = Z_2$.

4.3 Erlang Traffic

If, for the number of outlets, a very-large number of traffic sources have unrestricted access to the switching network to be investigated, the transition $q \to \infty$ can be effected in (12), (Erlang traffic)

$$R_0 = \lim_{q \to \infty} \frac{Aq}{[q - (y + c_v t_v)][1 - w(v_1)]}.$$

We obtain

$$R_0 = \frac{A}{1 - w(v_1)}, \quad \text{for } v_1 \leq \min(v_2, v_{Z_1}, v_{Z_2}, \cdots) \tag{15}$$

and

$$R_0 = A, \quad \text{for } v_1 > \min(v_2, v_{Z_1}, v_{Z_2}, \cdots) \tag{16}$$

because $w(v_1) = 0$ in (16). If several groups of outlets are connected to the switching network, v_2 denotes the total sum of all outlets, and $v_{Z1}, v_{Z2}, \cdots$ denote the number of links between the stages if a link arrangement is tested.

The probability $w(v_1)$ can be determined mathematically or, if the simulation program is arranged correspondingly, through the test itself. It is merely necessary to count the calling numbers that appear in the condition $x = v_1$. The simulated offered traffic A can be computed with (15) only on completion of the test.

When assuming $q \rightarrow \infty$, the inlets of the tested switching network are usually not connected to traffic sources. Instead, traffic has already been concentrated by means of other switching networks. The assumption of unlimited access of traffic sources to the inlets thus is seldom realized in practice because it would require a full-availability switching network as a connecting link. However, it is mandatory to assume such a full-availability switching link. Otherwise an entire switching office has to be simulated simultaneously. This assumption is also justified if the losses or the waiting times, respectively, are small in the preceding networks.

4.4 Engset Traffic with Overflow Provisions

When generating artificial telephone traffic, special attention must be paid to the treatment of rejected calls [11] if a finite number of traffic sources are to be simulated.

If a call is rejected by the tested group of trunks but is able to reach the destination via an overflow trunk, the mean duration of the rejected calls is $t_v = t_m$. The test result presents an overflow rate instead of a loss. To be able to set $t_v = t_m$, the overflow capacity must be sufficient to ensure that no loss occurs. In practice, this cannot be realized. Compared to the total traffic offered for the direction tested, however, losses are usually very small and the above assumption causes only negligible error.

If $y + c_v \cdot t_m = c_A \cdot t_m = A$, we derive from (12)

$$R_0 = \frac{Aq}{(q - A)[1 - w(v_1)]}, \qquad (17)$$

$$\text{for } v_1 \leq \min(v_2, v_{Z1}, v_{Z2}, \cdots)$$

and

$$R_0 = \frac{Aq}{q - A}, \qquad (18)$$

$$\text{for } v_1 > \min(v_2, v_{Z1}, v_{Z2}, \cdots).$$

4.5 Engset Traffic Without Overflow

If the calls rejected by the tested trunks are lost, the holding time t_v of the switching network inlets is normally very short. If t_v approaches zero [11] and if the unknown traffic-handling capacity is set at $y = A\ (1 - B)$, we derive from (12)

$$R_0 = \frac{Aq}{[q - A(1 - B)][1 - w(v_1)]}, \qquad (19)$$

$$\text{for } v_1 \leq \min(v_2, v_{Z1}, v_{Z2}, \cdots)$$

and

$$R_0 = \frac{Aq}{q - A(1 - B)}, \qquad (20)$$

$$\text{for } v_1 > \min(v_2, v_{Z1}, v_{Z2}, \cdots).$$

The loss B is not known before the test has been carried out. Thus, an R_0 has to be computed first for the estimated loss and, on completion of the test, the actually simulated offer A, using (19) and (20).

Equation (20) is important for testing subscriber switching networks because the finite number of subscribers must be taken into consideration and because definitely $q = v_1 > v_2$.

4.6 Performance of Traffic Tests

Since a simulation program should be written for as many applications as possible, the freely selectable parameters must first be determined for a particular test. The following information

must be supplied to the computer in addition to the program:

(A) Size of the switching network to be tested.

(B) Type of traffic to be simulated.

(C) Traffic offered.

(D) Traffic distribution to the inlets of the network.

The duration of the test depends, for example, on the number of calls offered. Since a statistical equilibrium has to be simulated, it is also necessary to achieve a mean state of occupancy before beginning to count calls handled and calls rejected.

With the aid of one of the equations (15) through (20), it is possible to compute, at least approximately, the factor R_0 for a predetermined offered traffic A. This simultaneously establishes the limits for the quantities of releasing and calling numbers (13). If the inlets of the switching network are divided into a number of subgroups, the quantity of calling numbers must be divided similarly, to be able to determine to which subgroup an originating call will be directed. If the traffic offered is distributed unequally to the subgroups, R_{0i} and the corresponding values q_i, A_i, $w(v_{1,i})$, and B_i must be computed using (15) through (20) for every subgroup. The calling number subquantities then are

$$[v_2, v_2+R_{01}), \quad [v_2+R_{01}, \quad v_2+R_{01}+R_{02}), \cdots,$$

$$[v_2+\sum_{i=1}^{m-1} R_{0i}, \quad v_2+\sum_{i-1}^{m} R_{0i}). \quad (21)$$

The modulus of limit changes is also different for each subgroup, that is, R_{0i}/q_i. The loss B_i must be determined separately for each subgroup so as to be able to compute, using R_{0i}, the actual simulated offered traffic A_i.

4.7 Reliability of Test Results

The simulation of busy-signal systems as described here presents, as the principal result of a test, the relative frequency of loss, expressed by the quotient of rejected calls c_v and offered calls c_A

$$B = c_v/c_A. \quad (1)$$

The limit of relative frequency for $c_A \to \infty$ would yield the true loss $\bar{B}$. The test, however, must be terminated after a finite time. The result achieved is only an approximation whose accuracy increases with the number of rejected calls c_v.

In sampling theory, it is customary to give so-called confidence intervals for measured results $(B - \Delta B, \ B + \Delta B)$ within which, with a certain statistical probability, the true value of the measured term B lies. This procedure can be used to indicate the degree of reliability of the test results.

The number of calls rejected during a test is liable to deviate randomly from the true mean value. To determine the loss deviation ΔB, the variance of c_v must be known. It can be determined theoretically for full-availability switching networks [12]. For the normally encountered case of switching networks with limited availability, the empirical variance is computed by registering during a test n intermediate results $c_{v,i}$ $(i = 1, 2, \cdots, n)$ after an equal number of calls offered c_A/n, and deriving n losses B_i therefrom. With these intermediate results, it is possible to calculate the empirical variance

$$\sigma^2 = \frac{1}{n-1} \sum_{i=1}^{n} (B_i - B)^2 \quad (22)$$

with

$$B = \frac{1}{n} \sum_{i=1}^{n} B_i = \frac{c_v}{c_A}. \quad (23)$$

With the aid of Student's t-distribution, it is possible to determine the deviation ΔB under certain conditions. The true loss $\bar{B}$ is then expressed with the statistical probability S by

$$B - \frac{1}{n^{1/2}} \sigma t_{S,n} \leq \bar{B} \leq B + \frac{1}{n^{1/2}} \sigma t_{S,n}. \quad (24)$$

The factor $t_{S,n}$ derived from the t-distribution depends on the required statistical probability

S and the number of intermediate results n. The limits of the confidence interval approach almost linearly to the mean value B, with quadratic extension of the test time.

The confidence intervals may be determined for the total loss as well as for subgroup loss.

5. Computer as a Traffic Machine

The individual elements of a switching network must be simulated in the computer to the extent that they influence the traffic flowing through the network. The outlets, and perhaps also the links, of the network are represented by storage cells into which digital free or busy signals can be entered. The duties of registers and markers that hunt for free links or outlets and occupy them, if available, are performed by a subprogram. We have to distinguish between two cases.

(A) If the establishment of connections between switching stages (Figure 3) and the order in which the outlets of a multiple are occupied follow rules that apply to a large group of switching networks, these rules will be programmed.

(B) If connections are apparently arbitrary, however, or if the outlets of several switching multiples are combined by grading that may vary widely, the connecting or grading pattern—in addition to the normal program—will be entered in suitable form into the computer. This is accomplished by writing the addresses of the links or outlets accessible from the following switching stage into the storage cells representing the links. In both cases, the subprogram must determine the route through the network that would be used for the call, at the same state of occupancy, in technically realized arrangements. If a call is to be released on a specific outlet, the busy criteria must be cancelled in the corresponding storage cell and in the cells associated with the links used. Besides the busying criteria, the subprogram must therefore write into the storage cells the links used to reach the respective outlets.

During the test, the total number of calls offered c_A and of calls rejected c_v must also be counted for each subgroup, if required. At the same time, loss B and loss deviation ΔB are determined. If Engset traffic has been simulated, the actual offered traffic A is computed using one of the equations (17) through (20).

Consider the relatively simple example in Figure 3 of a 2-stage link arrangement without directional selection (relay-type register-finder). The simulation program should be able to generate optionally either Erlang or Engset traffic, and the size of the network should also be selectable within certain limits.

The program establishes the number of subgroups, of links per subgroup, and of outlets in each switching multiple of the second stage. The type of link arrangement shown in Figure 3 is thus fixed. The connecting pattern is so easy and regular that it is worthwhile to program it. The outlets are tested one after the other from a home position for their states of occupancy.

Figure 4 is the block diagram for this simulation process and gives an outline of the program structure. A detailed description would

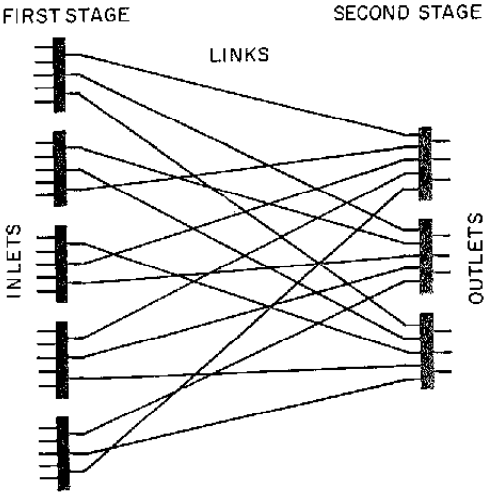


Figure 3—Two-stage link arrangement using 25 inlets, 15 links, and 9 outlets.

not be pertinent as it must be given for a specific computer model.

When designing a simulation program, there are additional considerations. It is desirable, for example, to work out programs for as large a group of switching networks as possible and to provide for all types of traffic; on the other hand, the computer time for a test should be minimized. A compromise must be found for these two contradictory demands.

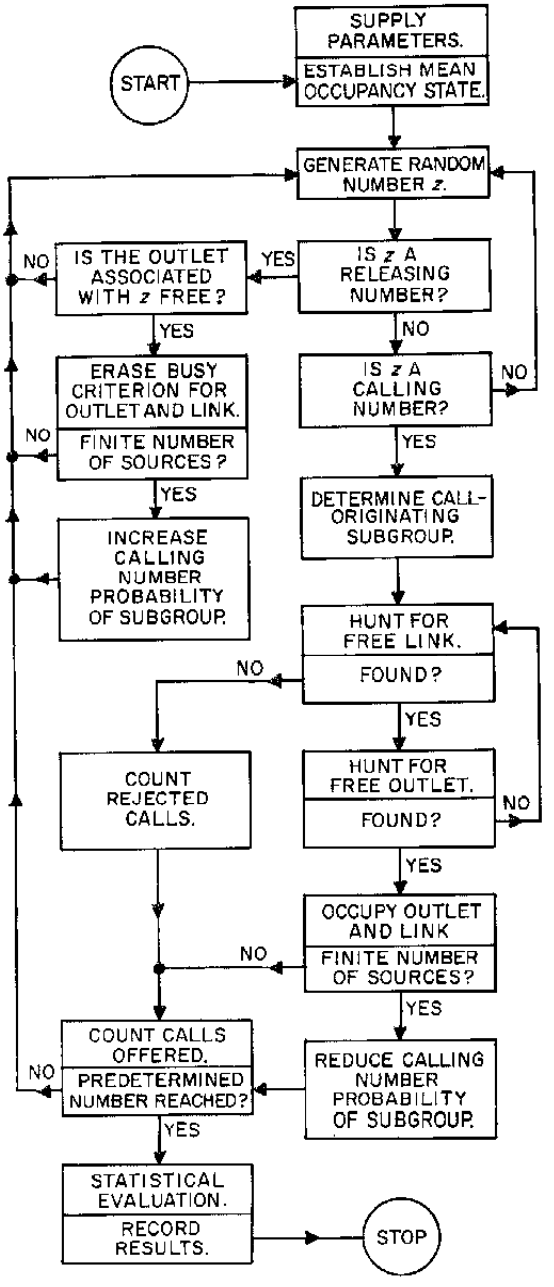


Figure 4—Block diagram of simulation program for 2-stage link arrangements.

6. Test Results

Test results were obtained with the program described in Section 5, using the *ER56* electronic computer. These results are only samples and were influenced by the method of generating the pseudo-random numbers and by the state of occupancy of the switching network at the beginning of the test. The program and the computer, however, did not influence the results. The switching network to be tested was chosen to permit a comparison of the test results with theoretical results verified in practical operation. The program for 2-stage link arrangements has been used here to test a 1-stage full-availability switching network having $v_2 = 27$ outlets and Erlang traffic (16). The first stage of the link arrangement in this case is 1 switching multiple with 27 outlets, each connected to only 1 outlet of a switching multiple of the second stage. This link arrangement theoretically corresponds to a 1-stage full-availability switching network.

The loss probability may be theoretically determined with the aid of the Erlang equation $B = E_{1,v2}(A)$ or may be obtained from [13].

In Table 2, test results and losses according to

TABLE 2

LOSS OF A FULL-AVAILABILITY SWITCHING NETWORK WITH $v_2 = 27$ OUTLETS (THEORETICAL AND TEST VALUES)

Offered Traffic A in Erlangs	Theoretical Loss $E_{1,27}(A)$ in Percent	Traffic Test	
		Loss $B \pm \Delta B$ in Percent	Number of Offered Calls ϵ_A
22.0	5.137	5.515 ± 0.742	20 000
19.5	2.205	2.140 ± 0.297	30 000
18.5	1.421	1.403 ± 0.290	30 000
17.5	0.852	0.862 ± 0.148	40 000
15.0	0.160	0.168 ± 0.067	50 000

[13] are compared for 5 different offers. Despite the relatively low number of calls offered, the results correspond well. The theoretical loss is never beyond the confidence interval and is usually very close to the mean value B.

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Günter Dietrich was born at Liebichau, Kreis Pr. Stargard, Poland, on 25 July 1932. In 1955 he received a diploma in mathematics from the University of Rostock.

In 1955, he joined Wissenschaftlich-Technisches Büro für Gerätebau in Berlin, where he was engaged in work on the development of special electrical measuring equipment. In 1958, he joined Standard Elektrik Lorenz, and is now chief of the department for telephone traffic theory.

Heinz Wagner was born in Künzelsau, Germany, on 26 January 1927. After serving as a public-school teacher, he studied mathematics at the Technische Hochschule in Stuttgart from which he received a diploma in 1958.

In 1957, he joined Standard Elektrik Lorenz where he is now engaged in studies of telephone traffic theory.

Automatic Switching for Telex–Operation

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The 7E rotary switching equipment [1] was selected for inclusion in the New York and San Francisco exchange centers of American Cable and Radio Corporation to provide both expanded facilities and increased automation. Fully automatic operation had already been established for direct subscriber-to-subscriber service between the United States, United Kingdom, and France.

It was necessary in the over-all design and adaptation of the 7E equipment to maintain compatibility with the other United States telex * carriers and with the administrations of some 70 directly or transit-connected overseas countries. Numerous problems pertaining to mandatory or desirable operating features were resolved during the course of manufacturing the equipment and prior to the official cut-over on 1 September 1962. The more-important features of the installations are briefly described.

1. Prefix Numbering Plan

To provide for the recording of both fully and semiautomatic calls in the 7E registers, it was necessary to develop an international prefix numbering plan having an indicator for each country that provides telex service. The basic plan [2] was developed in cooperation with the other United States carriers.

A 3-digit prefix precedes the subscriber's assigned telex number. The 1st digit indicates the continent and the 2nd and 3rd digits are general location and country or city indicators, respectively. In countries served by more than one carrier, the 2nd digit also identifies the carrier. For economy and other reasons, the 1st digit, the continent indicator, is presently omitted for traffic exclusively within the continental United States.

* Telex indicates private subscriber-to-subscriber teleprinter operation over the switching and transmitting systems of the public telephone and telegraph networks.

2. Overseas Trunks and Signalling

Trunks to overseas networks are provided through frequency-modulated voice-frequency telephone cable or multiplex radio channels with automatic error correction utilizing a constant-ratio code between terminals of the various continents.

Incoming trunk calls require keyboard selection and *A*-type [3] signalling. The various signal modes of outbound trunk calls are designed to conform with the requirements of overseas administrations with keyboard or dial selection and *A* or *B* type of signalling, respectively.

Incoming trunk calls to subscribers are routed automatically through the use of the subscriber's listed number and the assigned 3-digit prefix.

A transit call to a subscriber of an overseas administration that is provided with automatic switching is routed through the *7E* facilities without intervention of the telex operator. Alternatively, calls directed to networks without such facilities are automatically routed for completion to the operator of the gateway city of the terminating network.

3. Adaptation to United States Domestic Telegraph and Telephone Networks

Special adapter equipment is provided for connecting trunks of the dial-operated domestic telephone system, as these are normally channelled through the conventional telephone switching systems and terminate in special applique apparatus installed in the New York and San Francisco exchanges.

The operating characteristics of the telephone network as later described do not correspond with the specifications of the Comité Consultatif International Télégraphique et Téléphonique for telex service as employed by overseas administrations. Translators for interpreting variations in a number of keyboard characters, special convertors to accept speed character-

istics of alternative 45- and 50-baud systems, and many other devices of associated equipment had to be provided.

Of particular importance, special arrangements are provided for signalling capable of converting the conventional telephone tone signals to suitable control signals compatible with the international specifications. This permits overseas operators, without intervention of local operators, to complete calls directly into the telephone networks. An intercept board is provided for incompleted calls due to reception of telephone tone signals (busy, no answer, or no such number) that cannot be interpreted, the intercept operator informing the caller accordingly. Provision is also made so that this operator can re-establish the call if desired. If the overseas caller does not wish to select the telephone subscriber's number, the call is automatically routed to the intercept operator for completion.

Connections to the domestic telegraph system are provided for dial pulse and *B*-type signalling.

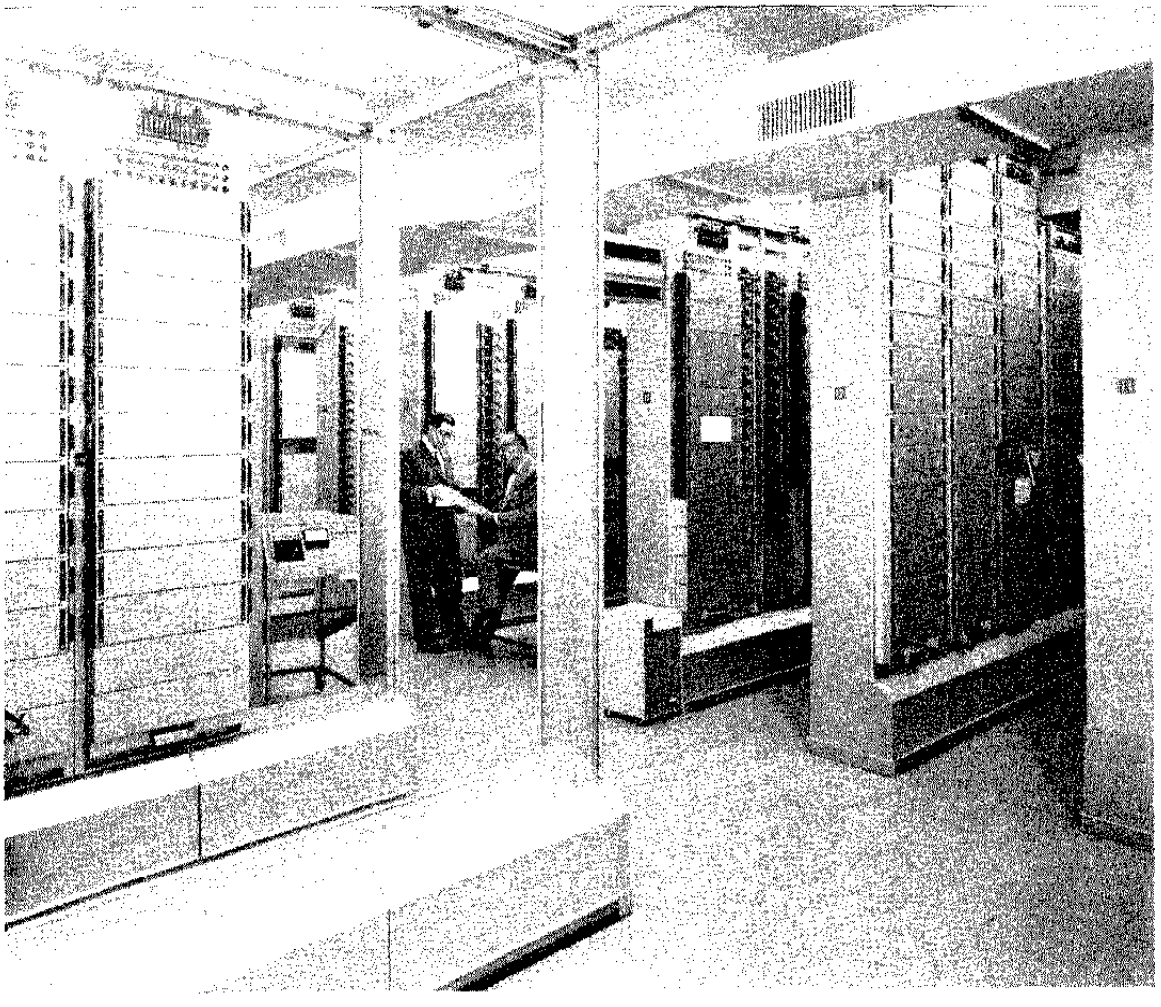


Figure 1—A view of the switching room at New York with the author (right) and Mr. T. Buyssens, project engineer for Bell Telephone Manufacturing Company (Antwerp).

On seizure of an incoming trunk from the telephone or telegraph system, no further selection is required by the calling party. Such calls are routed directly to telex operators, without register operation, for completion.

4. Trunk-Status Board

A trunk-status panel board, visible from all telex operator and overflow positions, is provided with nameplates for all connecting networks and their assigned trunk numbers. Two lamps, connected to each trunk as signals for out-of-order or busy conditions, serve as a ready indicator of trunk conditions.

5. Telex Operator Positions

Each telex operator position is equipped with 2 printers and 6 connecting circuits, each capable of handling 2 calls simultaneously and of establishing up to 6 calls. Incoming calls are automatically distributed and connected to idle positions. During intermittent peak traffic conditions when all operator positions are busy, further calls are automatically routed to overflow positions. The telex operators extend calls through the switching system by keyboard selection. These cordless positions are equipped with control keys and the indicating lamps necessary for operation and supervision.

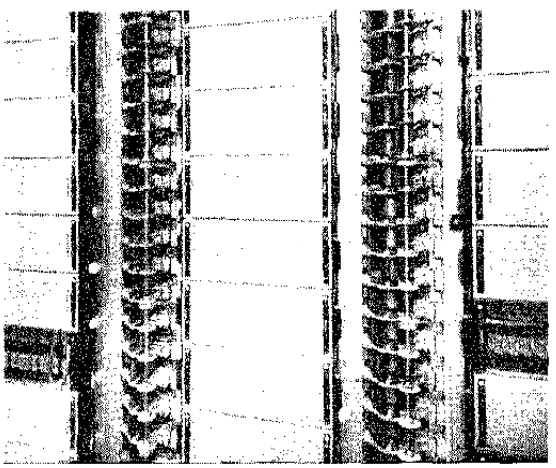


Figure 2—A typical rack of machine-driven selectors with their associated banks of relays.

Each position is fitted with electrically operated elapsed-time indicators. At termination of a call and release of the called party, provisions are made for the operator to transmit the charges to the calling party without setting up for a new call. In the arrangements for telephone and telegraph subscribers, however, where the equipment facilities prevent holding of the calling party, a device is provided that accepts a time-and-charge request from those subscribers in the form of an assigned telegraph code signal (5 times upper-case *M*) to alert the operator and for the release of the called party.

On a fully automatic call, the charges are automatically sent to the calling party at the termination of the connection without intervention of an operator.

6. Overflow Facilities

The overflow positions are similar to the regular operator positions. They are manned by supervisory personnel, are additionally provided with printers for each connecting circuit, and have arrangements to set up conference calls, to provide information to subscribers, and to provide other special services in addition to the handling of overflow calls, which are automatically diverted from fully occupied regular operating positions.

7. Message Traffic Facilities

All subscribers have facilities to use the telex printers for sending and receiving conventional overseas telegraph messages, which are billed at the established per-word rates. Message receiving and message sending positions are provided for this service.

7.1 Receiving Positions

The message receiving positions, each equipped with one printer to an associated trunk and normally unattended, can be reached by any

subscriber selecting the assigned code number. The calling party, receiving an automatically sent answer-back signal from an idle position, then proceeds to send his message. The calling party's number is also recorded for billing purposes. These positions are divided into 2 groups to facilitate routing of traffic in conformance with the requirements of alternative class of service, as specified.

7.2 Sending Positions

The message sending positions are provided with individual connecting circuits and, by means of keyboard selection, the operator may complete calls to any subscriber in the continental United States.

8. Subscribers' Equipment

Telex printer stations in subscribers' offices are equipped with control boxes containing a push button for initiating and releasing calls. Since keyboard selection is provided, service indicators, such as proceed to select, trunk line busy, called subscriber busy, trunk out of order, et cetera, are handled by typed messages. Some subscribers have provisions for off-line operation necessary for tape preparation. Such operation is interrupted in favor of incoming calls.

The proceed-to-select signal, containing an indicator for identification purposes, is terminated by upper-case *D*, which trips the subscriber's answer-back device. The *A/B* code contains the subscriber's number, which is extracted and stored in the *7E* register for billing purposes. The subscriber selects the called party's number, including the prefix. The register extends the call directly to the called subscriber of a fully automatic system or the call will be routed for completion to the telex operator of the gateway city.

9. Time and Date

The time-and-date information is transmitted to the calling and called subscribers on completion of a call. The data contains the progressive number of the day of the year; the carrier indication, which is essential for calls to telephone and telegraph subscribers to identify the carrier that routed the call; and Greenwich Mean Time.

10. Reperforator Equipment

Reperforators are provided to receive data in the 5-unit telegraph code from the memory circuits associated with the fully automatic connecting circuits. At the termination of each such call, which originates from any subscriber handled by the New York, San Francisco, or Washington offices, the memory circuit presents the following data to an idle reperforator: calling party's number, called party's number, time, date of call, and elapsed billable time.

11. Automatic Billing

Data from the punched tapes from the above reperforators is used for accounting and billing.

12. Monitoring Facilities

Banks of printers, each equipped with an automatic paper take-up spool arrangement, are connected to active trunk circuits to provide monitoring facilities for supervisory personnel. This provides for periodic visual inspection of operating conditions with hard copies of traffic for ready and future reference in case of errors.

13. Test Facilities

Automatic transmitters are provided for sending the "quick brown fox" sentence in biased or unbiased form. This test signal is automatically connected to any calling trunk or subscriber's line on reception of a special service code number assigned to this facility.

Mobile test sets equipped with printer, transmitter-distributor, and distortion-analyzer equipment are available in the automatic

switchrooms for plug-in connection to any trunk, subscriber's line, or operator's position for analyzing faults that may occur.

An automatic routine test arrangement, incorporated in the *7E* switchroom apparatus, provides for fully automatic circuit checking with an associated paper-tape device for recording faults and their location. This equipment will be operated periodically under maintenance schedule.

14. Conclusions

This system is intended to provide ultimately connections between subscribers in the United States and all telex subscribers overseas without the intervention of operators. This service may be extended to the domestic telephone- and telegraph-network telex subscribers.

Arrangements have been completed to extend these automatic switching facilities through the New York exchange to subscribers in Washington, District of Columbia.

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Henry L. Horwitz was born on 30 March 1901 in Berlin, Germany. He received a telecommunication engineering degree from the Institute of Technology in Berlin in 1923.

He worked on automatic telephone switching systems in Europe including 15 years in Italy. In 1944, he joined Federal Telephone and Radio Corporation. In 1959, he was transferred to ITT Communication Systems and came to American Cable and Radio Corporation the following year as a specialist on automatic switching.

Mr. Horwitz holds 23 patents. He is a Member of the Institute of Electrical and Electronics Engineers.

Automatic Switching for Telex–Equipment

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As mentioned in the first part of this paper, the New York and San Francisco exchange centers of the American Cable and Radio Corporation use the *7E* rotary teleprinter switching system.

In calling, the need for a telephone-type dial is avoided by using the teleprinter itself. The numbers sent from its keyboard are stored in a register that supervises the subsequent switching. The search is under electronic control at the register, which performs several functions that in other designs are handled by line finders or selectors, with simplification of that corresponding apparatus.

The position of the single-motion selectors is under direct control of the register. A 450-hertz source produces 12 phases each 30 degrees apart at two voltages, 12 and 20 volts, providing 24 distinguishable conditions. Each such condition identifies a different group of outlets in the group-selector arcs and in the final selectors identifies the first digit of a decimal group and the individual outlets.

Single-motion selectors are mechanically simpler and are easier to maintain than two-motion switches. They provide greater flexibility in assignment of lines as their outlets are not divided rigidly into, say, 10 levels of 10 outlets each, but can be electrically divided into any number of groups with differing numbers of outlets in the groups.

The routing diagrams for the three types of calls, incoming, outgoing, and transit, are shown in Figures 1 through 3.

The power-driven single-motion rotary switch shown in Figure 4 is employed both as line finder and selector. It has 100 sets of outlets and 7 double brushes. Neither the selector nor the finder is provided with a home position, but remains on the terminal to which it was last positioned. This reduces hunting time as the switch positioned nearest to the new-call outlet will accept the assignment with minimum motion, thus reducing wear on all the hunting switches. A universal 2-relay circuit is associated with each switch regardless of whether it is used as a group or as a final selector.

The wire-spring relay shown in Figure 5 is similar to that developed by Bell Telephone Laboratories in the United States. Its efficient

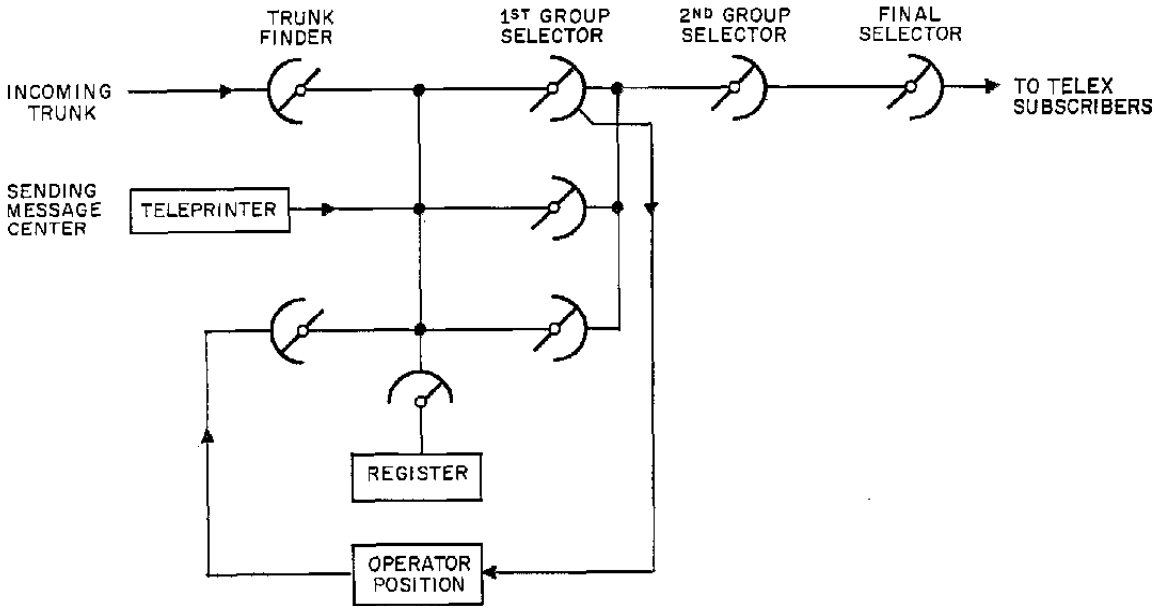
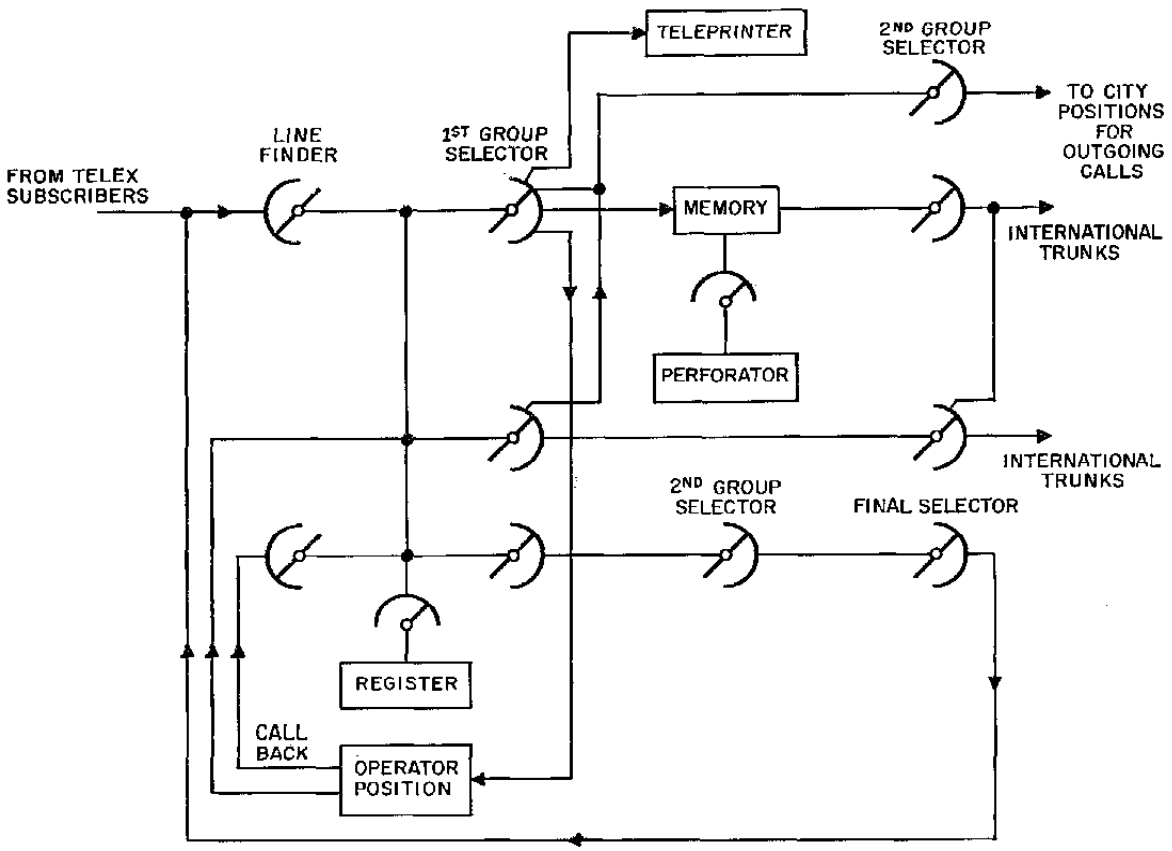


Figure 1—Routing diagram for incoming calls to telex subscribers.

Figure 2—Routing diagram for outgoing calls from telex subscribers.



magnetic circuit and pretensioned wire springs combined with card operation provide long life and fast operation with low actuating power.

Its excellent contact performance stems from the use of palladium, the equalizing of pressure on all contacts, complete independence of the twin contacts from each other, the special shape of the contacts, and the provision of contact covers. Palladium is used for the fixed contacts and palladium-gold for the moving contacts to improve operation under different load conditions.

Relays are mounted in completely dustproof enclosures. Those associated with the selector circuits are of the jack-in type to permit variation of the capacity of a group to meet traffic requirements.

A single switch rack accommodates equipment for 300 subscriber-station lines including the 300 line circuits, 3 groups of connecting circuits with repeaters, and 3 groups of final selectors. The interconnecting cables are in the base of the rack. An enclosed cable rack on the floor of the switchroom provides for interconnecting the racks.

Each subscriber's line circuit consists of 2 wire-spring relays and 2 adjustable resistors. The relays for 10 line circuits are grouped in one apparatus unit. The adjustable resistors are mounted at the top of the bay. These resistors control the line and bias currents of the repeater that is part of the connecting circuit and they must be adjusted before a line is put in service to compensate for the line characteristics.

Figure 3—Routine diagram for transit calls.

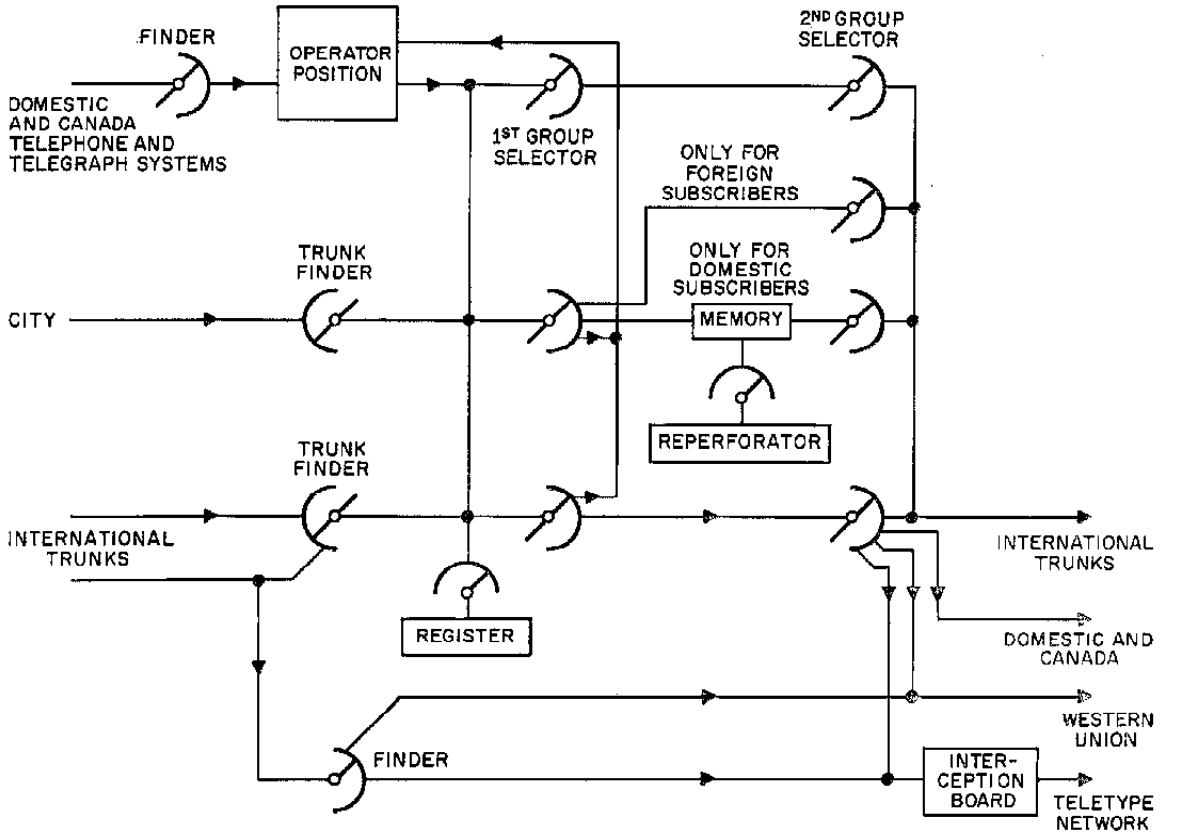


Figure 4—Rotary switch having 7 double-ended non-homing brushes and 100 sets of outlets.

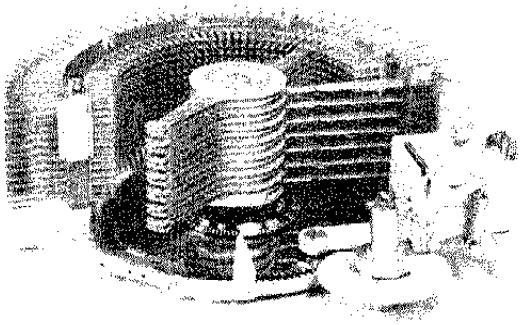
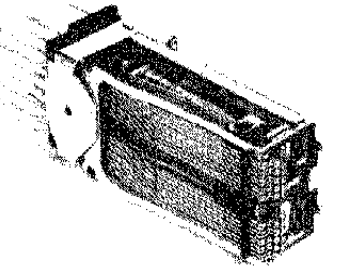


Figure 5—Wire-spring relay and contact cover.



P. J. Clemens was born in Utrecht, The Netherlands, on 3 May 1899.

In 1921, he joined Bell Telephone Manufacturing Company, Antwerp, as a student engineer. He was placed in charge of design and development of telegraph switching systems in 1936.

Mr. Clemens is now chief of the telegraph department.

High-Power Precision-Stabilized Rectifiers

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Certain physical experiments carried out in a strong magnetic field put stringent demands on the stability of the field and consequently on the current in the magnet coil. These demands are so rigid that formerly they could be met only by using batteries.

To avoid the disadvantages of batteries with regard to space, maintenance, efficiency, requirement of auxiliary equipment for precise regulation, et cetera, more-convenient solutions have been developed.

For example, there have been fairly successful attempts to provide motor-generator equipment with automatic control to obtain the constant current required. In this way it is possible to reach stabilities better than 1 in 10^3, but many problems of space, efficiency, vibration, and ripple remain unsolved.

Using electronically controlled rectifier equipment did not seem very attractive at first, since the electron tube operates essentially at a high impedance level and is not suitable for currents of more than a few amperes directly.

The advent of power transistors offered important possibilities. These transistors are suitable for application as control devices because of their low-impedance characteristics, and they can be inserted directly in the main current path. A final development concept was reached based on the use of power transistors, silicon power diodes, and transistors in all associated electronic circuits.

As a result, a stability of 5 parts in 10^5 of the selected current for an 8-hour period or 5 in 10^6 for a period of several minutes with an effective ripple of 5 in 10^6 can be guaranteed for equipments now being built.

In these equipments, the control circuit can cope with ±10-percent slow voltage variations, ±3-percent transient variations, and also resistance variations in the magnet coil.

Rectifiers are now being built with ratings from 1.4 kilowatts at a maximum of 8 amperes through 112.5 kilowatts at a maximum of 450 amperes. One of the 112.5-kilowatt units will be used to operate the famous "big magnet" in the Kamerling Omnes Laboratories at Leiden.

The essential problem in developing high-stability equipment of this type is not the circuit design, which is quite conventional, but the necessity to construct every detail with utmost care and to pay attention to minor considerations that are normally neglected.

Figure 1 shows the simplified circuit of the rectifiers. The 3-phase supply is connected to the motor-driven regulator transformers, which provide a voltage adjustable from zero to a level slightly above the supply voltage.

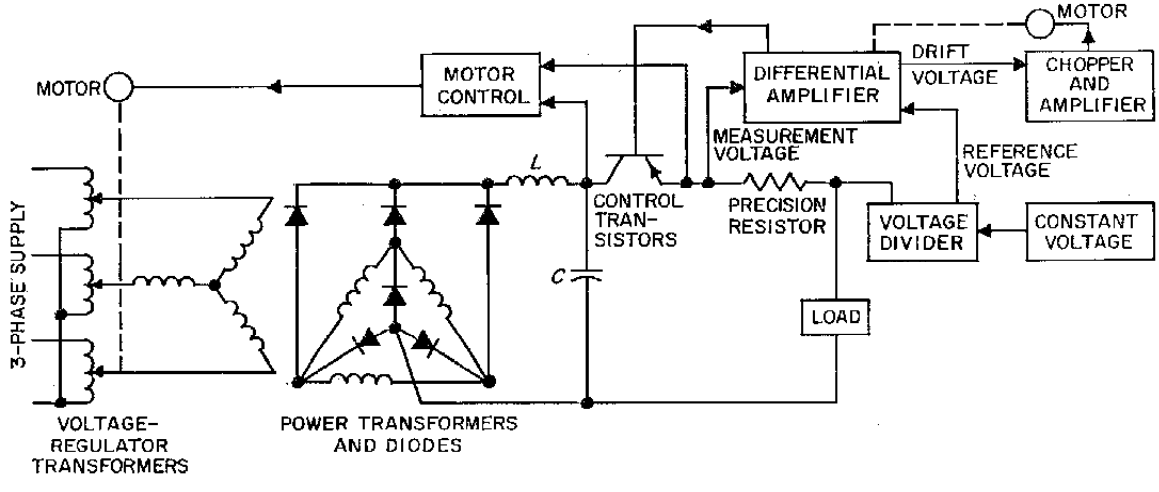


Figure 1—Simplified circuit of rectifier and control system.

This voltage is converted by the power transformers to the desired level and then rectified by water-cooled silicon diodes. The ripple is reduced to 1 percent effective by a smoothing filter consisting of inductance L and capacitance C. The rectified current flows through control transistors, a precision resistor, and the load.

When the control circuit is inoperative, the output current is a function of the supply voltage and the load resistance. Stabilization of load current takes place as follows. The precision resistor provides a voltage proportional to the load current. This voltage is compared with the reference voltage obtained from a precision voltage divider operating from a constant-voltage source. The difference between the two voltages indicates the deviation of the measured value of the load current from the desired value. This difference is amplified and used to bias the control transistors to provide a limited control range capable of coping with fast variations. For economy, the number of transistors is related to the load current. Slow variations caused by selected changes in output current and by slow supply-voltage variations are corrected by simple motor-driven control of the voltage-regulator transformers. The average value of the transistor voltage is thus stabilized.

Proper construction of the precision resistor is very important to maintain current stability. Its resistance value should stay constant under all load conditions and it should be isolated from induced extraneous voltages. It is constructed as follows. Two copper bars separated by a thin layer of mica are firmly clamped together. The ends of a circularly bent sheet of manganin, which forms the actual shunt, are soldered to the narrow side of each bar. Thus the temperature difference across the manganin-copper junctions is minimized because of the very small heat resistance between the bars. As a result, the thermoelectric potential difference will not exceed 0.3 microvolt from zero to full load (1 volt at 450 amperes) across the resistor. The voltage terminals are mounted on the side of the shunt in direct contact with the manganin. One terminal wire runs along the inside of the manganin and then enters a common shield with the wire from the other terminal. In this way they are not exposed to induction fields. Furthermore, advantage is taken of the fact that the resistance-temperature characteristic of manganin follows a parabolic curve with its top at about 30 degrees centigrade (86 degrees fahrenheit). The whole unit is immersed in oil that is thermostatically controlled within ± 0.5 degree centigrade (± 0.9 degree fahrenheit) at this optimum temperature. Forced oil circulation keeps the temperature difference between oil and manganin within this tolerance. Surplus heat is extracted by water cooling. After careful adjustment and aging, the precision resistor can maintain a stability of 1 in 10^6 for many hours.

Another direct influence on the stability of the output current is the reference voltage to which the shunt voltage is compared. This reference voltage, which can usually be adjusted between 0 and 1 volt, is obtained from a voltage-divider circuit connected to a very-stable voltage source consisting of 3 zener-diode stages in cascade that reduce supply-voltage variations to 1 in 10^6. The ripple on the reference voltage is also 1 in 10^6.

The only other serious problem is the fact that the parameters of the semiconductors, especially those in the final stabilization circuit, are temperature dependent. Three remedies effectively reduce this disadvantage.

(A) Careful compensation for positive and negative thermal effects.

(B) Enclosing parts within a copper bar to equalize their temperature variations.

(C) Placing the whole circuit in a thermostatically controlled oven to keep temperature variations within ± 0.5 degree centigrade (± 0.9 degree fahrenheit).

Thus the output voltage enjoys a stability in some units of 1 in 10^6 for many hours. It is also possible to keep the ripple voltage from exceeding a few microvolts by rigid shielding and careful arrangement of parts.

Adjustment of the voltage-divider circuit controlling the reference voltage for various output currents may be based on the following considerations: stability, possibility for remote control, resolving power, accuracy of resetting, and absolute value of the setting. Ordinary multiturn wire-wound potentiometers are unsatisfactory because the sliding contact may touch more than one turn at a time and cause variations of resistance with vibration and shock.

The voltage-divider network used consists of a number (usually 4) of high-quality 10-position step switches with wire-wound precision resistors. Thus the output current can be digitally adjusted in 10 000 equal steps. For instance, selection of the number 3785, for a power supply with a maximum output of 100 amperes, equals an output current of 37.85 amperes. The output current deviation does not exceed 5 units in the last digit, or 50 milliamperes. It should be noted, however, that this concerns only the current-setting accuracy and has nothing to do with the actual stability of the output current, which could be of the order of 2 milliamperes.

The entire voltage-divider circuit is placed in a thermostatically controlled oven. The switches are motor-driven using simple servo circuits. The control switch for these servo circuits can be installed at the most convenient spot with regard to the experiment for which the stable current is needed.

Without precautions, the output current could make enormous jumps (10 amperes) because of the fast response of the control system if the digital switch is adjusted indiscriminately. This is avoided by preventing any change in output current until selection of the new current setting has been completed. The output current then smoothly adjusts itself to the newly selected value. This method of resetting also permits slow and continuous scanning of a given current range.

Finally, the differential amplifier is of critical importance to the stability of the output current. This direct-current amplifier suffers from drift and a gain characteristic that slopes downward toward the high frequencies, where a certain amount of phase shift occurs. Very-high gain is required. A typical input-signal variation of 10 microvolts across 1000 ohms produces 0.01 microampere and 0.1 picowatt to control output values across the control transistors of 4.5 volts and 450 amperes, or 2 kilowatts, indicating a power amplification of 10^{16}. It is necessary, for example, to use as many as 4 voltage- and 5 current-amplification stages in cascade.

The required bandwidth is from direct current to 20 kilohertz, the limiting factor being the 10-millisecond time constant of the smoothing filter in the main rectifier circuit. As a result, higher frequencies in the output voltage influenced by a sudden supply variation are considerably attenuated. Current components containing frequencies higher than 5 kilohertz are sufficiently attenuated to reduce any variations in the output current considerably even without special means of control. Accordingly, the phase shift in the amplifier must be negligible from direct current to a few kilohertz, while at 20 kilohertz a 10-to-20-degree phase shift combined with a few decibels reduction in gain is acceptable. Therefore, the required characteristic can be met without too much difficulty because of the relatively small bandwidth, although the parts and wiring require careful arrangement.

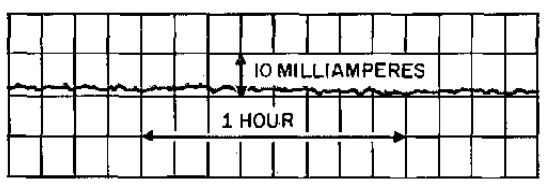


Figure 2—Rectifier stability test results at 200 amperes.

An additional difficulty in keeping the control system free of oscillation is that these rectifiers must be capable of energizing various types of magnet coils and other loads having a wide variety of impedances. The direct-current cable capacity may also be a problem. To guarantee proper operation under all conditions, the system has been provided with several feedback loops. One of these is obtained by re-injecting the ripple component of the output voltage via a simple network into the amplifier, resulting in effective ripple suppression.

The deviation of the output current from its average value under the influence of quick disturbances never exceeds ± 5 parts in 10^6.

With regard to drift, the amplifier output voltage should have a certain absolutely constant value under ideal conditions when the differential input terminals are short-circuited. In reality there are two deviations, drift and amplifier noise, the first causing a variation in the direct-current level and the second being an alternating voltage containing a certain frequency spectrum. The latter factor is negligible since the noise level at the amplifier input is less than 1 microvolt. To correct for drift, it is necessary to provide chopper stabilization for the amplifier. The differential-amplifier direct-voltage drift of several microvolts is chopped at 50 hertz by a precision switch and converted into a small alternating voltage.

The phase of the alternating voltage depends on the polarity of the drift voltage. After high amplification, it is used to control a small servomotor coupled to an adjustable resistor in the differential amplifier. If drift occurs, the servomotor drives the adjustable resistor in the direction to cancel the drift voltage. The drift of the differential amplifier is thus held to less than 1 microvolt. The cancellation rate is relatively slow and has a bandwidth of only a few hertz.

By summing the maximum errors of all individual circuits, the expected stability of the

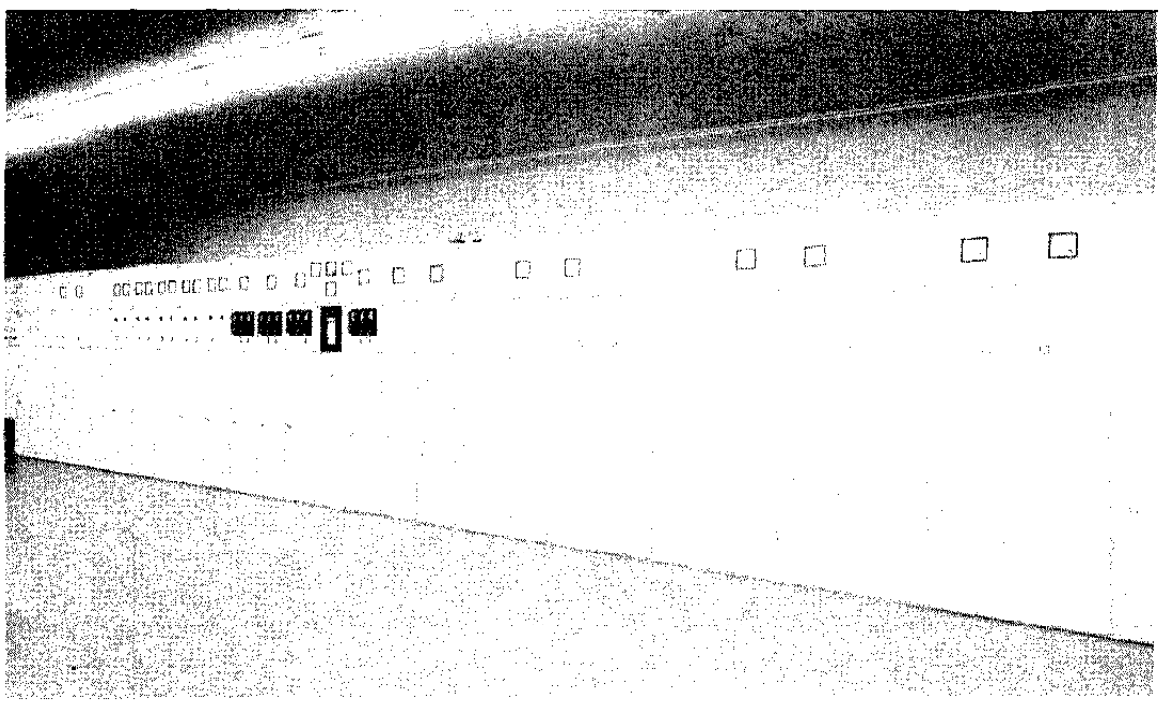


Figure 3—High-stability rectifiers located in the Physical Laboratory of the University of Groningen.

output current can be predicted. If we consider that in a certain rectifier installation the compared shunt and reference voltages vary between 0.2 and 1 volt, we find the following maximum errors: shunt, 1 microvolt; reference voltage, 3 microvolts; voltage divider, 3 microvolts; and differential amplifier, 1 microvolt. The total error is less than 10 microvolts or a stability of 1 in 10^5 at the maximum output current, or 5 in 10^5 at 0.2 of the maximum output current. An additional safety margin is that it is highly improbable that all errors are both maximum and additive.

Figure 2 shows part of a stability test record of one of the rectifiers. A distribution room in the Physical Laboratory of the University of Groningen, housing a series of high-stability rectifiers, is shown in Figure 3.

Acknowledgement

The author is greatly indebted to his assistants, J. de Graaf and A. Cools, who have contributed substantially to the realization of the equipment outlined in this article. He also thanks Mr. W. A. Hansen of The Netherlands Government Building Authority and the directors of the various research institutes for their full cooperation and confidence.

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After several years of work on automatic traffic controls, he joined Nederlandsche Standard Electric in 1952. In 1954, he was placed in charge of circuit engineering for the rectifier department. Since 1957, Mr. Kemper has been responsible for engineering and development for the rectifier and automatic-controls department.

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material of the same conductivity type to avoid formation of barrier layers. At least two of the layers have temperature coefficients of conduction of opposite sign.

Intelligence Storage Equipment

3 051 931

P. W. Lennox

A magnetic storage unit for building up a matrix of rows and columns, with a control access magnetic core, is the subject of this patent. The unit provides a series of magnetic cores, which form a matrix row, around the periphery of a somewhat larger core that serves as a switch to control access to the individual components of the storage array.

Sealed Contact Device

3 056 869

B. Dal Bianco and M. Scata

This is a sealed relay switch, operable by magnetic effects. It consists of two conductors of magnetic material and a magnetic conductive element to bridge between the two conductors. The bridging element is pivoted on one of the conductors, and in the absence of an operating magnetic force opens the switch to an extent determined by the enclosing envelope.

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Intelligence Storage Equipment With Independent Recording and Reading Facilities

3 048 827

E. P. G. Wright, D. S. Ridler, and R. Grimmond

This is a storage arrangement that receives and stores a block of data consisting of a number of words, each of which is received in binary code in parallel fashion over a number of channels equal to the number of elements in the word. This block of data is stored and then read out either immediately or subsequently, at a predetermined rate, which might be different from the rate at which it is stored. The recording and reading means are entirely separate so that both processes can be in progress concurrently.

Circular Antenna Array System Scanning Switch

3 048 843

F. Steiner

A rotating-pattern Doppler-type circular antenna array is provided with a scanning switch to selectively actuate the antennas of the array to produce a rapid oscillation of the pattern that progresses at a slow rotation rate around the circular array. This is accomplished by a variation in the spacing of stator and rotor coupling elements of the scanner. A vernierlike incidence of coupling is provided.

Telephone Systems

3 049 594

F. P. Gohorel

This switching circuit arrangement, designed for the Pentaconta system, allows a considerable simplification in the design of the registers, multiplying circuits, and supervisory pulse senders. The system features common and individual devices for controlling selection and certain auxiliary functions controlled by trunk bundles designed to associate the devices temporarily. The number of these trunk bundles is relatively small because of their short holding time.

Electrical Condenser

3 054 029

M. Wagner and M. Kolb

This is a capacitor having an anode of a base material such as aluminum or tantalum with a coating of aluminum or tantalum oxide and a plurality of layers between the coating and the cathode. The layers are of semiconductor

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