Dr R.A. Bones

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Microwave Radio System for Multichannel Telephony and Television in the 6-Gigahertz Range (5 Parts)—Each system is capable of handling 1800 telephone channels or a television program and 8 such systems can be operated together between 5925 and 6425 megahertz. Solid-state devices have replaced electron tubes except for the traveling-wave tube in the output stage of each transmitter.

For relaying, the system operates with the signal through-connected at the intermediate frequency of 70 megahertz, the same radio equipment being used at repeater and terminal stations.

The transmit local-oscillator chain starts with a quartz crystal in the 50-megahertz range. A series of varactor multiplier stages using lumped circuits increases this frequency to 800 megahertz, followed by multiplication to 6400 megahertz in a varactor multiplier using waveguide and coaxial techniques. About 50 milliwatts are available to drive the output mixer. The receive local-oscillator chain is similar except that transistors replace some of the varactors as the required output is only 2 milliwatts. This reduces the number of adjustments that must be made for optimum noise performance.

The intermediate-frequency amplifiers must raise the incoming received signal from as low as -55 to about $+5$ decibels referred to 1 milliwatt. If the input signal carrier fails, a locally generated carrier is automatically substituted to maintain the following equipment in normal operation.

A regulating amplifier for automatic gain control and an output amplifier are coupled through delay and amplitude equalizers. The pass band of 20 megahertz centered at 70 megahertz is established by a low-pass filter and the input transformer.

The antenna must operate from 5.9 to 6.45 gigahertz and permit simultaneous transmission and reception. A cassegrain system is used. For transmission it is excited from a circular waveguide terminated in a horn. This irradiates a hyperboloid that reflects the energy to the main paraboloid, which is pointed at the distant cooperating station. The signal path is reversed for reception. The paraboloid is 13 feet (4 meters) and the hyperboloid is 6 inches (15 centimeters) in diameter. The antenna gain is 45 decibels and the efficiency is above 60 percent. The 3-decibel-down beamwidth is 0.8 degree.

The circular symmetry permits operation with waves having crossed polarization planes. Operating at the same frequency, routes at azimuth angles to each other exceeding 15 degrees can operate with crossed polarization and at angles exceeding 85 degrees with the same polarization.

A section of circular waveguide, its horn, and the hyperboloid reflector are mounted rigidly on a fiberglass cone, which is in turn bolted to the paraboloid.

The remote control and supervision of the many repeater stations in a chain are done through auxiliary equipment that provides up to 24 telephone channels. It operates over the same antennas and in the same general frequency range as the main system. The 10-milliwatt transmitters use only solid-state active components. A crystal-controlled signal is frequency modulated and then multiplied to increase the amplitude, frequency, and modulation deviation. In the receiver a 4-section printed-circuit filter determines almost completely the 1.5-megahertz pass band of the 35-megahertz intermediate-frequency system.

Carrier Telephone System for Mine Railroads—The *Z1G* carrier telephone system operates between 35 and 135 kilohertz with narrow-band frequency modulation and transmitter powers of about 1 watt. The voice-frequency band is from 500 to 2400 hertz to suppress the high noise levels found in mines and give maximum intelligibility for the power used.

The central dispatching station is directly connected to a 2-wire transmission line that is

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branched through hybrid circuits; the ends of runs are terminated in the characteristic impedance of the transmission line. The mobile stations in mine locomotives are coupled to the transmission line through loop antennas. Two transmission frequencies are used, one for the central station to which all mobile receivers are fixed tuned, and the other for all mobile transmitters to which the central-station receiver is fixed tuned.

The transmitter at the central station operates continuously and is heard by all the mobile stations. Any mobile station can call the central station and work duplex on a press-to-talk basis. By connecting its receiver output to transmitter input, the central station can serve as a relay to permit mobile stations to interwork on a simplex basis. During such operation, a mobile installation at the central station permits a normal dispatch of business.

Theory and Design of an Adjustable Equalizer—An equalizer is often required that has an attenuation characteristic varying as a function of a single resistor (for example, a thermistor). This equalizer is often used in regulated systems; it is known as the Bode equalizer or adjustable equalizer. Here the mathematical relations are developed for the linear passive network required, particularly dealing with a special type of equalizer. A ladder of bridged-T networks and all-pass filters determines the attenuation. Accurate approximation of the attenuation characteristics calls for the all-pass filters as phase-correcting components. While the output of this ladder network is terminated with the adjustable resistor, the input is connected into a series branch where the fan-out of the attenuation characteristic must take place. Of course, the relation for a dual arrangement can conveniently be derived.

The method to obtain an adjustable equalizer is demonstrated by an example. The necessary numerical calculations were made with an electronic computer.

Transmission-Line Mismatches and System Noise Figure—Improvement of even a fraction of a decibel in the noise figure of some amplifier systems can result in a significant increase of operating capability.

Equations are developed to help the design engineer make accurate predictions on the noise figures of highly sensitive systems, once the performance of the individual components is known.

Computer Assistance to Pentaconta Engineering—Newly designed frames of Pentaconta telephone switching equipment are now being wired under normal manufacturing conditions from instructions produced by a computer. Substantially less time is needed for the computer than for manual design of wiring, and this shortens the overall manufacturing time for new equipment.

The high density of the Pentaconta equipment requires that much of the internal wiring within a frame be run separately and not in harnesses. Thus, two wiring methods must be used and wiring documents prepared for each. The process is as follows.

An engineer prepares a wiring schematic drawing. From it, a technician lays out the equipment for each frame and the site of each relay or other component is identified according to a plan. He also prepares a drawing of the outlet fields of the frame plus the connections that go to each. All wiring instructions are developed from these three drawings. All these data are punched on cards, which are then read into the computer.

A program is stored on magnetic tape for the computer, as is a catalog of all additional data covering equipment that will be needed. The computer processing is done in a core memory. Output is by printer at 1000 lines per minute.

Several documents are issued automatically: a list of separate wires including length and path information, a semioptimized list for harness construction with code colors, a connecting list

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for the harness, and a complete wiring document for factory testers and for installers and customers. Each of these documents is in the most-suitable form for the user.

Power-Frequency Induction on Coaxial Cables With Application to Transistorized Systems—Using equations developed for the purpose, computer calculations have been made of the currents and voltages induced from adjacent power systems into small-diameter coaxial cables over which direct-current operating power is supplied to transistor repeaters. The induced currents and voltages between coaxial inner and outer conductors are smallest if these conductors are insulated from earth and if the exposed region of cable is as far as possible from the center of the coaxial power-feeding section. The highest values of current in the coaxial inner-to-outer-conductor circuit occur near the center of the power-feeding section regardless of the position and length of the exposed region, the voltage being highest at the two ends of the section.

Representative curves are given for a 300-channel system using *174*-type coaxial cable with nominal repeater spacing of 3.6 miles (5.8 kilometers) and a direct-current power-feeding section of 40 miles (64 kilometers). The system conforms with the recommendations of the Comité Consultatif International Télégraphique et Téléphonique.

Photo-Etching for Reliability in Electron-Tube Manufacture—A chemically clean sheet of metal is coated with a material that is exposed to light through a mask made photographically. Where exposed to light, the material resists the action of a mordant that etches away the unprotected areas. Practically, the minimum width of alternate metal bars and etched regions should not be less than the thickness of the sheet being etched. In work on sheets 0.05 millimeter (0.002 inch) in thickness, perforations of that dimension were made to tolerances of ± 10 microns.

Production quantities of sheet-metal components of electron tubes can be photo-etched in one-tenth the time required for conventional press-tooling and without trouble from burrs, contamination, and reduced uniformity resulting from tool wear. This contributes to the quality and reliability of the tubes. Prototype parts and later modifications can be produced quickly to allow time for operational testing of new designs.

Selenium Rectifiers in 1965—Present theories of the electrical properties of selenium are surveyed. The laws governing electrical conduction in selenium differ considerably from those for silicon and germanium so that the extensive research in these latter semiconductors is not directly applicable to selenium.

Despite the limitations of present theoretical knowledge, it has permitted substantial improvements in commercial selenium rectifiers. Power ratings have increased by two orders and with improved efficiency, termination of life because of a steadily increasing forward voltage drop and consequent heating has been eliminated, physical dimensions have decreased, and longer life is assured.

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Recent Achievements

New Factories in Spain—In October 1964, His Excellency Generalissimo Franco inaugurated two new factories at Villaverde, about 9 kilometers (5.6 miles) from Madrid. Other honored guests, including ministers of the government and officers and directors of the Compañia Telefónica Nacional de España, were received by the officers of Standard Eléctrica and by Messrs. H. S. Geneen, E. W. Stone, and F. J. Dunleavy of the International Telephone and Telegraph Corporation, who came to Madrid especially for the event (Figure 1).

The two identical plants (Figure 2) have a total floor area of 40 000 square meters (430 000 square feet) and the site is of 125 000 square meters (31 acres). Both factories were in operation with about 3800 employees.

The first plant began operation in 1963 and now produces about 400 frames of Pentaconta switching equipment per week. A scheduled increase to 500 frames will permit the full requirements of the Spanish Telephone Administration to be met on a continuing basis with some exportable surplus.

Figure 1—At the inauguration of the new Villaverde factories near Madrid, the Chief of State, his Excellency Generalissimo Franco signed a commemorative document. From left to right were: Count of Casa de Loja, Head of the Casa Civil; M. Fraga Iribarne, Minister of Information and Tourism; G. Lopez Bravo, Ministry of Industry; J. Sánchez Cortés, Undersecretary of Finance; H. S. Geneen, President and Chairman of the Board of International Telephone and Telegraph Corporation; G. Azcoitia Muesca, member of the Board of Directors of Standard Eléctrica; and M. Márquez Mira, President of Standard Eléctrica.

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The second plant went into service in September 1964 and will produce 120 000 lines of rotary equipment per year. About a third of its area will be used for making transmission systems. This equipment will be based on the Standard Equipment Practice for ITT Europe, which was recently adopted by the Spanish Telephone Administration.

Standard Eléctrica
Spain

International Direct Dialing Initiated in France—With the inauguration by Jacques Marette, Minister of Post and Telecommunications, of Cadet, an international outgoing toll center, foreign calls may be dialed directly from the French telephone network. Figure 3 shows some of the honored guests. The first calls were placed by Mr. Marette to his colleagues in West Germany and in England, in which countries the event was reported on television.

This Pentaconta equipped toll center is reached from Paris central offices by dialing 19 followed by the 1-, 2-, or 3-digit prefix identifying the wanted country and then the full number of the called party. Initially equipped for 650 voice channels, it has an ultimate capacity of 1600 channels and 90 service lines.

In handling international transit traffic, Cadet cooperates with Cinat, the first Pentaconta incoming and transit toll center put into service last year.

Le Matériel Téléphonique
France

Figure 2—The two new Villaverde factories near Madrid.

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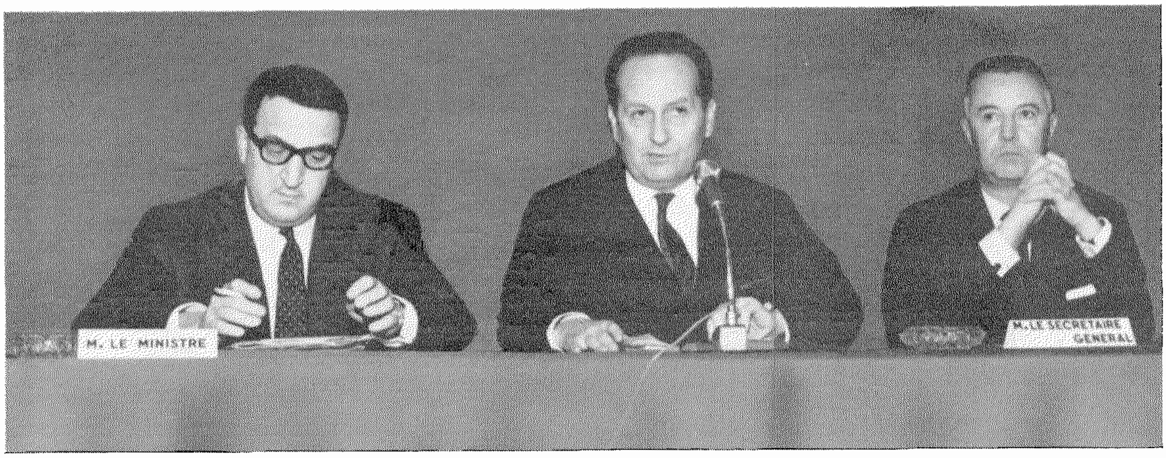
Figure 3—At the inauguration of the French international direct dialing exchange in Paris were, from left to right, J. Marette, Minister of Post and Telecommunications, R. Croze, General Director of Telecommunications, and I. Cabanne, General Secretary of Post and Telecommunications.



Málaga Factory Inaugurated—In December 1964 a new factory in Málaga on the southern coast of Spain was inaugurated by G. Lopez Bravo, Minister of Industry, and other government officials.

The combined factory and office building has a floor area of 13 300 square meters (140 000 square feet), is on a plot of 37 000 square meters (9 acres), and is in operation with 650 employees. Planned entirely for the manufacture of the Assistant type subscriber set, the present output of 10 000 sets per week is about two-thirds of the final capacity. It is capable of supplying the national needs of the Spanish Administration and also providing for export. A section of the assembly line may be seen in Figure 4.

Compania Internacional de Telecomunicacion y Electronica
Spain

Figure 4—Inauguration of factory in Málaga, Spain. Immediately behind the worker are M. Marquez Mira, president of the company, and G. Lopez Bravo, Spanish Minister of Industry.

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Air Traffic Control—The air-traffic-control console shown in Figure 5 was demonstrated at the biennial convention of the Guild of Air Traffic Control Officers in Bournemouth, England, in October 1964. Using recordings on audio tape, radar pictures were presented on the plan-position indicators, and alphanumeric data, normally obtained from symbol generators under control of a computer, were presented on a tabular display.

In this digital radar system, the returned echoes (the hits) are correlated so that only one pulse is produced for each target. Echoes from the ground, rain clouds, et cetera, however are to a large extent discarded. Digital encoded azimuth position of the radar antenna is used as the base for the generation in real time of digital X and Y coordinates. The output target pulses release the X and Y coordinates to produce target information for transmission over a telephone line and/or a display on plan-position indicators. A standard telephone channel can transmit about 50 such target reports per second.

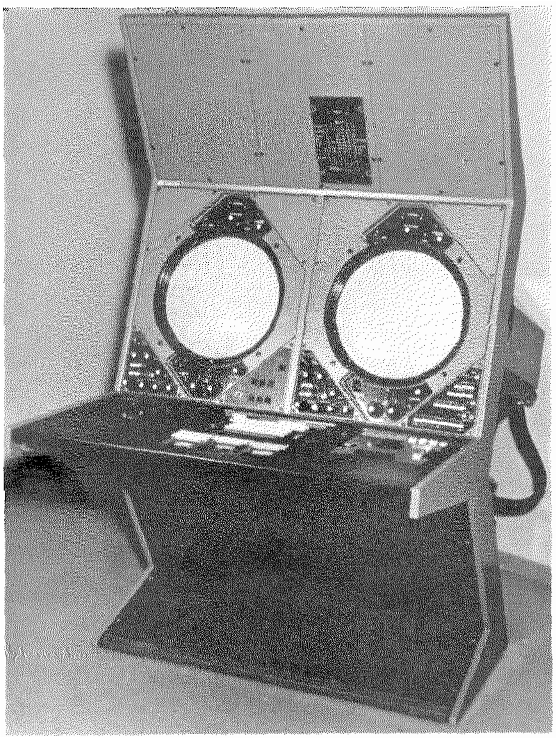

Figure 5—Air-traffic-control console operable over a standard telephone channel.

Standard Radio & Telefon
Sweden

New Headquarters in London—A new office building in the Strand opposite St. Clement Danes—The Wren church famous for its "Oranges and Lemons" chimes and now the official church of the Royal Air Force—is the new headquarters of Standard Telephones and Cables. Known as STC House, it is 10 stories high, fully air conditioned, and provides 47 450 square feet (4400 square meters) of office floor. There is underground parking for 40 cars.

Standard Telephones and Cables
United Kingdom

Push-Button Subscriber Sets Installed in Germany—Late in 1964 the first large private automatic branch exchange in Germany equipped for push-button dialing was inaugurated by the Bayerische Gemeindebank with its new office building in Munich. The exchange has a capacity of 67 public lines, 800 extensions, and 80 links. Initially 600 extensions were equipped and are provided with push-button subscriber sets as shown in Figure 6.

About a year and a half earlier, a central office equipped for push-button dialing was opened in Stuttgart, so that both types of switching centers are now in public service.

Standard Elektrik Lorenz
Germany

Refractometer Cures Radar Mirage—Variations in the refractive index of atmospheric regions can create optical mirages in which distant objects appear to the eye to be incorrectly placed. Similar bending of radio waves can produce inaccurate radar target positions.

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An absolute microwave refractometer has been delivered to the United States Navy to monitor the direction of radio waves through the atmosphere and to detect conditions that might cause radar waves to bend or to depart from straight paths in their travel. It measures the amount of bending and provides a signal that can be used by the computers that track the targets.

ITT Federal Laboratories
United States of America

Microwave System Installed in Short Time— Known as Big Rally 2, a transportable communication system linking military centers in Italy, Greece, and Turkey to the United States was turned over to the United States Air Force recently. The trunk system was completed in 144 days, during which time sites were acquired and prepared and equipment was procured, transported, installed, and tested.

The system includes 8 tropospheric-scatter links and 15 microwave spurs. As installed it has a capacity of 24 voice channels.

Federal Electric Corporation served as prime contractor under supervision of the Electronic Systems Division of the Air Force Systems Command. The using command is the Air Force Communication Service, and the system will also be utilized by both the Navy and the Army.

Federal Electric Corporation
United States of America

British Atomic Research Establishment Modernizes Its Communications—The largest private telephone system of its kind in Great Britain to use cordless operator switchboards was cut over at the Harwell plant of the Atomic Energy Research Establishment. Replacing a veteran exchange made by us, the new 2500-line private automatic branch exchange consists of a parent and two satellite exchanges at different locations in the Harwell establishment. They operate as a single system. About 60 lines connect to the public telephone network and there are 55 lines direct to associated facilities in the country.

Standard Telephones and Cables
United Kingdom

Oslo Crossbar Exchanges Cut Over—Two electronically controlled crossbar-switch telephone central offices have been put in service in Oslo. These *8B* exchanges use 100-point multi-switches.

The 4000-line Centrum-4 *8B* exchange can carry the heavy traffic of 4.4 equated busy-hour calls per subscriber line and can be extended to 6 such calls. The Kastellet *8B* exchange has an initial installation of 10 000 lines and a capacity of 20 000 lines.

Standard Telefon og Kabelfabrik
Norway

Infrared Photodiodes for Laser Applications— A series of photodiodes sensitive in the infrared region have been developed for investigation of the properties of lasers and for communication and radar systems using lasers.

Three tubes have opaque photocathodes with *S1* spectral response. The *F4018* is 1.25 inches (32 millimeters) in diameter, the *F4000* is 2.25 inches (57 millimeters), and the *F4015* is 5

Figure 6—About 600 of these push-button subscriber sets have been installed in a private automatic branch exchange in Germany.

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inches (127 millimeters) in diameter. Peak linear outputs exceeding 0.5, 5, and 30 amperes, respectively, with a rise time shorter than 5×10^{-10} second, can be directly coupled to a 50-ohm coaxial cable. The plane parallel electrodes have high voltage ratings and low dark current.

A 0.75-inch (19-millimeter) design, *F4014,* has an opaque photocathode of *S4* spectral response and will produce an output of 0.1 ampere with a rise time of 1×10^{-10} second. The *FW114A* has *S20* spectral sensitivity and is particularly applicable to high-energy pulsed ruby lasers.

ITT Industrial Laboratories
United States of America

Gerona Telephone Network—The first installation stage in the telephone network of the Spanish Province of Gerona comprises 17 exchanges that were put in service between July and September 1964. As is evident from Figure 7, it connects the main resort points of the Costa Brava with the capital of the province, Gerona, which has a transit center in addition to an urban exchange. About 12 000 lines of newly installed Pentaconta *1000* switching equipment serves all but the Figueras 2000-line sector center, which is an existing exchange using *7D* rotary switching.

Automatic switching to other provinces is planned but such calls are still being handled

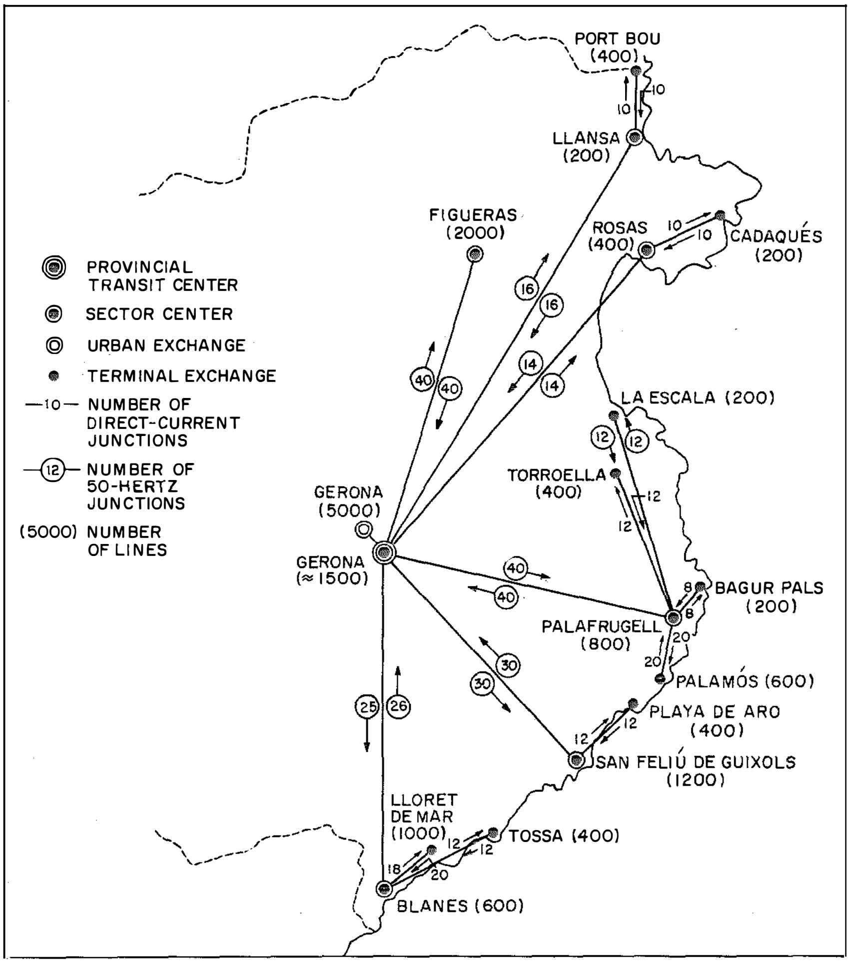

Figure 7—Gerona, Spain, provincial telephone network recently placed in service.

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manually. Selection signals are of the multi-frequency type except between Gerona and the *7D* exchange at Figueras, which uses 50-hertz signaling. The toll board at Gerona uses multi-frequency key sets.

Metering of local calls is by impulsing the subscriber's meter after the called number answers. Toll calls are metered every 3 minutes by a number of pulses related to the distance tariff.

Standard Eléctrica
Spain

Static Frequency Converter—Figure 8 is an equipment, using silicon controlled rectifiers, that converts 100–120 volts at 50 hertz to 75 hertz with a capacity of 110 volt-amperes. The output is stabilized for changes in load and in mains voltage and frequency. The load may be inductive or capacitive and may be short-circuited without damaging the equipment. It may be operated between −15 and +50 degrees Celsius.

The converter was designed for use in an automatic train control system for the Netherlands railroads.

Nederlandsche Standard Electric Maatschappij
The Netherlands

Pneumatic Tube for Schiphol Airport—The new Schiphol airport in Amsterdam will be equipped with an automatic pneumatic tube system having 46 stations and an ultimate capacity of 70 stations. Its 7000 meters (23 000 feet) of 75-millimeter (3-inch) tubing and magnetic sensing for switching carriers will make it one of the largest installations in the world.

Nederlandsche Standard Electric Maatschappij
The Netherlands

Mobile Radio System for Austrian Police—An installation of 40 fixed and 145 mobile very-high-frequency radiotelephone stations is part of a modernization program for the Austrian police system. Another 40 mobile sets will be added.

The *SEM 37/47* vehicle sets are designed for 10 channels. All but the driver and power stages of the transmitters use transistors. A telephone handset is supplied in addition to a loudspeaker. Mobile stations may communicate with each other as well as with the fixed stations.

Standard Elektrik Lorenz
Germany

Cathodic Protection of Ship's Hull—The corrosion that had been occurring to the hull of an ocean liner was arrested by a new cathodic protection apparatus, whose rectifiers and control equipment were supplied by us.

During a year of testing in which the ship sailed in practically all parts of the world, the continuous log of the operation of the system showed that the large variations in the current required to maintain the planned protection level could be met by the equipment.

Standard Telephones and Cables
United Kingdom

Crossbar Distributor for Television and Radio Signals—The distribution panel shown in Figure 9 permits 20 input television picture-and-sound channels requiring up to 6 conductors each to be connected to any of 20 output cables. The connecting cables are soldered to the panel terminals but the crossbar type of switching is done by inserting the plug shown in Figure 10 at the crosspoint of the two desired channels.

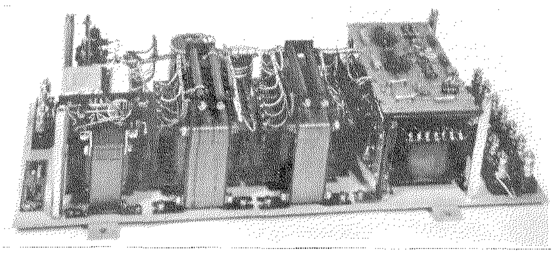

Figure 8—Static frequency converter using silicon controlled rectifiers.

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Crosstalk is so low that 9 such panels can be arranged to accommodate 60 input and 60 output channels. A lamp in the head of the plug can be provided to indicate active channels. Psophometric potentials from the signal circuit are negligible.

The space requirements have been greatly reduced over previous models and are only 520 by 490 millimeters (20.5 by 19.3 inches).

Standard Téléphone et Radio
Switzerland

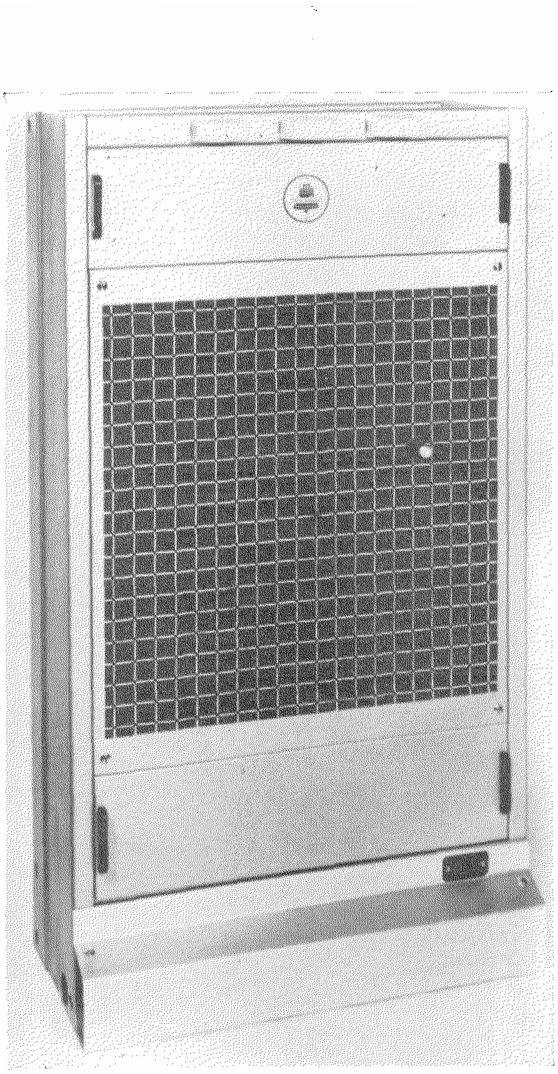

Figure 9—Crossbar-type distributor for television signals.

Line Concentrator for Use With Carrier Equipment—The *CM 12* Pentaconta crossbar line concentrator serves 52 subscribers over up to 12 voice-frequency channels of a carrier system. If the carrier operates over a physical 2-wire loop that may be used for signaling, all 12 carrier channels may be used for speech, otherwise one of the channels must be reserved for signaling.

Signaling is by one out of four frequencies: 900, 1350, 2000, and 3000 hertz. The maximum attenuation permitted is at the highest frequency and is 2 nepers (17.4 decibels) relative to 1 milliwatt referred to a point of zero relative level. Subscriber supervisory signals and 50-hertz ringing current are provided so that any carrier with *E* and *M* leads may be used.

The remote unit may be wall mounted and can be swung out for access to the rear. Power is from a sealed cadmium-nickel battery of 3.5-ampere-hour capacity charged from the alternating-current mains through a 75-volt-ampere rectifier. The 52 subscriber 2-wire lines are connected to plug-in sockets. The dimensions are 145 by 66 by 20 centimeters (57 by 26 by 8 inches).

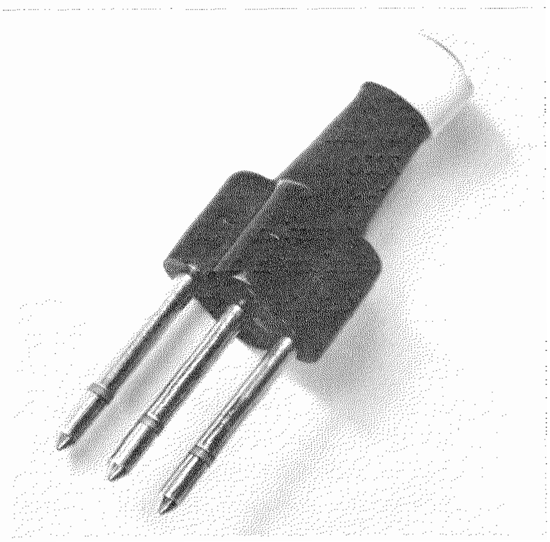

Figure 10—Plug for crossbar television distributor.

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The central-office unit will operate with any system that has a third wire available in the subscriber-line circuit. Normal power supply is −48 or −60 volts. Housed in a frame with removable covers, the equipment is normally mounted in a rack in the room with the main distributing frame. Dimensions are 127 by 76 by 20 centimeters (50 by 30 by 8 inches). Both equipments are shown in Figure 11.

Le Matériel Téléphonique
France

Precipitators Reduce Air Pollution—At a recent installation of an electric precipitator in the chimney of a power generating station, the thick gray smoke changed to a clear heat haze in 6 seconds. As much as 4 long tons (4100 kilograms) of dust can be removed each hour from the smoke of a single power station.

Our high-voltage rectifier equipment is used in these installations, which charge the dust particles at one polarity and collect them by attraction to plates maintained at the opposite polarity.

Standard Telephones and Cables
United Kingdom

Selective Calling for *SEM 47* Mobile Radio—The narrow panel attached to the bottom of the *SEM 47* mobile radio set shown in Figure 12 permits each of 45 different stations to be called selectively from the central transmitter. A pilot lamp, a buzzer, and the loudspeaker, which may be switched off if not needed, indicate receipt of the call code for the station.

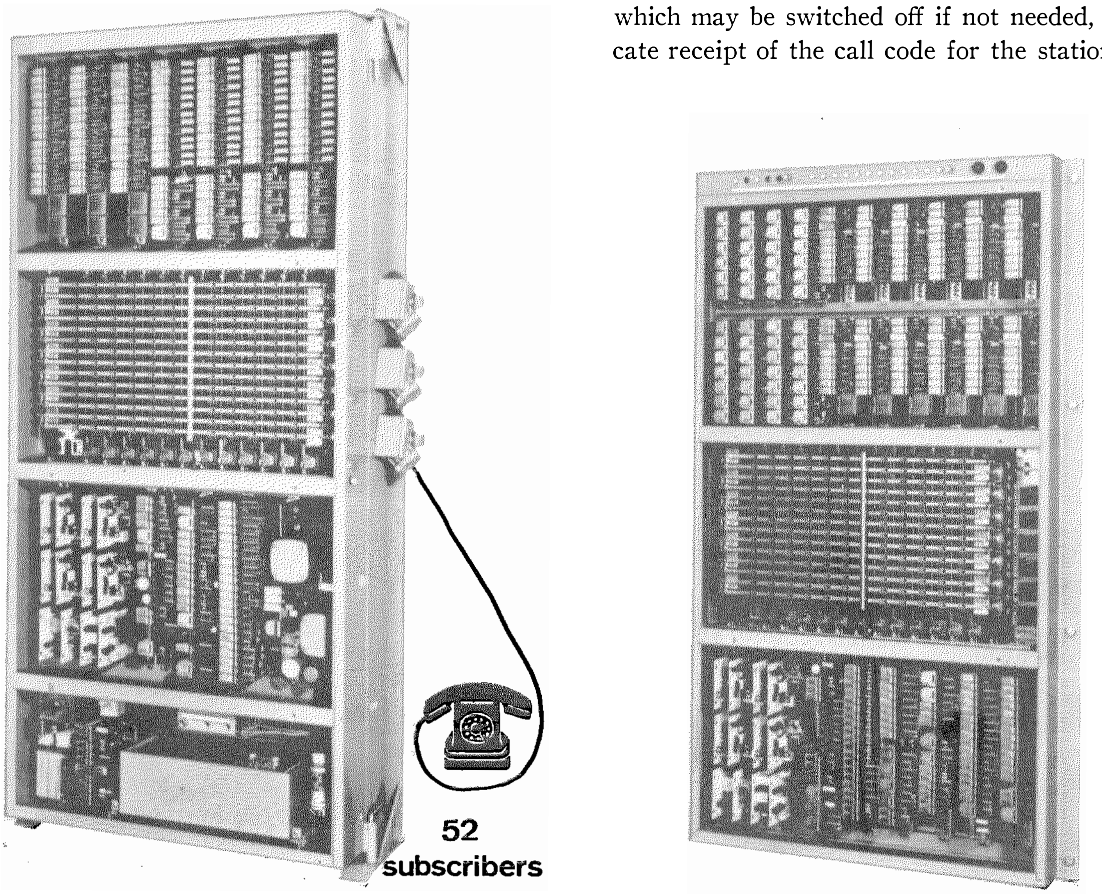

Figure 11—Remote (left) and central-office (right) units of the *CM 12* line concentrator.

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Changes may be made in the calling code of a receiver by resoldering capacitors. The loudspeaker in the receiver will be muted until its code signal is received so as not to reproduce other traffic on the same channel.

Standard Elektrik Lorenz
Germany

Figure 12—Mobile radio set *SEM 47* with selective call attachment mounted beneath the cabinet.

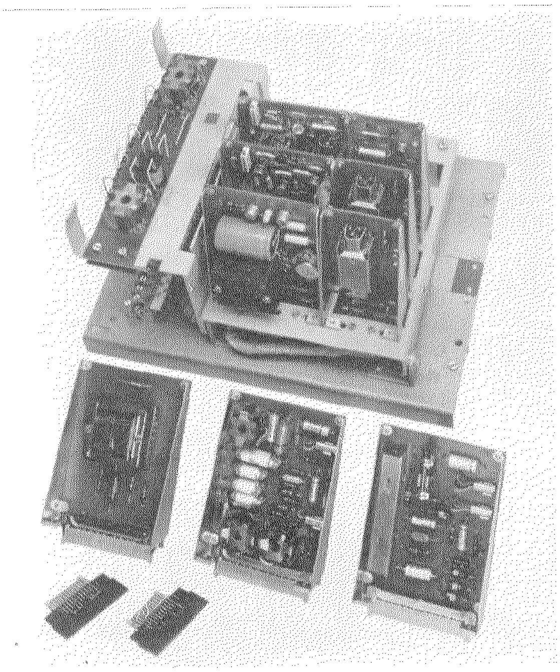

Figure 13—Interface equipment for connecting *SEM* and *SEF 27* to *47* radio installations to the public telephone network.

Radio-Wire Exchange Equipment—The use of transistor mobile sets for fixed radio stations makes necessary new interface equipment for connecting the radio system to the public telephone network. The extension of the voice path offers no problem but the 12-volt power source of the radio equipment is not suitable for the telephone plant. Direct-current conversion is used to produce the required higher voltage.

Figure 13 shows the small wall-mounted equipment and plug-in panels that can be selected to meet the needs of each installation. Without wiring changes in the radio equipment, panels are available for 1-way or 2-way operation, 2-wire or 4-wire switching, and tone calling. The *SEM* and *SEF 27* to *47* radio equipment for which the interface was designed is fully

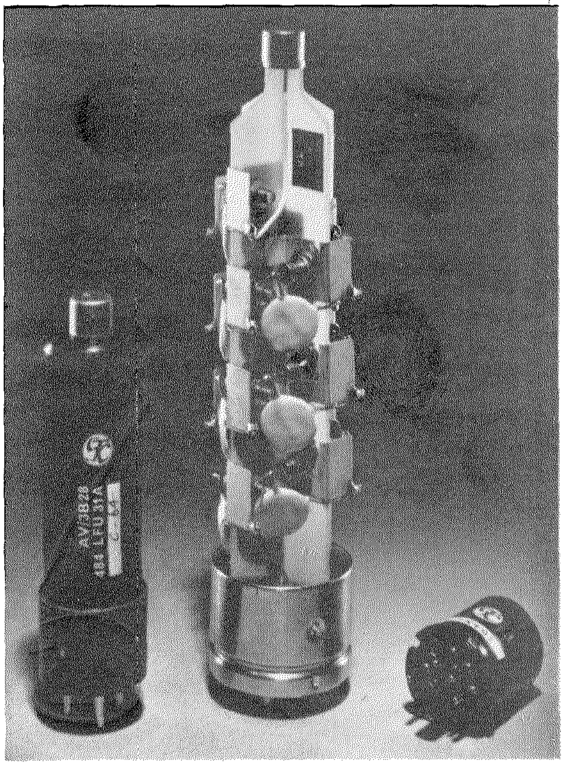

Figure 14—Silicon rectifiers for extra-high tension are mounted on conventional thermionic-rectifier bases for direct replacement.

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matched to the 600-ohm telephone line impedance and also provides for level compensation.

Standard Elektrik Lorenz
Germany

Silicon Rectifiers on Valve Bases—Extra-high-tension silicon avalanche rectifiers are now available on conventional vacuum-tube bases as direct replacements for thermionic rectifiers used in radio and television equipment. The tube-filament power supply is no longer needed and may be left without load. Power is available immediately as there is no warm-up period. Typical units are shown in Figure 14.

Standard Telephones and Cables
United Kingdom

Export Paystation—The export model telephone paystation shown in Figure 15 is suitable for local and automatic toll calls. It can be equipped with an electronic ringer for incoming calls. Either a nonequalized or an equalized type of circuit can be provided.

Figure 15—Export model telephone paystation.

It can be equipped with one, two, or three coin slots that accept coins between 15 and 35 millimeters (0.6 and 1.4 inches) in diameter. The diameter, thickness, weight, and magnetic properties are automatically checked. The cashbox is in a separate compartment having a cylinder lock. The housing is designed to resist vandals and burglars. The handset cord is protected by a metallic spiral sheath.

The switchhook is elastically coupled to the inner mechanism to avoid damage if violently agitated, and a retarding device delays closure of the line contacts. The coin channels are plug-in to facilitate maintenance, and unused coins are sent to the covered return outlet when the handset is restored.

Bell Telephone Manufacturing Company
Belgium

Diode-Capacitor Memory—A memory element consisting of a capacitor and two diodes has been developed for pulse-code-modulation telephone switching systems. This temporary store has a write and read cycle of 0.5 microsecond, which can be made shorter as it depends on controllable elements—the speed of the diodes and access circuits and on the time constant of the capacitor.

Figure 16 shows a memory for 72 words of 24 binary digits each. The control current is only

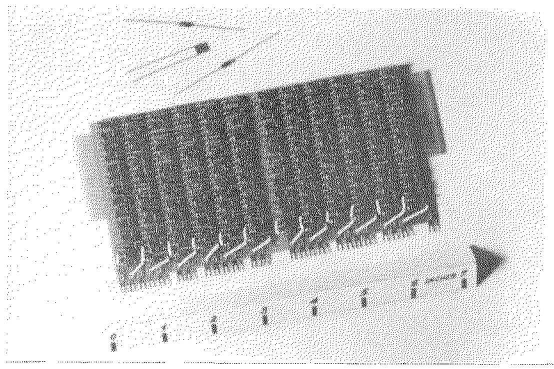

Figure 16—Diode-and-capacitor memory for pulse-code-modulation telephony.

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5 milliamperes per digit; the output signal of 100 nanoseconds duration is 2 volts and can be increased to 12 volts with a single stage of amplification. Operation from −25 to +70 degrees Celsius is within ±30 percent on voltages.

Laboratoire Central de Télécommunications
France

Long-Life Sensitive Switch—Capable of being operated by a force of only 0.5 ounce (15 grams), the *ES-4* switch can operate up to 300 times per minute for counting, signaling, or controlling equipment, such as the machine whose output it is monitoring. Life tests have exceeded 200 million operations.

Employing a torsion spring, the operating point may be adjusted from 24 to 59 degrees from the rest position and overtravel to a total of 69 degrees will be accommodated. The operating arm may be positioned and shaped to meet the specific application. See Figure 17.

ITT General Controls
United States of America

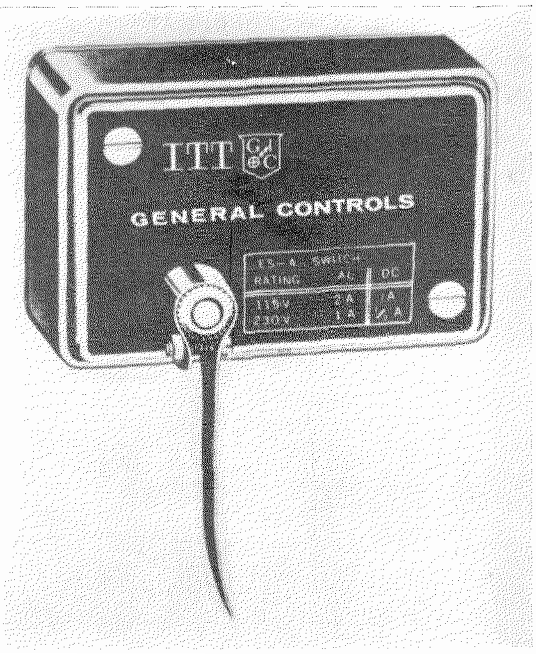

Figure 17—Sensitive switch.

Police Teleprinter Center in Hamburg—Recently installed in the Hamburg Police Headquarters is a combined rotary dial and push-button teleprinter switching center (Figure 18). Priority connections are provided for. Broadcasts may be made to several other stations with automatic calling of recorded numbers. Outgoing messages are assigned a number, date, and time automatically. A monitor teleprinter records a caller's telex number, identification, and time of call.

Standard Elektrik Lorenz
Germany

Magnetic Thin-Film Experimental Memory—An experimental magnetic memory employing thin films is shown in Figure 19. It accommodates 64 words of 16 bits each with a read-only delay of 60 nanoseconds and read-write or read-restore cycle times of 200 nanoseconds. The address decoding matrix uses 64 transistors and 8 word drivers. The memory matrix has a density of 200 bits per square inch (32 bits per square centimeter). The thin-film planes have been thoroughly tested for immunity to creeping.

The word current is 400 milliamperes and the digit current is 150 milliamperes. The minimum sense output is ±1 millivolt for a 20-nanosecond rise time of the word current. The typical digit noise is ±3 millivolts. All unwanted signals have been carefully balanced to

Figure 18—Teleprinter switching center in Hamburg Police Headquarters.

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reduce the dead recovery time of the sense amplifier.

Laboratoire Central de Télécommunications
France

Static Inverters—Rotating machines and vibrating switches are traditionally used for changing direct current into alternating current. Thyristors, in the form of silicon controlled rectifiers, are advantageously used as such inverters. They have no moving parts and require no maintenance.

Inverters may supply the load continuously or be on standby for automatic connection in case of mains failure.

The model shown in Figure 20 has both voltage and frequency stabilized within 1 percent. Its frequency is controlled by a 50-hertz multivibrator that can be synchronized from the alternating-current mains.

Standard Elektrik Lorenz
Germany

Equalizer for Subscriber Telephones—An equalizer has been developed for installation in existing subscriber sets using type *W 48* circuits. It is suitable for 0.5-millimeter (24 American Wire Gage) subscriber lines of up to 5.5 kilometers (3.4 miles) long having loop resistances up to 1000 ohms. The sensitivity of both receiver and microphone capsules must be within certain tolerance limits. Thermistors and diodes are used as the nonlinear elements.

In the transmit direction, the voltage for the control circuit is obtained from the microphone supply current. For loop resistance up to 500 ohms, full equalization of loss caused by feed-current variations is achieved in addition to partial equalization of line losses. For 500 to 1000 ohms it compensates for feed-current losses to maintain the transmitting reference equivalent at a constant value.

For the receive direction, the control circuit voltage comes from the loop current. Reference equivalent is maintained almost constant over the rated maximum line length.

Standard Elektrik Lorenz
Germany

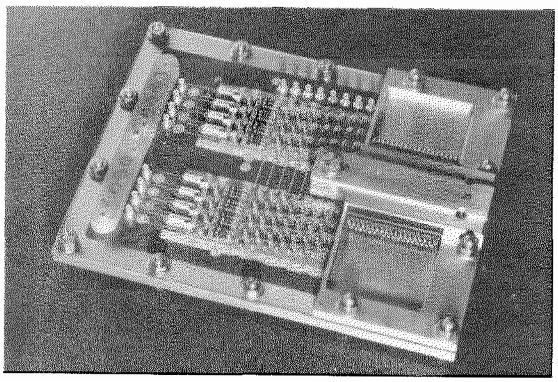

Figure 19—Experimental magnetic thin-film memory.

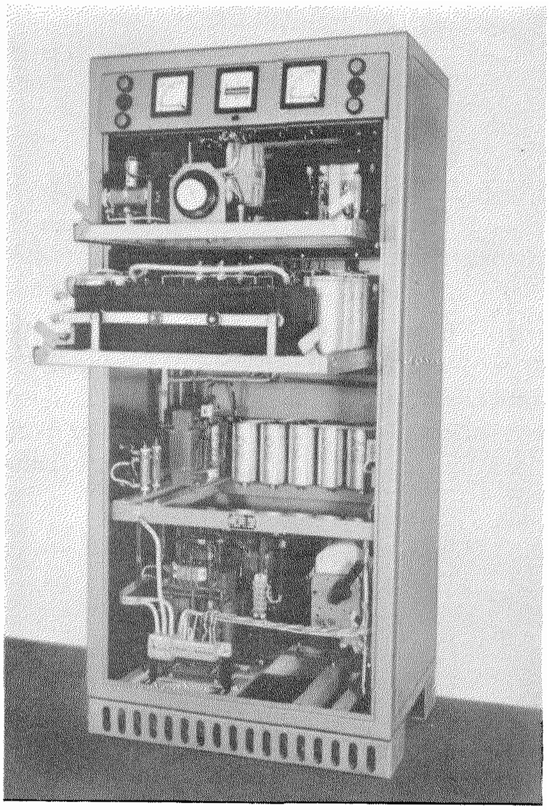

Figure 20—Static inverter using thyristors to change direct current into 50-hertz alternating current.

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Speech Quality of Pulse Code Modulation for Telephony—In developing an integrated telephone system in which both switching and transmission are done with pulse code modulation, it is necessary to demonstrate the effects of various operating conditions on speech quality.

In the equipment shown in Figure 21, the center box between the two attenuators and subscriber sets permits the number of pulse code bits to be changed from 1 through 7. The sampling rate also may be set at 8 or 10 kilohertz. The equipment can handle 24 channels.

With a 7-bit code, no degradation of speech is evident even over an input amplitude range of 36 decibels. This performance is aided by a digital compressor that is constant over a wide temperature range and is substantially independent of variations in component characteristics. It requires no matching in manufacture or service.

Laboratoire Central de Télécommunications
France

French Railroad Information Center—To handle telephoned inquiries from the public, the Société Nationale des Chemins de Fer Français recently placed in service at Pont Cardinet, Paris, the central information office shown in Figure 22.

A single telephone number gives access to 70/94 (initially equipped/ultimate capacity) local exchange lines. There are 72/96 operator positions among which incoming calls are automatically distributed. They are divided as follows: 40/55 for passenger travel in France and neighboring countries, 10/15 for long-distance international passengers, 2/2 for inquiries in foreign languages, 10/12 for freight matters, and 10/12 for requests for transport of luggage from passengers' premises to their trains.

Two monitor positions have signals to indicate the free or busy status of each operator position and if the answer to a call has been delayed. The monitor may listen to a conversation and may assist by entering into it or by speaking with the operator without the outside caller hearing them. These latter conversations are automatically recorded on magnetic tape.

In addition, there are two listening positions that have the same listening facilities as the monitor positions. They record simultaneously on graphic registers the traffic to 20 information positions.

Compagnie Générale de Constructions Téléphoniques
France

Award for Quality of Components for Pershing Missile—A special award for outstanding quality performance was presented at the annual convention of the Association of the United States Army to ITT Industrial Products Division by The Martin Company, the prime contractor for the Pershing missile. It signified best performance among 80 suppliers delivering major functional items and is part of a

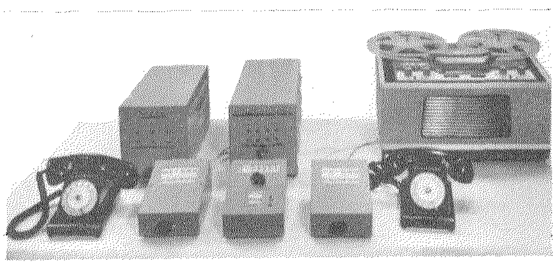

Figure 21—Equipment for demonstrating speech quality for pulse-code-modulation telephony.

Figure 22—Information center of the French railroads.

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program to achieve zero defects in the Pershing missile.

ITT Industrial Products Division
United States of America

Pentaconta Private Automatic Exchanges for Carrier Networks—In addition to the large-capacity telephone exchanges developed to meet the communication requirements in operating the electric-power distributing grids of Électricité de France, two smaller-capacity equipments have been developed.

The medium size is suitable for tandem switching and has capacity ranging from two 4-wire interexchange lines and twenty 2-wire local extensions to ten 4-wire interexchange lines and four 2-wire local extensions. Local communication is possible. It may have 1 or 2 registers and 2 to 4 junctors. The smaller size, suitable for terminal stations, is of the all-relay type with 1 incoming line and 10 local extensions that cannot be interconnected.

Operable with either 2- or 4-wire circuits, they are suitable for transmission by carrier over high-voltage power lines or over other carrier systems. Tolerance of wide climatic conditions is evident from installations in France, Italy, Spain, Chile, Venezuela, Morocco, Tunisia, and the Sahara.

Compagnie Générale de Constructions Téléphoniques
France

Printed-Circuit-Board Connectors—A new multiple-contact male connector having a service life of more than 10 000 insertions, a wide temperature range of operation, and high insulation has been developed for use with the female connectors that are part of the ITT Standard Equipment Practice.

With standard contact spacing of 2.54 millimeters (0.1 inch), the terminals extend through the board for access to both sides and permit either soldered or wrapped connections. Connectors have 11, 25, or 33 contacts, and units of different sizes can be fitted to large boards.

The brass contacts are first silver plated and then gold plated. They are molded in plastic with chamfered guide projections for protection and to assure alignment with the floating female contacts. A mechanical keying arrangement permits only the intended male and female assemblies to be mated to each other. Examples are shown in Figure 23.

Bell Telephone Manufacturing Company
Belgium

Thermistor for Television Deflection Circuit—Thermistor *KS 46 W* has a negative temperature coefficient that is matched to the change in resistance with temperature of the vertical deflection coil of a television receiver. A resistor is not needed in parallel to the thermistor.

Standard Elektrik Lorenz
Germany

Pumps for Chemical Processing—Over a hundred electrically driven pumps have been delivered to a new chemical plant in Durgarpur, India.

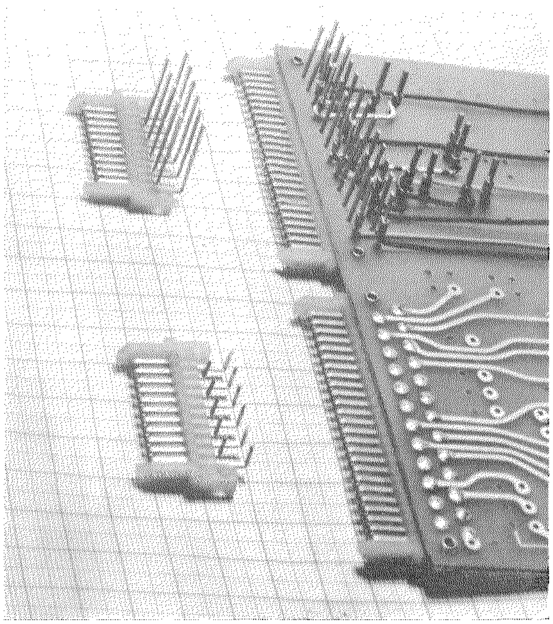

Figure 23—Male connectors for printed-circuit boards.

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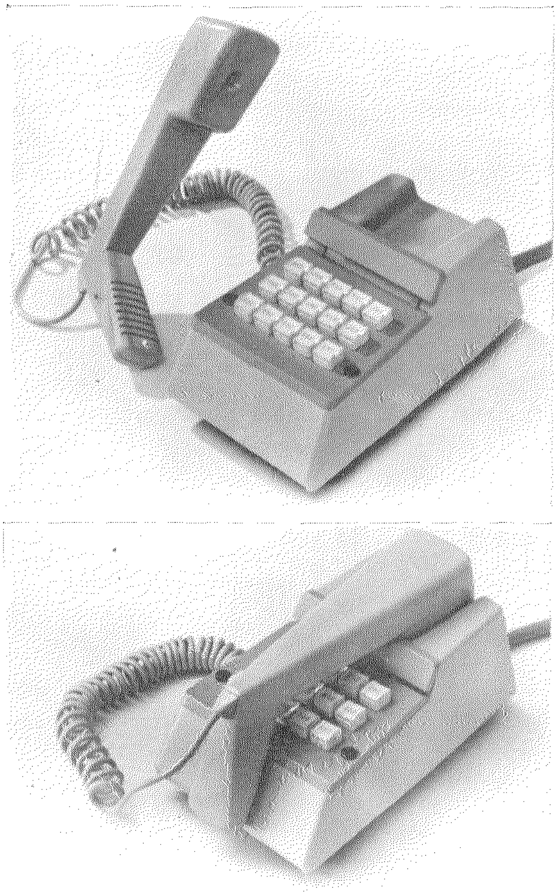

Figure 24—Lightweight push-button Deltaline intercommunication set.

For driving heating fluids at 380 degrees Celsius at flow rates of 130 cubic meters (4500 cubic feet) per hour to a height of 60 meters (200 feet), the pumps are of cast steel with cooled mechanical differential seals. For corrosive fluids, suitable grades of stainless steel are used. For naphthalene, heating envelopes are fitted to ensure fluidity, particularly during start-up.

Le Matériel Téléphonique
France

Lightweight Push-Button Intercommunication Set—The Deltaline Interphone is a new internal telephone system accommodating up to 15 extensions with push-button calling. A warbling tone electronically generated within the telephone replaces the conventional bell.

The handset weighs only 4 ounces (113 grams) and has a concealed speaking tube leading to the microphone; the microphone and earphone are in one piece, close to the ear. A set is shown in Figure 24.

Standard Telephones and Cables
United Kingdom

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Microwave Radio System for Multichannel Telephony and Television in the 6-Gigahertz Range
Part 1—Basic Features

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1. Introduction

Standard Telephones and Cables and Standard Elektrik Lorenz have been active in the field of radio relay systems since the 1930's, when initial installations were made of narrow-band systems. After World War 2 both companies contributed to the development of wide-band microwave systems, and in 1952 Standard Telephones and Cables installed the first commercial system that used a traveling-wave amplifier. This system is still transmitting monochrome television between Manchester and Edinburgh. By the late 1950's both companies had independently designed 4-gigahertz wide-band radio relay systems to transmit either frequency-division multiplex of up to 960 telephone channels or a television picture signal, with or without sound and color. These systems have been supplied and installed in many countries throughout the world, providing national and international circuits that meet international standards.

As the demand for traffic capacity continued and the 4-gigahertz band became increasingly congested, the International Radio Consultative Committee established the basic frequency plan and operating parameters for radio relay systems in the 6-gigahertz band. It provides for 8 radio channels, each capable of accommodating up to 1800 telephone channels or a television program. Large systems of this type require independent circuits for remote control and supervision of the equipment, and independent radio channels have been allocated for this purpose in the 6-gigahertz band. Bell Telephone Manufacturing Company of Belgium has had long experience in the development and manufacture of smaller-capacity systems suitable for this application.

In 1961, all 3 companies joined in developing a new 6-gigahertz system that accommodates 1800 speech channels or 1 television channel for home and export markets. The design criteria were very-good reliability, low initial and maintenance costs, and low operating costs.

This paper describes the basic features of the *1800/TV* system for operation at 6 gigahertz.

2. Impact of Solid-State Techniques

The design of communication equipment as a whole is presently taking a tremendous step forward, characterized by the fact that semiconductors are superseding thermionic cathodes and solids are replacing evacuated bodies. This change from vacuum to solids is accompanied by a considerable reduction in space requirements, not only because of the reduction in size of the semiconductor, but also because of the possibility of different mechanical arrangements. As elevated temperatures are no longer necessary to produce controllable electron streams, the power consumption and, more important, the dissipated power, are reduced considerably. This reduction of dissipated power permits more-compact packaging.

The electron emission from red or yellow hot bodies, which takes place in vacuum tubes, is always accompanied by a certain consumption of cathode material. Vacuum tubes, therefore, necessarily have a finite life. No such consumption takes place in semiconductors, and this makes for much-greater reliability. In addition to size, dissipation, and reliability, effects such as voltage-dependent capacitance in semiconductors give rise to new ways of signal processing.

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In recent years, semiconductors have slowly entered the radio-link field. In fact, transistors and diodes are already used in certain areas of the existing 960-channel systems where the product of power and frequency squared is low. When the development of the 1800-channel system started, tubes were still considered mandatory for the signal path and for microwave generation, but it was realized early that such a conservative design would be obsolescent by the time the system was completed.

With the decision to introduce solid-state devices into the signal path, many problems arose. These were aggravated by the fact that the capacity of a single broad-band channel was being increased from 960 to 1800 telephone channels. The first step was to use transistors in the baseband amplifiers. The design of resistance-capacitance-coupled amplifiers with a bandwidth of 10 megahertz, flat within 0.1 decibel, is quite straightforward; the only difficulties to be solved were the output stages, which must deliver adequate voltage drop across low impedances with sufficiently low distortion.

The next step was to use transistors in the 70-megahertz intermediate-frequency amplifiers. It took much study to learn that it is not sufficient to use transistors with an f_T of 7 or 10 times the working frequency, but that a determining factor with respect to bandwidth and temperature stability is the backward attenuation or the time constant $r_{bb}' C_c$. These parameters until recently were not even specified in most transistor data sheets. The requirements of the system included low-noise input stages, gain regulation without introducing prohibitive intermodulation or thermal noise, the correct choice of interstage coupling, and output stages that drive power mixers or discriminators. How these requirements were met is discussed in a separate paper [1].

The last step in adapting very-high-capacity microwave radio equipment to transistors, at least for the next several years, is to replace the vacuum-tube circuits in the primary microwave power sources by all-solid-state circuits. This appears to be not only the most-advanced step but also the most difficult. Some competitive designs, in fact, use semiconductors only in the frequency-stabilizing circuits, while tubes are still used for the generation proper. In this part of the equipment, namely the local-oscillator chains, not only are transistors with the highest product of power times frequency-squared used, but variable-capacitance diodes are used as passive frequency multipliers.

For these diodes, theories are available to predict the efficiencies obtainable, but practical results fall considerably short of theory where the power limit is concerned. The same discrepancy between theory and practice applies to the thermal noise in the oscillator chains. It took considerable investigation before designs were achieved that met the requirements. Several approaches to the problem were pursued and one is discussed in an associated paper [2].

The heading "solid state" includes not only semiconductors but also other solid materials having inherent electron and field interactions. For instance, ferrite elements combining high-quality material and precision manufacturing are employed in this system to a greater extent than ever before and help to improve both performance and layout.

3. Functional Specifications and Performance

A radio-link system performs the same function as a coaxial cable system; it accepts signals, transmits them over hundreds or thousands of kilometers, and delivers them to the receiving points without changing them by more than specified amounts. The admissible signal distortion, levels, and other important requirements are well covered by various recommendations [3]. The 6-gigahertz band provides a maximum of 8 wide-band signals, each of which may be either the frequency-division multiplex of 1800 (or 1860) one-way telephone channels or a one-way television signal.

The 1860-telephone-channel signal, consisting for example of one supergroup (60 channels)

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and two super-master groups (900 channels each), covers frequencies from 60 to 8204 kilohertz. The video band of a 625-line picture extends from virtually zero to 5 megahertz, with a sound carrier at 7.5 megahertz. In both cases a continuity pilot at 9.023 megahertz is required, so that a baseband 10 megahertz wide must be transmitted over the system.

In a modulator equipment this baseband signal is amplified, pre-emphasized, added with or without a pilot to a sound carrier, and frequency modulated on the 70-megahertz intermediate-frequency carrier. This signal, having a bandwidth of about 40 megahertz, is passed to the transmitter section of the radio-frequency equipment, where it is amplified, amplitude-modulated on the radio-frequency carrier, single-sideband selected to retain the bandwidth of 40 megahertz, amplified, and radiated from the antenna.

At the next station the received signal is reconverted to the 70-megahertz intermediate frequency and amplified. This station retransmits the signal to the following station at a different radio frequency, and so on over as many links as necessary. In the final station the receiver output is demodulated, the baseband is recovered by a discriminator, de-emphasized, and amplified to the specified level. Figure 1 shows one direction of this signal flow for a single channel. A completely equipped system requires 8 of these channels operating independently in each direction. Common antennas may be used.

A communication system, however reliable, can still be improved considerably by the use of automatic channel switching. It is therefore advisable to have one or two channels on standby, to which the traffic can be switched in the rare event of failure of a working channel. Switching may be done at the baseband or at the intermediate frequency; the first has the advantage that the modulator and demodulator are also protected, the second permits protection sections to be chosen independently of modem sections (see Figure 1). Either switching system can be supplied with the 6-gigahertz 1800-channel system.

In using one standby channel to protect several working ones, as is recommended for the 6-gigahertz system, information must be communicated from one end of a protection section to the other to initiate the switchover. In addition, it may be necessary to provide an order wire for maintenance, remote control, and communication. An auxiliary radio-link equipment has been developed for the 6-gigahertz band and can be supplied with the 1800-channel system to provide these requirements [4].

The 6-gigahertz radio relay system as presently discussed follows recommendations 383 and 389 of the International Radio Consultative Committee frequency allocation plan [3]. This plan is based on a module $M = 14.8259$ megahertz, the center frequencies in the lower half of the 5925-to-6425-megahertz range being odd multiples and those in the upper half being even multiples of M, leaving a gap of $3M$ in between.

At any station, transmitters are always in one half of the band and receivers in the other half. This arrangement ensures that third-order mixing of the high transmitter powers, which may

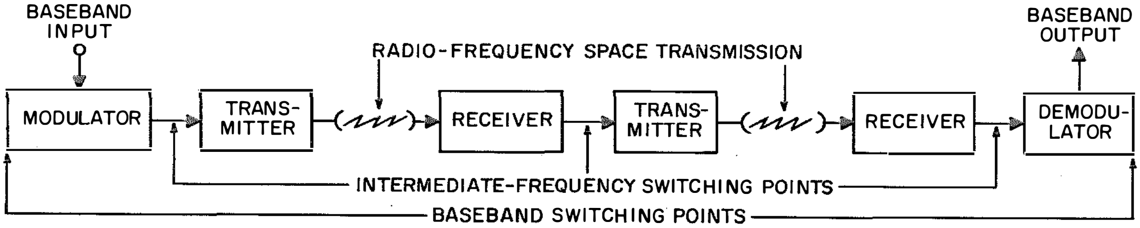

Figure 1—Diagram of a wide-band radio-link system.

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occur in a common antenna or even outside the antenna, cannot interfere with the weak received signals. There are other interfering signals that must be taken into account, for example those at image frequencies, oscillations from transmitters that imitate the local oscillators of receivers, and multiples of receive-transmit-shift frequency.

The occurrence of interfering signals is also intimately connected with the frequency plan used. The international recommendations [3] permit a choice between two intermediate-frequency center frequencies, namely 74.13 (5*M*) or 70 megahertz. In the latter case some of the selectivity requirements are of the order of 50 decibels more stringent than in the former. However, 70 megahertz is preferred by most European administrations because the same modems are used for a number of radio relay systems and because of existing switching and test equipment. This frequency has therefore been adopted despite the greater difficulties.

The use of 6-section filters in the branching arrangement, antennas with a high degree of polarization decoupling, and separate generation of all local-oscillator frequencies make it possible to accommodate 4 transmit and 4 receive channels on the same antenna. The antennas play an important part in preventing cross-channel interference. A separate paper [5] deals with the design of new microwave antennas.

A basic international noise requirement for long-haul communication systems is that the psophometrically-weighted mean power shall not exceed 3 picowatts per kilometer (4.8 picowatts per mile) in any hour for any telephone channel. The *1800/TV* system is so designed that half this power is independent of load, as it is of thermal origin, and the other half is added by intermodulation at full 1800-channel loading. With the design figures of 140 kilohertz root-mean-square unpre-emphasized channel deviation, International Radio Consultative Committee recommended pre-emphasis, 10-watt transmitter power, and 10-decibel receiver noise figure at deep-fading conditions (14.5 decibels at nonfading condition, including transmitter and local-oscillator noise), the thermal share of the requirement is met with 3 to 4 decibels of system margin (that is, simultaneous fading margin) for a 62-decibel total path loss between the transmitter-power and receiver-noise-figure reference points.

To meet the intermodulation part of the noise requirement (plus other requirements such as baseband-to-baseband response and differential gain and phase for television transmission) in the presence of the above-mentioned selectivity requirements, not only did each unit have to be designed for optimum flat transmission characteristic, but also an effective equalizer had to be developed and applied to the system.

If the same system is loaded with only 960 channels, the improvement in thermal and intermodulation noise performance makes it likely that the 6-gigahertz equipment will be suitable for an intercontinental communication system with reduced picowattage per kilometer. Such a system will be discussed by the International Radio Consultative Committee.

4. Equipments

The *1800/TV* main system consists of the radio equipment with its associated channel branching filters, the modem, and the protection switching equipment.

The system operates with the signal through-connected at intermediate frequency in repeater stations. The equipment design, however, is such that the same radio equipment is used at terminals and repeaters. The mechanical embodiments of the system developed by Standard Telephones and Cables and by Standard Elektrik Lorenz are not identical because of local requirements. Each embodiment has been designed to produce an integrated system; the equipments have similar appearance and dimensions and use common components and mechanical parts wherever possible. In both cases the mechanical design, in particular the radio-

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frequency channel filters for multiple use of antennas, allows for gradual extension of the system from 1 or 2 to the full number of available radio-frequency channels. All the equipments may be operated from batteries or from alternating-current mains.

4.1 Radio Equipment

On the transmit side of the radio equipment, the intermediate-frequency signal is converted to radio frequency and then amplified by a traveling-wave tube. On the receive side, conversion takes place from radio to intermediate frequency. Subsequent signal amplification in the receiver provides the level required for through-connection to the associated transmitter at a repeater station or to the demodulator at a terminal. The radio-frequency channel branching arrangement for each transmitter and each receiver is part of the radio equipment.

As illustrated in Figure 2*A*, the received power goes from the branching circulator and the channel filter through an isolator into the receiver mixer, which is supplied with local-oscillator power from a temperature-controlled

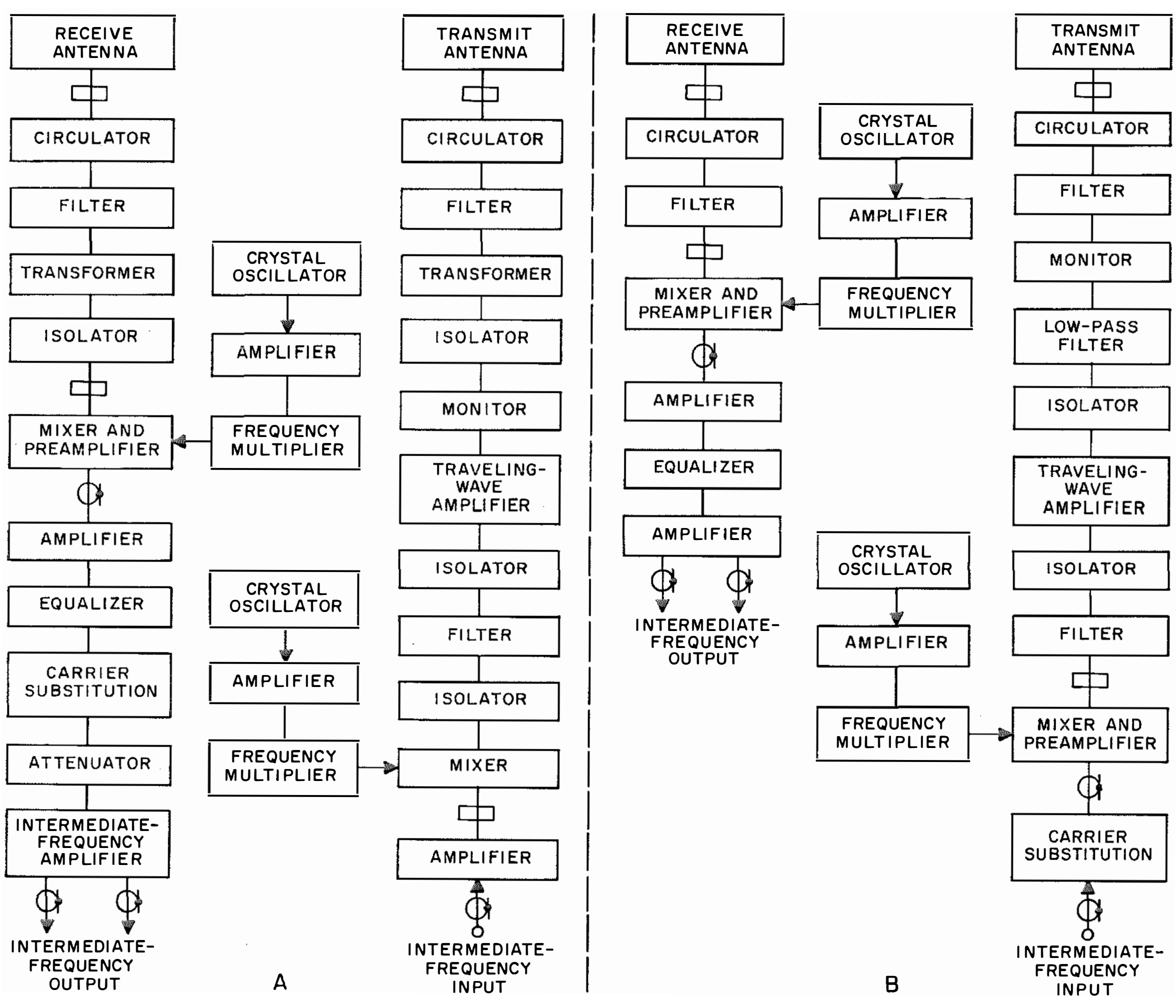

Figure 2—Diagram of the *1800/TV* radio equipment. *A* is the Standard Elektrik Lorenz version and *B* is the Standard Telephones and Cables version.

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crystal oscillator, transistor amplifier, and varactor-multiplier chain. The intermediate-frequency signal passes through the preamplifier and the main amplifier, automatic gain control being applied to both. The carrier-substitution circuit interrupts the transmission path in case of a fault and replaces the signal by an auxiliary 70-megahertz carrier that is generated in it. This unit is followed by an amplifier that provides two equal output signals. (Figure 2*B* shows the carrier-substitution circuit incorporated in the transmitter section.)

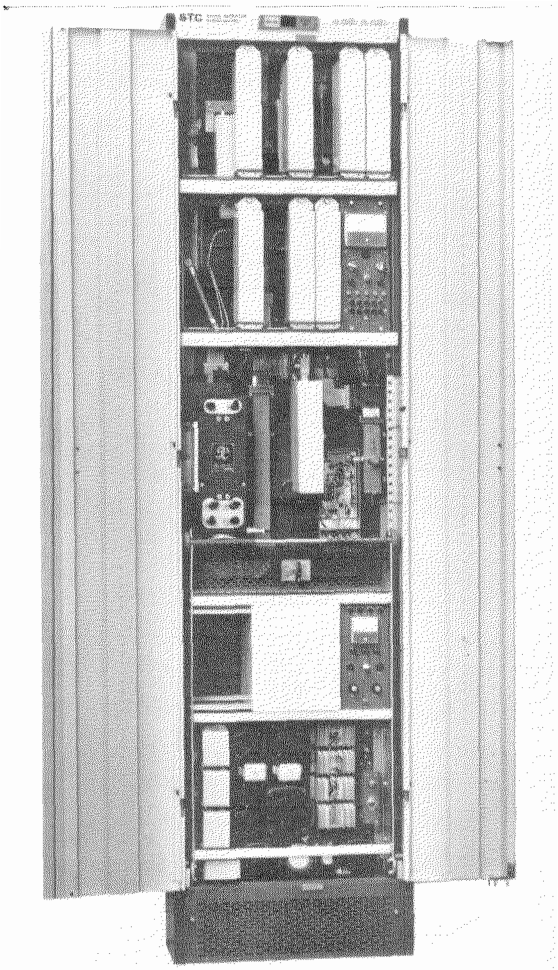

Figure 3—Standard Telephones and Cables cabinet housing 1 transmitter, 1 receiver, and the branching filter arrangement. The cabinet is 206.4 × 52 × 22.5 centimeters (81.3 × 20.5 × 9 inches) in size.

Returning to Figure 2*A*, in the transmit section the intermediate-frequency power amplifier drives the mixer for the signal conversion to radio frequency. Local-oscillator power is again derived from a crystal-controlled oscillator with subsequent amplification and multiplication by a varactor chain. The radio-frequency signal passes through an isolator for image matching, a filter, and another isolator. It is then amplified in the traveling-wave tube, from where it passes through a further isolator and the branching assembly to the antenna.

The radio equipment has been designed using solid-state techniques with the exception of the transmitter power stage, for which a traveling-wave amplifier is employed. Silicon transistors and diodes are used almost exclusively for stability at high temperatures. The solid-state technique made it possible to reduce equipment dimensions considerably compared with earlier vacuum-tube equipment, so that the required floor space for a station has been sharply decreased.

Power dissipation in the radio equipment has also been reduced, so that forced air cooling and a separate blower are not required. The major heat generators are the transmitter power supply, the traveling-wave amplifier, and the high-power transistor stages in the local-oscillator chains. The transmitter power supplies have been designed for high efficiency to reduce dissipation. The traveling-wave amplifier, which is focused by periodic permanent magnets, and the local-oscillator chains are cooled by conduction and convection.

Except for the traveling-wave amplifier, all active units in the transmitter are part of the intermediate-frequency transmission chain that uses solid-state techniques exclusively. The realization of this amplifier chain is covered in detail by a separate article of this series [1].

The progress of solid-state semiconductor techniques makes it possible to have separate oscil-

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lator chains in transmitter and receiver. This means that the same equipment arrangement is used for terminal and repeater stations. Information on the local-oscillator chains is given in a separate article of this series [2].

Wide-band passive components in the radio-frequency transmission path include the isolators and the circulators. They have been designed for the full bandwidth from 5925 to 6425 megahertz and are manufactured by precise methods that guarantee high accuracy and uniformity in production. In one version electroforming is used for this purpose. The insertion loss of an isolator is less than 0.25 decibel, the reverse attenuation more than 30 decibels. Similarly, the forward loss from port to port of a circulator is considerably less than 0.1 decibel, the decoupling more than 30 decibels. All filters are made of plated invar material to reduce temperature effects.

The two versions of the radio equipment are shown in Figures 3 and 4. The cabinet shown in Figure 3 houses one transmitter and one receiver including branching filter arrangement. The upper framework includes the local-oscillator chain, the units of which are installed in hinged containers. Below these are the units of the intermediate-frequency chain. Above the power supply at the bottom of the cabinet, part of the waveguide structure (Type *R70* of the International Electrotechnical Commission) and the traveling-wave amplifier are visible.

In Figure 4 two transmitters and two receivers are mounted in a vertical structure. The channel branching filters are mounted above the cabinet. Inside the cabinet flat-type waveguide, (Type *F70* of the International Electrotechnical Commission) is used. Instead of housing two radio equipments side by side as shown in the illustration, 1 radio equipment and 1 modem may be mounted to provide a complete terminal within one cabinet.

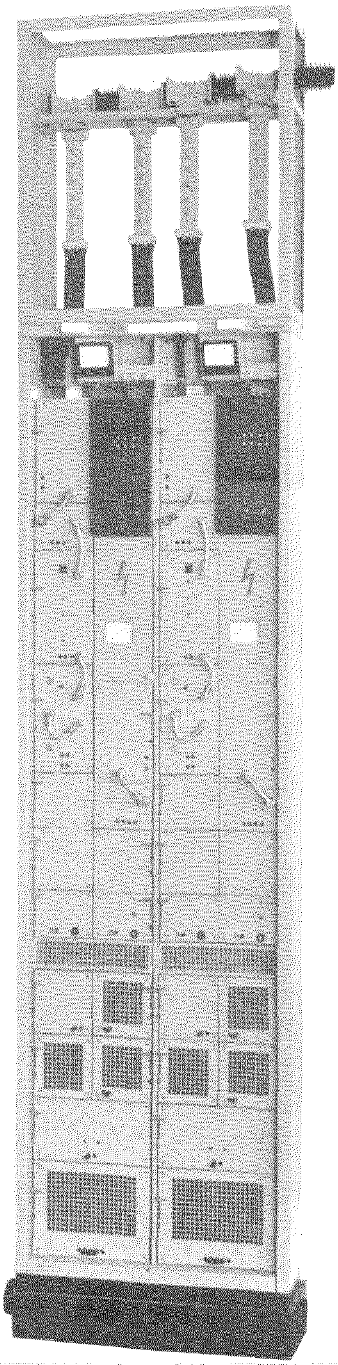

Figure 4—Standard Elektrik Lorenz cabinet housing 2 transmitters and 2 receivers. The cabinet is 206.4 × 60 × 22.5 centimeters (81.3 × 23.6 × 9 inches) in size.

4.2 Modem Equipment

The modulator unit includes a baseband amplifier, a pre-emphasis network, and the frequency

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modulator, the center frequency of which is stabilized by automatic frequency control. The modulator output supplies a frequency-modulated signal at intermediate frequency for through-connection to the radio transmitter.

The demodulator unit consists of an intermediate-frequency amplifier, limiter, discriminator, and baseband amplifiers. The points of injection and extraction of the continuity pilots depend on the type of protection switching used.

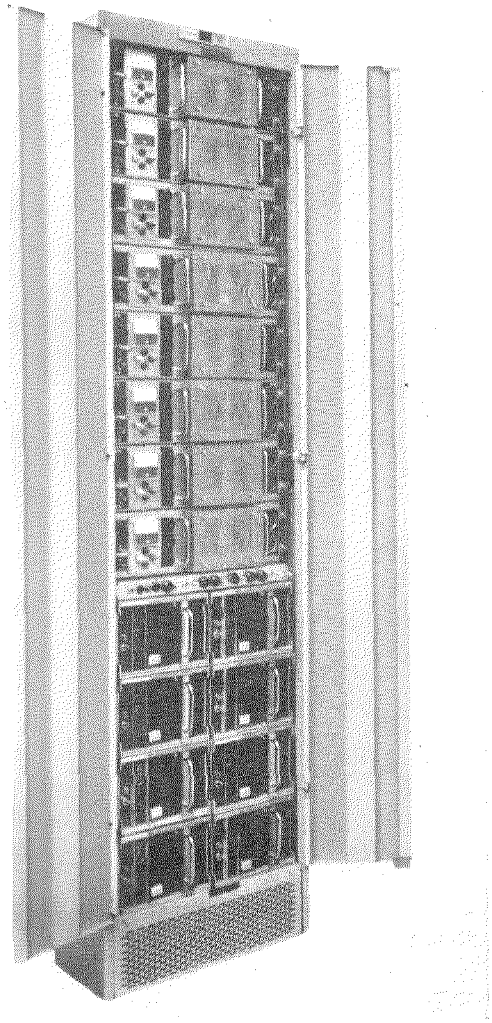

Figure 5—A cabinet the same size as in Figure 3 houses up to 4 complete modem equipments.

With a baseband switching system they are incorporated in the switching equipment, whereas with an intermediate-frequency switching system they are incorporated in the modem equipment.

In contrast to the well-known klystron heterodyne type of large-capacity modulator heretofore used that requires complex and expensive design, the type chosen is a variable-phase-shift direct frequency modulator. This considerably reduces space and power consumption. Figure 5 shows that a cabinet of the same size as that in Figure 3 houses up to 4 complete modem equipments, each consisting of 1 modulator panel, 1 demodulator panel, and 2 identical power panels. The modulator and demodulator panels are interchangeable mechanically so that various combinations may be fitted up to a maximum of 8 panels. Apart from 6 vacuum tubes in the modulator panel and 4 in the demodulator panel, transistors and diodes are used throughout as active elements.

Electron valves were included in the modem as an interim design until solid-state devices became available that would meet the stringent linearity and level requirements of the *1800/TV* system; recent developments show that this can now be done. Figure 6 shows a subrack mounting a modulator and a demodulator. The wide-band modulator is of the push-pull heterodyne type, using 250 and 320 megahertz

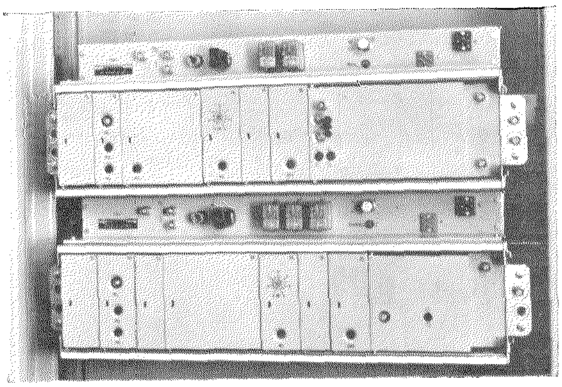

Figure 6—Subrack mounting a solid-state modulator and a demodulator.

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as oscillator center frequencies. The slope of the modulator characteristic as well as of the demodulator characteristic deviates from a constant figure by less than 1 percent over the band of ±12 megahertz referred to the center intermediate frequency. Four such modems, all fully independent regarding mounting and cabling, can be accommodated in one cabinet, with additional space available for ancillary units such as sound modems for television transmission or for recording facilities.

4.3 Protection Switching Equipment

Recommendation 305 of the International Radio Consultative Committee [3] proposes one or more standby channels in addition to the operating channels of a radio relay system. In case of a failure, traffic is automatically switched to these standby channels. The protection switching equipment has been designed for this purpose. Either 2 standby channels can be provided for up to 6 operating channels or 1 standby channel for up to 7 operating channels. It is also possible to allocate 1 standby channel to some of the operating channels and the other standby channel to the remaining channels. When not needed to replace operating channels, the standby channels can be used temporarily to transmit less-important programs or for long-term measurements, either of which can tolerate interruption if switching takes place.

The protection switching equipment monitors the operating channels on the receive side with regard to pilot level and noise. The switching criterion is derived in a logic system and converted into the switching command for the transmit side at the other end of the switching section. The command signal is transmitted through a separate line or channel, for example by means of the auxiliary system mentioned in Section 5. The local logic at the transmit side actuates parallel connection of the standby channel and the faulty operating channel. The standby switching at the receive end then follows.

Protection switching may be performed at baseband or at intermediate frequencies, depending on operational requirements. Figure 7 shows the baseband switching equipment that uses solid-state devices exclusively, printed wiring, and fast-acting reed switches in the transmission path. All the transmit and receive baseband switching devices for 8 channels are mounted in one cabinet of the same front area as the radio bay but twice as deep. Switch transit time is less than 3 milliseconds. For a system having more than one working channel, the total switching time is less than 35 milliseconds.

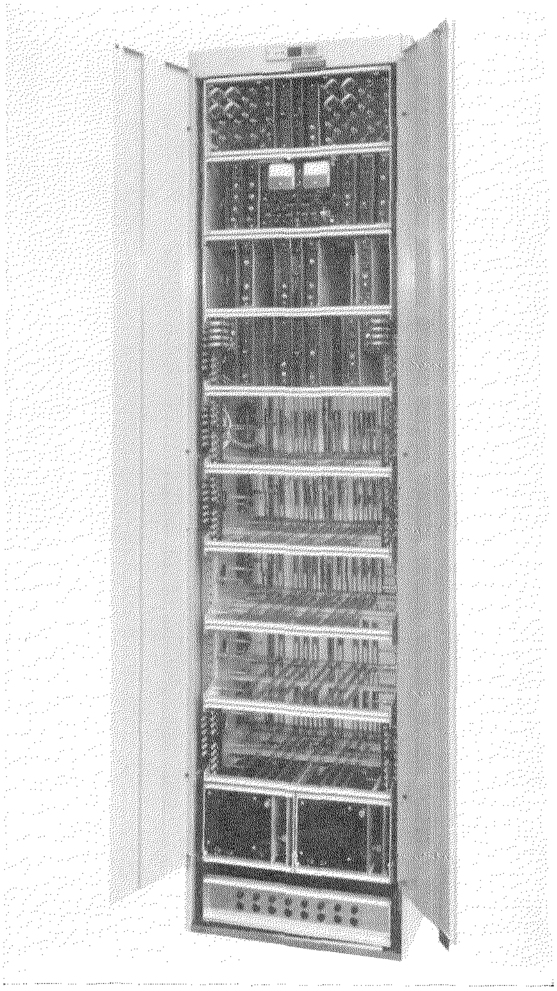

Figure 7—Baseband switching equipment.

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On long television routes, which may be divided into several switching sections without demodulating to baseband, or which require the dropping of radio channels at switching points, intermediate-frequency switching may be preferred. Transfer time is less than 10 microseconds with the use of intermediate-frequency diode switches. Figure 8 shows a 7-way selector using diode switches that is part of the intermediate-frequency switching equipment.

5. Auxiliary Radio System

In the operation of a microwave radio system, rapid communication between stations, both for speech and switching information, is essential. For protection in the event of failure of the main system, the means for providing these auxiliary circuits should be as independent as possible. There is often no alternative means of providing circuits other than along the microwave route, and for this purpose the auxiliary radio-link system has been developed.

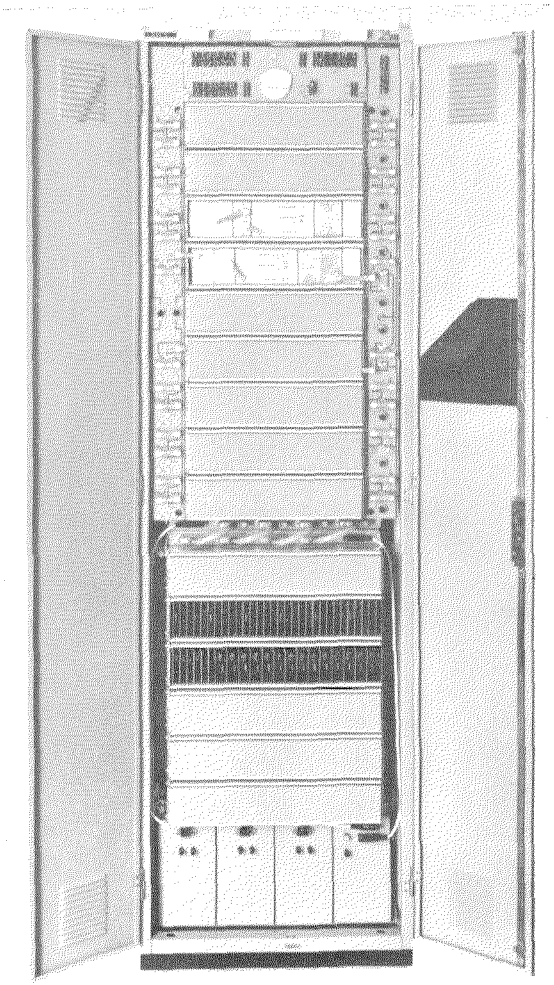

Figure 8—Intermediate-frequency switching equipment.

It operates over the same antenna and branching system as the main system and on frequencies in the same band, but in all other respects may be made completely independent. It provides up to six 4-kilohertz channels that may be used either as speech circuits or to provide narrow-band telegraph circuits for transmission of instructions for protection switching systems. The system is of the direct-modulation type employing phase modulation, and the signal is demodulated at each repeater since access to the baseband is required at these points.

The equipment employs all-solid-state techniques and, since the reliability of this system is vital to the successful operation of the main protection switching systems, this aspect has been given particular attention.

A description of this system is given in a separate article of this series [4].

6. Conclusion

This paper discusses the principal problems and steps that have been taken to successfully achieve the design of a 6-gigahertz system with solid-state devices to transmit 1800 simultaneous telephone conversations or a television program. Vacuum tubes have been replaced in all applications apart from the microwave power amplifier and an interim modem equipment. This has led to a considerable reduction in size and power consumption and is expected to improve reliability substantially.

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Dr. Müller has published articles on electron optics, magnetic circuits, waveguide circuits, tunnel diodes, and distortion theory.

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Microwave Radio System for Multichannel Telephony and Television in the 6-Gigahertz Range
Part 2—Solid-State Microwave Power Generators

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1. Introduction

For economy of space, reduced power consumption, lower cost, and increased reliability, solid-state devices are now replacing thermionic valves in practically all telecommunications equipment.

One of the areas least tractable to this progress is the generation and amplification of microwave power, but a major advance has been achieved by using a chain of varactor multipliers driven from a transistor oscillator to provide a microwave power source.

This article discusses the design requirements and the realization of these solid-state sources for use as local oscillators in the *1800/TV* radio relay system.

2. Design Requirements

In a repeater equipment two microwave generators are required. One operates at a power level of about 50 milliwatts and is used to convert the intermediate frequency to the required super-high transmitted frequency. This is referred to as the "transmit local-oscillator chain." The other generator operates at a power level of about 2 milliwatts and its output is mixed with the received signal to produce the required intermediate frequency. This is referred to as the "receive local-oscillator chain."

With the exception of power level, the performance requirements for both chains are identical and are very similar to those of a klystron or a coaxial-line valve that is frequency controlled to within ±100 kilohertz. The parameters to be specified for a multiplier chain are the output power, frequency stability, spurious frequency modulation, and the level of unwanted harmonics.

2.1 Power Levels

The transmit local-oscillator chain must provide the power to convert the intermediate frequency to super-high frequency. The power level required mainly depends on the required transmitted output power, the gain of the traveling-wave amplifier, and the conversion loss of the mixer. In the system under consideration, the minimum gain of the traveling-wave amplifier for the required output of 10 watts (+40 decibels referred to 1 milliwatt) is 36 decibels. The conversion loss of the mixer is 10 decibels. If allowance is made for the losses in the circuit that supplies the traveling-wave amplifier, a local-oscillator level of 32 milliwatts (+15 decibels referred to 1 milliwatt) is required at the mixer crystal. A circuit employing a circulator supplies the local-oscillator signal to the mixer, this method having the advantage that it minimizes the power level to be provided by the local oscillator. To allow for circuit losses and variations, the output from the multiplier chain was designed to be 50 milliwatts (+17 decibels referred to 1 milliwatt).

The receive local-oscillator chain, as previously stated, supplies power to the input mixer crystal. The local-oscillator level required at the mixer for optimum noise figure is 0.8 milliwatt (−1 decibel referred to 1 milliwatt). The input signal and the local-oscillator signal go to the input mixer through a circulator. To allow for circuit losses and aging, the chain should be designed to provide a power level of 1.6 milliwatts (+2 decibels referred to 1 milliwatt).

2.2 Frequency Stability

The inherent frequency stability of a multiplier chain depends on the fundamental frequency and the frequency stability of the oscillator.

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A high degree of stability can be produced by using a crystal reference, but this limits the frequency of the present oscillator circuit to less than 100 megahertz. The choice of frequency is most important to prevent interaction between the oscillator and the intermediate frequency from seriously deteriorating the system performance; frequencies in the range of 70 ± 10 megahertz must therefore be avoided. Although crystals above 80 megahertz are available, it is more difficult to produce a highly stable low-noise oscillator at these higher frequencies. It was therefore decided to use a range of oscillator frequencies below 60 megahertz. The range of local-oscillator frequencies to be covered is from 5875 to 6475 megahertz and this can be covered by using fundamental frequencies within the range of 46 to 51 megahertz with a multiplication factor of 128.

Using this multiplication factor, the frequency stability of the fundamental oscillator can be determined, being the allowable output frequency variation of ± 100 kilohertz divided by 128, requiring an oscillator stability of ± 15 parts per million.

2.3 Spurious Frequency Modulation

Any spurious frequency modulation on the oscillator signal causes noise or interference in the baseband, depending on the character of the modulation. Taking other parameters of the system into consideration, a reasonable noise allocation for the two local oscillators in a repeater on an 1800-channel system is about 20 picowatts, measured psophometrically, in any baseband channel between 300 kilohertz and 8.2 megahertz. Assuming an equal contribution, this means that each chain contributes 10 picowatts. Noise in the carrier can produce phase modulation when passed through an amplitude-limiting network. When detected by a frequency discriminator, it will give a noise level rising at 6 decibels per octave, the deviation being increased by the magnitude of the frequency multiplication after the introduction of the noise. Therefore any noise deviation on a 50-megahertz signal is increased by 42 decibels if it is multiplied 128 times to 6400 megahertz. This noise problem greatly influences the selectivity and the power levels in various stages of the chain.

Interference from external sources can be picked up on the supply leads or within the circuit itself and can produce a spurious modulation on the local-oscillator chain. This type of interference determines the degree of decoupling required in the supply leads and also the mechanical design required to provide effective screening. The system specification requires that a noise limit of 10 picowatts not be exceeded with the equipment subjected to a field of 1 volt per meter at any frequency from 10 kilohertz to 1 gigahertz. This limit must also be maintained if there is injected in the supply leads 100 millivolts root-mean-square from 50 hertz to 20 megahertz, or 1 volt root-mean-square from 20 to 300 megahertz.

2.4 Harmonic Levels

A limit must also be placed on the level of the sidebands about the output signal produced by other harmonics of the fundamental signal. The permissible level depends on the frequency plan used for the system as well as other system parameters. The acceptable limit for the 6-gigahertz *1800/TV* system is 75 decibels for sidebands occurring at 100 megahertz and below, and 100 decibels for those above.

3. Design of Local-Oscillator Multiplier Chains

3.1 Transmit Local-Oscillator Chain

This chain is divided into three parts, the division being determined by the maximum frequency to which a certain technique is limited. The first part of the chain consists of the oscillator and the drive section, which uses transistors and varactors and has an output frequency of 100 megahertz, a practical economic limit for transistors at the required

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power levels. The second part consists of varactors with an output frequency of approximately 800 megahertz, the maximum frequency at which lumped circuits can be conveniently used. The third part multiplies the frequency to 6400 megahertz and consists of a varactor multiplier using waveguide and coaxial techniques.

As previously stated, the output level for the transmit local-oscillator chain is 50 milliwatts (+17 decibels referred to 1 milliwatt) and, assuming that the conversion loss is proportional to the multiplication factor, a conversion loss of 21 decibels would be expected for a multiplication factor of 128. This requires a power level of 6.5 watts at the fundamental frequency.

To produce maximum efficiency, low-order multiples (doublers and triplers) should be used. If the chain were made completely of doublers, 7 stages in tandem would be required, involving 6 interstages between varactor multipliers with their associated problems of adjustment. Some efficiency must therefore be sacrificed to enable a more-economical circuit to be designed. Figure 1 is a block diagram of both chains indicating the stages used, the power levels, and the multiplication factors.

3.1.1 *Oscillator*

The design of the oscillator was mainly influenced by the need for frequency stability and low noise. The frequency stability required for the chain is ± 15 parts per million and this must be obtained over a temperature range from $+5$ to $+55$ degrees centigrade. To obtain this accuracy a crystal with a tolerance of ± 2 parts per million was used, and the oscillator circuit was placed in an oven to keep the temperature within 60 ± 2 degrees centigrade. The remainder of the allowable variation was reserved for aging.

To obtain this required stability, the power level at the crystal must be limited to 1 milliwatt (0 decibels referred to 1 milliwatt). The carrier-to-noise ratio required at the oscillator for 10 picowatts, psophometrically weighted, in a baseband channel is as follows.

$$\frac{C}{N} = 79.5 + M + F \text{ decibels,}$$

where

M = multiplication factor
= 128 times (42 decibels)
F = frequency-modulation disadvantage
= 20 $\log_{10}$ (baseband frequency/peak deviation).

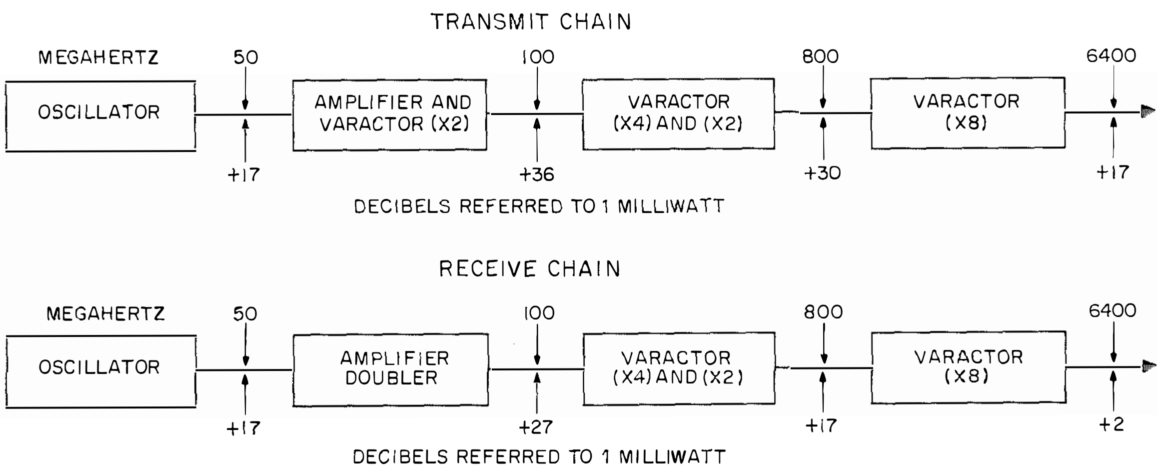

Figure 1—Block diagram of multiplier chains.

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Using the recommended deviation and pre-emphasis of the International Radio Consultative Committee [1] for an 1800-channel system, this gives C/N values of

149.4 at 5 megahertz
144.6 at 2 megahertz
139.2 at 1 megahertz
128.9 at 0.3 megahertz.

This is only 9.1 decibels above the thermal noise level at 300 kilohertz and less than the thermal noise level by 1.2 decibels at 1 megahertz, assuming a carrier level of 1 milliwatt (0 decibels referred to 1 milliwatt).

This means that the noise figure of the oscillator circuit must be better than 9 decibels and must have at least 11 decibels of selectivity at 1 megahertz to achieve the required 10 picowatts in the baseband, assuming no contribution from any other source. Since this is the lowest level in the chain and the lowest frequency with the highest multiplication factor, it is reasonable to allocate a large proportion of the noise allowance to this source.

The noise level at high frequencies can be improved by filtering, and a narrow-band filter with a 3-decibel bandwidth of 250 kilohertz is used to achieve an acceptable noise level.

3.1.2 *Driver*

The driver unit accepts the signal from the oscillator, doubles the frequency, and raises the level from 50 milliwatts (+17 decibels referred to 1 milliwatt) to 4 watts (+36 decibels referred to 1 milliwatt) to drive the varactor multipliers. The circuit therefore involves using high-frequency power transistors that have collector currents in excess of 100 milliamperes. As the collector current in a transistor is increased the noise figure is degraded, so that with amplifiers delivering a few watts of power noise figures of less than 30 decibels are difficult to achieve. At a power level of 4 watts (+36 decibels referred to 1 milliwatt), a noise figure of better than 30 decibels at 50 megahertz must be obtained to achieve the performance requirements, unless the amplifier is followed by a narrow-band filter similar to that required after the oscillator. However, the driver is required to supply a varactor stage and, as a very-narrow-band filter does not provide a very-stable source impedance, the filter bandwidth must not be less than ±1 megahertz. It was therefore decided to double the frequency before final amplification and filtering to achieve a 6-decibel noise advantage by reducing the multiplication factor following the amplifier by two.

The unit therefore consists of two 50-megahertz transistor amplifier stages, a varactor doubler, a 100-megahertz transistor output stage, and a filter designed in the form of a helical line [2] with a bandwidth of ±1 megahertz.

The power dissipated by the unit is of the order of 15 watts, hence the unit must be packaged so that this power can be dissipated without excessive temperature rise. Figure 2 shows the mechanical assembly including the cooling arrangement.

3.1.3 *Quadrupler-Doubler*

The frequency is multiplied by 8 in the quadrupler-doubler from 100 to 800 megahertz. This multiplication factor can be obtained in several ways. The solution that proved to be a practical compromise was the use of a quadrupler followed by a doubler. The quadrupler uses a varactor connected in series, which makes it possible to use a circuit that does not require additional networks to act as idlers. The doubler is conventional with the varactor in shunt. Using this circuit a maximum conversion loss of 6 decibels is obtained for the complete quadrupler-doubler.

The circuit uses lumped components mounted on a metal chassis with screens separating the interstage matching circuits to prevent interaction, as shown in Figure 2.

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3.1.4 *Transmit Octupler*

The transmit octupler is driven by 1 watt (+30 decibels referred to 1 milliwatt) at 800 megahertz as shown in Figure 1. The basic mechanical arrangement is shown in cross section in Figure 3.

Two coaxial stub tuners at $\lambda/15$ spacing form an input matching control of large admittance range with rapid convergence. They are followed by a 3-section coaxial low-pass filter with a cutoff frequency of 1200 megahertz, which isolates the varactor from the input circuit at all the harmonic frequencies. The filter forms part of a 50-ohm T-junction, one arm of which contains the varactor; the other is probe coupled into the *R40* waveguide (internal dimensions of 1.372 by 0.662 inches (3.5 by 1.7 centimeters)) output circuit, where the 8th harmonic is selected by a conventional 3-section band-pass filter (not shown). The varactor arm is enclosed by a removable end cap, which contains a resistor that permits the varactor to provide self-bias. The end cap also terminates the line for high frequencies by a capacitor using a disc of beryllium oxide. Since only the 6th and higher harmonics are propagated in the waveguide, the lengths of line between the varactor and the waveguide and between the varactor and the low-pass filter act as idler circuits up to the 5th harmonic; the distances are chosen empirically, along with the spacing of the band-pass filter from the coaxial waveguide junction, to optimize the conversion loss.

3.2 Receive Local-Oscillator Chain

The receive local-oscillator chain was designed to provide the power levels indicated in Figure 1. The design was the same as for the transmit chain except in the driver unit, where the frequency was doubled in one of the transistors

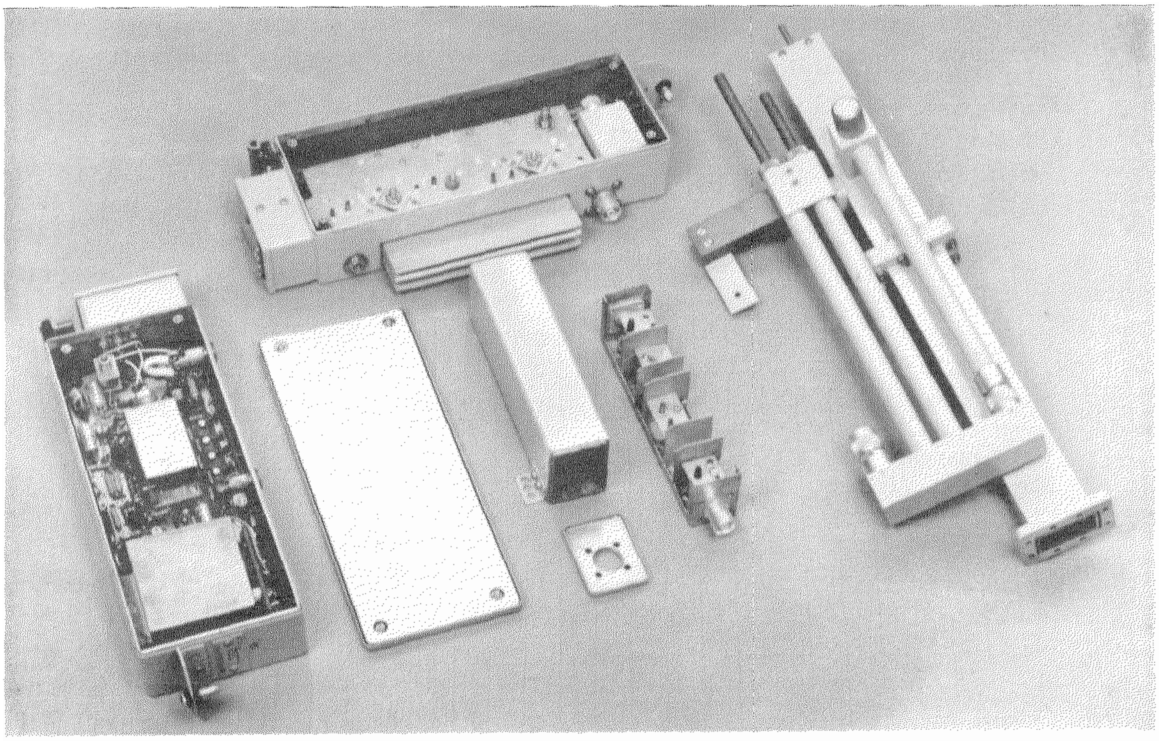

Figure 2—Multiplier chain showing oscillator with lid at left, driver at top, quadrupler-doubler with case at center, and octupler at right.

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instead of in a varactor. This economy was made possible by the lower output power required.

4. Measured Results

4.1 Power Levels

The transmit and receive chains have been adjusted to meet the required output levels with several sets of varactors.

Both octupler circuits can be adjusted to have a constant conversion loss, to be stable, and to be free from hysteresis over a range of input ±3 decibels from nominal.

4.2 Noise Measurements

Multipliers have proved to be a source of spurious frequency-modulation noise, and at an early stage noise measurements were carried out. Two methods of measuring noise were used in the laboratory, the choice of method depending on the noise level.

Figure 4 shows the test circuit for measuring the noise contribution of the octupler. The limit of resolution is set by the contributions from the filters, 800-megahertz source, and local-oscillator source at low baseband frequencies, and by the bandwidth of the second multiplier chain at the upper end of the baseband. It was possible to measure noise levels as low as 0.5 picowatt (for 1800-channel loading) over the baseband range from 1 to 4 megahertz, and the octupler contribution was found to be less than 3 picowatts at 1 megahertz and less than 1 picowatt at 4 megahertz.

Complete chains were measured using a standard demodulating technique without an additional multiplier and using a super-high-frequency vacuum tube as local oscillator. The basic noise was determined by substituting for the chain a second filtered super-high-frequency vacuum-tube oscillator, and this method permitted measurements down to 1 picowatt with an uncertainty of ±1 decibel from 200 kilohertz to 8.9 megahertz. The limit of this test circuit is the noise performance of the discriminator at low and high frequencies and the mixer noise at high frequencies. Using this circuit the chains were developed to produce a typical noise response as shown in Figure 5.

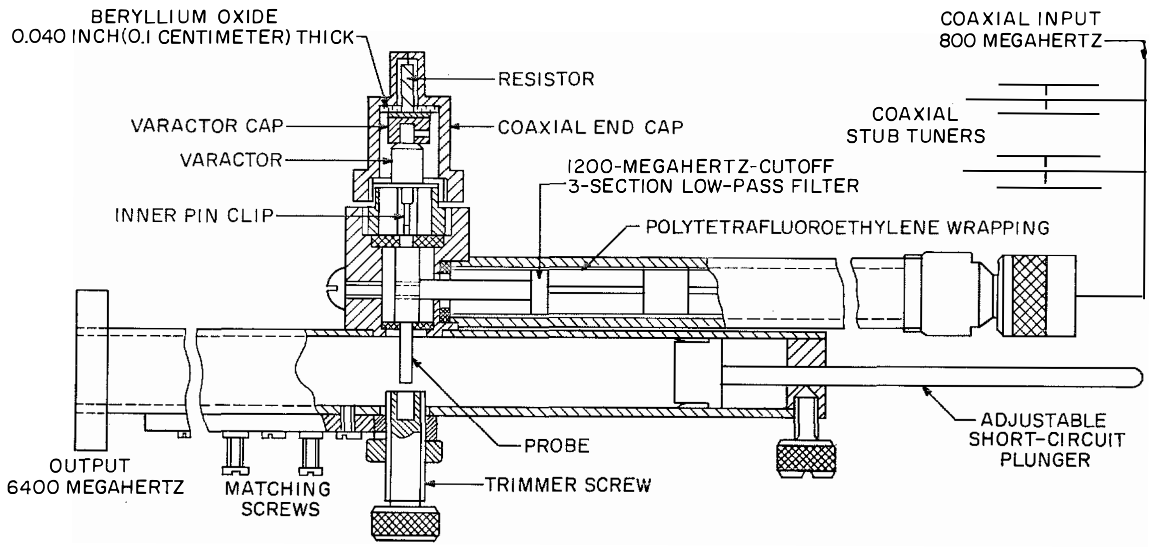

Figure 3—Octupler.

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4.3 Sideband Measurements

The output levels of unwanted sidebands from the complete chains were estimated from measurements of the sideband output levels of the chains at 800 megahertz and of separate measurements of the selectivity of the octuplers. These margins were found to be 120 decibels at ±100 megahertz and greater than 120 decibels at ±50 megahertz.

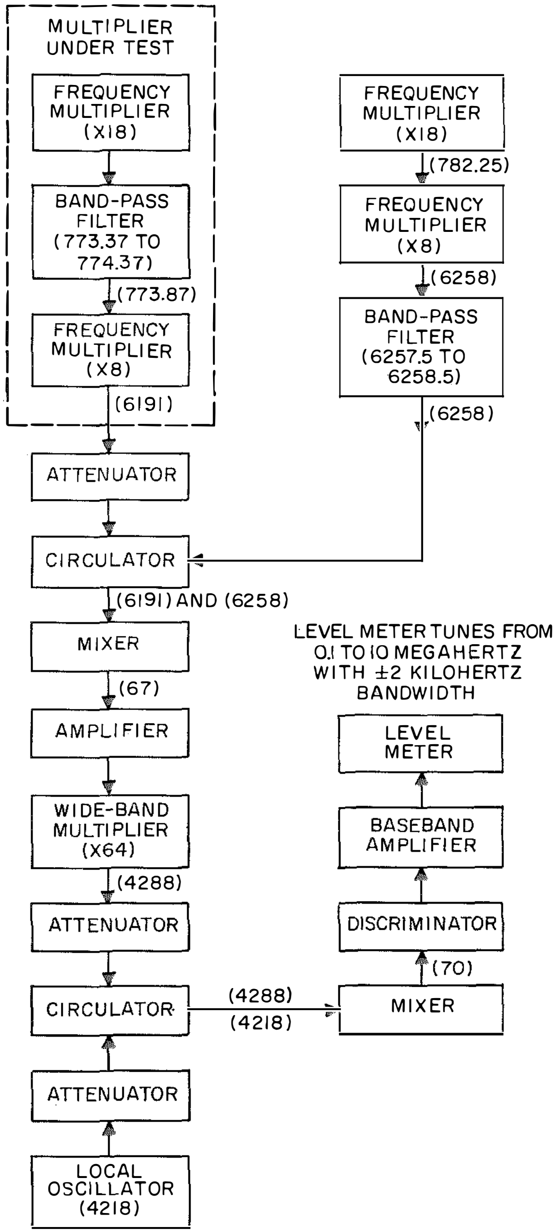

Figure 4—Noise-measuring circuit. The frequencies within parentheses are in megahertz.

5. Comparison with Other Sources

The microwave power sources described take the place of thermionic devices previously used. Thermionic devices require special frequency-control circuits or temperature-controlled ovens if a klystron or a coaxial-line oscillator is used, or these devices need a reference cavity if a microwave triode is used. They also require complex power supplies.

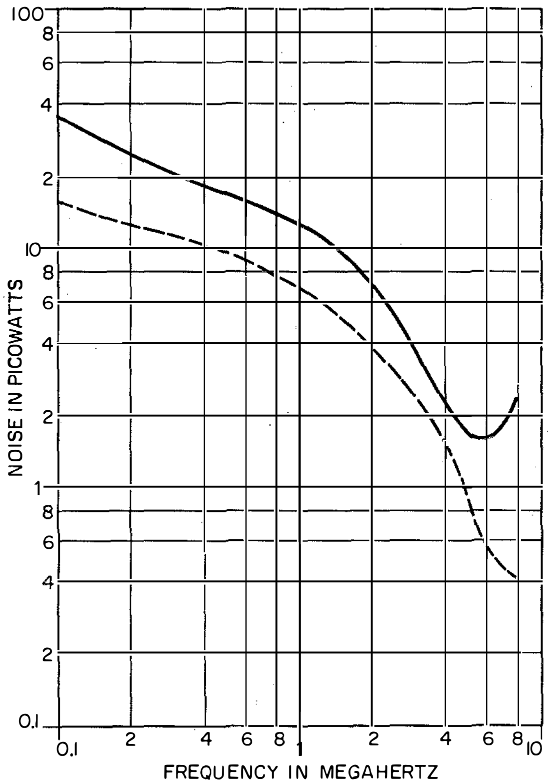

Figure 5—Noise level on transmit and receive multiplier chains. The broken line is the typical measured noise level, while the solid line is the specified noise limit.

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The lives of oscillator tubes in the microwave region have reached figures that compare very favorably with other components used in a radio system. The lives of solid-state sources have not yet been established by long field experience, but all indications are that long lives can be expected.

The main disadvantage of varactor multiplier chains compared with vacuum-tube sources is that to obtain an equivalent performance a number of adjustments must be made involving special equipment. This is particularly apparent when adjusting for optimum noise performance. This disadvantage may rapidly diminish with better components using high multiplication factors per stage. There is very little difference in power consumption between the average thermionic device and the transmit oscillator chain described, each dissipating about 20 watts. However, when the auxiliary circuits are included to obtain the necessary performance for thermionic devices, the power consumption is considerably increased.

6. Acknowledgments

The authors would like to thank all their colleagues associated with the work on the multiplier chains for valuable assistance and guidance.

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Microwave Radio System for Multichannel Telephony and Television in the 6-Gigahertz Range
Part 3—Wide-Band Intermediate-Frequency Amplifiers

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1. Introduction

With the introduction of modern transistors increasing use has been made of these components to replace amplifier tubes in commercial equipments. This transition has occurred gradually. In the radio relay field, systems were initially of mixed construction, with the intermediate-frequency amplifiers located in the transmission path equipped with tubes, while other amplifier units used an increasing share of semiconductors.

Improved methods of manufacturing semiconductors, and above all the advent of planar epitaxial transistors with cutoff frequencies of up to 1000 megahertz, made it possible to design intermediate-frequency amplifiers of wide bandwidth, even though these amplifiers must meet very-stringent conditions regarding their transmission characteristics when applied to wide-band radio-link systems. Long life, low power dissipation, and high gain-bandwidth product are strong advantages of transistors over tubes, even though major problems still remain.

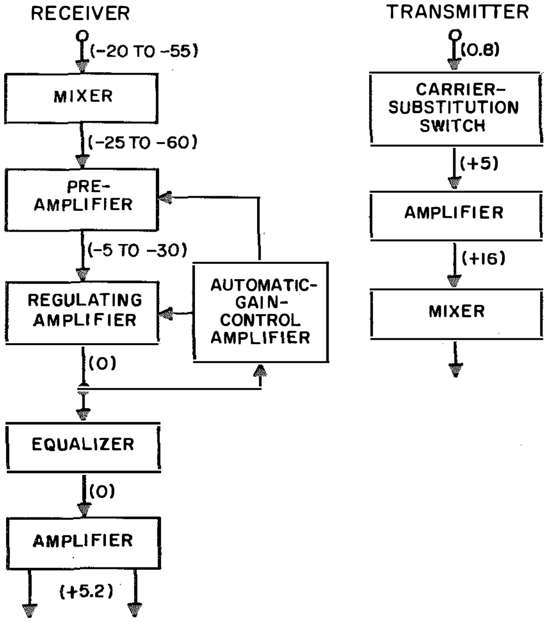

Figure 1—Diagram of the intermediate-frequency-amplifier chain. Signal levels in decibels referred to 1 milliwatt are in parentheses.

The dependence of the transistor parameters on frequency, temperature, and operating point has a considerable effect on the transmission performance of the amplifier and makes necessary a change from the conventional circuit design hitherto used in tube amplifiers. An example is the automatic gain control of the intermediate-frequency amplifier. Unlike tubes, transistors allow no control to be applied because of their high dependence on the above-mentioned factors, and specific regulating devices must be provided in the amplifier.

In this paper the all-semiconductor intermediate-frequency amplifier chain of the *1800/TV* radio-link system is described. The design of this chain was based on specifications for a radio-link system conforming with recommendations of the International Radio Consultative Committee.

2. General Description

2.1 Level and Block Diagram

Figure 1 shows the block diagram and signal levels of the intermediate-frequency chain. By subtracting from the transmitting power the losses in the filter and transmission line and the path attenuation including fading, the possible input power of −20 to −55 decibels referred to 1 milliwatt is obtained at the receiver mixer. This power must be raised in the receiver part of the chain to +5.2 decibels referred to 1 milliwatt, corresponding to 0.5 volt in 75 ohms, and made available at two equivalent outputs decoupled from each other.

The input level at the transmitter part of the chain is 0.8 decibel referred to 1 milliwatt, which corresponds to 0.3 volt in 75 ohms. A

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power amplifier raises this level to +16 decibels referred to 1 milliwatt (40 milliwatts) for driving the transmitter mixer (acting simultaneously as a limiter), with a carrier-substitution control circuit providing automatic injection of a temporary carrier in the event of failure of the signal carrier.

The actual units in the amplifier chain correspond with these functions. The receiver mixer and preamplifier as well as the transmitter mixer and power amplifier form single units, respectively, connected to the main amplifier and the carrier-substitution unit by cable. The main amplifier is divided into a regulating amplifier and an output amplifier to provide for the insertion of a delay and amplitude equalizer. The level is 0 decibels referred to 1 milliwatt at this point, so that a 5-decibel output amplifier with two decoupling outputs is available that may also be employed as a separating or distribution amplifier for protection-channel and distribution equipments.

The signal levels at the point of interconnection between the preamplifier and main amplifier should on the one hand be rather high to make the noise contribution of the succeeding amplifier negligibly small, while on the other hand excessive levels should be avoided to prevent overloading the semiconductor diodes in the regulating networks.

Taking into account the noise figure of the regulating amplifier, the maximum gain for the preamplifier and regulating amplifier has been set at 30 decibels. This makes it necessary to provide a regulating network in the preamplifier also, to avoid overdriving the regulating amplifier. Because of the delayed start of the action in the regulating network of the preamplifier, the preamplifier gain remains constant until the output level approaches −5 decibels referred to 1 milliwatt, and only then does the regulating action start. This ensures that a slight degradation of the receiver noise figure will occur only at high input levels.

2.2 Coupling Network

In vacuum-tube amplifiers, the coupling networks between the stages determine the selectivity of the amplifier. They normally consist of transitionally coupled band-pass filters. The amplifier gain is regulated by adjusting the grid bias, while compensating means must be provided in the cathode lead of the tube to prevent any essential change of the tube input and output impedances.

With transistors, the use of such circuit configurations is confined to narrow-band amplifiers, particularly if additional damping resistors are provided. Extensive research has therefore been done to realize coupling and regulating networks for wide-band amplifiers. Because the current gain of transistors in common-emitter circuits is frequency dependent for the frequency range considered and because of the resulting need for complementary frequency-dependent coupling networks, the use of this type of circuit is restricted to the input stage of the preamplifier and preference is given to the common-base configuration. In this type of circuit the current gain is less than unity, and amplification can be achieved only by transformer action between the stages.

Amplifiers with transformer-coupled stages of extremely wide bandwidths are well known, with the required selectivity being achieved by means of a passive filter network between the preamplifier and main amplifier. If the filter is provided with temperature compensation, such an arrangement is rather insensitive to temperature variations, although it has (because of the large bandwidth of the amplifier) the disadvantage of low stage gain. Moreover, there occurs a summation of all the sum and difference frequencies that are produced in the stages following the filter and that fall in the transmitted frequency band.

Measurements on such an amplifier have shown that the linearity and intermodulation requirements of the specification could not be met,

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and therefore the coupling networks are required to contribute to selectivity. This led to a network consisting of a transformer that, by the addition of a low-pass π section, has been made equivalent to a band-pass filter. The edges of the pass band are determined by the mutual inductance of the transformer and by the cutoff frequency of the low-pass section. In Figure 3, which shows part of the regulating amplifier, can be seen the circuit of such a network inserted between the first and second stages. A resistor-capacitor combination at the emitter input matches the transistor input impedance to the filter. A thermistor equalizes the temperature-dependent variation of the input impedance.

Two versions of this network are used. In the bandwidth-determining stages, an air-core transformer is employed to avoid spreads in inductance due to the manufacturing tolerance of ferrite cores. In the stages having larger bandwidth, the transformer contains a small double-hole ferrite core, which type of core enables a higher transfer ratio to be achieved. The stage gains are 8 and 12 decibels, respectively.

Of great importance for determining the coupling network and the resulting stage gain is the feedback conductance of the transistor employed. This conductance is both frequency and temperature dependent and can be expressed as a mismatch at the transistor input. The mismatch increases with increasing collector resistance (that is, the mismatch becomes greater as the transformed input impedance of the succeeding stage is made larger). The mismatch increases with increasing temperature, and this (since it is frequency dependent) increases the slope of the amplitude-frequency curve in such a way that higher frequencies are progressively attenuated. With a given feedback conductance of the transistor, the allowed maximum slope delineates the maximum transfer ratio and hence the stage gain. The stage gains mentioned above apply to the *2N918* transistor.

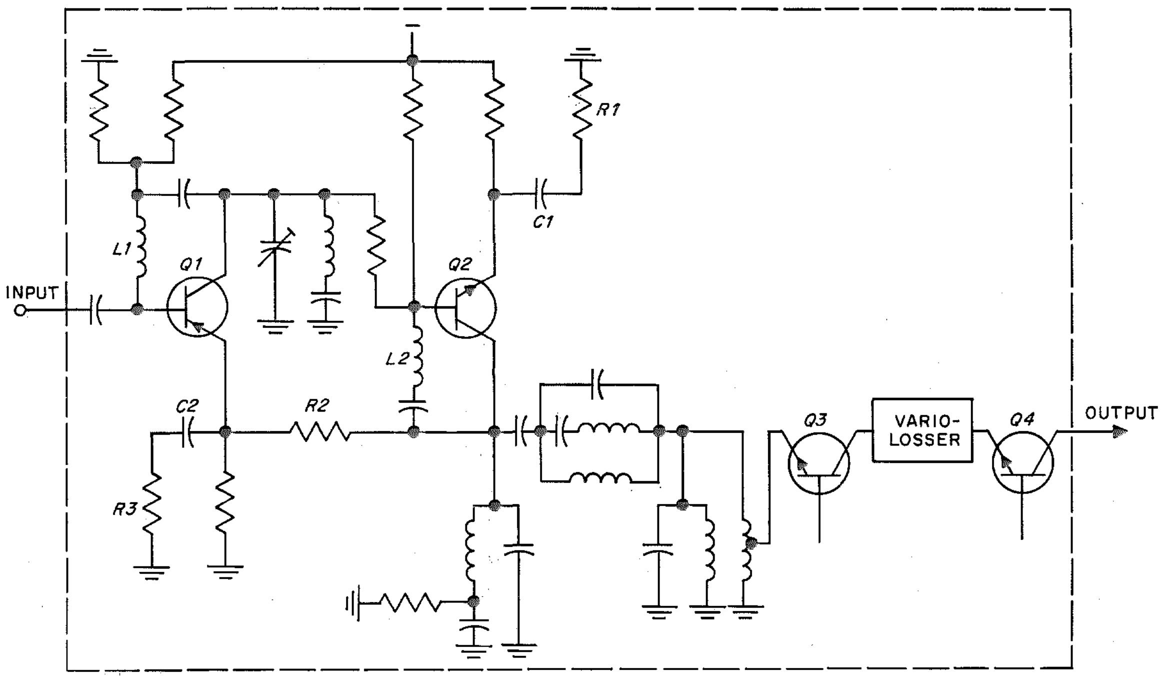

Figure 2—Preamplifier low-noise input stages with feedback.

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2.3 Regulating Network (Variolosser)

The small feedback conductance of the *2N918* transistor makes unnecessary a constant-impedance regulating network. Regulation is effected by parallel-connected silicon diodes. The circuit of the regulating network may be seen in Figure 3 between the second and the third stage. The regulating range of one network is 15 decibels without a noticeable change of the transfer characteristic.

3. Detailed Description

3.1 Preamplifier

The preamplifier of Figure 2 consists of 4 amplifier stages and a variolosser regulating network. To obtain high linearity and thus avoid possible cross-modulation in the input stages, the first two stages are combined into one unit with a negative feedback path. This unit is followed by a band-pass filter with zeros at ±29.5 megahertz from the center frequency to suppress the adjacent-channel carrier frequency. The band-pass filter is followed by an isolating stage, the regulating network, and the output stage. Between transistors *Q1* and *Q2*, a low-noise *p-n-p 2N2415* germanium transistor and an *n-p-n 2N918* silicon transistor, a negative feedback path is provided via the voltage divider *R2*, *C2*, and *R3*. *L1* and *L2* neutralize the collector-base capacitance. *C1* and *R1* are for correction of frequency and phase response. The negative feedback used increases the input impedance of the first stage considerably, which improves the match for minimum noise transfer from this stage to the mixer.

3.2 Main Amplifier

As already mentioned, the main amplifier includes a regulating amplifier and an output amplifier, with delay and amplitude equalizing networks between them. This is a better way to achieve an optimum system noise figure than to insert the equalizers between the preamplifier and the main amplifier.

The regulating amplifier contains 8 transistors,

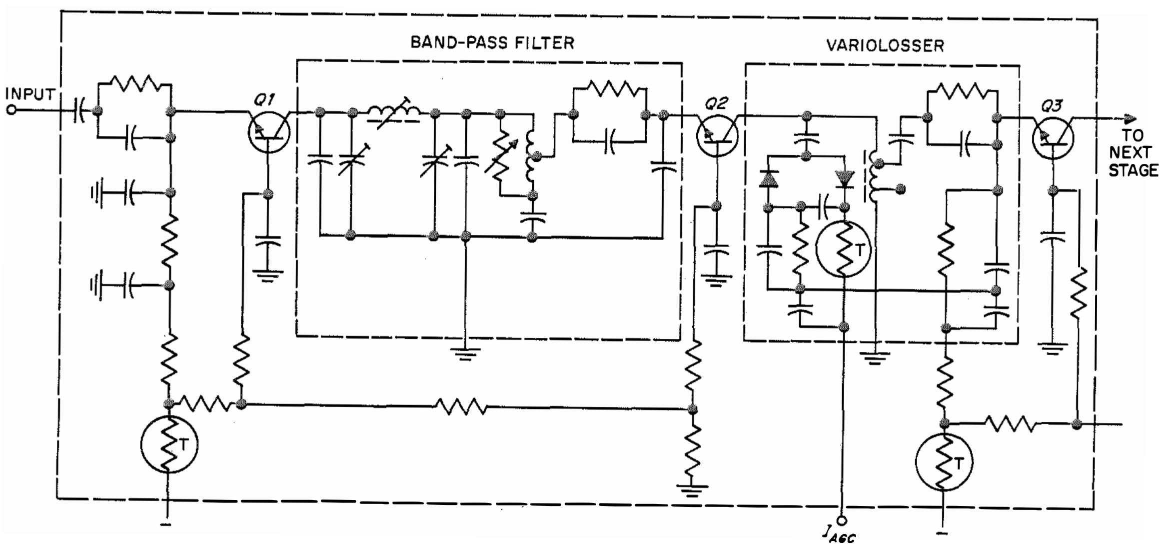

Figure 3—Input, band-pass, and regulating stages of the main intermediate-frequency amplifier.

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one of which is arranged in a branching stage for the automatic-gain-control circuit. Figure 3 shows the circuit of the input stage, the succeeding band-pass-filter stage, and the regulating stage. Another band-pass-filter stage and regulating stage follow. The signal then goes to an isolating stage driving two transistors with their inputs connected in parallel. One of these transistors supplies the output signal level of +0.8 decibel referred to 1 milliwatt, while the other produces automatic-gain-control voltage by detection of the intermediate-frequency level. To save current, two stages receive operating power in series, as may readily be seen in Figure 3.

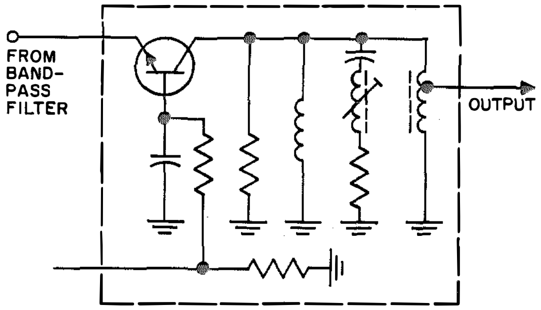

Figure 4—Output circuit of main amplifier.

The automatic-gain-control voltage drives a direct-current amplifier consisting of 4 transistors, 2 of which are in a temperature-compensated bridge circuit while the other 2 provide control current to the regulating networks. The onset of the control current is somewhat delayed in one of the control transistors by a preceding zener diode; this current controls the regulating network in the preamplifier.

The output amplifier contains 5 transistors and supplies a signal level of +5.2 decibels referred to 1 milliwatt to each of two decoupled outputs. Its input stage drives 2 amplifiers,

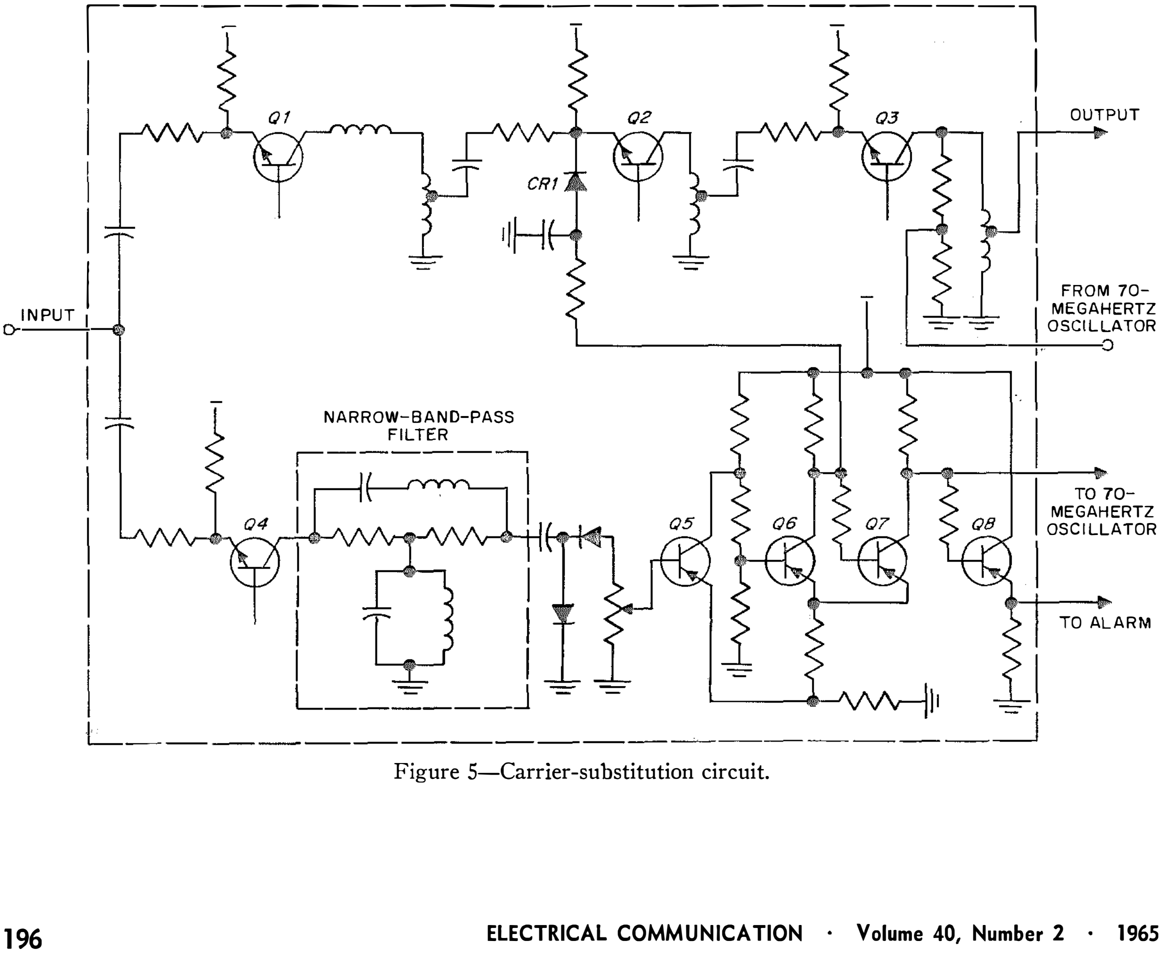

Figure 5—Carrier-substitution circuit.

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each having 2 transistors. This arrangement is relatively expensive but it ensures an isolation of more than 40 decibels between the two outputs. To obtain a high return loss, compensating networks are connected to the two outputs as shown in Figure 4. The tapped output transformer and coil, together with the collector capacitance, form a parallel-tuned circuit, which is compensated by the series-tuned circuit damped with the matched characteristic impedance. The coupling networks between the stages are identical with that of the band-pass filter in Figure 3.

3.3 Carrier-Substitution Unit

When no carrier is being received, the carrier-substitution unit interrupts the signal path and simultaneously injects a temporary carrier to drive the transmitter. Figure 5 shows the schematic diagram. In parallel with the input of a plug-in amplifier is a narrow-band amplifier, *Q4*, centered at 70 megahertz to detect the mean value of the input signal. To avoid any reaction on the input of the circuit, the narrow-band filter is designed as a bridged-*T* network with constant impedance.

While a carrier is being transmitted, the rectified intermediate-frequency level is high enough to keep the following flip-flop cut off so that the amplifier is through-connected and the 70-megahertz oscillator is switched off. Assuming that the effective bandwidth of the automatic-gain-control circuit is at least as large as the useful bandwidth of the amplifier chain and that the power delivered by the main amplifier does not exceed +5.2 decibels referred to 1 milliwatt even when fully driven and including noise, the rectified-intermediate-frequency level after the narrow-band amplifier will be proportional to the ratio of the bandwidth of the narrow-band amplifier to the bandwidth of the intermediate-frequency chain.

A level reduction of this order (or zero level in the case of no signal) causes the flip-flop to switch. Diode *CR1*, which has been nonconducting, is then connected to ground and passes current from transistor *Q2*. Since the base voltage of this transistor is kept constant, the transistor becomes nonconducting. The 70-megahertz oscillator is likewise cut in by a diode and supplies a signal to the power amplifier via the output network of the amplifier.

3.4 Power Amplifier

The power amplifier is a 3-stage amplifier using in successive stages *2N918*, *2N3137*, and *2N2950* transistors. The coupling networks are similar to that shown in Figure 3. The peak output power is approximately 120 milliwatts at 1-decibel compression, so that sufficient linearity is ensured at 40 milliwatts normal level.

4. Measured Results

Figure 6 shows the amplitude and group-delay responses as a function of the frequency of the amplifier chain. The measurements were made with a dummy mixer simulating the impedance of the receiver mixer and with a dummy load in place of the transmitter mixer. The maximum change of the amplitude curve through ±10 megahertz due to equivalent input level variations from −20 to −55 decibels referred to 1 milliwatt at the receiver mixer is less than 0.2 decibel. Figure 7 shows the amplitude-frequency

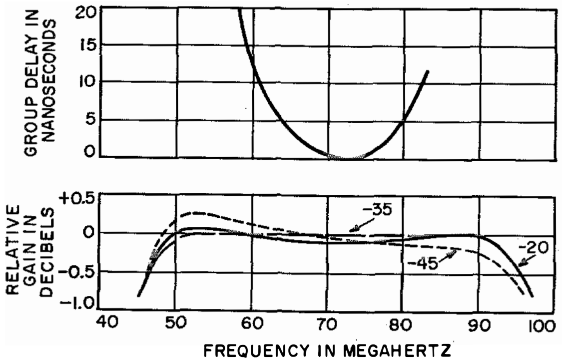

Figure 6—Amplitude and group-delay responses of the intermediate-frequency amplifier as a function of frequency. The numbers on the curves are in decibels referred to 1 milliwatt.

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curve as a function of temperature. In this case also the maximum change through ± 10 megahertz at an equivalent input power between -20 and -40 decibels referred to 1 milliwatt is less than 0.2 decibel. Only at an amplifier setting corresponding to input power of -55 decibels referred to 1 milliwatt does the deviation rise to between 0.2 and 0.3 decibel.

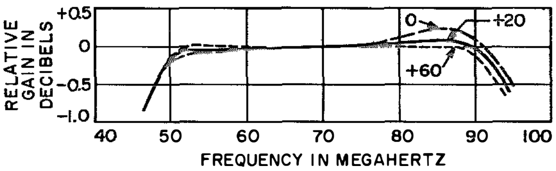


Figure 7—Amplitude-frequency response of the intermediate-frequency amplifier as affected by the temperatures indicated in degrees centigrade.

In Figure 8 the noise figure of the intermediate-frequency chain is plotted, both alone and with the receiver mixer, as a function of the receive level, which corresponds to a given gain. During all measurements, the automatic gain control was disconnected and the amplifier regulated manually to the particular nominal value.

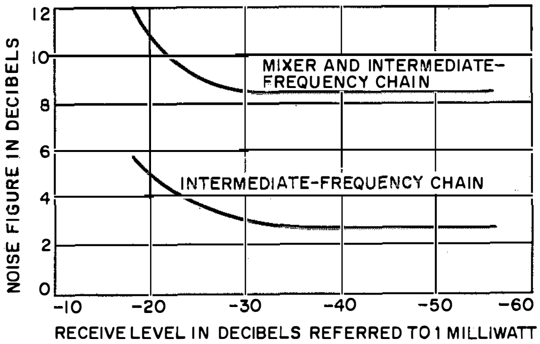


Figure 8—Noise figure.

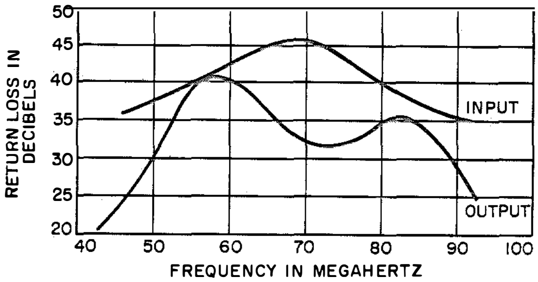
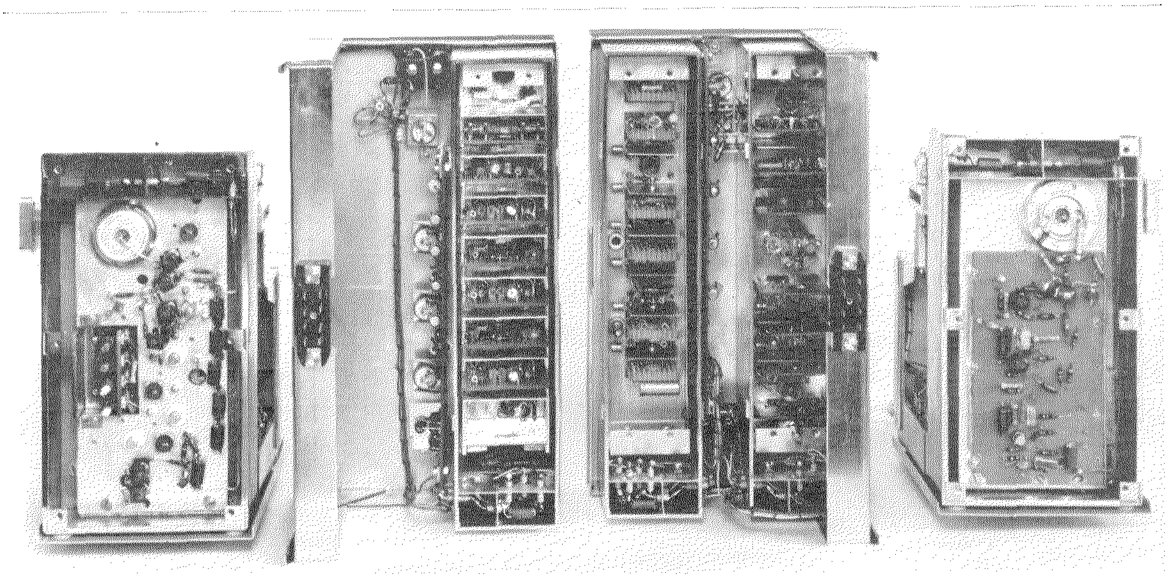


Figure 9—Return loss.

Figure 10—Intermediate-frequency-amplifier chain.



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The return loss of the amplifier inputs and outputs is shown in Figure 9. Figure 10 is a photo of the entire amplifier chain.

5. Conclusion

New design techniques have been evolved that have made practicable the achievement of a completely solid-state wide-band intermediate-frequency-amplifier chain to the stringent requirements of the radio repeater and terminal equipment in the 6-gigahertz *1800/TV* radio relay system.

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Microwave Radio System for Multichannel Telephony and Television in the 6-Gigahertz Range
Part 4—Cassegrain Antenna

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1. Introduction

The need for a high-performance low-cost aerial and feeder system was appreciated from the start of the development of the *1800/TV* system. None of the existing designs fully met the requirements and it was decided that a cassegrain type of paraboloidal aerial might provide the optimum design. This paper describes the design of the 4-meter (13-foot) diameter bipolar aerial for the 6-gigahertz system and compares its performance with existing designs.

2. Theory of Cassegrain Antenna

The cassegrain principle was first used for optical astronomical telescopes. In this case a paraboloidal reflector produces an image of a star at its focal point. A smaller secondary hyperboloidal reflector is placed near the focus of the paraboloid and produces a real image near the apex of the paraboloid. The eyepiece is then arranged to view this image through a suitable opening in the paraboloid. This construction is convenient for optical purposes, especially if the focal length of the paraboloid is large.

The cassegrain idea has been applied to antennas for super-high-frequency use in recent years with considerable benefit. The principal advantages are flexibility in design; improved performance in terms of efficiency, polar diagram, and bandwidth; and simplicity of primary-feed construction.

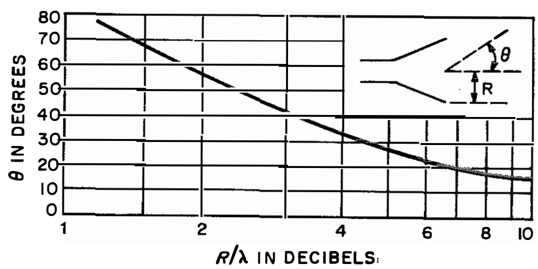

Figure 1—Beamwidth of conical horns.

2.1 Radiation from Primary Feed

A horn may be used to introduce energy into a cassegrain system. The beamwidth of a circular horn decreases as the size of the horn increases, as shown in Figure 1. To efficiently use the energy radiated from the horn, the hyperboloidal reflector must intercept energy up to, say, 10 decibels below the maximum. It is therefore possible to consider the use of a small horn with a reflector subtending a large angle, say 60 degrees, at the horn. Alternatively, a larger horn can be considered with a reflector subtending a smaller angle, say 20 degrees. Paper designs based on geometric optics show similar efficiencies for a very-wide range of designs. Practical designs obey the geometric laws only very approximately because of the small ratio of diameter to wavelength of the hyperboloidal reflector. It is therefore necessary to rely on measurement to determine the optimum design.

2.2 Radiation from Hyperboloidal Reflector

The hyperboloidal reflector generates an image of the primary horn that can be considered as the basic feed to the main paraboloidal reflector. For super-high frequency, it is convenient to use a convex hyperboloidal reflector as it results in immediate divergence of the signal and consequently reduces the energy reflected back into the primary feed. The image is therefore virtual, lying behind the reflector, and is reduced in size compared with the size of the primary feed.

This last point is important in obtaining an efficient design, as phase errors are reduced proportionately with the image size.

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Figure 2 shows a family of hyperbolae and can be used as a design chart for cassegrain antennas. The waveguide feed aperture is located at the source, and the ratio of reduction of the image varies as shown from unity for the flat reflector (number 1) up to 4 times for hyperboloid number 5.

Radiation of signals from the hyperboloidal reflector is modified by diffraction, which makes it impossible to achieve a sharp transition from high to low intensity. The intensity of illumination in the shadow region is shown in Figure 3. This diagram can also be used by application of Babinet's principle to give the illumination from the hyperboloidal reflector outside the geometric region. The diagram also shows that better control over the radiated energy can be achieved if the reflector size is large compared with a wavelength.

Reflection of energy into the primary feed is greatest if the feed is close to the hyperboloid, but is quite small if the distance is increased. This reflection can be reduced by recourse to an apex plate matching technique on the hyperboloid.

This method causes cancellation of energy propagating toward the axis of the paraboloid and is effective over a broad frequency band. The region of strong illumination of the paraboloid is modified by this method and becomes an annulus in which the maximum field strength occurs at an angle of about 45 degrees from the axis.

This pattern of illumination leads to high efficiencies, but also increases the level of the first side lobe.

3. Development of Cassegrain Antenna

Preliminary trials on 4-foot (1.2-meter) paraboloids of 90-degree aperture showed that a high efficiency could be obtained by the use of

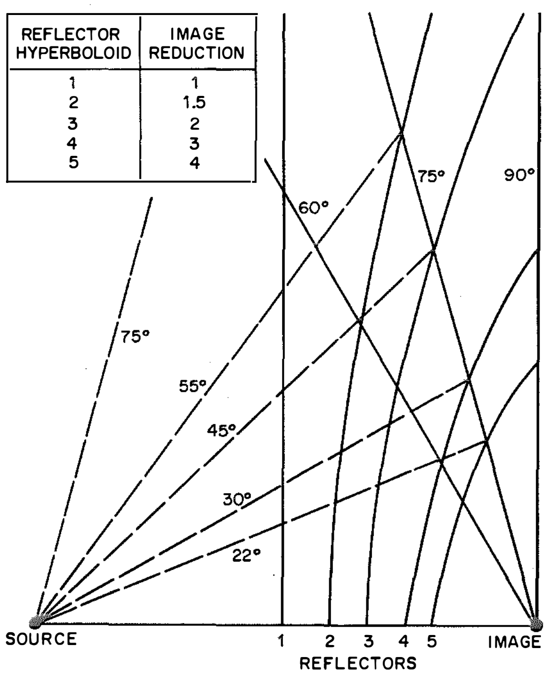

Figure 2—Hyperboloid family. Solid lines are intercept lines for paraboloids indicated in degrees.

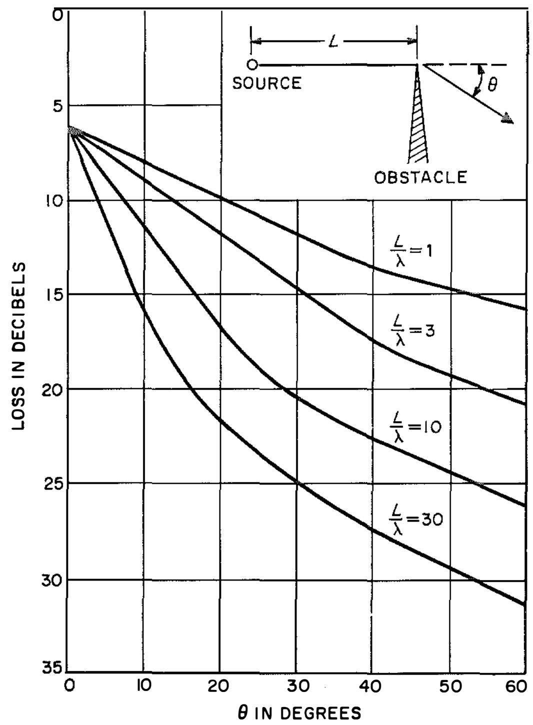

Figure 3—Diffraction at a straight edge.

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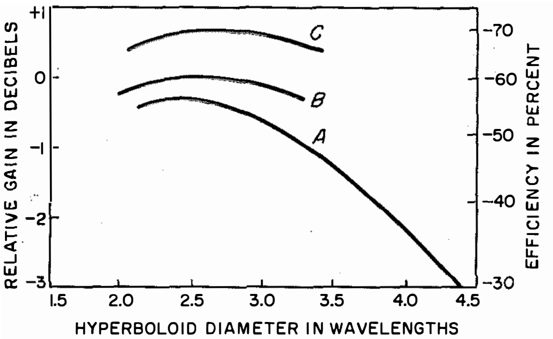

CURVE	HYPERBOLOID APERTURE AT HORN IN DEGREES	HORN MOUTH DIAMETER IN INCHES (CENTIMETERS)
A	50	1.75 (4.4)
B	50	2.2 (5.6)
C	60	2.2 (5.6)

Figure 4—Dependence of cassegrain antenna on hyperboloid design for gain and efficiency. For the example shown, paraboloid diameter = 4 feet (1.2 meters), aperture = 90 degrees, and frequency = 6.2 gigahertz.

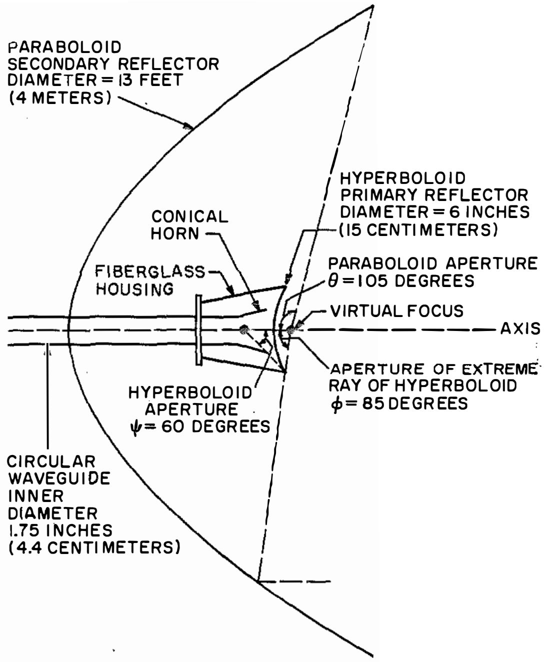

Figure 5—Geometry of the cassegrain antenna. The parts are not drawn to scale.

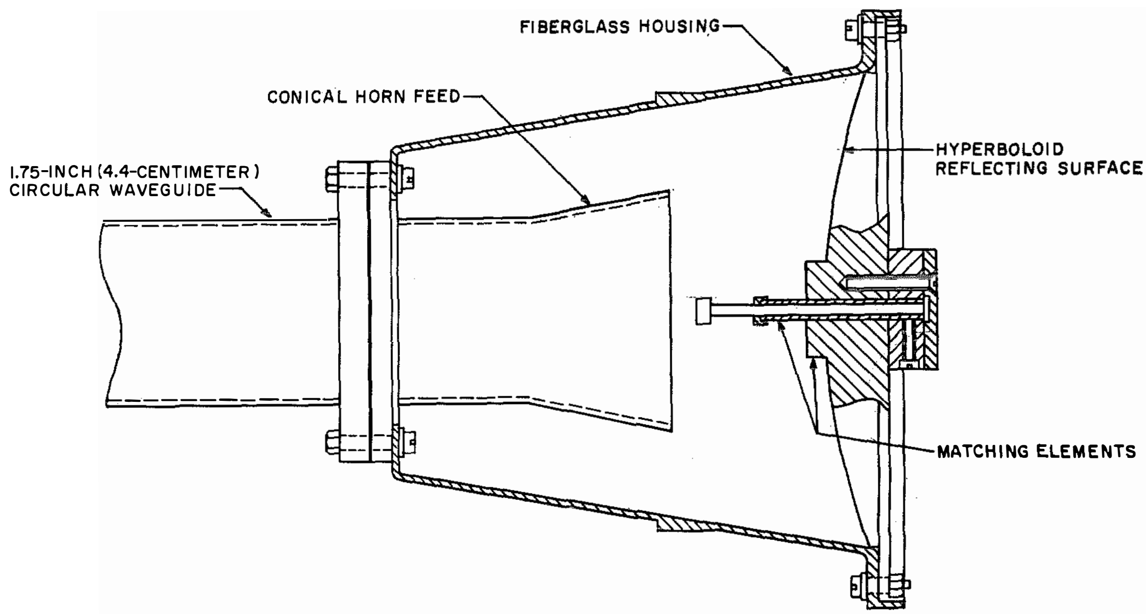

Figure 6—Cassegrain horn feed.

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a small hyperboloid (see Figure 4) mounted close to the primary horn as shown in Figure 5. It was found necessary to extend the aperture of the main reflector to 105 degrees to obtain a good polar diagram in the region of 90 degrees off axis.

4. Cassegrain Horn Feed

The experimental work on the 4-foot (1.2-meter) antenna was carried out with the hyperboloidal reflector supported by a Perspex housing mounted on a flange attached to the waveguide feed. This arrangement was retained for the engineered version except that the material for the housing was changed from Perspex to fiberglass. Figure 6 shows the final arrangement with the matching elements in position. Figures 7 and 8 are photographs of the disassembled and assembled horn. Minor changes were made to the shape of the housing to obtain a broad-band impedance match. Adjustable rods enable a small reflection of variable amplitude and phase to be introduced to compensate for the residual reflection of the assembly and for variations in manufacture. The whole assembly can be sealed against moisture and is robust enough to withstand any rough treatment occurring during transport and assembly.

The horn assembly is supported on a short section of circular waveguide that bolts to the apex of the paraboloid. The whole assembly is rigid and can withstand maximum wind loading without further support.

Due to the wide aperture angle of the paraboloid, the horn lies inside the rim of the dish and is not exposed to damage from falling ice.

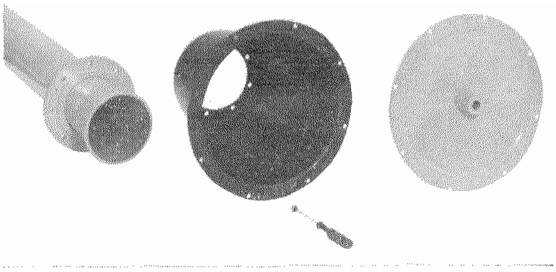

Figure 7—Disassembled horn.

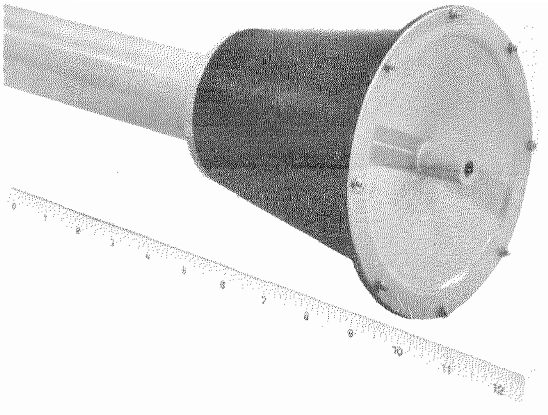

Figure 8—Assembled horn.

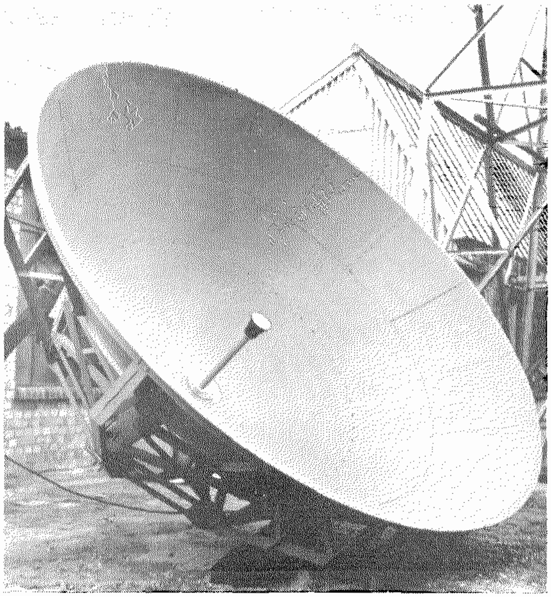

Figure 9—Close-up of cassegrain aerial.

5. Paraboloid

5.1 General

This 13-foot (4-meter) 105-degree paraboloid was constructed by adding a sheet-metal extension to a paraboloid 10 feet (3 meters) in diameter and 90 degrees in aperture. The accuracy of profile was plus or minus one-eighth

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to one-sixteenth of a wavelength over the extended portion, but was much better over the central portion. A close-up photograph appears in Figure 9 and the complete aerial mounted on an adjustable tower is shown in Figure 10.

5.2 Performance

The antenna gain was measured at 5.9, 6.2, and 6.4 gigahertz over two different paths and was found to be greater than 46 decibels. The efficiency is thus greater than 60 percent.

Polar diagrams were plotted for *E*- and *H*-plane operation at 5.9, 6.2, and 6.4 gigahertz and also for the crossed-polar properties. Figures 11 and 12 are representative polar diagrams. These show that the crossed-polar properties are very good, as would be expected from the symmetry of the structure, and reach a level 68 decibels below the main beam at 20 degrees off axis. The normal polar diagram is also good, reaching 68 decibels at 90 degrees off axis. The beamwidth to the 3-decibel-down points is 0.8 degree in the 6-gigahertz band.

5.3 Coupling Between Adjacent Antennas

Side-to-side coupling of cassegrain antennas by horn-to-horn coupling has been calculated to be 86 decibels for aerial centers 15 feet (4.6 meters) apart. Side-to-side coupling can also be caused by obstacles close to the aerial site and near the direction of shoot or within the main beam at distances up to 4 miles (6.4 kilometers).

Calculation for a level site shows side-to-side coupling of 89 decibels for aerials 100 feet (30 meters) above ground. This assumes isotropic scattering from the ground that may occur on tree-covered terrain. Back-to-back coupling is estimated at 130 decibels.

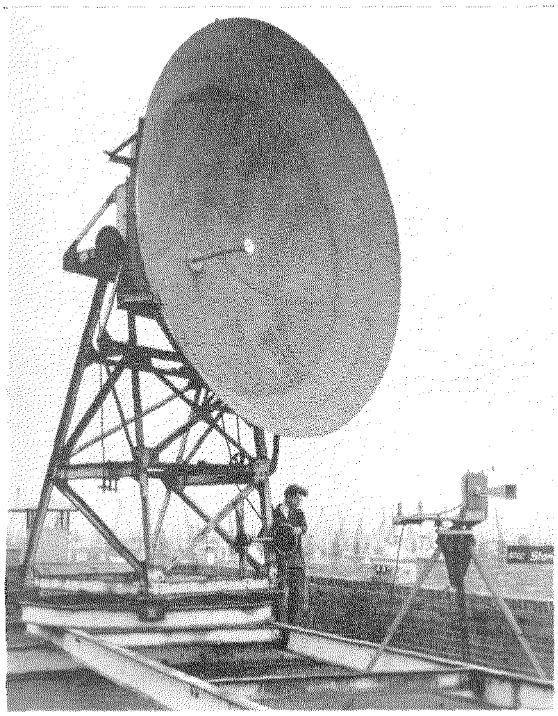

Figure 10—Aerial mounted on adjustable tower.

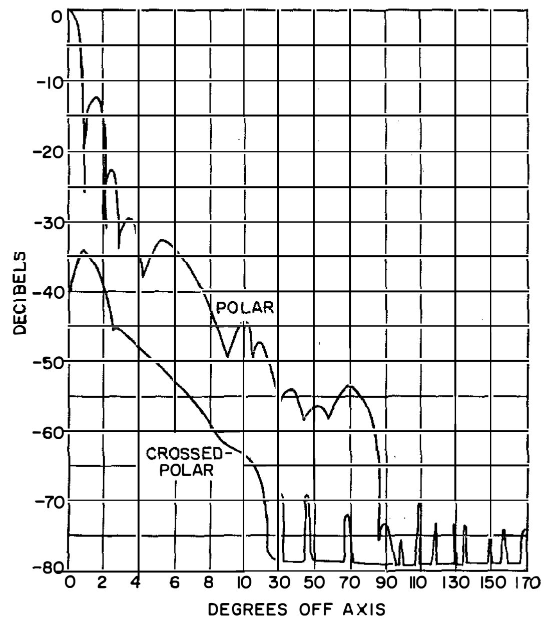

Figure 11—Polar and crossed-polar diagrams in *E* plane at 6.2 gigahertz. The paraboloid is 13 feet (4 meters) in diameter. The gain is 46 decibels and efficiency is 60 percent. The abscissa scale is expanded below 10 degrees.

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5.4 Impedance and Bandwidth

The reflection coefficient of the complete horn and reflector assembly was determined in the laboratory for the frequency range from 5.9 to 6.8 gigahertz and is shown in Figure 13. From 5.9 to 6.45 gigahertz the reflection is less than 1 percent, but it rises to 4 percent at 6.75 gigahertz. The reflection of the antenna complete with apex plate and horn feed was shown to agree with Figure 13.

5.5 Design

An engineered version of the cassegrain antenna has been produced. It comprises an aluminum spinning 9 feet (2.7 meters) in diameter surrounded by pressed aluminum sectors. The whole is mounted on a strong aluminum framework as shown in Figure 9.

The focal length of the antenna is 27.56 inches (70 centimeters). The effective diameter is 12 feet 4 inches (3.76 meters). The overall diameter is 12 feet 8 inches (3.86 meters). The gain predicted is 45 decibels; this allows a reduction of 0.6 decibel from the measured figures because of its smaller size, and a further reduction of 0.4 decibel for possible errors in the profile of the main reflecting surface.

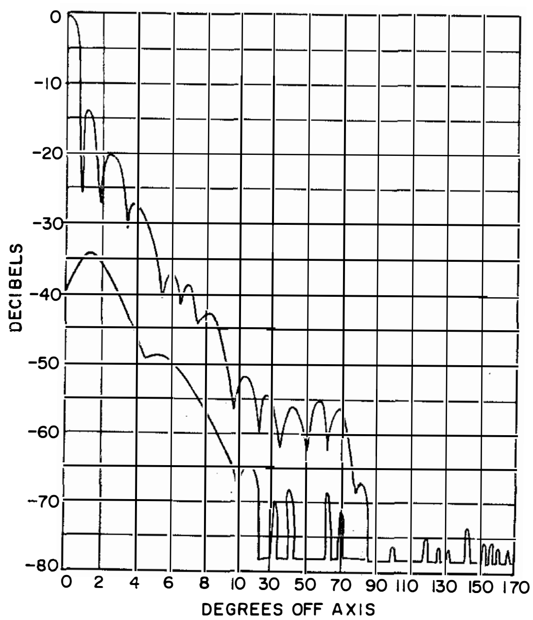

Figure 12—*H*-plane plot corresponding to Figure 11.

6. Performance Comparisons

The cassegrain antenna has been developed for use on super-high-frequency links as an alternative to existing antennas, of which the principal ones in general use are the horn-fed paraboloid and the horn reflector antenna.

6.1 Polar Diagram

Polar diagrams for cassegrain, paraboloid, and horn types are illustrated in Figures 11, 14, and 15, respectively, for designs giving 45 or 46 decibels of gain. It will be seen that the horn reflector aerial of Figure 15 is best for the normal polar diagram in the region from 30 to 90 degrees and the conventional paraboloid (Figure 14) is the worst. However, the cassegrain is best for crossed-polar properties at all angles, as a consequence of its circular symmetry. This property of the cassegrain antenna can be put to good use on crossing routes. For instance, decoupling of 68 decibels is achieved at all angles exceeding 20 degrees, thus enabling spur and crossing routes to be engineered to work on the same frequency as the main route.

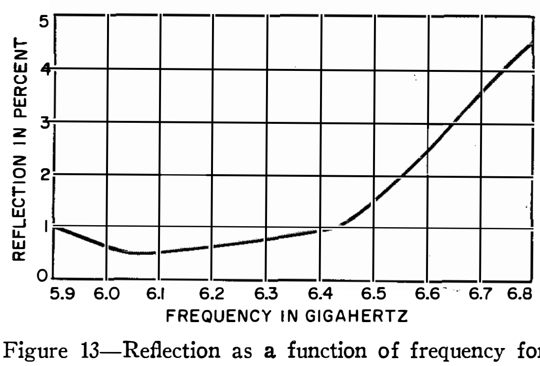

Figure 13—Reflection as a function of frequency for the complete cassegrain aerial.

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TABLE 1
COMPARISON OF CASSEGRAIN AND HORN REFLECTOR ANTENNAS

Parameter	Cassegrain	Horn Reflector
Gain in decibels	45	45
Radiating area in square feet (square meters)	120 (11.2)	110 (10.2)
Windage area in square feet (square meters)	126 (11.7)	192 (17.8)
Height in feet (meters)	12.6 (3.8)	25 (7.6)
Width in feet (meters)	12.6 (3.8)	14 (4.3)
Depth in feet (meters)	5 (1.5)	11 (3.6)
Weight in pounds (kilograms)	920 (417)	1700 (771)
Wind load in pounds (kilograms) at 100 miles (160 kilometers) per hour*	5000 (2268) at 0 degrees 6250 (2835) at 45 degrees 3100 (1406) at 180 degrees	6800 (3084) at 0 and 180 degrees 8100 (3674) at 45 degrees 4500 (2041) at 90 degrees
Operating band in gigahertz	5.9–6.45	3.8–4.2 5.9–6.45 10.7–11.7
Maximum reflection coefficient in percent	2	1

* The wind load on the paraboloid at 0 and 45 degrees is reduced if a radome is used.

6.2 Efficiency

The efficiency of the cassegrain antenna is 60 percent, which is a little better than the conventional paraboloid with 90-degree aperture. The efficiency of the normal paraboloid can be improved without degradation of the wide-angle side lobes by increasing the focal length and adding an absorbing cowl around the periphery of the paraboloid. However, this approach is expensive and adds to the weight and the area of the aerial subject to the wind. The efficiency of the horn reflector antenna is 65 percent, but the overall size is much greater for a given gain than that of a paraboloid because of the large pyramidal cone.

6.3 Impedance

The reflection of the model cassegrain antenna was less than 1 percent over the 6-gigahertz band. It is expected that production can proceed with a permissible test limit of 2 percent. This is better than is currently achieved on swan-neck paraboloids. However, the reflection of the horn reflector aerial is better, being less than 1 percent over the band from 4 to 11 gigahertz.

6.4 Summary

The main parameters of the engineered version of the 6-gigahertz cassegrain are compared with the horn reflector antenna in Table 1.

7. Utilization on *1800/TV* Links

7.1 Antenna and Feeder Arrangement

The 6-gigahertz cassegrain antenna has been designed with a 1.75-inch (4.45-centimeter) horn feed of circular aluminum waveguide. The same circular waveguide is used for the main feeder to the aerial from ground level, thus providing a convenient low-loss bipolar feeder arrangement. The connection from feeder to aerial is by means of a large-radius bend in the circular waveguide. This arrangement enables the total reflection at the masthead to be kept at about 2 percent, which is adequate for use on 1800-channel systems.

The cassegrain aerial and feeder may be used in single or bipolar operation in any of the combinations permitted by the recommendations of the International Radio Consultative Committee.

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7.2 Aerial Arrangements at City Centers

If numerous routes converge on a city center, the aerial side-lobe performance is important in determining the maximum number of radial directions permitted.

For 1800-channel systems operating on the same frequency, it is estimated that antenna discrimination of 65 decibels is required. Using the 6-gigahertz cassegrain antenna, it is then possible to operate two routes on one polarization if they are at angles greater than 85 degrees to each other and with other routes on the opposite polarization at angles of 15 degrees or more. In this way a total of 8 radial routes is possible.

With the horn reflector antenna, little advantage is obtained by cross-polarizing adjacent directions. The minimum spacing between routes, according to Figure 15 for horn reflectors, is thus 50 degrees for the *H* plane and 70 degrees for the *E* plane. Hence a maximum of 7 radial routes can be operated in the *H* plane, but only 5 routes in the *E* plane.

8. Relative Costs

Since the cassegrain antenna is circularly symmetrical, the reflecting surface can be made by spinning it in one piece. The technique minimizes tool costs, is satisfactory for diameters up to 13 feet (4 meters), and avoids all problems of sealing joints.

By contrast, the horn reflector antenna consists of an offset sector of a paraboloid, enclosed in a rectangular pyramid. The reflecting surface cannot be spun, but is made by stretch-forming from sheet metal or by metal spray on a fiberglass surface. It is essential to provide a large

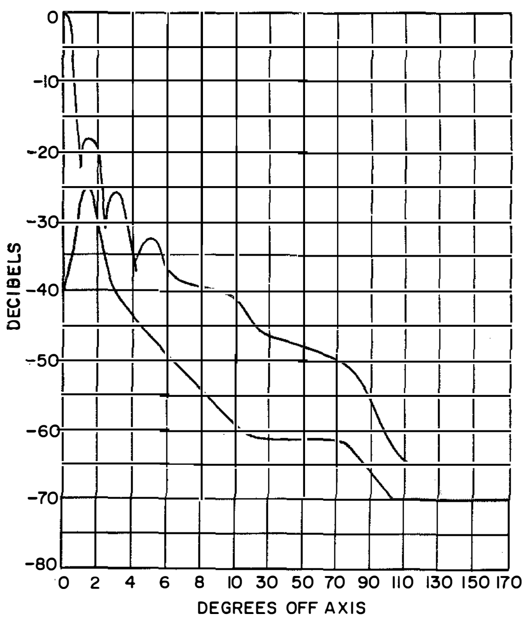

Figure 14—Polar and crossed-polar plots of paraboloid with swan-neck feed in *E* plane at 6.2 gigahertz. The gain is 45 decibels and efficiency is 50 percent. The paraboloid is 13 feet (4 meters) in diameter. The aperture is 90 degrees.

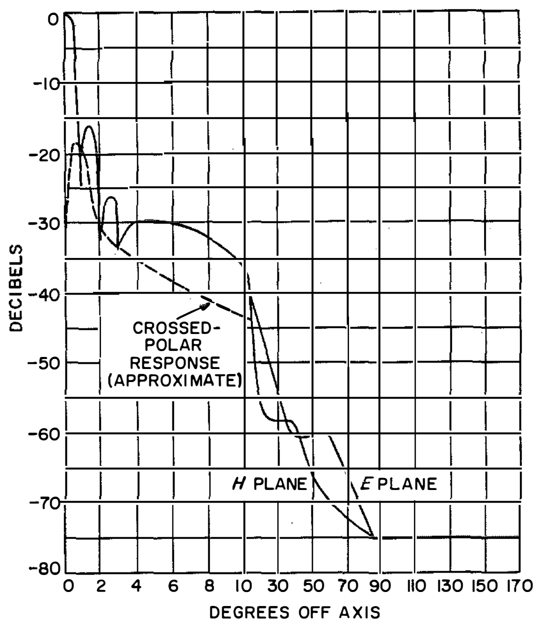

Figure 15—Polar and crossed-polar plots of horn reflector antenna in *H* and *E* planes at 6 gigahertz. The gain is 45 decibels and efficiency is 65 percent. The height is 25 feet (7.6 meters) and the width is 14 feet (4.3 meters).

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transparent window in the horn reflector to prevent ingress of rain or dirt. The body must also be made in a number of sections to facilitate transport. Sealing the joints constitutes a major problem, as the jointing material must operate electrically and also seal against rain and air. Prevailing manufacturing costs for the horn reflector antenna are some 3 to 4 times those of the cassegrain type. In addition, the shape of the horn reflector makes special demands on the tower, which must be designed from the start for such equipment. The cassegrain antenna, on the other hand, can be mounted on any existing tower with little modification to the tower.

9. Conclusions

The new cassegrain aerial represents a significant improvement over previous paraboloidal antennas, including the horn paraboloid, and is designed for use on trunk routes carrying 1800-channel traffic. The aerial is efficient and economical and should replace existing designs of paraboloids by virtue of its improved performance and lower cost.

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He worked on the development of transmitting tubes and microwave devices at Mullards for 6 years and was then at the Patent Office for a year.

Mr. Ware joined Standard Telephones and Cables transmission division in 1955 and is responsible for development of microwave aerials and feeders for use on radio relay systems. He has also designed ferrite isolators and circulators.

Graham C. Stemp was born in London, England, in 1936 and received an Honours degree in Physics from the University of Bristol in 1960.

He joined Standard Telephones and Cables in 1960 as a graduate apprentice and spent 3 years on the development of aerials, feeder components, and ferrite devices. At present he is responsible for the design and development of input and output mixers for communication systems.

Mr. Stemp is a Graduate Member of the Institution of Electrical Engineers.

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Microwave Radio System for Multichannel Telephony and Television in the 6-Gigahertz Range
Part 5—Auxiliary Radio Relay System *BFM24/6000*

A. LIEKENS
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1. Introduction

Experience with large-capacity long-haul radio relay systems over a number of years has led, with respect to the requirements for remote control and supervision, to the concept of an auxiliary radio relay system integrated with the main system. This auxiliary system uses the same antenna branching network and works within the same frequency plan, but is completely independent of the main line equipment. It has been used successfully in the 4-gigahertz band. The International Radio Consultative Committee has also recommended for the 6-gigahertz band a frequency plan for such an integrated system.

This article describes the auxiliary radio relay equipment developed for the remote control and supervision of the *1800/TV* radio relay system in the 6-gigahertz band.

2. General

The design of an integrated auxiliary radio relay system according to the frequency plan recommended by the International Radio Consultative Committee (recommendation 389), is strongly influenced by the requirement for interference-free operation in a completely equipped system using all 8 radio-frequency channels. The level of interference between the auxiliary channel and the main channel must be reduced below an acceptable value, both under normal field conditions and under fading conditions that can occur in a well-planned system.

Economic considerations, plus the fact that some of the main line system parameters are fixed, are additional constraints that have required the optimization of some parameters of the auxiliary system as, for example, the transmitter power and degree of radio-frequency filtering.

The number of telephone channels generally required ranges from a few up to 12 channels. For the 6-gigahertz system, 4 channels are assigned as follows.

Channel 1: Express order-wire speech channel between control stations equipped with voice-frequency signaling.

Channel 2: Omnibus order-wire speech channel between stations within a control section, with party-line voice-frequency signaling facility.

Channel 3: Voice channel available as a 4-wire circuit for high-speed protection-channel switching.

Channel 4: Voice channel available as a 4-wire circuit for transmission of voice-frequency sub-multiplexed remote-indication and remote-control signals.

Generally one or two additional channels are desirable to meet special requirements of the customer.

The radio-frequency equipment *BFM24/6000* provides for a maximal capacity of 24 telephone channels spaced at 4-kilohertz intervals and multiplexed in the frequency band from 6 to 108 kilohertz. For requirements of up to 12 channels, the band from 60 to 108 kilohertz is preferred to avoid the need of additional translating equipment in the multiplex process.

The finally adopted nominal transmitter power of 10 milliwatts made it possible to design an equipment that uses only semiconductors. All transistors used are of the silicon planar epitaxial type. Their operating levels are kept well below the highest allowed ratings to achieve a high degree of reliability. The all-solid-state approach has led to a small and compact design and low power consumption.

3. Radio Equipment

The radio-frequency equipment uses direct modulation, and the signal is demodulated at

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each repeater to provide baseband access in all repeater stations. The baseband signal phase modulates a crystal-controlled frequency. Subsequently, this modulated signal is amplified and multiplied in frequency. The required output power is obtained through high-efficiency varactor multipliers. In addition to the stability from starting with a crystal-controlled oscillator, the high order of multiplication after modulation avoids the nonlinearity distortion associated with the direct production of a large modulation index. The selected system results in noncritical adjustments and easy maintenance.

3.1 Technical Characteristics

Transmitter channel allocation in megahertz	6171.9 6423.9 5926.1 6178.1
Local-oscillator frequencies in megahertz	6136.9 6388.9 5961.1 6213.1
Cumulative frequency error in kilohertz on a terminal pair	±300
Transmitter power in milliwatts	10
Phase excursion for 0 decibels relative to 1 milliwatt referred to a point of zero reference level	1 radian root-mean-square at 100 kilohertz
Pre-emphasis in decibels per octave	3
Receiver intermediate frequency in megahertz	35
Intermediate-frequency selectivity in megahertz at 3-decibel-down points	1.5
Receiver noise figure in decibels	12
Fading margin in decibels	30
Image rejection in decibels	70
Radio-frequency input and output impedance in ohms	50
Baseband characteristics:	
Frequency range in kilohertz	6 to 108
Input level in decibels referred to the level at the 2-wire point of origin	−45
Output level in decibels referred to the level at the 2-wire point of origin	−15
Impedance in ohms (unbalanced)	75
Power consumption at usual alternating voltages or at 24 direct volts	average 30 volt-amperes
Weight in kilograms (pounds)	16 (35)

Measurements made with a white-noise loading of +4.5 decibels referred to the level at the 2-wire point of origin, corresponding to 24-channel loading, show that the total intermodulation plus thermal noise in one channel is between −60 and −62 decibels relative to 1 milliwatt referred to a point of zero reference level, psophometrically weighted, depending on the channel number for a normal path attenuation of 62 decibels.

The equipment is suitable for operation in the temperature range from −25 to +55 degrees centigrade with a relative humidity of 90 percent.

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3.2 Circuit Description

3.2.1 *Transmitter*

In Figure 1, the crystal-controlled oscillator in the transmit chain that operates between 46.3 and 50 megahertz is temperature stabilized to meet the stringent stability requirements. This oscillator is followed by a buffer amplifier and then a phase modulator consisting of a circuit similar to a frequency discriminator. In the modulator two orthogonal signals that are amplitude modulated according to the baseband signal amplitude are added to a carrier signal to produce a phase-modulated signal at the carrier frequency. This signal is further amplified. The modulating baseband signal is applied through a pre-emphasis network that gives a 3-decibel-per-octave characteristic (from 6 to 108 kilohertz).

The phase-modulated signal is amplified, limited, and then multiplied by 4 through 2 doubler stages to about 200 megahertz with an output increased to about 1 watt in the driver.

The 200-megahertz signal is further converted to 800 megahertz in the varactor quadrupler unit, which uses a varactor mounted in a strip-line configuration incorporating the necessary idler circuits. The efficiency of this unit is approximately 40 percent.

Final multiplication by 8 is achieved in the varactor octupler and filter unit using a combination of coaxial and waveguide mounts. The coaxial part contains a low-pass input filter and the varactor, whereas the output band-pass filter consists of a 3-section waveguide structure that selects the 8th harmonic of the input frequency. The efficiency of this octupler is better than 15 percent.

The total multiplication factor for the complete transmitter chain is 128.

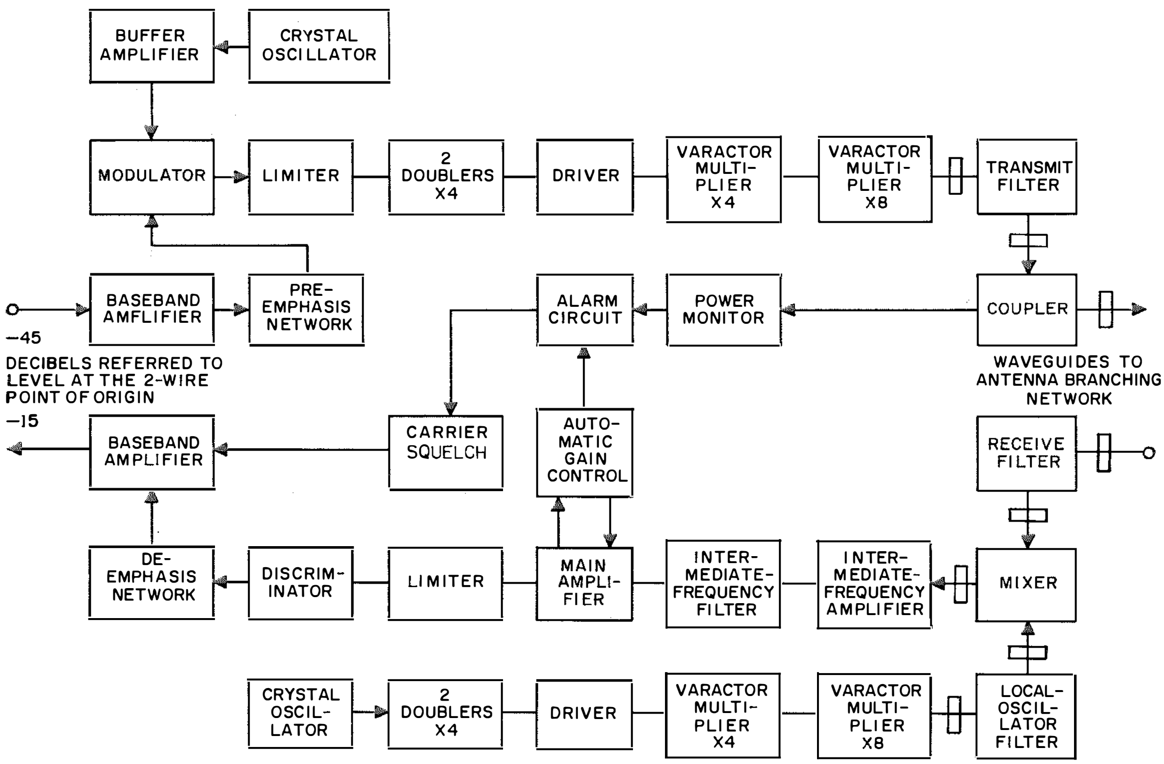

Figure 1—Block diagram of *BFM24/6000* equipment.

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The output signal to the transmitting-antenna waveguide goes through a circulator that is part of the antenna branching network.

3.2.2 *Receiver*

The connection to the receiving antenna waveguide is made through a circulator from the antenna branching network.

The local-oscillator unit is similar to the transmitter chain except that the modulation circuit is omitted.

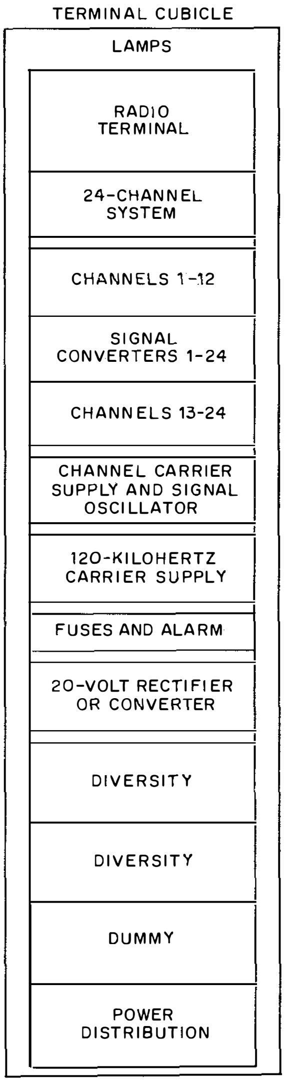

Figure 2—Typical 24-channel auxiliary-link-terminal rack.

The received super-high-frequency signal is filtered by a 5-cavity preselector, which suppresses images and prevents local-oscillator radiation. The desired signal is passed to a waveguide balanced mixer terminated by a low-noise 35-megahertz intermediate-frequency amplifier, which has a gain of approximately 20 decibels. The waveguide mixer and amplifier form one mechanical unit.

A 4-section filter, placed between the intermediate-frequency amplifier and the main amplifier, determines almost completely the selectivity characteristics of the receiver. It consists of a printed circuit having a 1.8-megahertz bandwidth; combined with the selectivity of the main amplifier, an overall 1.5-megahertz bandwidth results. The 35-megahertz main amplifier consists of 5 amplifier stages in cascode arrangements. As is well known, such arrangements give excellent isolation between adjacent stages, which allows straightforward noncritical alignment and assures good stability of the circuit. Automatic gain control is used.

The main amplifier is followed by a limiter-discriminator. The necessary limiting to assure proper frequency demodulation is performed in two cascaded stages operating into a conventional discriminator. The demodulated signal is applied to a de-emphasis network identical to the pre-emphasis network of the modulator in the transmitter. These two 3-decibel-per-octave circuits compensate for the 6-decibel-per-octave characteristic resulting from the demodulation of the phase modulation by the frequency discriminator.

The demodulated signal is further amplified to the nominal output level by the baseband amplifier for connection to the multiplexing equipment. The output level can be adjusted in steps of 0.5 decibel by attenuator pads over a range of ± 3 decibels from the nominal level.

3.2.3 *Power Supply, Monitoring, and Alarm Units*

All the auxiliary functions for a complete transmitter-receiver are concentrated on a common

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panel. This panel includes the controls for a regulated power supply that operates either on mains or on batteries, and for the alarm circuits controlled by the transmitter output power level and the automatic-gain-control voltage of the receiver. It also mounts the monitoring meter with associated multiswitch for checking all values important for equipment maintenance.

4. Multiplex Equipment

The frequency-division-multiplex equipment associated with the radio equipment uses the normal channel allocation recommended by the Comité Consultatif International Télégraphique et Téléphonique; that is, the bands from 12–60 or 6–54 and from 60–108 kilohertz for the two groups of channels.

In many cases only a few telephone channels are required, and for economic reasons channel carriers are then provided by individual crystal oscillators that can be built into an oven if high stability is required for voice-frequency telegraph transmission.

For repeater equipment, conference networks are provided to allow through connections as well as local access to other stations.

5. Mechanical Layout

The mechanical standard used is the modular technique followed in the standard equipment practice for ITT Europe [1].

The subunits necessary for the auxiliary radio relay equipment are mounted in a rack (Figure 2), which is in line with the racks of the main line equipment. This rack contains 1 or 2 radio-frequency subracks, 1 multiplex equipment (normally with 6 channels), and conference networks for a repeater station. Additional space is available for remote-control and indication equipment.

The rack is also equipped with a panel that gives the alarm condition of the units in the rack.

5.1 Radio Equipment

The radio equipment is divided mechanically into a subrack that contains the power supply and all units with maximum frequency of 800 megahertz, and a platform that mounts all waveguide structures (see Figure 3).

All units mounted in the subrack are die cast and of the plug-in type. Each is a well-screened box with a multiconnector at the rear for direct-current connections, while signal connections are made at the front through subminiature coaxial connectors. This modular construction is particularly suitable for maintenance and testing.

5.2 Multiplex Equipment

The multiplex equipment is contained in plug-in modules mounted on shelves. Signal access is to the fixed part of the shelves rather than directly to the plug-in units.

6. Conclusion

An all-solid-state auxiliary radio relay equipment has been designed for the 6-gigahertz band that meets all the operational requirements for interworking with the *1800/TV* radio relay system. The auxiliary equipment is electrically and mechanically compatible with the main line equipment.

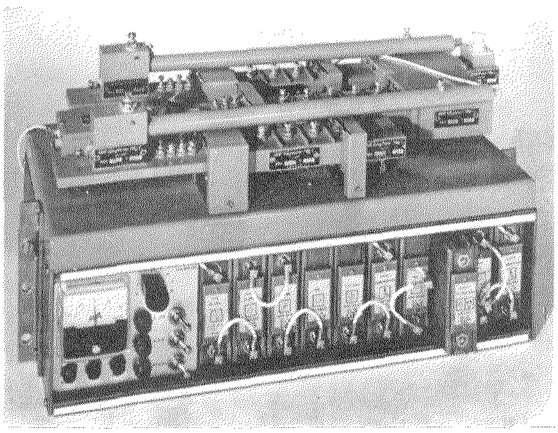

Figure 3—Mechanical design of the radio equipment.

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7. Reference

1. F. Beerbaum, J. Evans, and F. Leyssens, "Standard Equipment Practice for ITT Europe," *Electrical Communication,* volume 39, number 2, pages 199–211; 1964.

Albert Liekens was born in Kieldrecht, Belgium, on 19 January 1921. He obtained the degree of electrical and mechanical engineer at the De Nayer Institute of Mechelen in 1941. In 1961 he graduated in mathematics at the University of Leuven.

In 1941 he joined the Bell Telephone Manufacturing Company and in 1963 was appointed chief engineer of the radio and line transmission division.

Étienne Reygaerts was born in Enghien, Belgium, on 7 July 1927. He graduated as an electrical and mechanical engineer from the University of Brussels in 1950.

In 1951 he joined the Bell Telephone Manufacturing Company and since 1962 has been in charge of the radio-link and system-planning group of the radio and line transmission division.

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Carrier Telephone System for Mine Railroads *

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E. REMPP

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A communication system that permits all locomotive motormen in a mine to be reached from a central point contributes to an easier, safer, and less-expensive flow of all traffic. The dispatcher, usually stationed in the vicinity of the main shaft, should be able to communicate with the locomotive drivers in an extensive branched mine railroad regardless of distance and train speed. As the free uniform propagation of radio waves found above ground does not occur underground, radio communication of this type is possible in most cases only over line-of-sight paths that are usually quite short [1].

Although experiments [2–6] had been made in communicating over the catenary of power wires that supply energy to the usual electrically driven machines in mines or via the compressed-air and water distributing pipes, these systems still were unable to meet all requirements under various operating conditions. For this reason the type *Z1G* carrier telephone system for mine railroads was developed.

1. Considerations

The system should provide satisfactory communication over a distance of 20 kilometers (12.5 miles). To transmit the signal, the central station is directly coupled to a 2-wire transmission line to which the locomotive stations are inductively coupled via loop antennas.

The frequencies used are between 35 and 135 kilohertz, within which range the attenuation of the 2-wire system is low enough for the distances involved. Below 35 kilohertz, the noise level is increased by harmonics of the power supply and by interference from rectifiers and commutator-type electric machines. The upper frequency limit is set by the regulations governing all radio services.

All equipment was designed for both simplex and duplex operation and for alternative use of a telephone handset or a microphone and loudspeaker. Duplex operation requires a separate carrier frequency in each direction. This type of operation is preferable for mines as no special training of personnel is needed. It provides communication between the central and locomotive stations but not between two locomotive stations. However, there are mines where the drivers should be able to intercommunicate. In such a case simplex operation is possible by using the central station as a relay station.

Frequency modulation provides an excellent means to care for even-large level fluctuations. Considering the level differences between transmitter and receiver, the two carriers are spaced by 40 kilohertz. The low transmission frequency restricts the modulation frequency deviation so that the full benefit of wide-band frequency modulation is not attained. However, the amplitude limiting saves a costly control circuit that would be required in amplitude modulation.

The *Z1G* system is applicable to all types of mine railroads and road vehicles, independent of the type of propulsion. In most mines, mixed operations using overhead wires, storage batteries, diesels, and compressed-air locomotives are encountered. The central station is usually supplied from the power line, while power supplies have been designed for connection to available converters, generators, or dynamos in the mobile stations.

The design was governed by the conditions found in rough mining operations and by the safety specifications of the mining authority. The units use only transistors as active elements, extending life and virtually eliminating maintenance.

* Revision of a paper originally published in German in *SEL-Nachrichten,* volume 10, number 4, pages 179–187; 1962.

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2. Configuration of the 2-Wire Transmission Line

The 2-wire transmission line had to cover the whole railroad network. In the frequency range from 35 to 135 kilohertz, inductive coupling is the preferred way to link the wire line and the mobile locomotive station. Metallic coupling is out of the question because of the mechanical difficulties, and capacitive coupling was not used because the achievable coupling capacitances are small and require therefore a high input resistance.

A 2-wire transmission line terminated in its characteristic impedance offers much-better and more-uniform transmission characteristics than a single wire against ground, because of the undefined nature of the "ground" and the adverse influence of other long metal structures, such as cables and pipes. The ambient rock has a much-smaller effect on the properties of the 2-wire line, which can be laid along the roof or the side walls of the railroad tunnel.

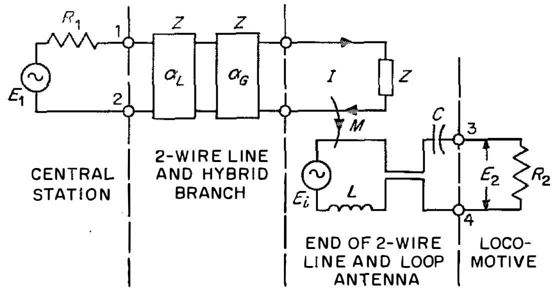

Figure 1—Equivalent circuit between central station and locomotive.

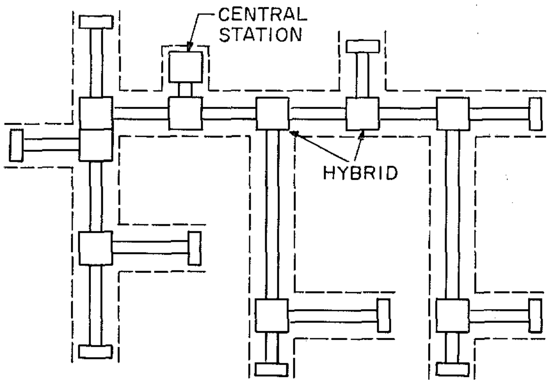

Figure 2—Diagram of 2-wire system in a large mine. All lines are connected to hybrids at branching points or to terminating impedances at their ends.

The transmission path between central station and locomotive is shown schematically in Figure 1. The central station is directly coupled to the 2-wire line and applies an alternating voltage E_1 via its internal resistance R_1 to input terminals 1 and 2 of the wire network. The internal resistance R_1 approximately equals the characteristic impedance Z of the line represented by the 4-pole networks with the line loss α_L and the branching loss α_G. Here α_L results from the length of line and the associated attenuation constants, while α_G represents the attenuation of the hybrid circuits (differential transformers) inserted into the branching joints of the line. At its end, the line is terminated with its characteristic impedance Z.

Assume a locomotive station coupled by a loop antenna to the line close to its end. The current I induces, via the mutual inductance M, an alternating voltage

$$E_i = \omega_0 MI \qquad (1)$$

in the loop antenna. The inductance L of the loop antenna and the tuning capacitor C form a series-resonant circuit for the carrier frequency f_0. This circuit is attenuated by the station input resistance R_2 (Figure 1), thus determining the bandwidth of the transmission path.

It is possible to define, between terminals 1–2 and 3-4, a passive 4-pole network presenting the same attenuation α_B in forward and backward directions of the carrier f_0. This attenuation corresponds to the ratio of the transmitter power output P_1 to the power received P_2.

$$\alpha_B = \ln\,(P_1/P_2)^{1/2} = \ln\left(\frac{E_1^2 R_2}{4R_1 E_i^2}\right)^{1/2}. \qquad (2)$$

Substituting (1) in (2) and defining

$$I = \frac{E_1}{2}\, e^{-\alpha_L}\, e^{-\alpha_G}\, \frac{1}{Z} \qquad (3)$$

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we obtain the overall attenuation between terminals 1–2 and 3–4 in both directions.

$$\alpha_B = \alpha_L + \alpha_G + \ln \frac{(ZR_2)^{1/2}}{\omega_0 M}. \qquad (4)$$

It is convenient to overlook the line loss α_L and the branching loss α_G of a 2-wire transmission line (or line as it will be called hereafter). The last term of the sum in (4) is the coupling attenuation α_k between the loop antenna and the line. It changes with the mutual inductance as varied by local conditions affecting the spacing between antenna and line. In some cases, for instance, it is impossible to mount the line to the roof of a tunnel because air or gas pipes are already there. The two wires then must be laid along the two faces at the sides of the railroad track.

Propagation conditions underground have been thoroughly investigated [7]. It was found that the coupling attenuation α_k may vary between 3 and 8 nepers (26 and 70 decibels).

Figure 2 shows a mine railroad system with lines connecting a central station to each heading and crosscut. Hybrids at each joint decouple the two branches. If a short circuit or the like occurs in one branch, the other branch remains unaffected. Any trouble can thus be conveniently located. A calculated attenuation of 0.4 neper (3.5 decibels) can be assumed for each joint. If 5 hybrids in tandem are used—sufficient in most cases—the branching loss amounts to 5 times this or 2 nepers (17.5 decibels).

Both mechanical and electrical considerations are involved in the choice of the line. The greatest distance between central station and the farthest locomotive, 20 kilometers, must be bridged. Considering a line loss α_L of 2 nepers (17.5 decibels) as the maximum permissible, the attenuation constant for the operating frequency is 0.1 neper per kilometer (1.4 decibels per mile). Figure 3 shows attenuation constants of various conductor types; type D, developed especially for rough operations as in mining, is most suitable for this application. It is constructed of 7 strands, each 0.8 millimeter (0.03 inch) in diameter, 3 of which are of copper and 4 of steel. It has high tensile strength and is flexible at the same time. The insulating material is polyvinylchloride.

Considering the line loss, branching losses, and coupling attenuation, a maximum overall loss α_B of 12 nepers (104 decibels) must be anticipated in dimensioning the transmission equipment. If one of the attenuation components can be reduced, the tolerances of the others may be relaxed. If, for instance, careful arrangement of the loop antenna results in a coupling attenuation not exceeding 6 nepers (52 decibels), then a distance of 40 kilometers (25 miles) can be covered even with 5 hybrids in the transmission path. Thus there are means to meet even extreme specifications.

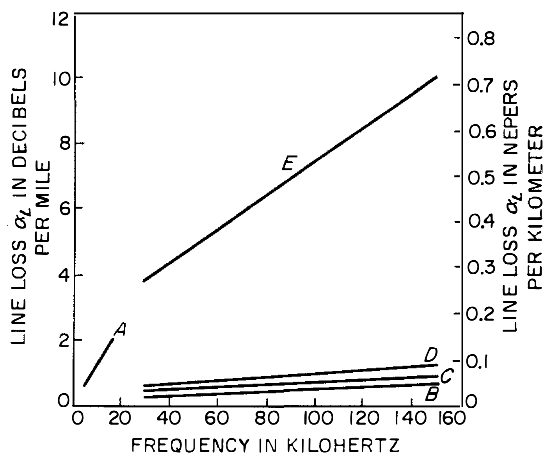

Figure 3—Attenuation constants of 2-wire lines. Curve A is for an open-wire steel line 2 millimeters (0.1 inch) thick, dry weather, with wire spacing of 0.3 meter (1 foot). Curve B is for an open-wire copper line 3 millimeters thick, rainy weather, with the same wire spacing. Curve C is for a line below ground with a cross-sectional area of 1.5 square millimeters (0.002 square inch), with wire spacing of 1.3 to 2 meters (4 to 6.5 feet). Curve D is for a line below ground of 7-stranded wire (3 copper, 4 steel including center strand), each strand 0.8-millimeter (0.03-inch) thick, with wire spacing of 2 meters (6.5 feet). Curve E is for a line below ground of 7-stranded wire (center strand of copper, 6 steel), cross-sectional area of each strand 1.5 square millimeters (0.002 square inch), with wire spacing of 2 to 4 meters (6.5 to 13 feet).

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The range of regulation of the equipment is determined by the lowest and highest overall attenuation α_B. Assuming a locomotive close to the central station and an optimum coupling attenuation of 3 nepers (26 decibels), the overall operating attenuation α_B can be between 3 and 12 nepers (26 and 104 decibels), from which it follows that the range of regulation should be 9 nepers (78 decibels).

3. Types of Communication

The *Z1G* carrier telephone system provides not only duplex telephone traffic between central station and mobile stations, but also simplex operation between mobile stations. This is a substantial improvement over the equipment thus far used with underground railroads in mines.

In duplex, a separate carrier is used for each direction (2-wire operation). Transmission and reception are thus possible at the same time. In simplex operation, the direction of transmission is changed as agreed by the parties to a conversation; any station can either receive or transmit at a particular time, therefore only one carrier is necessary. Equipment for duplex is costlier because of the selective filters and branching filters required to decouple the transmitter from the receiver; on the other hand, duplex may be required in some applications.

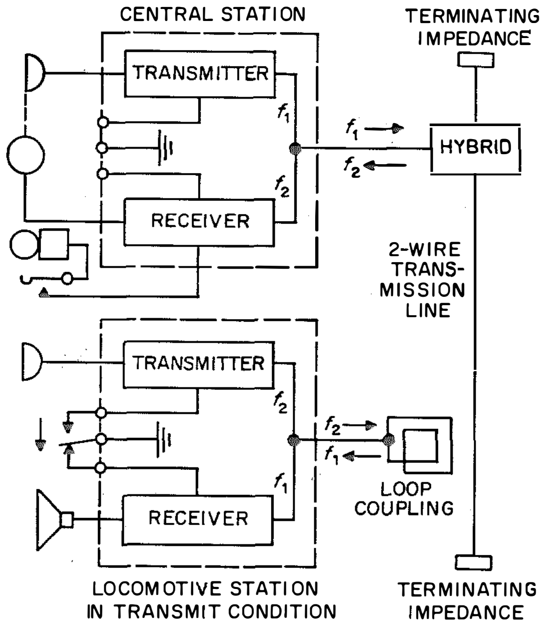

Figure 4—Duplex operation between central station and mobile station.

3.1 Duplex

The most-frequent application is probably duplex conversation between operators of a central station and a mobile station in deep mining, surface working, crane operating, et cetera. The principle of duplex may be seen from Figure 4. The central station is directly connected to the line network, the ends of which are terminated by resistors corresponding to the characteristic impedance of the line. Branches are joined via hybrids. Mobile stations are inductance coupled to the line via loop antennas.

The central station continuously emits its carrier f_1, and each locomotive driver is able to identify an interruption in this carrier by a signal appearing on his equipment. The central station in most cases uses a handset hung on a hook-type gravity switch, while the mobile stations are usually equipped with key-operated hand microphones and pneumatic (pressure chamber) loudspeakers although, of course, all these transducers are interchangeable.

To call the central station, the mobile-station driver operates his microphone key and thus transmits frequency f_2, which appears in the central receiver and operates a ringer until the operator lifts the handset. If a dispatcher wants to talk to a driver, he simply calls the latter by name. All drivers hear the dispatcher's announcement, but not the reply. The drivers cannot call or talk to each other.

3.2 Simplex

In some cases it is desirable that drivers communicate with each other. Figure 5 shows the simplex arrangement then used. The central station, directly coupled to the line network,

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serves as a relay repeater. It receives frequency f_2 from the transmitting mobile station and demodulates the voice-frequency signal. This signal is applied to the central transmitter and modulates carrier f_1, which is supplied to the line and reaches all mobile stations coupled to the line by loop antennas. The parties talk alternately; each one operates a changeover switch to reverse the direction of communication. All drivers may listen in.

Generally the central station is unattended in the repeater mode. However, normal dispatching may still be required and the central-station operator may monitor the traffic and send messages to drivers through an extra mobile installation that is directly coupled to the line.

4. Station Details

Since the *Z1G* system was planned for universal application, the equipment was divided into units: transceiver, power supply, and accessories. This permits effective use of the limited space in the cab of a mine railroad locomotive, direct adaptation to the locomotive power supply, and simplified maintenance.

4.1 Electrical Design

If a carrier is limited to 1 watt and is modulated by voice, the maximum noise power of 5×10^{-8} watt existing at the highest permitted frequency of 135 kilohertz results in a signal-to-noise ratio within a tolerable limit.

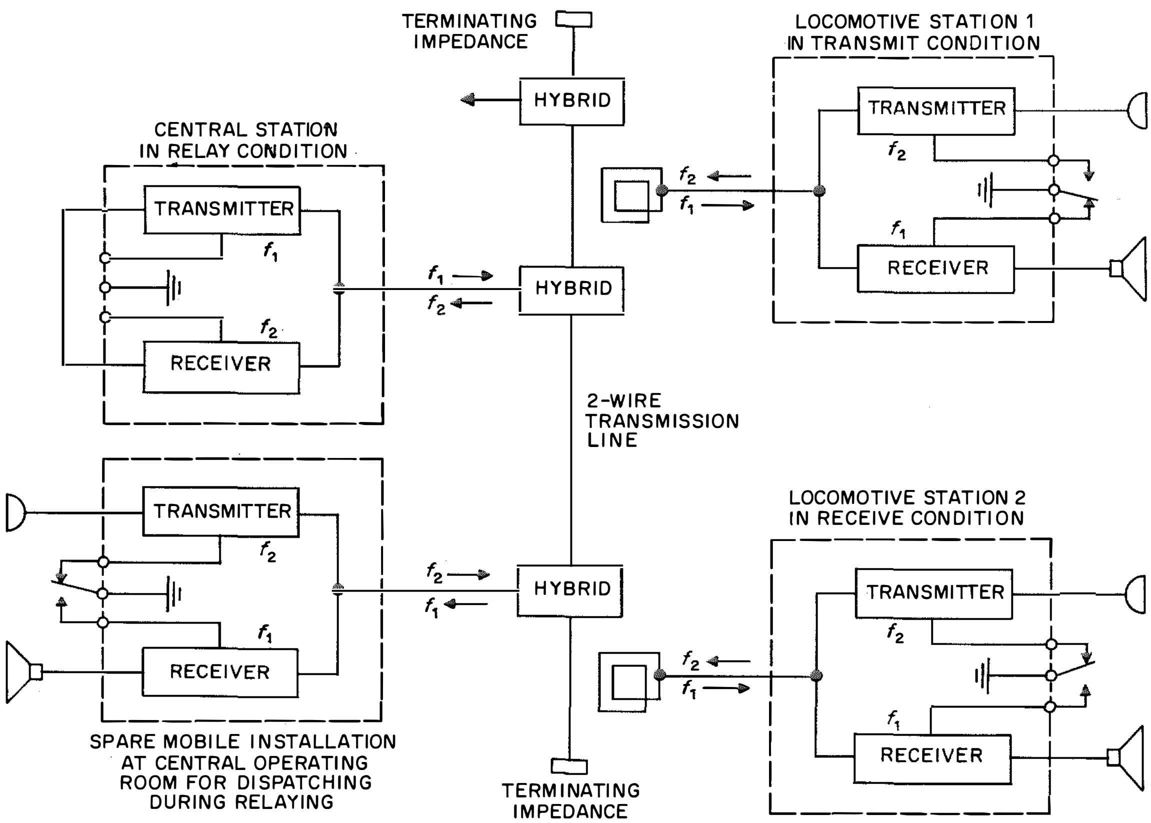

Figure 5—Simplex operation between mobile stations.

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Government regulations to prevent interference among all radio services have limited the highest carrier frequency for these mobile transmitter stations to 110 kilohertz. The second carrier frequency is determined by the receiver selectivity that can be achieved at reasonable cost. The transmitter at any station must not interfere with the simultaneous reception of the weakest station, usually the most-distant one.

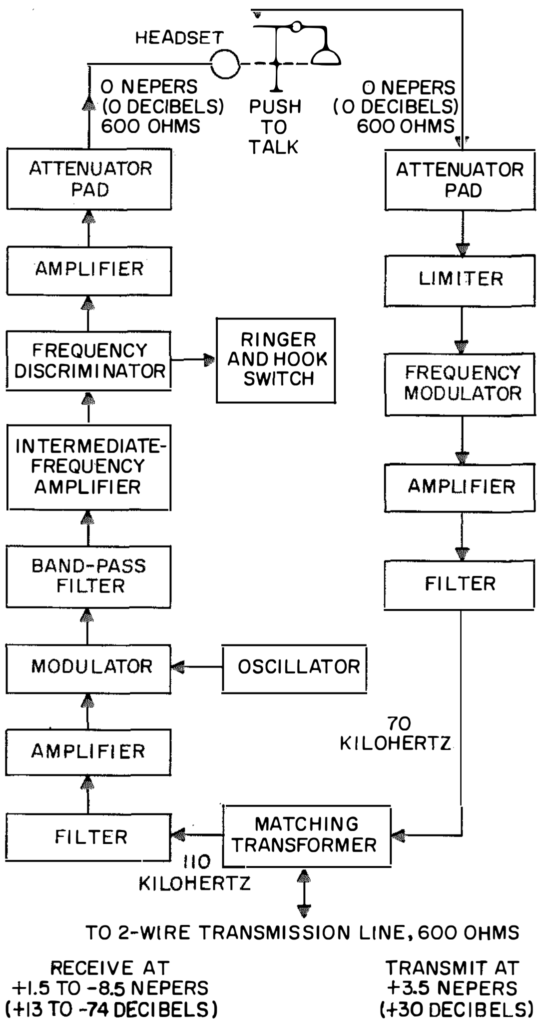

Figure 6—Diagram of a central station. In a mobile station, the handset is replaced by a separate microphone and loudspeaker, the two attenuator pads are replaced by amplifiers, the transmitting and receiving frequencies are interchanged, and a different type of line coupling is used.

These requirements are met if the second carrier is spaced 40 kilohertz from the first carrier. In this specific installation the transmit frequency of the central station is 70 kilohertz and the locomotive stations operate at 110 kilohertz.

Figure 6 shows the transceiver of a central station; that of a mobile station has the transmit and receive frequencies interchanged and employs a different type of line coupling.

The voice-frequency end in Figure 6 is arranged for connection to a handset having transmitter and receiver capsules and a push-to-talk switch. However, this can be replaced by a noise-compensated dynamic microphone having a sensitivity of 0.15 microvolt per microbar and a loudspeaker. The two attenuator pads are then replaced by amplifiers. The voice-frequency band is restricted to the range from 500 to 2400 hertz as the best compromise between performance and cost.

The transmit level chosen is +3.5 nepers (+30 decibels), corresponding to transmitter power of 1 watt, and the permissible receive-level range is −8.5 to +1.5 nepers (−74 to +13 decibels).

Figure 7 shows the levels of a voice-frequency signal with a modulation deviation of 2 kilohertz compared with the weighted noise of a

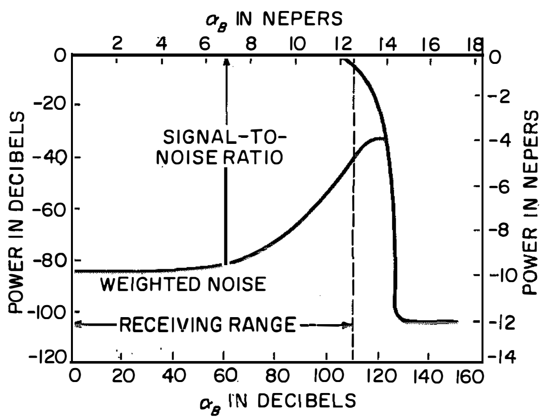

Figure 7—Signal-to-noise ratio of a receiver as a function of transmission attenuation α_B related to a signal with a frequency swing of 2 kilohertz.

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receiver without modulation, plotted as functions of the overall attenuation α_B between transmitter and receiver. Only for attenuations beyond 12.7 nepers (110 decibels) does the signal-to-noise ratio become poorer than 2.3 nepers (20 decibels).

Measurements of the frequency-dependent net loss between a central station and a mobile station are plotted in Figure 8. As has been indicated, voice frequencies below 500 hertz are not transmitted so that ambient noise at low frequencies is largely suppressed. This enhances the intelligibility. The frequencies above 2400 hertz do not improve intelligibility much but do make the voice sound more natural. However, this point is unimportant in mine communications.

The total distortion coefficient measured at 800 hertz at the normal signal level corresponding to a frequency swing of 2 kilohertz is less than 10 percent in the traffic between the central and a mobile station.

4.2 Mechanical Design

The problem was to provide an explosion-proof transceiver that conformed to the applicable German standards *VDE 0170/171* while having a tolerable weight and size. This led to a mechanical design that is entirely new in this field of application.

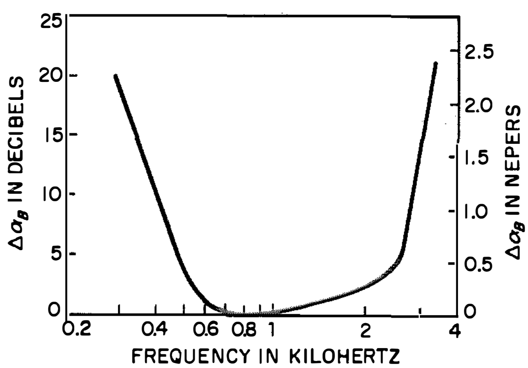

Figure 8—Attenuation distortion as a function of frequency in the transmission range between central station and mobile station.

The equipment has an outer and an inner cabinet. The outer cabinet is watertight and provides mechanical protection; it contains the cable lead-ins and covers the terminals. The inner cabinet is explosion-proof; after the printed-circuit units have been mounted, interconnected, and tested, this cabinet is filled with annealed quartz of suitable granulation. The antenna, loudspeaker, and microphone outputs are intrinsically safe. This construction was tested and approved by the mining authority. The equipment is usable in the temperature range from −20 to +40 degrees Centigrade.

Figure 9 shows how the equipment is mounted in a mine locomotive. The loop antenna is on the cab roof, and the transceiver is suspended below the roof. The pneumatic loudspeaker is oriented toward the head of the driver.

4.3 Power Supply

The all-transistor transceiver of the *Z1G* carrier telephone system operates on direct current at 24 volts ±5 percent. Power consumption is

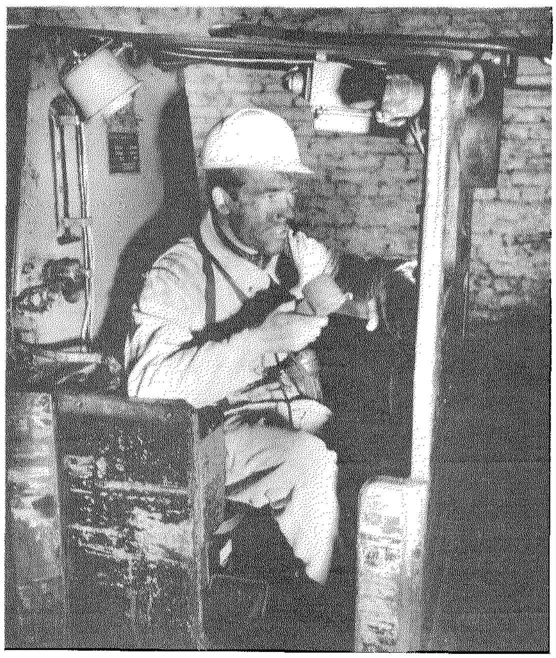

Figure 9—Arrangement of units in a locomotive cab.

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9 watts. This voltage must be maintained by the associated power supply in all modes of operation (transmit, receive, standby) despite the differences in current drain.

Universal application of the system requires that it be operable from any voltage source available underground. Hence, there are the following types of power supplies.

(A) Power supply for direct connection to 220 volts +10, −20 percent, 50 hertz.

(B) Power supply for air-driven locomotives. An air-driven generator supplies 12 volts +10, −20 percent at a frequency between 100 and 600 hertz.

(C) Power supply for storage battery of 12 volts ±20 percent.

(D) Power supply for diesel locomotives. A generator coupled to the diesel engine supplies a voltage between 17 and 42 volts at a frequency between 100 and 225 hertz.

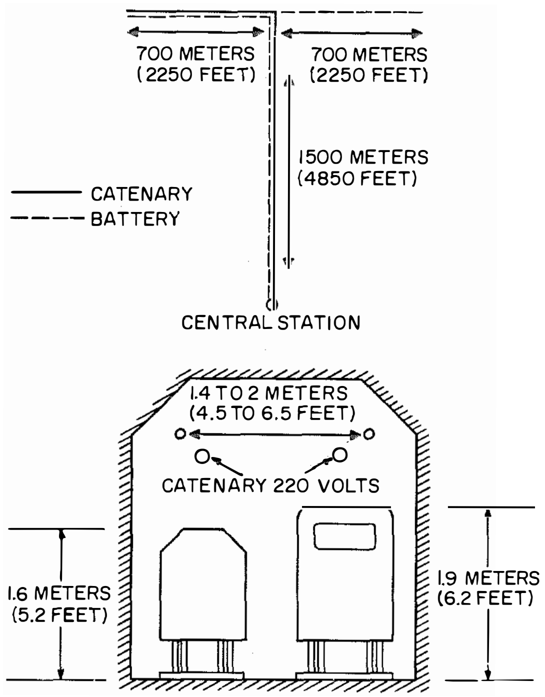

Figure 10—Plan of the railroad and gallery cross section of the Germania mine in Dortmund.

(E) Power supply for catenary locomotives. An all-transistor voltage divider uses the catenary direct voltage of 180 to 280 volts and supplies the rated voltage to the transceiver.

4.4 Accessories

The loop antenna for a locomotive station consists of a flexible multiconductor cable. It is mounted with the matching network on the locomotive or vehicle at a suitable point. Generally the length of the loop antenna is 4 meters (13 feet), but this can be reduced and the antenna adapted to various types of locomotives. The antenna inductance must be determined experimentally because it is affected by the locomotive shape. The matching network is aligned to the antenna inductance by a small series inductance. A cast-iron junction box contains the network and the antenna terminals.

The branching joints of the 2-wire transmission-line system consist of a hybrid symmetrically designed with two pot-core transformers [8]. The hybrid and a balancing circuit are accommodated in a junction box. The branching loss amounts to 0.4 neper (3.5 decibels).

The terminating resistance network of 600 ohms at each end of the line is likewise mounted in a cast-iron housing.

The accessories also include the waterproof handset with hook-type key, the dynamic microphone with receptacle, and the pneumatic loudspeaker. These units, specially designed for mining operations, have been used in underground communication systems for several years.

5. Trial Operations

Two trial operations were carried out in early 1960 in the Germania mine near Dortmund and the Emscher-Lippe mine in Datteln, to study underground propagation conditions and the performance of the *Z1G* system.

Figure 10 shows the layout of the railroad in the Germania mine. Both catenary and storage-

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battery locomotives are used on the railroad lines tested, which have a total length of 3 kilometers (1.8 miles). The insulated 2-wire transmission-line conductors (copper with a cross-sectional area of 1.5 square millimeters (0.002 square inch)) were laid 1.4 to 2 meters (4.5 to 6.5 feet) apart and temporarily mounted on the catenary supports. Curve *D* of Figure 3 shows the attenuation of this conductor.

In the other trial operation (Figure 11), a stranded-wire construction (6 steel and 1 copper wires) was used for the line, the attenuation of which is shown in curve *E* of Figure 3. This line was adequate for the trial but is unsuitable for greater distances. Moreover, air ducts and gas pipes prevented the line from being mounted to the mine roof, so it had to be fixed to the faces on both sides.

In the Germania trial operation, a characteristic impedance of 840 ohms was found at 50 kilohertz; in the other trial, the characteristic impedance had a real component of 530 to 980 ohms at frequencies from 50 to 150 kilohertz and amounted to 700 ohms plus a reflection coefficient of 20 percent.

The coupling attenuation in the first case varied between 6.1 and 8.5 nepers (53 and 74 decibels), depending on the instantaneous location of the locomotive; in the second case it varied between 5.4 and 8 nepers (47 and 70 decibels) (Figure 11). These values had been expected and corresponded to theoretical values. Small corrections in the physical layout of the line could improve the coupling attenuation, but were not carried out because of the short length of the line.

No interference from bow contacts to overhead (catenary) power supply conductors, commutator-type machines, rectifiers, et cetera, was noticed in the two trials although the mines were in full operation. The frequency modulation employed reduced interference.

In deep mining, the relatively high local acoustic noise is a problem. It can be suppressed in the central station so that the choice between the microphone-loudspeaker combination and the handset may be governed by purely operational considerations. As for the mobile station, however, the locomotive running at high speed or a train coming from the opposite direction generates strong noise that enters the microphone and is superimposed on the speech. This was particularly noticeable in the Germania mine where the speech band was 300 to 2400 hertz. Suppression of frequencies below 500 hertz and the use of a noise-compensated microphone markedly improved the speech transmission even in the presence of strong noise.

The microphone is superior to the transmitter capsule in the handset on the mobile station where local noise is considerable. Contrariwise, the handset pressed to the ear excludes ambient noise better than the loudspeaker although the signal-to-noise ratio of the latter is improved by its directivity. Thus the most-advantageous solution for the mobile station would be an

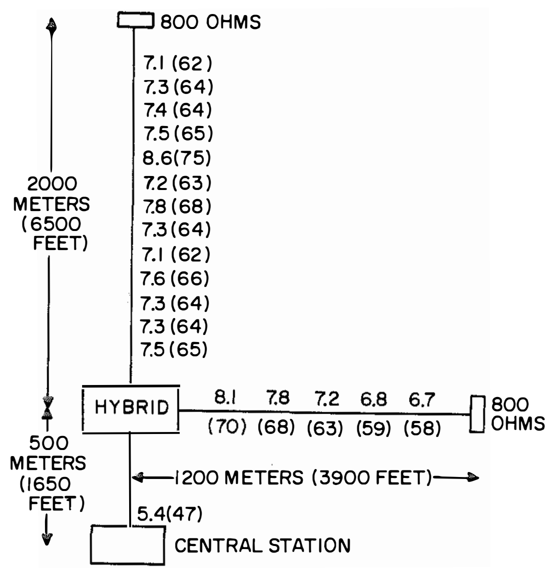

Figure 11—Plan of the railroad in the Emscher-Lippe mine. The coupling attenuations, measured at 140 kilohertz at different places between line and loop antenna, are in nepers (decibels).

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earset and a noise-compensated dynamic microphone. For operational reasons the combination of microphone and loudspeaker is preferred, as the operator should have both hands free.

The *Z1G* system fulfilled expectations in the trial operations and confirmed all theoretical considerations. A large number of other trials in various coal- and ore-mining installations confirmed the value of the following features: duplex operation, simplex operation between mobile stations, hybrid branching of the transmission line, and restricted speech band to reduce interference even at the high noise levels generated by the direct-current motors of locomotives supplied with power through sliding contacts to catenary cables and silicon rectifiers.

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Theory and Design of an Adjustable Equalizer *

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1. Introduction

Automatic regulation is required in carrier-frequency systems to equalize the changes in attenuation of cables and open-wire lines that result from variations in temperature and climate. An essential part of such regulation equipment is an adjustable equalizer, the attenuation characteristic of which can be changed by adjusting a single resistor (for instance, a thermistor). In the case under discussion, a special type of adjustable equalizer will be investigated. The network that determines the attenuation is composed of bridged-T networks and all-pass filters, thus forming a ladder network. The input terminals 2-2′ of this ladder network are connected to the series or shunt branch of a voltage divider as shown in Figures 1 and 2. The output terminals 3-3′ are connected to the adjustable resistor. Figure 3 shows the typical fan-out attenuation characteristics.

Contrary to the conventional method in which the attenuation-determining ladder network comprises bridged-T networks, all-pass filters are used here as correcting components. These filters are necessary to closely approximate the specified attenuation characteristic. There may be many cases in which such close approximation is unnecessary so that the bridged-T networks do not need filters.

The maximum spread or fan-out of the characteristics $a_{B\max} - a_o \leq a_o \geq a_o - a_{B\min}$ depends on the structure of the network, on the basic attenuation a_o, and on the technically available thermistor magnitudes $x_{\min} \leq x \leq x_{\max}$. Here it will be assumed that the attenuation curves $a_v(x, \omega)$ are located symmetrically to the basic attenuation a_o. It will readily be seen that a constant input resistance is obtained across the input terminals 2-2′ if the terminals 3-3′ are terminated with the normalized resistance $x = 1$. The curves of maximum fan-out are obtained when the resistance values are $x = 0$ or $x = \infty$.

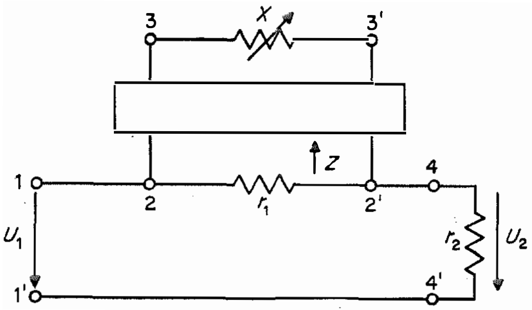

Figure 1—Adjustable equalizer with ladder network in series branch.

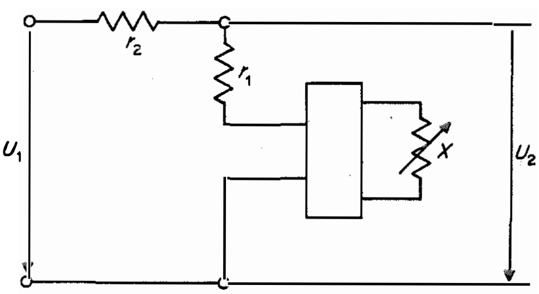

Figure 2—Adjustable equalizer with ladder network in shunt branch.

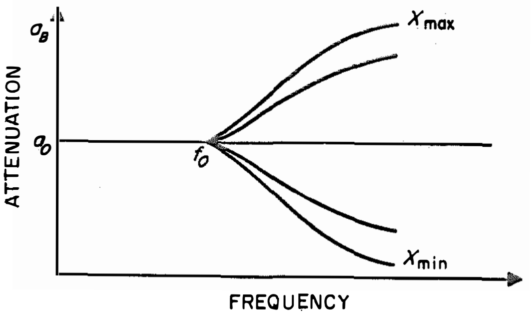

Figure 3—Typical fan-out attenuation characteristics.

* Translated and reprinted from *Archiv der Elektrischen Übertragung*, volume 17, page 217; 1963.

A distinction is made here between variable and adjustable. The electrical properties of a transmission line are variable; they are a function of its environment over which man has little or no control. The electrical properties of the equalizer are adjustable as they can be set to any value within a range to compensate for the variability of the transmission line.

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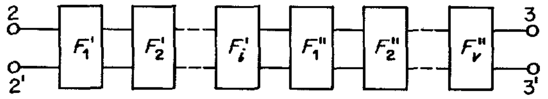

Figure 4—Ladder network of the 4-poles.

In the following sections, the mathematical relations are established between the propagation factor of the attenuation-determining network, connected to the voltage divider by its terminals 2-2′, and the propagation factor $(U_1/U_2) = e^{g_v}$ between terminal pairs 1-1′ and 4-4′. Here g_v is the propagation constant and U_1 and U_2 are the input and output voltages, respectively.

2. Mathematical Relations

All physical quantities such as resistance, inductance, capacitance, and frequency are normalized by the well-known method. Intermediate calculations are omitted if they are not necessary to comprehend the formulas.

From Figure 1 we obtain the ratio of the input voltage to the output voltage

$$\frac{U_1}{U_2} = e^{g_v} = 1 + \frac{r_1 Z}{r_2(r_1 + Z)}. \qquad (1)$$

Here r_2 is the load resistance, r_1 the so-called balancing resistor, and Z the input impedance of the attenuation-determining network.

First it is necessary to determine the total propagation factor of the network between the pairs of terminals 2-2′ and 3-3′. This is a ladder network comprising both bridged-T networks and all-pass filters. These 4-poles have a constant input resistance if they are terminated in the characteristic impedance. The total propagation factor of such networks can be represented as the product of the individual propagation factors

$$F_G(p) = \prod_i F_i'(p) \prod_v F_v''(p). \qquad (2)$$

Here $F_i'(p)$ are the propagation factors of the bridged-T terms and $F_v''(p)$ the propagation factors of the all-pass filters.

The propagation factor $F_i'(p)$ of the ith bridged-T network is

$$F_i'(p) = e^{g_i'} = 1 + r_i. \qquad (3)$$

$g_i' = a_i' + jb_i'$ is the propagation constant with the attenuation a_i' and the phase b_i'. The symbol r_i is a 2-pole function.

In the same manner the relation can be given for the propagation factor $F_v''(p)$ of the vth all-pass filter comprising reactances only

$$F_v''(p) = e^{g_v''} = \frac{r_v + 1}{r_v - 1} \qquad (4)$$

where $g_v'' = jb_v''$. The propagation constant g_v'' of the all-pass filter contains only the phase b_v'' because $|F_v''(p)| = 1$ is to apply to all frequencies. As is known from the 4-pole theory, the iterative matrix of the ith bridged-T has the appearance

$$(A_i') = \begin{bmatrix} \dfrac{r_i^2 + 2r_i + 2}{2(r_i + 1)} & \dfrac{r_i(r_i + 2)}{2(r_i + 1)} \\[2ex] \dfrac{r_i(r_i + 2)}{2(r_i + 1)} & \dfrac{r_i^2 + 2r_i + 2}{2(r_i + 1)} \end{bmatrix} = \begin{vmatrix} \cosh g_i' & \sinh g_i' \\ \sinh g_i' & \cosh g_i' \end{vmatrix}. \qquad (5)$$

Likewise, the following is true of the vth all-pass filter.

$$(A_v'') = \begin{bmatrix} \dfrac{r_v^2 + 1}{r_v^2 - 1} & \dfrac{2r_v}{r_v^2 - 1} \\[2ex] \dfrac{2r_v}{r_v^2 - 1} & \dfrac{r_v^2 + 1}{r_v^2 - 1} \end{bmatrix} = \begin{vmatrix} \cosh g_v'' & \sinh g_v'' \\ \sinh g_v'' & \cosh g_v'' \end{vmatrix}. \qquad (6)$$

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The right halves in (5) and (6) are obtained when the relations of (3), namely $r_i = e^{g_i'} - 1$, and of (4) solved for r_v, namely $r_v = \coth(g_v''/2)$, are substituted in the left halves of (5) and (6). The total iterative matrix of i bridged-T networks and v all-pass filters may be written

$$(A_G) = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} = \begin{bmatrix} \cosh\left(\sum_i g_i' + \sum_v g_v''\right) & \sinh\left(\sum_i g_i' + \sum_v g_v''\right) \\ \sinh\left(\sum_i g_i' + \sum_v g_v''\right) & \cosh\left(\sum_i g_i' + \sum_v g_v''\right) \end{bmatrix}. \tag{7}$$

The total propagation factor can be formulated somewhat differently with the aid of (3) and (4)

$$F_G(p) = e^{(\sum_i g_i' + \sum_v g_v'')} = e^{\sum_i a_i' + j(\sum_i b_i' + \sum_v b_v'')}. \tag{8}$$

For the arguments of the hyperbolic functions in the matrix of (7) it is expedient to introduce the abbreviation

$$\phi_z = \sum_i g_i' + \sum_v g_v''. \tag{9}$$

Thus a very-simple presentation is obtained for the four matrix elements

$$\begin{aligned} a_{11} &= a_{22} = \cosh \phi_z, \\ a_{12} &= a_{21} = \sinh \phi_z. \end{aligned} \tag{10}$$

The input impedance of the network across input 2-2′ if output 3-3′ is terminated by the resistance x obeys the relation

$$Z = \frac{x + \tanh \phi_z}{1 + x \cdot \tanh \phi_z}. \tag{11}$$

The resistance x may vary after any function; therefore it is permitted to write $x = \tanh y$. Hence, (11) assumes the form

$$Z = \tanh(\phi_z + y). \tag{12}$$

If (12) is substituted into (11), a relation is obtained between the propagation factor $(U_1/U_2) = F_v(p) = e^{g_v(p)}$ of the fan-out 4-pole and the propagation factor ϕ_z of the network whose output terminals 2-2′ are connected to the series branch of the voltage divider and whose output is terminated with an adjustable resistance.

$$F_v(p) = e^{g_v} = 1 + \frac{r_1}{r_2}\frac{1}{1 + r_1 \coth(\phi_z + y)} = \frac{e^{-2(\phi_z+y)}\left(1 - \frac{r_1 + r_2}{r_1 r_2}\right) + \left(1 + \frac{r_1 + r_2}{r_1 r_2}\right)}{e^{-2(\phi_z+y)}\left(1 - \frac{1}{r_1}\right) + \left(1 + \frac{1}{r_1}\right)}. \tag{13}$$

The attenuation characteristics are to be symmetrical about a frequency-independent basic attenuation a_o. For $x = 1$, that is $\coth(\phi_z + y) = 1$, it is

$$a_o = \ln\left(1 + \frac{r_1}{r_2}\frac{1}{1 + r_1}\right); \quad r_1, r_2 > 0, \text{ resistor}. \tag{14}$$

To achieve symmetric characteristics, the propagation factor $F_v(p)$ should be an odd function in $e^{-2(\phi_z+y)}$. Therefore if $e^{-2(\phi_z+y)}$ is replaced by $-e^{-2(\phi_z+y)}$, then $\coth(\phi_z + y)$ in (13) is transferred

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into tanh $(\phi_z + y)$. Hence

$$\left[1+\frac{r_1}{r_2}\frac{1}{1+r_1\coth(\phi_z+y)}\right]\left[1+\frac{r_1}{r_2}\frac{1}{1+r_1\tanh(\phi_z+y)}\right]=\left[1+\frac{r_1}{r_2}\frac{1}{1+r_1}\right]^2. \quad (15)$$

This relation must be valid for all ϕ_z, therefore also for $(\phi_z + y) = 0$

$$\left[1+\frac{r_1}{r_2}\right]=\left[1+\frac{r_1}{r_2}\frac{1}{1+r_1}\right]^2. \quad (16)$$

After some calculation, we finally obtain

$$r_2 = \frac{r_1}{r_1^2 - 1}. \quad (17)$$

For the symmetrical fan-out of the characteristics, (14) and (17) give

$$r_1 = e^{a_o}$$
$$r_2 = \frac{1}{2}\frac{1}{\sinh a_o}. \quad (18)$$

When a_o is given, resistors r_1 and r_2 are determined; contrariwise, the basic attenuation is determined when r_1 and r_2 are given. The conditions of (18) should now be substituted in (13). Instead of the hyperbolic functions, the exponential function can again be introduced, thus providing a presentation of the relation between the propagation factor e^{g_v} of the fan-out 4-pole and the propagation factor of the network inserted in the series branch, permitting further simplification of the equation.

$$\left.\begin{aligned} e^{g_v} &= e^{a_o}\frac{1+\alpha e^{-2\phi_z}}{1-\alpha e^{-2\phi_z}} \\ \alpha &= \tanh\frac{a_o}{2}e^{-2y} \\ &= \tanh\frac{a_o}{2}\cdot\frac{1-x}{1+x}. \end{aligned}\right\} \quad (19)$$

The transmission constant g_v is obtained by looking up the logarithms

$$g_v = a_o + \ln\frac{1+\alpha e^{-2\phi_z}}{1-\alpha e^{-2\phi_z}}. \quad (20)$$

This equation can be simplified by the progression

$$\begin{aligned} g_v - a_o &= \ln\frac{1+\alpha e^{-2\phi_z}}{1-\alpha e^{-2\phi_z}} \\ &= 2\left[\alpha e^{-2\phi_z} + \frac{\alpha^3 e^{-6\phi_z}}{3} + \cdots\right]. \end{aligned} \quad (21)$$

The series converges for $|\alpha e^{-2\phi_z}| < 1$. In almost all cases of application, the first term of the progression will prove satisfactory. The approximation equation thus obtained is

$$\begin{aligned} g_v &\approx a_o + 2\alpha e^{-2\phi_z} \\ &= a_o + 2\tanh\frac{a_o}{2}\frac{1-x}{1+x}e^{-2\phi_z}. \end{aligned} \quad (22)$$

In the first approximation, the error Δg_v can be estimated by computing the next term in the progression.

$$\Delta g_v \approx -\tfrac{2}{3}\alpha^3 e^{-6\phi_z}. \quad (23)$$

In (22) the propagation constant $g_v = a_v + jb_v$ can be split into the attenuation a_v and the phase b_v; likewise, the propagation constant $\phi_z = a_z + j(b_z + \varphi_z)$ can be split into the attenuation a_z, the phase b_z of the bridged-T, and φ_z, the phase of the all-pass filter.

$$\begin{aligned} a_v + jb_v &= a_o + 2\alpha e^{-2[a_z + j(b_z+\varphi_z)]} \\ &= a_o + 2\alpha e^{-2a_z}[\cos 2(b_z+\varphi_z) \\ &\qquad - j\sin 2(b_z+\varphi_z)]. \end{aligned} \quad (24)$$

The attenuation $a_v = \ln|U_1/U_2|$ and the phase $b_v = \sphericalangle U_1 - \sphericalangle U_2$ between the pairs of terminals 1-1′ and 4-4′ follow from (24).

$$\begin{aligned} a_v &= a_o + 2\alpha e^{-2a_z}\cos 2(b_z+\varphi_z) \\ b_v &= -2\alpha e^{-2a_z}\sin 2(b_z+\varphi_z). \end{aligned} \quad (25)$$

The error Δa_v is to be determined from (23) with the aid of (25).

$$\Delta a_v \approx -\tfrac{2}{3}\alpha^3 e^{-6a_z}\cos 6(b_z+\varphi_z). \quad (26)$$

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For many practical purposes this error is very small and may be neglected.

In designing adjustable equalizers, the fan-out characteristics are always specified on the basis of system planning. Hence the explicit form of a_z and $b_z + \varphi_z$ is wanted. If (25) is solved for these quantities, the following relations are obtained.

$$\left.\begin{aligned} a_z &= \tfrac{1}{4} \ln \frac{4\alpha^2}{(a_v - a_o)^2 + b_v^2} \\ b_z + \varphi_z &= -\tfrac{1}{2} \arctan \frac{b_v}{a_v - a_o}. \end{aligned}\right. \qquad (27)$$

Generally, only the attenuation curve a_v is available. In particular, the limiting curves denoting the maximum fan-out at top and bottom are known. To determine the attenuation and phase of the network in the series branch, it is necessary to know also the phase b_v. As is well known, the attenuation $a_v = \ln|U_1/U_2|$ comes into being by voltage division between the pairs of terminals 1-1′ and 4-4′. This voltage divider, which contains only linear passive components, cannot be more than a minimum-phase network. This gives a relation between attenuation a_v and phase b_v.

3. Attenuation and Phase of Minimum-Phase Network

As is known from the theory of functions, the Cauchy-Riemann differential equations

$$\left.\begin{aligned} \frac{\partial A}{\partial \sigma} &= \frac{\partial B}{\partial \omega} \\ \frac{\partial A}{\partial \omega} &= -\frac{\partial B}{\partial \sigma} \end{aligned}\right\} \qquad (28)$$

are valid between the real and the imaginary portions of an analytical function $f(p) = A(\sigma, \omega) + jB(\sigma, \omega)$, which is regular throughout in a zone G of a complex variable $p = \sigma + j\omega$.

Assuming further that the partial derivatives of the second order also exist in G and are continuous, the so-called Laplace differential equations

$$\left.\begin{aligned} \frac{\partial^2 A}{\partial \sigma^2} + \frac{\partial^2 A}{\partial \omega^2} &= 0 \\ \frac{\partial^2 B}{\partial \sigma^2} + \frac{\partial^2 B}{\partial \omega^2} &= 0 \end{aligned}\right\} \qquad (29)$$

follow from the Cauchy-Riemann differential equation, that is, neither the real nor the imaginary portion of $f(p)$ can be chosen at will. Rather, each should obey the Laplace and both together should obey the Cauchy-Riemann differential equations. The Cauchy-Riemann equations lead to the relation frequently used in the theory of potentials

$$B(\sigma, \omega) = \int \left(\frac{\partial A}{\partial \sigma} \mathrm{d}\omega - \frac{\partial A}{\partial \omega} \mathrm{d}\sigma \right) + K. \qquad (30)$$

The imaginary portion $B(\sigma, \omega)$ is determined by the real portion apart from a constant K. The Cauchy integral says nothing else

$$f(p) = \frac{1}{2\pi j} \int^{(\mathfrak{L})} \frac{f(\xi)}{\xi - p} \mathrm{d}\xi. \qquad (31)$$

In other words, if the function $f(p)$ is regular in the region G and its values are known along a closed double-point free path $(\mathfrak{L})$ in G, its values for any point inside G must necessarily be known. We restrict the above equation to functions admitted in the network theory, that is, only to the relation along the imaginary axis (real frequencies). G. Wunsch has given the relation

$$b_v(\omega) = \frac{1}{\pi} \int_{-\infty}^{+\infty} \frac{a_v(\omega + t)}{t} \mathrm{d}t \qquad (32)$$

where $t = \omega_o - \omega$ and $b_v(\omega) = B(0, \omega)$; $a_v = A(0, \omega)$.

Here a_v and b_v again mean attenuation and phase in accordance with (26) and (27).

This results in a graphic method suitable also for use in analog and digital computers.

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4. Graphic Method for Determining Phase b_v from Given Attenuation a_v

As shown in Figure 5, we subdivide the given characteristic $a_v(\omega)$, the image of which is desired at the ordinates, into an arbitrary number of sections $\Delta\omega$. If now the phase $b_v(\omega_1)$ is to be constructed for any point of the frequency ω_1, we draw a reference parallel to the $a_v(\omega)$ axis at a distance C (conveniently an even multiple of $\Delta\omega$).

All connecting lines from ω_1 to the intersections of the curve $a_v(\omega)$ with the straight lines of the sections $\Delta\omega$ cut the reference parallel. The distance l_v from the abscissa to the intersection is measured each time for the reference parallels, and this value is inserted as a summation term in (34). It will be seen from Figure 5 that the following geometric relations apply.

$$\frac{l_v}{C} = \frac{a_v(\omega_1 + v\Delta\omega)}{v\Delta\omega} \tag{33}$$

and this results in

$$b_v(\omega) \approx \frac{1}{\pi}\sum_{v=-n}^{v=+n} \frac{a_v(\omega_1 + v\Delta\omega)}{v\Delta\omega} = \frac{\Delta\omega}{\pi C}\sum_{v=-n}^{v=+n} l_v. \tag{34}$$

As a limit, we again arrive at (32).

The attenuation a_v and the phase b_v of the fan-out 4-pole are thus known. With the aid of (27) and assuming a basic attenuation a_o, this gives the attenuation and phase curves a_z and $b_z + \varphi_z$ of the network in the series branch of the fan-out equalizer. These calculated curves are now approximated with the aid of bridged-T's and all-pass filters. Initially the attenuation a_z required can be approximated with voltage dividers containing RLC 2-poles in the series branch. This can be done with the accuracy desired. From this, the bridged-T's are obtained by the well-known method. The approximation methods for the series impedance of a voltage divider will not be discussed here. It may be mentioned that families of curves, tables, or variations of component values with the aid of an electronic computer may lead to the desired results for certain structures.

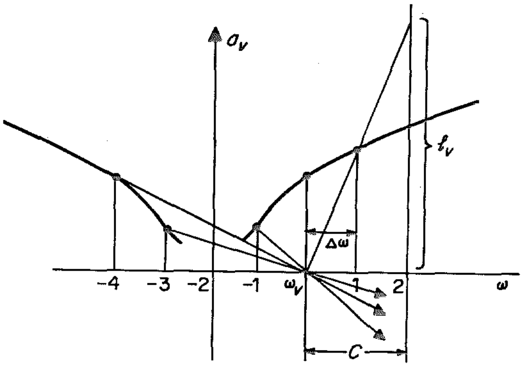

Figure 5—Graphic determination of the phase b_v from given attenuation a_v.

The method of Unbehauen is also known. It may be assumed that a_z had been approximated with the aid of a 2-pole without regard to other conditions. The phase of the 2-pole or of the bridged-T generally will not coincide with the required phase $b_z + \varphi_z = b_z'$. The phase belonging to a_z may be assumed as b_z. To obtain the desired phase b_z', a phase-correcting network can be inserted so that $b_z + \varphi_z = b_z'$ applies. From this the equation for the all-pass filter is obtained: $\varphi_z = b_z' - b_z$. The all-pass filter corrects the phase without changing the desired approximation of the attenuation a_z.

5. Instructions for Realizing Practical Circuits

Naturally there are many practical applications in which the error from not exactly

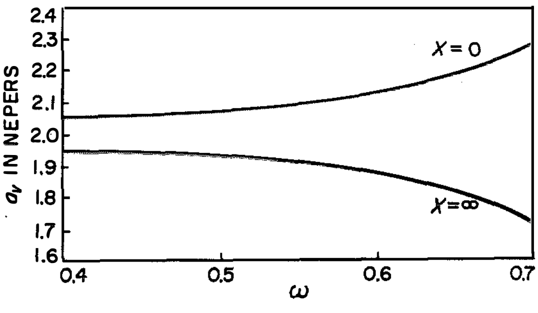

Figure 6—Maximum values of attenuation.

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meeting the phase requirements is so small that no phase-correcting network is required.

It is important to recognize that it is unwise to require a very-complex 2-pole to provide for a complex curvature of a_v, simply to limit the design to only one bridged-T in the series branch of the equalizer. These complex networks are more difficult to control than simple 2-poles that may be connected in series in the bridged-T's to produce the same desired summation curve. In practice it has been found unfavorable to go beyond a 2-pole of the third order. To achieve an approximation that is as accurate as possible, the phase must be corrected as indicated by (25) and (27). From the attenuation curve a_v of the fan-out, the phase is determined by the method described. To do this, it is necessary to know not only the curvature within the frequency range stated, but also beyond this range. This curvature depends on the network to be determined.

In the first place, only the probable characteristic can be assumed on the basis of limit considerations. The phase b_v therefrom is a first approximation. This gives also a_z and b_z. Through the approximation, the attenuation curvature a_v is now corrected outside the limits, starting with a_z and b_z. These new values give a better approximation through the phase b_v. Through repeated iterations the desired curvature of a_v is finally obtained with satisfactory accuracy. These iterative methods also are particularly suitable for the use of electronic computers. The special problems of programing will not be discussed here.

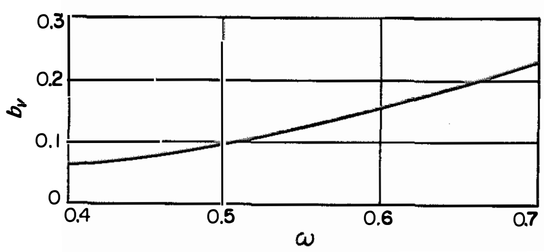

Figure 7—Phase as a function of frequency.

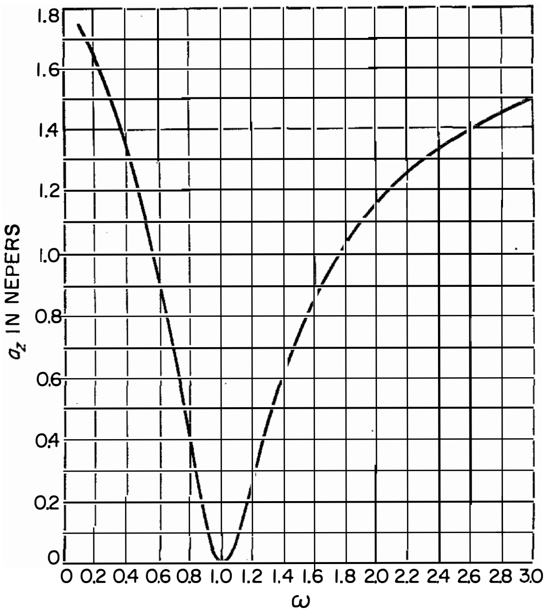

Figure 8—Attenuation characteristic of the network in the series branch.

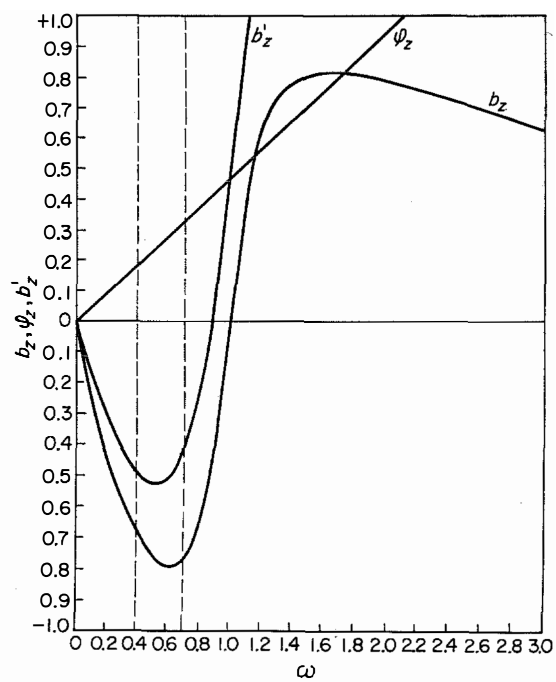

Figure 9—Phase characteristics of the network in the series branch. The dashed lines delineate the desired frequency range.

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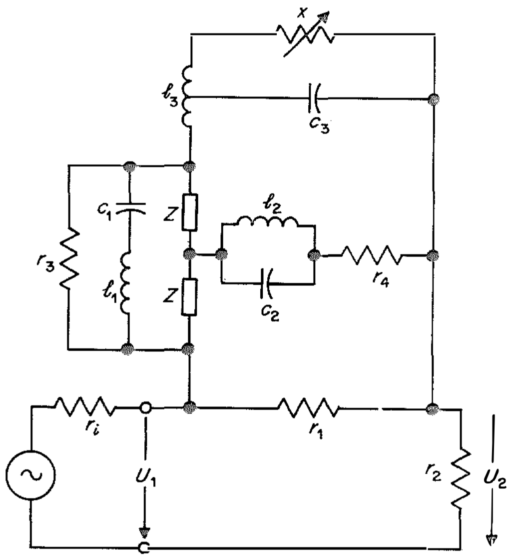
Figure 10—Circuit of the adjustable equalizer. Normalized values follow.

$c_1 = 0.5$	$l_3 = 0.5$
$c_2 = 2$	$r_1 = 7.4$
$c_3 = 0.5$	$r_2 = 0.137$
$l_1 = 2$	$r_3 = 5$
$l_2 = 0.5$	$r_4 = 0.2$



In practice it is often specified that all fan-out curves intersect the same frequency point (see f_o in Figure 3). This point is used as the pilot frequency. Unfortunately it is impossible to exactly satisfy this requirement. Considering (25) or (27), it is readily seen that this is possible only when $a_z = \infty$. This condition can never be realized exactly with lossy circuits. Normally, only the two pairs of curves symmetric to each other intersect on the straight line a_v = constant = a_o at one point. This point migrates on the frequency axis depending on the degree of fan-out. The only thing that can be demanded is that the attenuation a_v be within certain limits at a given frequency for all x terminations.

6. Example

Figure 6 shows the given curves a_v for the extreme values of the adjustable resistance x. In accordance with the graphic method described in Section 4, the associated phase may be constructed point by point. The result for the frequency range of interest, $\omega = 0.4$ to 0.7, may be found in Figure 7. This gives the values necessary to compute the desired attenuation curve a_z and phase b_z' of the fan-out-determining network. The curves a_z and b_z are shown between the frequency points desired in Figures 8 and 9, respectively. The curvatures were extended beyond the desired frequency range to give an idea of the actual curvature of the equalizer that can be realized electrically. The difference of $b_z' - b_z = \varphi_z$ is realized by the all-pass filter, the circuit of which is shown in Figure 10 with normalized values.

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Transmission-Line Mismatches and System Noise Figure

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1. Introduction

The advent in recent years of high-sensitivity receivers with very-low noise figures created the problem of designing the total receiving system to take full advantage of them. This is important in ground receivers for satellite communication, for example, in which threshold extension is normally used. As a consequence, these receivers present a very-sharp threshold characteristic and improvement of even a fraction of a decibel in the system noise figure could significantly increase the length of communication time during the satellite orbit.

The importance of mismatches along the transmission line from antenna to receiver on the noise performance of the total receiving system deserves careful evaluation.

2. Power Transferred from Antenna to Receiver

Figure 1 shows an antenna with internal impedance Z_A, connected by a transmission line having characteristic impedance Z_0 and propagation constant $\gamma = \alpha + j\beta$ to a receiver R through a passive quadrupole Q.

Assume (Figure 2) that output terminals CC of the quadrupole are closed on a resistive load Z_L equal to Z_0 and that the voltage standing-wave ratio at input terminals BB is S_Q. The power P_o delivered under these conditions to Z_L (demonstrated in the Appendix, Section 6) is given by

$$P_o = P_M \exp(-2\alpha l_1) \frac{4S_A}{(1+S_A)^2} \frac{4S_Q}{(1+S_Q)^2} \times \frac{r}{|1 - \rho_A\rho_Q \exp(-2\gamma l_1)|^2} \quad (1)$$

where r is the ratio between P_o and the power P_Q entering quadrupole Q through terminals BB, P_M is the available power at the antenna terminals, and the other symbols are defined as in the Appendix. In particular, ρ_Q is the reflection coefficient at terminals BB *when quadrupole Q is loaded with a resistive impedance equal to Z_0.* It is possible to express r as a function of the insertion loss A of the quadrupole and of S_Q.

A in Figure 3 is defined as the ratio of the power that would be delivered to $Z_L = Z_0$ when the load resistance is connected directly across terminals BB, to the actual power P_o^* delivered to $Z_L = Z_0$.

Introducing the available power P_{BB} at terminals BB leads immediately to

$$P_{BB} = AP_o^*.$$

Using (18) to calculate the power entering quadrupole Q in the conditions of Figure 3,

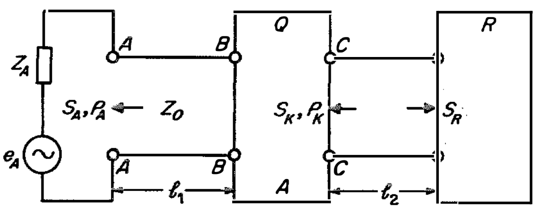

Figure 1—Antenna connected by transmission line to a receiver through a quadrupole.

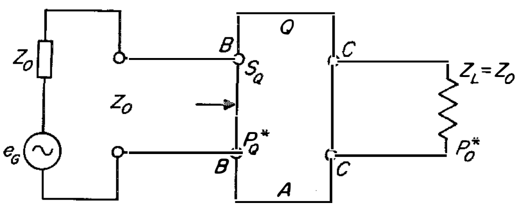

Figure 2—Antenna and quadrupole loaded with $Z_L = Z_0$.

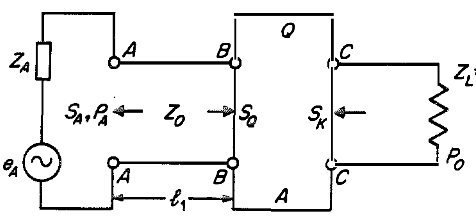

Figure 3—Generator and quadrupole loaded with $Z_L = Z_0$.

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$P_Q{}^*$, as a function of P_{BB}

$$r = \frac{P_o{}^*}{P_Q{}^*} = \frac{P_{BB}/A}{P_{BB}4S_Q/(1+S_Q)^2} = \frac{1}{A}\,\frac{(1+S_Q)^2}{4S_Q}.$$

This value for r is not a function of the input power to quadrupole Q. Therefore, it can generally be used independently. The available power P_{CC} at terminals CC of Figure 1 may now be determined. In fact, using (18)

$$P_o = P_{CC}\,\frac{4S_K}{(1+S_K)^2}$$

with the voltage standing-wave ratio S_K measured at terminals CC looking toward Q. From this equation

$$P_{CC} = P_o\,\frac{(1+S_K)^2}{4S_K} = P_M\,\frac{\exp(-2\alpha l_1)}{A}\,\frac{4S_A}{(1+S_A)^2}\,\frac{(1+S_K)^2}{4S_K}\,\frac{1}{|1-\rho_A\rho_Q\exp(-2\gamma l_1)|^2} \tag{2}$$

and finally, the signal power P_R delivered to the receiver is

$$P_R = P_{CC}\exp(-2\alpha l_2)\,\frac{4S_K}{(1+S_K)^2}\,\frac{4S_R}{(1+S_R)^2}\,\frac{1}{|1-\rho_K\rho_R\exp(-2\gamma l_2)|^2}.$$

Using (2), P_R can be written as

$$P_R = P_M\,\frac{\exp[-2\alpha(l_1+l_2)]}{A}\,\frac{4S_A}{(1+S_A)^2}\,\frac{4S_R}{(1+S_R)^2}$$
$$\times\,\frac{1}{|1-\rho_A\rho_Q\exp(-2\gamma l_1)|^2}\,\frac{1}{|1-\rho_K\rho_R\exp(-2\gamma l_2)|^2}. \tag{3}$$

Equation (3) can be extended immediately to the case where two quadrupoles Q_1 and Q_2 are inserted in the transmission line, as indicated in Figure 4. The available power P_{DD} at terminals DD is still given by (2). Rewritten

$$P_{DD} = P_M\,\frac{\exp(-2\alpha l_1)}{A_1}\,\frac{4S_A}{(1+S_A)^2}\,\frac{(1+S_{K1})^2}{4S_{K1}}\,\frac{1}{|1-\rho_A\rho_{Q1}\exp(-2\gamma l_1)|^2}.$$

The power delivered to the receiver P_R is given by (3), in which P_M must be replaced by P_{DD}, (l_1+l_2) by (l_2+l_3), S_A by S_{K1}, ρ_A by ρ_{K1}, ρ_Q by ρ_{Q2}, ρ_K by ρ_{K2}, and A by A_2. Thus

$$P_R = P_M\,\frac{\exp[-2\alpha(l_1+l_2+l_3)]}{A_1A_2}\,\frac{4S_A}{(1+S_A)^2}\,\frac{4S_R}{(1+S_R)^2}$$
$$\times\,\frac{1}{|1-\rho_A\rho_{Q1}\exp(-2\gamma l_1)|^2}\,\frac{1}{|1-\rho_{K1}\rho_{Q2}\exp(-2\gamma l_2)|^2}\,\frac{1}{|1-\rho_{K2}\rho_R\exp(-2\gamma l_3)|^2}.$$

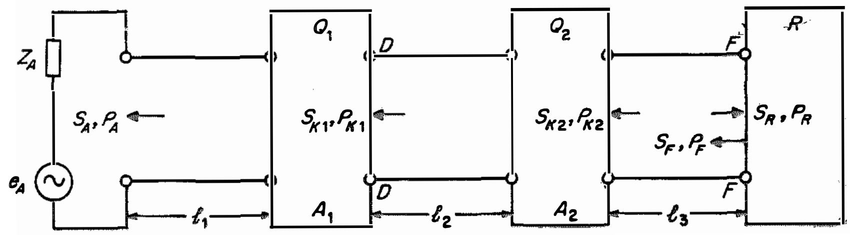

Figure 4—Two quadrupoles inserted in transmission line between antenna and receiver.

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This procedure clearly can be extended to the case where $n - 1$ quadrupoles are inserted in the transmission line. Generally

$$P_R = P_M \frac{\exp(-2\alpha l)}{A_1 A_2 \cdots A_n} \frac{4S_A}{(1+S_A)^2} \frac{4S_R}{(1+S_R)^2} \prod_{j,i=1}^{n} \frac{1}{|1 - \rho_i \rho_j \exp(-2\gamma l_i)|^2} \tag{4}$$

where l is the total length of the transmission line and ρ_i, ρ_j are the reflection coefficients as previously defined at the ends of the ith section of transmission line. Equation (4) can be used to determine the noise power P_{nA} delivered from the antenna to the receiver in a small bandwidth Δf, provided P_M is interpreted as the available noise power $kT_A\Delta f$ in this bandwidth at the antenna terminals. Of course, the result thus obtained is valid only if the frequency-dependent quantities $S_A, S_R, A_1, A_2, \cdots$, do not vary appreciably in Δf. If that is not true, the noise power delivered to the receiver is given by

$$\int_o^{\Delta f} P_{nA} df$$

but the evaluation of the integral requires knowledge of the variation with frequency of P_{nA}.

This is never the case in system design and the most-fruitful approach in noise-figure calculations appears to involve only P_{nA} (spot noise figure).

The total noise power $P_{n\,\mathrm{tot}}$ delivered to the receiver can be obtained in a more-direct way whenever the equivalent noise temperatures of all components of the receiving system (antenna included) are equal to the reference ambient temperature T_o.

In this case, the available noise power P_{FF} at the input terminals of the receiver (see Figure 4) is simply $kT_o\Delta f$ because it can always be considered as generated by the resistive component of the impedance measured between terminals FF looking toward the antenna. Therefore

$$P_{n\,\mathrm{tot}} = kT_o\Delta f \frac{4S_R}{(1+S_R)^2} \frac{4S_F}{(1+S_F)^2} \times \frac{1}{|1 - \rho_R \rho_F|^2}$$

with obvious meaning for S_F and ρ_F. Of this total noise power, the fraction

$$P_{nA} = kT_A\Delta f \frac{\exp(-2\alpha l)}{A_1 A_2 \cdots A_n} \frac{4S_A}{(1+S_A)^2} \frac{4S_R}{(1+S_R)^2} \prod_{j,i=1}^{n} \frac{1}{|1 - \rho_i \rho_j \exp(-2\gamma l_i)|^2} \tag{5}$$

where $T_A = T_o$, is coming from the antenna; the noise power P_{nl} delivered to the receiver from the transmission line is therefore

$$P_{nl} = P_{n\,\mathrm{tot}} - P_{nA}. \tag{6}$$

If the equivalent noise temperature of the antenna is T_A, the noise power P_{nA} from the antenna is given by (5), in which $T_A \neq T_o$. However, the noise power P_{nl} from the transmission line *is still given by (6)*.

3. System Noise Figure

The system noise figure F_{sys} can be calculated from

$$F_{\mathrm{sys}} = \frac{(P_{nA} + P_{nl})G_R + (F_R - 1)kT_o\Delta f G_R \dfrac{4S_R}{(1+S_R)^2} \dfrac{4S_F}{(1+S_F)^2} \dfrac{1}{|1 - \rho_F \rho_R|^2}}{P_{nA} G_R}$$

$$= 1 + \frac{P_{n\,\mathrm{tot}} - P_{nA}}{P_{nA}} + \frac{kT_o\Delta f (F_R - 1)}{P_{nA}} \frac{4S_F}{(1+S_F)^2} \frac{4S_R}{(1+S_R)^2} \frac{1}{|1 - \rho_R \rho_{K(n-1)} e^{-2\gamma l_n}|^2} \tag{7}$$

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which is the ratio of the total available noise power at the receiver output to that portion of the total available noise power attributable to the noise generated by the antenna. Consequently, G_R is defined as the ratio of the available power at the receiver output to the input signal power to the receiver that generates it. The expression

$$(F_R - 1)kT_o\Delta f G_R \frac{4S_R}{(1+S_R)^2} \times \frac{4S_F}{(1+S_F)^2} \frac{1}{|1-\rho_F\rho_R|^2} \quad (8)$$

represents the fraction of the total available noise power at the receiver output due to the noise generated in the receiver. The value F_R represents the actual noise figure of the receiver measured with terminals FF closed on an impedance equal to that presented by the transmission line at the standard temperature T_o. Substitution of (6) and (5) in (7) leads to

$$F_{\text{sys}} = 1 - \frac{T_o}{T_A} + \frac{T_o}{T_A} A_{\text{tot}} F_R \frac{(1+S_A)^2}{4S_A} \times \frac{4S_F}{(1+S_F)^2} \prod_{j,i=1}^{n-1} |1 - \rho_i\rho_j e^{-2\gamma l_i}|^2 \quad (9)$$

where $A_{\text{tot}} = A_1 A_2 \cdots A_n \exp(2\alpha l)$ = total insertion loss of the transmission line. From the definition, it follows that the quantity $kT_A\Delta f F_{\text{sys}}$ is proportional to the available noise power at the output of the receiver, the proportionality constant being the available power gain of the whole system from antenna terminals to receiver output. In system design, it is customary to refer all noise levels to the available noise power from a resistor at the reference temperature $T_o = 290$ degrees Kelvin. This suggests the introduction of an equivalent system noise figure $F_{\text{sys eq}}$ defined by

$$F_{\text{sys}} kT_A\Delta f = F_{\text{sys eq}} \cdot kT_o\Delta f$$

from which

$$F_{\text{sys eq}} = (T_A/T_o) F_{\text{sys}}$$

or

$$F_{\text{sys eq}} = \frac{T_A}{T_o} - 1 + F_R A_{\text{tot}} \frac{(1+S_A)^2}{4S_A} \times \frac{4S_F}{(1+S_F)^2} \prod_{j,i=1}^{n-1} |1 - \rho_i\rho_j e^{-2\gamma l_i}|^2. \quad (10)$$

The value of $F_{\text{sys eq}}$ can never be negative although it can be less than 1. In this case, the noise power at the receiver output is less than that available if the equivalent noise temperature of the antenna is T_o.

As repeatedly stated the values for the ρ_i appearing in (10) are those measured at the input terminals of each quadrupole when its output terminals are closed on a resistive load equal to Z_0. On the other hand, the values for the ρ_j are those measured at the output terminals of the quadrupoles looking toward the antenna in the actual conditions of the transmission line.

The fact that S_R is not in (10) appears strange at first sight; however, its effects clearly are included in the value of F_R which has been defined as the noise figure of the receiver measured under the actual condition of input loading.

Another concept widely used in system noise calculations is the so-called equivalent system noise temperature T_{sys}. By definition

$$T_{\text{sys}} = F_{\text{sys}} T_A = T_A - T_o + T_o F_R A_{\text{tot}} \frac{(1+S_A)^2}{4S_A} \times \frac{4S_F}{(1+S_F)^2} \prod_{j,i=1}^{n-1} |1 - \rho_i\rho_j e^{-2\gamma l_i}|^2.$$

From the definition, it follows that the quantity $kT_{\text{sys}}\Delta f = kT_o\Delta f F_{\text{sys eq}}$ is proportional to the available noise power at the output of the receiver, the proportionality constant being the available power gain of the whole system from antenna terminals to receiver output.

Both T_{sys} and $F_{\text{sys eq}}$ are employed in system calculations; it is clear from the preceding that the use of either leads to the same results and is a question of personal preference.

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3.1 Particular Case

Where no quadrupole is inserted in the transmission line, it is sufficient to consider a piece of line of infinitesimal length as a fictitious quadrupole. Then

$$F_{\text{sys eq}} = \frac{T_A}{T_o} - 1 + F_R A_{\text{tot}} \frac{(1+S_A)^2}{4S_A} \frac{4S_F}{(1+S_F)^2} \quad (11)$$

where $4S_F/(1+S_F)^2$ could be calculated as a function of S_A through the relation

$$\frac{4S_F}{(1+S_F)^2} = 1 - \left(\frac{S_A-1}{S_A+1}\right)^2 \exp(-2\alpha l) \quad (12)$$

obtained in standard transmission-line theory.

If the system is assumed to be perfectly matched, we find from (10) and (11)

$$F_{\text{sys eq}}^* = \frac{T_A}{T_o} - 1 + A_{\text{tot}} F_R \quad (13)$$

which is a standard equation used in system design.

3.2 Example

It is interesting to compare the value of $F_{\text{sys eq}}$ given by (13) and (10) in an example where two quadrupoles, each having an insertion loss of 0.5 decibel, are inserted at equal distances along a transmission line that has a total loss of 1 decibel (see Figure 5).

Assume that $T_A = 60$ degrees Kelvin, $F_R = 1.58$ ($\equiv$ 2 decibels), $S_A = S_F = 1.5$ (S_F does not have to satisfy (12) in this case), and that the absolute value of all reflection coefficients involved is equal to 0.2. The expression $|1 - \rho_i\rho_j \exp(-2\gamma l_i)|^2$ can assume values between a maximum and a minimum given by

$$(1 + 0.2 \times 0.2 \times 0.925)^2 = 1.037^2 = 1.075$$

and

$$(1 - 0.2 \times 0.2 \times 0.925)^2 = 0.963^2 = 0.927.$$

Therefore

$$1.075^2 \geq \prod_{j,i=1}^{n-1} |1 - \rho_i\rho_j \exp(-2\gamma l_i)|^2 \geq 0.927^2.$$

Substituting these values in (10)

$$(F_{\text{sys eq}})_{\max} = \frac{60}{290} - 1 + 1.58^2 \frac{2.5^2}{4 \times 1.5} \frac{4 \times 1.5}{2.5^2} \times 1.075^2$$
$$= 0.207 - 1 + 2.5 \times 1.15 = 2.1 (\equiv 3.25 \text{ decibels})$$

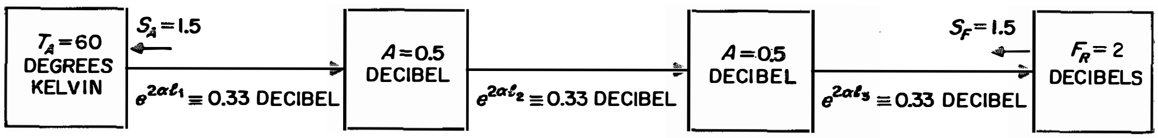
in one case and

$$(F_{\text{sys eq}})_{\min} = \frac{60}{290} - 1 + 1.58^2 \frac{2.5^2}{4 \times 1.5} \frac{4 \times 1.5}{2.5^2} \times 0.927^2$$
$$= 0.207 - 1 + 2.5 \times 0.86 = 1.35 (\approx 1.3 \text{ decibels})$$

in the other. Using (13)

$$F_{\text{sys eq}}^* = 0.207 - 1 + 1.58 \times 1.58 = 1.7 \; (\equiv 2.3 \text{ decibels}).$$

Figure 5—Two quadrupoles, each having an insertion loss of 0.5 decibel, inserted at equal distances along a transmission line that has a total loss of 1 decibel.



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4. Conclusion

It is apparent from the particular case considered that the difference between the actual system noise figure and that calculated, assuming matched conditions for the transmission line, can be significant. Admittedly, if reflection coefficients of lower values than those assumed in Section 3.2 are found in a system, the difference between (10) and (13) tends to vanish. This is also true if the circuit attenuations are larger.

However, this analysis is intended to provide the designer with a few workable equations to enable him to make accurate predictions on the noise figure of highly sensitive systems once the performance of the individual components is known. This is particularly helpful in practical cases to assess the possibility of further improvement. It should be realized that when working with amplifiers having noise figures on the order of a fraction of a decibel (for example, 0.1 decibel is typical of maser amplifiers) in conjunction with low-noise-temperature antennas (15 to 20 degrees Kelvin for a good horn antenna pointing at zenith), the system noise figure could be limited by the associated circuits.

5. Reference

1. W. Jackson, "High-Frequency Transmission Lines," John Wiley & Sons; 1953: page 62, equation (4.6) with $x = l$.

6. Appendix

It is often necessary to determine the power delivered by a radio-frequency generator to a load through a length of transmission line. Standard equations require knowledge of the series or parallel components of the equivalent circuit for the load and for the generator.

An equation is derived here for the transmitted power as a function of more-readily measured quantities such as standing-wave ratios and reflection coefficients.

With reference to Figure 6, consider a generator with an internal impedance $Z_G = R_G + jX_G$ delivering power to a load through a length l of transmission line with characteristic impedance Z_0. The voltage between the load terminals LL is given, in complex notations [1], by

$$V_L = e_G \frac{Z_0}{Z_0 + Z_G} \frac{(1 + \rho_L) \exp(-\gamma l)}{1 - \rho_L \rho_G \exp(-2\gamma l)} \quad (14)$$

where

$\gamma = \alpha + j\beta$ = propagation constant of the transmission line

ρ_L = reflection coefficient at the load

ρ_G = reflection coefficient at the generator.

From (14), introducing the resistance component $1/R_L$ of the equivalent load admittance, the power transmitted is given by

$$P_L = \frac{|V_L|^2}{R_L} = e_G^2 \left| \frac{Z_0}{Z_0 + Z_G} \right|^2 \frac{|1 + \rho_L|^2}{|1 - \rho_L \rho_G \exp(-2\gamma l)|^2} \frac{\exp(-2\alpha l)}{R_L}. \quad (15)$$

But, with reference to Figure 6

$$1 - |\rho_G|^2 = 1 - \frac{|Z_G - Z_0|^2}{|Z_G + Z_0|^2} = \frac{[(Z_0 + R_G)^2 + X_G^2] - [(R_G - Z_0)^2 + X_G^2]}{|Z_G + Z_0|^2}$$

or

$$1 - |\rho_G|^2 = \frac{4 Z_0 R_G}{|Z_G + Z_0|^2}$$

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from which, introducing the standing-wave ratio S_G presented by the generator impedance to the transmission line

$$\frac{Z_0}{|Z_0 + Z_G|^2} = \frac{1 - |\rho_G|^2}{4R_G} = \frac{1 - \left(\frac{S_G - 1}{S_G + 1}\right)^2}{4R_G} = \frac{1}{4R_G}\frac{4S_G}{(S_G + 1)^2}.$$

Substituting this expression in (15) and letting $e_G^2/4R_G = P_{\max}$ (= maximum power the generator is capable of delivering), we find

$$P_L = P_{\max} \exp(-2\alpha l) \frac{4S_G}{(1 + S_G)^2} \frac{|1 + \rho_L|^2}{|1 - \rho_L\rho_G \exp(-2\gamma l)|^2} \frac{Z_0}{R_L}. \tag{16}$$

The ratio Z_0/R_L can be expressed as a function of ρ_L as follows. First, using for convenience the admittances $Y_0 = 1/Z_0$ of the transmission line and $Y_L = (1/R_L) + (1/X_L)$, we see that

$$1 - |\rho_L|^2 = 1 - \frac{|Y_0 - Y_L|^2}{|Y_0 + Y_L|^2} = \frac{\left[\left(Y_0 + \frac{1}{R_L}\right)^2\left(\frac{1}{X_L}\right)^2\right] - \left[\left(Y_0 - \frac{1}{R_L}\right)^2 + \left(\frac{1}{X_L}\right)^2\right]}{|Y_0 + Y_L|^2}$$

$$= \frac{4Y_0}{|Y_0 + Y_L|^2}\frac{1}{R_L}$$

or

$$\frac{1}{R_L} = \frac{1 - |\rho_L|^2}{4Y_0}|Y_0 + Y_L|^2. \tag{17}$$

Second, from the definition

$$\rho_L = \frac{Y_0 - Y_L}{Y_0 + Y_L}$$

it follows that

$$Y_L(\rho_L + 1) = Y_0(1 - \rho_L) \quad \text{or} \quad \frac{Y_L}{Y_0} = \frac{1 - \rho_L}{1 + \rho_L}.$$

Therefore

$$|Y_0 + Y_L|^2 = Y_0^2\left|1 + \frac{1 - \rho_L}{1 + \rho_L}\right|^2 = Y_0^2 \frac{4}{|1 + \rho_L|^2}$$

and introducing this value in (17)

$$\frac{Z_0}{R_L} = \frac{1 - |\rho_L|^2}{|1 + \rho_L|^2} = \frac{1}{|1 + \rho_L|^2}\frac{4S_L}{(1 + S_L)^2}$$

where S_L is the voltage standing-wave ratio measured at the output terminals of the transmission line. Equation (16) can now be written in the simpler and symmetrical form

$$P_L = P_{\max} \exp(-2\alpha l) \frac{4S_G}{(1 + S_G)^2} \frac{4S_L}{(1 + S_L)^2} \frac{1}{|1 - \rho_G\rho_L \exp(-2\gamma l)|^2}. \tag{18}$$

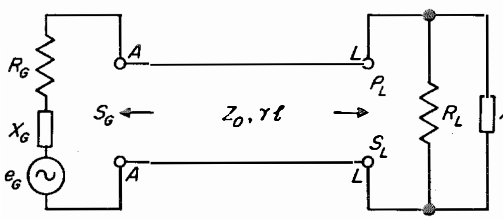

Figure 6—Generator delivering power to a load through a length l of transmission line having a characteristic impedance Z_0. The generator has an internal impedance $Z_G = R_G + jX_G$.

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From (18), we see that the influence on P_L of a small variation in the length of the transmission line around an average value l_o is essentially contained in the term

$$\frac{1}{|1 - \rho_G \rho_L \exp(-2\gamma l)|^2}.$$

The denominator of this expression oscillates between a maximum given by

$$[1 + |\rho_L| |\rho_G| \exp(-2\alpha l)]^2$$

and a minimum given by

$$[1 - |\rho_L| |\rho_G| \exp(-2\alpha l)]^2.$$

Since

$$|\rho_L| = \frac{S_L - 1}{S_L + 1}$$

and

$$|\rho_G| = \frac{S_G - 1}{S_G + 1}$$

we conclude that, as a function of the length of the transmission line, the power delivered to the load varies between two limits practically given by

$$P_L = P_{\max} \exp(-2\alpha l) \frac{16 S_G S_L}{[(S_L + 1)(S_G + 1) \pm (S_L - 1)(S_G - 1) \exp(-2\alpha l)]^2}. \tag{19}$$

As a practical application of (19), consider the situation represented in Figure 7, where a generator with internal impedance Z_0 delivers power to a resistance Z_0 and to a transmission line of characteristic impedance Z_0 with a 2-decibel transmission loss terminated by a purely resistive component $Z_0/2$. Let $P_{\max} = e_G^2/4Z_0$ be the maximum power the generator is capable of delivering to the resistance Z_0 connected between terminals AA when the transmission line is disconnected. The maximum power the generator is capable of delivering to the transmission line is $P_{\max}/2$, as can be seen immediately from Thévenin's theorem applied to the circuit to the left of terminals AA (see Figure 8). In our case $S_G = 2$, $S_L = 2$, and $\exp(-2\alpha l) = 0.634$. Substituting these values in (19), we find

$$P_L = \frac{P_{\max}}{2} 0.634 \frac{16 \times 2 \times 2}{(3 \times 3 \pm 1 \times 1 \times 0.634)^2}$$

$$= P_{\max} 0.634 \frac{32}{(9 \pm 0.634)^2}.$$

Therefore

$$P_{L\,\min} = P_{\max} \times \frac{32}{93} \times 0.634 = 0.22 P_{\max}$$

and

$$P_{L\,\max} = P_{\max} \times \frac{32}{70} \times 0.634 = 0.29 P_{\max}.$$

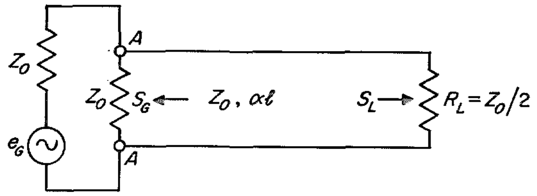

Figure 7—Generator with internal impedance Z_0 delivering power to a resistance Z_0 and to a transmission line of characteristic impedance Z_0 with a 2-decibel transmission loss terminated by a purely resistive component $Z_0/2$.

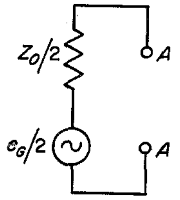

Figure 8—Equivalent circuit of Figure 7 to the left of terminals AA.

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From 1949 to 1953, he worked for Fabbrica Italiana Macchine Elettriche and for the

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following four years for Fabbrica Italiana Apparecchiature Radio in Milan.

From 1957 to 1959, he was with Westinghouse Electronics Division in the United States working on wide-band directive antennas.

In 1959, Mr. Imboldi joined ITT Federal Laboratories, where he has been engaged in various phases of satellite ground-station engineering.

He is a Member of the Institute of Electrical and Electronics Engineers.

Richtfunkverbindung (Radio Links)

Helmut Carl of Standard Elektrik Lorenz has presented in this book a concise survey of radio links including system planning, main equipment, accessory equipment, and operating requirements. It is based on his extensive experience in this branch of communication engineering supplemented by service on committees of the International Telecommunication Union. The reader will be aided by a glossary of technical terms and numerous references to the literature.

The book is 158 by 233 millimeters (6.2 by 9.2 inches) and has 210 pages. It is available from Berliner Union Verlag in Stuttgart at DM 38.

Computer Assistance to Pentaconta Engineering

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1. Introduction

At the beginning of 1961, it was proposed to use computers to prepare wiring information and thus speed the mass production of Pentaconta equipment in our various manufacturing companies. Although the scope of this proposal was questioned as being too limited, the matter was evaluated and a complete investigation was then made. After a favorable report was submitted at the end of 1961, a group of engineers and technicians from several of our companies began the programing for a Bull Gamma *30* computer.

The first version of the program was ready at the end of 1962, and prototype lists were prepared for use by our manufacturers in France. The program was then improved with the cooperation of the users. Successive versions of the program were named *LDF-0, LDF-1,* and *LDF-2,* and the latter was placed in operation in April 1963 by the French companies of ITT Europe; it is now beginning to be used by our Belgian, Swiss, and Spanish companies, and almost 200 different frames have already been completed. Figure 1 illustrates the process.

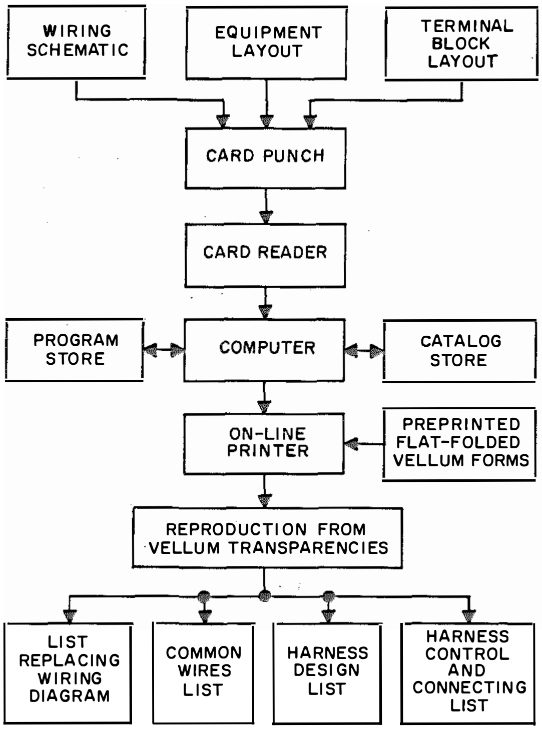

Figure 1—Computer processing of Pentaconta wiring.

To understand clearly the value of using a computer to solve wiring problems, the complexity of these problems within the Pentaconta system must be considered. In crossbar switching systems, all connecting and control functions in the selecting stages are carried out by relays, without using sequence switches, step-by-step switches, et cetera. Therefore the average number of relays per line is higher than in systems using rotary switches under the same conditions of traffic and capacity, and the average number of contact springs per relay is also higher. These facts, plus the greater compactness of the equipment, lead to an increased number of terminals per square inch; this number is 5 times higher in a Pentaconta register than in the corresponding rotary *7A* circuit.

It is obvious that such a concentration of terminals complicates the preparation of wiring documents as well as the actual connection of the wires. The latter task was simplified by the use of wrapped connections without soldering. In contrast, the time required for the manual preparation of wiring documents led to serious delays in supplying technical information to the workshops during the early manufacture of Pentaconta systems. This bottleneck was eliminated, however, by the computer.

2. Pentaconta Equipment

The basic Pentaconta unit is the frame, a rigid box in which the selectors and relays are mounted as shown in Figure 2. The front and rear are entirely accessible and may be closed by covers. The frames are mounted one above another to form bays; an installation consists of a number of parallel rows or suites of bays.

The frames come in 2 sizes—the smaller size 1 meter (39 inches) and the larger 1.290 meters

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(50.8 inches) wide, and the dimensions of the bays vary accordingly. The normal height of a frame is 0.390 meter (15.4 inches), enough to hold a crossbar switch or a vertical row of 11 relays, each carrying 2 spring pileups. Rarely, a frame only 0.235 meter (9.3 inches) high may be used in which 5 relays with 2 spring pileups each may be mounted in a vertical row. All frames are 0.200 meter (7.9 inches) deep.

The relays are mounted on metal strips one above another in vertical columns. The small and large frames have a maximum capacity of 15 and 20 strips, respectively, of relays having not more than 2 springs per pileup. The average figures are 10 and 14.

The relay strips are not installed individually within the frame, but are preassembled in groups of from 2 to 4 by means of a pair of right-angled supports, upper and lower, which form a subframe called a "platine". A platine (Figure 3) is essentially a wiring unit that does not necessarily constitute a circuit unit. It can hold 2 or even several circuit units of the same

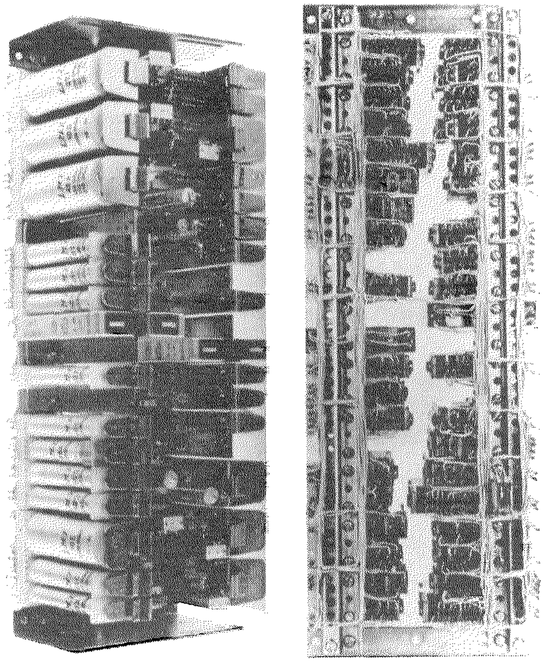

Figure 3—Platine equipped with relays before mounting on the frame. Front and rear views are at left and right, respectively.

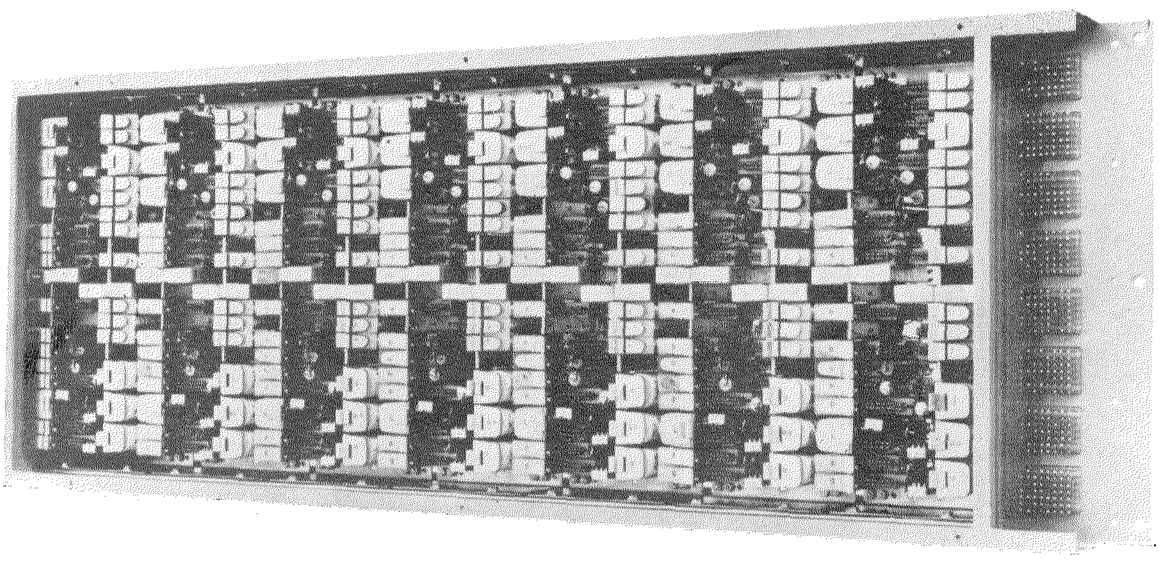

Figure 2—Front view of Pentaconta frame.

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type, and it can also support only part of a larger circuit unit that is distributed as logically as possible.

Fields of terminals are generally arranged at the left and right ends of the frame, allowing for connecting internal to external wiring. Some frames, however, carry only one terminal field; these terminals have one end accessible at the front and one at the rear. There are 8 vertical blocks of terminals in each field as shown in Figure 2. Each block is composed of 7 horizontal rows with 8 terminals to a row. The horizontal rows are known as "octuples." Thus 2 fields of 448 terminals are obtained (896 terminals for the whole frame). The internal wiring is connected to the terminals at the rear, the front being reserved for the external wiring.

2.1 Internal Wiring

The internal wiring is done in 2 separate operations at the workbench. This stems from the considerable number of connections required within a Pentaconta frame. The average number of connections per frame is approximately 3000 for the small model and 4000 for the large model, while the maximum number may be three times as high. With so many wires, it is impossible to follow the wiring method used for the rotary system. If applied to the Pentaconta system, the rotary arrangement would require wires of different colors to be laced together as a harness. The wires at each end would be connected to the terminal fields. The wires for components mounted to each strip would be run in an extension of the main harness as a smaller harness that parallels the mounting strip. In a small frame with 3000 connections (6000 wire ends to be connected), this system would involve an average of 500 wires for each smaller harness behind the strips, or more than 20 wires per lacing stitch or per outlet from the harness. Making connections by this method would be extremely difficult. Also,

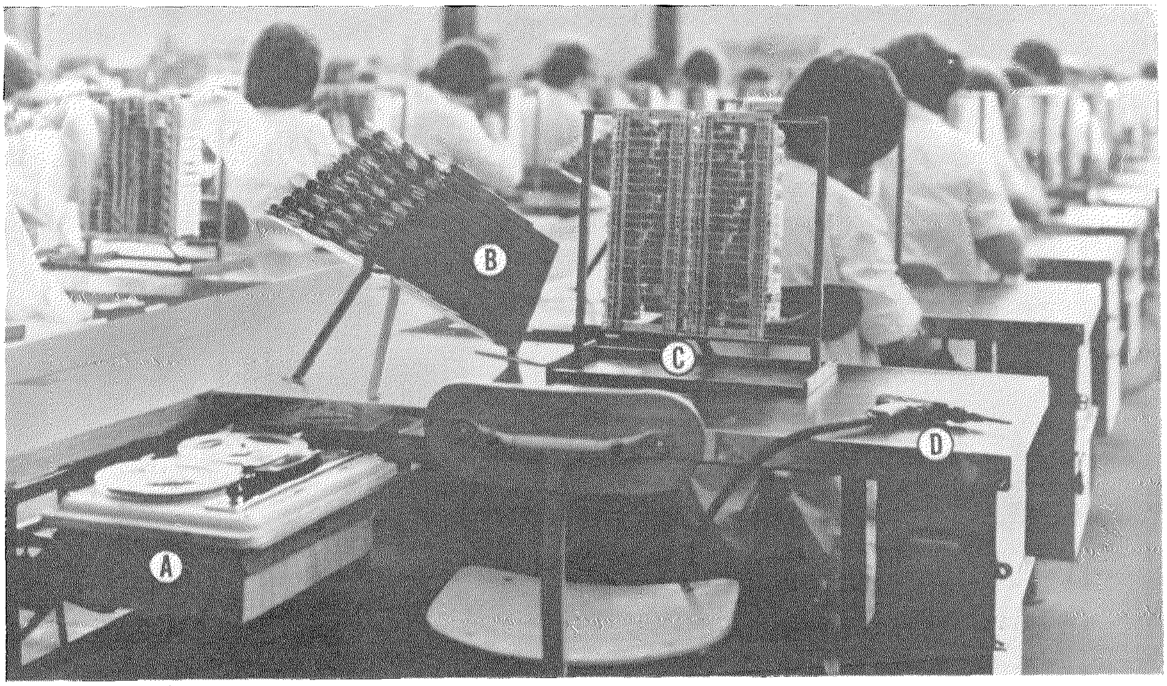

Figure 4—Cabling benches for common wires. A = magnetophone, B = rack with prepared wires, C = platine to be wired, D = wrapping tool.

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the color code cannot accommodate enough combinations for easy identification of the wires. The wiring side of the frame would have a tangled bunchy appearance, and access to connecting terminals for repairs would be virtually impossible.

The solution adopted limits the use of this harness to those wires connecting the octuples to the platines plus those between the platines, excluding the wires within any platine. The number of wires constituting the harness is thus reduced to a third and the approximate maximum is 1000 wires.

The internal connections of a platine are made wire by wire from one terminal to another along the bases of the components and along the mounting strips to each pileup. Not being laced into cables, these wires need no differentiation and are therefore all yellow.

Each terminal to be connected is designated by its "wiring name." This name comprises letters that denote the strip, the position on the strip of the relay, and the number of the terminal within the relay. For a specific connection the information ends with the length of wire to be used and the path the wire must follow. For convenience yellow wires are cut to different standard lengths and their ends are skinned; they are then stored by length in numbered compartments.

The path information shows plainly the changes in direction the wire must take. These elements of information take the form of lists, one line for each connection. As shown in Figure 4, the lists may be recorded on magnetic tape and reproduced as oral instructions so the operators need not shift their eyes repeatedly between the terminals and the lists. In these lists the wires are arranged to allow rapid and easy manipulation. The first type of list is the common-wiring list and may be supplemented by a checklist. Checking may consist of verifying the number of connections to each terminal; this number will be 0, 1, or a maximum of 2. Next the platines equipped with their common wires are assembled side by side in the frame as shown in Figure 5.

Meanwhile the harness has been prepared on a framework that can be adjusted for the exact siting of all wire terminations for any type of frame; the framework also indicates the placement of the harness outlets and the lacing.

The harness is made according to a list giving the color and terminations of both ends of each wire. This list is more or less optimized; those wires whose ends are contiguous, which may therefore be assembled together, are grouped to reduce manufacturing time. Similarly, all wires of the same color are listed successively to give the operator as few changes of wire as possible. The use of 28 colors and color-combinations permits each wire leaving the harness at a particular lacing stitch to be uniquely identified.

The harness checklist gives the number and colors of the wires at each harness outlet. The

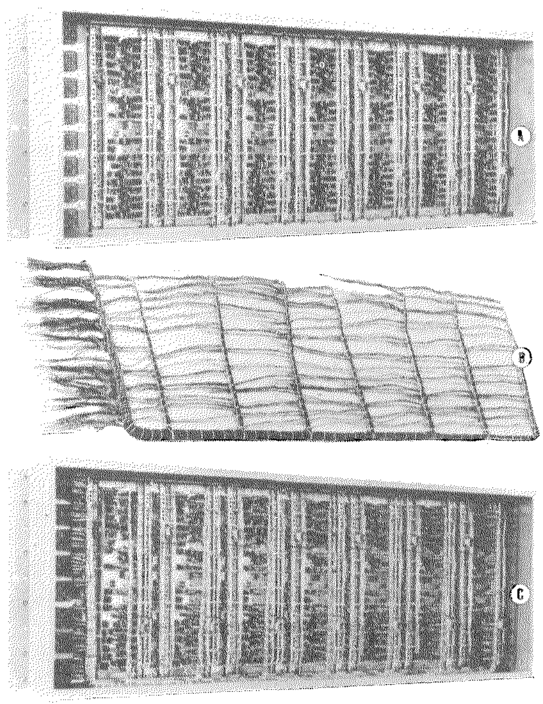

Figure 5—Cabling arrangements for Pentaconta frame. A = frame without harness, B = harness, C = frame with harness connected. Note that the frame has a field of terminals only at the left.

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harness is then placed on the assembled frame and attached according to the harness connection list, which gives the numbers of the terminals to which each wire in the harness is to be connected.

2.2 Preparation of Technical Information

In designing a new frame, an engineer draws the electrical circuit. This goes to a technician who designs a layout in which all the relays are represented. In this layout each relay is sited by indicating its strip and the mounting position on the strip; the strips are indicated by capital letters and the positions by lowercase letters. As there are 22 positions per strip, they are assigned the letters *a* through *v*. The technician then draws on a prepared form another diagram that gives the layout of the terminal fields to which the external wires are connected. On this form he identifies the use of each terminal of the different frame outlets. Wiring documents must always be prepared from these 3 drawings whether the method of construction is to be manual or automatic.

If the manual method is used, the preparation of wiring information starts with the drawing of a wiring plan, intended for the use of frame-terminal-control personnel, installation personnel, and the customer. This plan shows the wiring sides of the relays, selectors, and other elements, giving their physical characteristics concisely and the connecting terminals in full detail. These terminals are numbered, each having identification letters for the appropriate strip and position. The most-recent version of this plan consists of a series of paper sheets about 20 by 30 centimeters (8 by 12 inches), on each of which 2 strips and 6 relay positions may be shown; thus 4 sheets are needed for a complete diagram of 1 or 2 strips.

Above or below each terminal, the wiring name of the other terminal to which it must be connected is marked; also the color of the wire if it is to be in a harness. This plan serves as a basis for preparing all the required wiring lists.

It may be a drawback of the manual method that the first plan drawn is the last to be used; manufacturing cannot begin until the common-wiring and harness lists are ready. It may also be thought unfortunate that the advantages of the modular concept of Pentaconta equipment have not been more-fully realized. But the principal disadvantage is the length of time necessary to draw plans and prepare wiring lists.

The wiring plan is drawn by a wiring draftsman, while the lists are prepared by draftsmen or technicians who specialize in this area. The arrangement reduces the cost of technical information but involves extra delay in that different specialists act successively on the documents.

Considering the variety of frames, it is impossible to closely estimate the time necessary to prepare the documents or the wiring information; a total of 25 hours of drafting per strip, however, is generally accepted, with the overall period varying from 1 to 3 months after issuance of the schematic.

Using a computer to produce these documents serves primarily to reduce the time necessary for this work and also reduces the cost of technical information.

3. Description of Computer

The computer chosen for this work is the Bull Gamma *30*, shown in Figure 6. The central processing unit of this computer has a magnetic core memory for 20 000 alphanumeric characters, the location of each being individually addressed. The memory cycle, the time taken for selection and reproduction of a character, is 7 microseconds. A block of 6 magnetic tape units is connected to the central processing unit; each tape unit is capable, under program control, of reading in the forward or reverse direction and of writing in the forward direction at a rate of 10 000 characters per second.

A card reader is used for information input; this reads and translates 80-column punched cards at a rate of 600 per minute.

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Output documents are prepared on the line printer, which prints at 1000 lines per minute. A card punch that produces 100 cards per minute can also be used. This equipment represents the minimum needed for the program described in this paper; in particular, the process could not be done with fewer than 6 magnetic tape units.

This computer can be programed to perform a wide variety of operations. Such a program consists of a series of instructions, each of which is placed in the core memory of the central processing unit when it is needed. The program is carried out either in the order of storage or according to the most-recent instruction.

4. *LDF-2* Program

If an instruction requires 10 characters and the central memory can accommodate only 20 000 characters, the maximum number of instructions that may be stored is 2000; the operating maximum is even smaller since certain parts of the memory are reserved for other information or as processing zones.

The *LDF-2* program, designed for the issuance of wiring information, involves some 7000 instructions, to which must be added others concerning standard subprograms for sorting processes or other auxiliary functions. Consequently, the whole *LDF-2* program is divided into several segments, each of which comprises an information block. These blocks are stored on magnetic tape and are separated from each other on the tape by blanks. The information blocks are read one by one from the magnetic tape into the central memory and executed; thus all segments of the program are carried out successively, each one providing for a separate and distinct function.

During the execution of the program, 1 of the 6 tape units handles the program tape while the other 5 are used to read from or write on tapes used as auxiliary memories or for processing (as in sorting operations).

On the program tape, for convenience, the catalog of Pentaconta components has been stored at a location suitable for its use. This general catalog is composed of the lists of the various

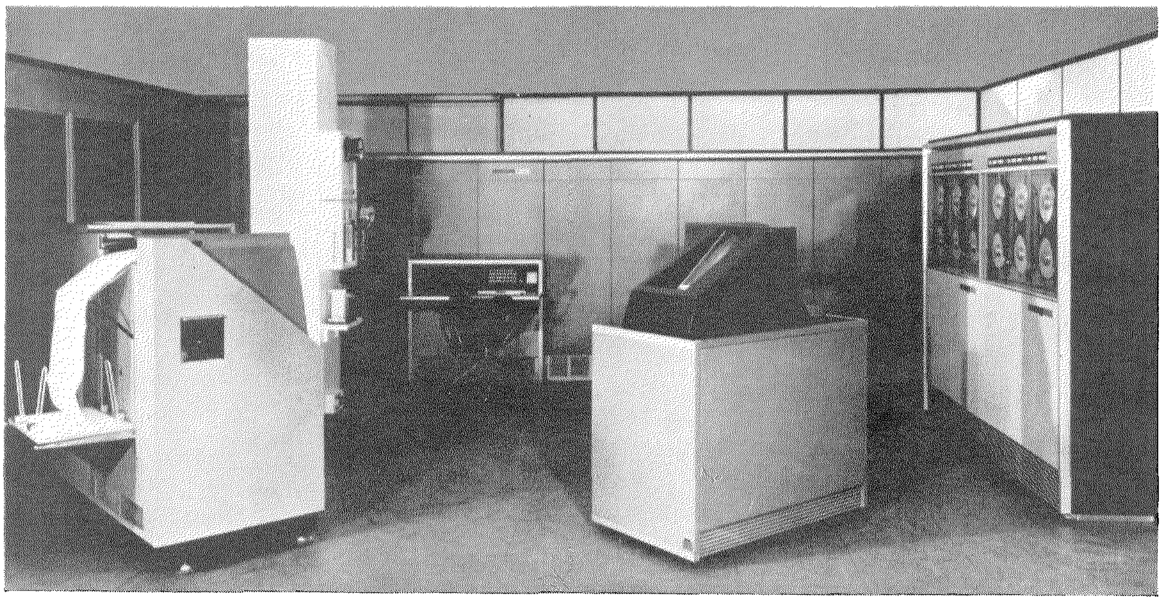

Figure 6—Gamma *30* computer.

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Pentaconta relay coils, relay pileups, and other components.

Each of these elements, identified by its code number, constitutes an information block on the magnetic tape; the various classes of elements are stored successively in the numerical order of their code numbers, which are used as the address of the information. Each block contains all necessary information for any terminal of the element.

In a preliminary step, information is taken from the general catalog and recorded on a processing tape to form a specific catalog concerning the elements used in a particular frame.

The tape on which the program and the general catalog are written contains the general information that enables the computer to issue wiring information, but this general information must be complemented by specific information concerning the frame to be handled. This specific information must be taken from the 3 basic documents and transferred to the computer by means of punched cards chosen especially for this function. It is therefore necessary to translate the 3 basic documents into data suitable for the computer and capable of being transcribed on the punched cards.

The most-satisfactory method for this transcription would be direct punching by the operator reading the 3 documents; however, the documents are not suitable for direct punching in their conventional form, and a complete reorganization to overcome this problem would not be acceptable.

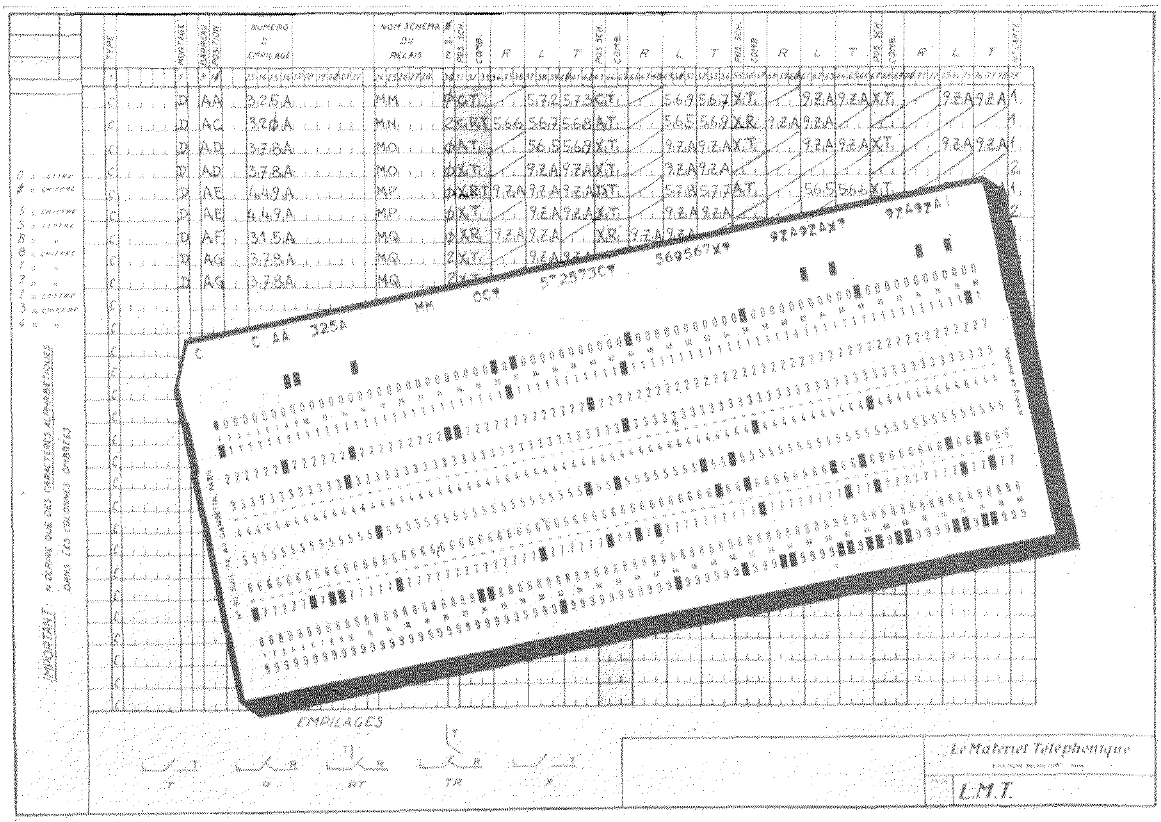

Figure 7—Input data on standard form and a punched card corresponding to the item entered on the first line of the form. This particular form covers relay pileups.

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A system of uniformly coding the information contained in the basic documents has therefore been drawn up; this can easily be transcribed on special printed forms. A different form is used for each type of element; one for relay coils, one for relay pileups, et cetera. Each form is printed with a grid that divides it into rows and columns.

There are 80 columns on a form (Figure 7) corresponding to the 80 columns of the punched card; each row on the form corresponds to a single element and consequently to a punched card. If an element has too many items, several rows and consequently several cards are used.

This process is extremely well suited for card punching; it also greatly reduces the possibility of errors or omissions. Each column corresponds to the same element for each item and any square on the form can contain only one element of information. A square left empty in error is easily noticed. These characteristics make self checking much easier.

The forms are filled in by experienced wiring personnel who have the opportunity to detect any errors in the basic document at this time. The duration of this operation is about 6 hours for a strip of 22 average-size relays.

Figure 8 shows some typical elements of a Pentaconta circuit. Three things should be noted.

(A) The relays are drawn with detached coils and contacts to permit short and straight connections.

(B) Each item on the drawing is labeled by its "schematic name," which identifies all the detached items that make up a single element, permitting reassembly of the detached parts.

(C) Information concerning each relay is summarized in a table shown in Figure 8. Each

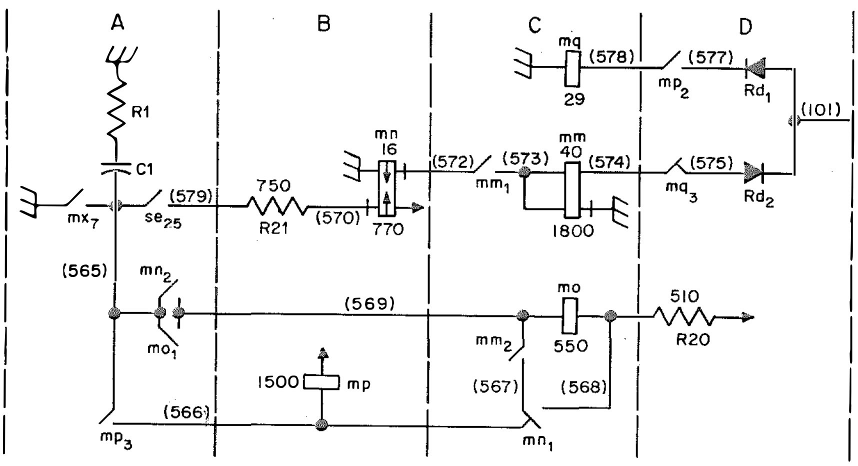

Figure 8—Part of a typical schematic diagram and relay table.

Relais	Contacts/Positions												Empil	Bobines
Position	1	2	3	4	5	6	7	8	9	10	11	12	Palet	Fonct
mm	T	T	T	T									325A	60G
C	C	C	X	X									U4	B22/B11
mn														
B														
	RT	T	R										320A	21K
	C	A	X										D6	B27/B20
mo	T	T	T	T	T	T							378A	50D
C	A	X	X	X	X	X							U2	21
mp	RT	T	T	T	T	T							449A	50X
B	X	D	A	X	X	X							U4	22
mq	R	R	R										315A	10K
C	X	X	D										D2	59
	T	T	T	T	T	T							378A	
	X	X	X	X	X	X								

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relay pileup occupies a double line. In the upper part are written from left to right: the schematic name of the relay, the type of each contact combination (*T* for a front or make contact, *R* for a back or break contact, *RT* for a transfer or changeover, *TR* for a continuity transfer or make before break, *X* for an early make), and finally the code numbers of the pileup and of the coil. (Not all designations are shown in the figure.) A location index is written under the schematic name of each relay and under each contact combination. This location index makes it easy to find each item on the schematic diagram, which is divided into corresponding columns. The operating current and the type of armature are indicated under the code numbers.

The schematic name of a coil generally has a minimum of 2 letters, the first indicating the function of the circuit, the second giving the chronological order, expressed alphabetically, of the relay function. Exceptions to this principle are possible; other letters may be added, and digits may be used as indexes. The contacts have the same schematic name as their coil, but the number of each contact combination is added as a subscript.

Components are connected by links, which may connect 2 or more terminals. These connections are known as "lines." A circuit diagram uses many such lines and these are numbered as shown on the diagram of Figure 8 between parentheses. For instance, the connection between resistor *R20* and the front contact of mn_1 is numbered (*568*). Lines numbered 1 and 2 are kept for battery and ground, respectively, which are multipled on a large number of terminals.

After all the lines are numbered, the various forms may be filled in by the technician. He begins by reading the diagram and particularly its relay table. First he takes from the table all data concerning the relay pileups in the order that they are listed; he writes them on a special pileup form, filling in one line of the form per pileup, or more if necessary, but beginning a new line or set of lines for each pileup.

Figure 7 shows a form for pileups prepared from the relay table of the schematic diagram. The code number of the pileup is put into columns 13–16 and the designation of the relay into columns 24–28. A 0 or 2 goes into column 30 depending on whether the relay has one or two pileups. The columns to the right of column 30 are for contact combinations at intervals of 12 columns per pileup. The first 3 columns of each block of 12 are shadowed to identify these blocks. The 3 shadowed columns carry the location index and the type of contact combination. The 9 remaining columns of each group of 12 are used in groups of 3 in which is written the number of the line connected to the terminal of the spring member of the contact combination.

Forms for relay coils and for other components are filled in similarly.

The technician then uses the equipment layout to determine the way each relay or component is to be mounted (right or left) and the location (letters identifying the relay strip and the place in the strip). He adds these data to the forms previously filled in from the diagram information.

The cards punched from the data on the forms are known as type *1* cards; they carry all the wiring information obtainable from the schematic drawing and equipment layout.

Several supplementary elements are also supplied to the computer by auxiliary cards (known as type *2* entry cards). These include the distribution of the strips on the platines (*CAD* cards), the attachment of the frame to the lateral terminal blocks (*ETI* cards), the ultimate repetition of identical circuits within the equipment (*REP* and *LII* cards), and the instructions and elements of dimensional information needed to calculate wire lengths and determine paths (*CHE* and *IND* cards).

After all the cards are punched, the work of the computer can start. This consists of dividing the lines of the schematic diagram into

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elementary connections each linking *2* terminals; of distributing these connections among the various wiring lists and giving for each elementary connection the wiring names of the terminals; and of calculating the length needed for common wires, specifying the path to be followed, and choosing the colors for harness wires. Afterward, the computer controls the printing of the wiring lists by the line printer.

All this is done according to the standard wiring rules included in program instructions and by reference to the general catalog stored on the program magnetic tape.

5. Computer Operation

The first segment of the program is brought to the central memory from the program tape by the only manual operation at the control panel. Thereafter the whole program progresses automatically, each segment as it finishes calling up the next one.

The computer operation flow is represented in Figure 9.

5.1

In the first program operation, the type *1* entry cards are read and then written on tape at the rate of 1 block of tape per card. Stored data are then examined for faults (misplaced columns, numbers in the letter columns, incorrect layout of lines, et cetera). If no errors are found the next segment is brought up; if errors are detected the list of faulty cards is written out on the printer with a list of the faults, and the program comes to a halt. This permits minimum loss of machine-hours for entry errors.

5.2

In the next operation the blocks of entry-information tape are sorted in ascending order of component code numbers. These are then compared with the stored general catalog, which is also arranged in ascending order of component code numbers.

The result of this comparison is a tape known as *T23,* which is the specific catalog of the components in the frame and contains an article of information for each component connecting point. This article gives the schematic name of the terminal (for example, *1 R1* = terminal *1* of resistor *R1* and *S NO* = coil outlet of relay *NO;* the location and size of the terminal within the frame; the wiring name of the terminal, which is the only appellation known by the worker to identify a terminal (this element of information and the preceding one are learned by consulting the catalog); and the number of the line to which the terminal is connected.

Such an article requires 28 alphanumeric character locations.

5.3

A sorting operation follows, in which the terminals of the different lines of the schematic are regrouped. Moreover, within any given line the terminals are sorted in the conventional equipment order (from top to bottom and from left to right of relay strip *A*, and then of the other relay strips in alphabetic order).

All the lines, in numerical order, are successively sorted in this way, and the result is the full list of the wires of the circuit.

5.4

In the next operation the type *2* input cards are read (with a similar check for errors), after which occurs automatic repetition (if several identical circuits are equipped in the same frame); plus the addition to each line, if necessary, of the corresponding outlet point on the lateral terminal blocks.

5.5

In the next operation, from among the wires sorted in Section 5.3, those are singled out that will constitute the harness (that is, those that do not have both terminals on the same platine). Each selected wire is color coded

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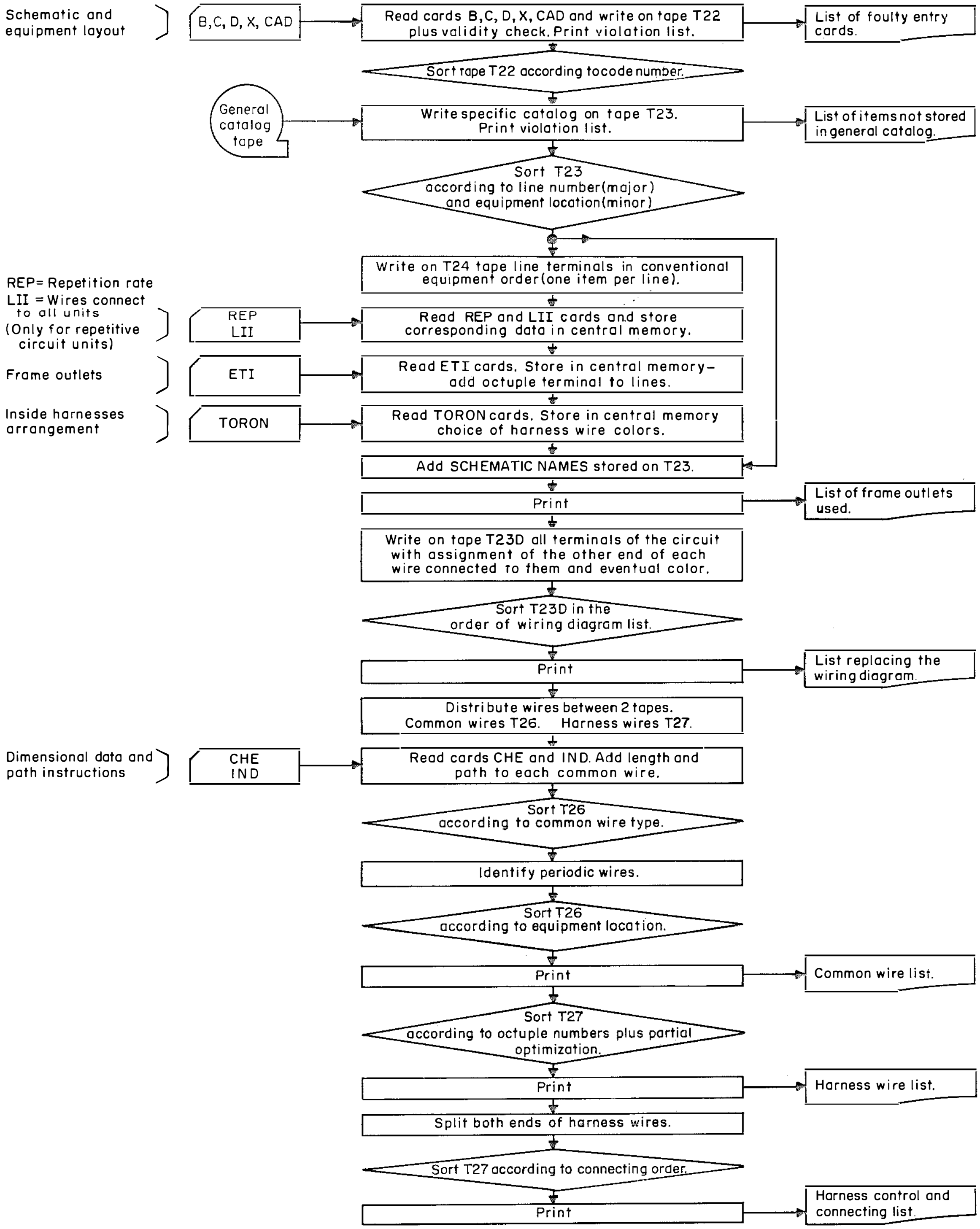

Figure 9—Flow chart.

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without permitting the same color twice in the same bunch. This is a fairly complex operation, since a harness may contain up to 1000 wires and only 28 color combinations are available; it is carried out entirely in the central memory.

5.6

Since the terminal data for Sections 5.4 and 5.5 are given as abridged designations because of limited space in the central memory, it is therefore necessary to complete the data by reference to the *T23* tape.

The next operation gives the first output. It provides a recapitulation list of the octuple outlets plus, for each wire connected to one of them, the wiring name of the terminal to which its other end will be connected as well as the wire color. (All these wires, of course, are harness wires.)

5.7

The later program operations prepare and print the list replacing the wiring plan. This lists all terminals and all elements of the frame in the conventional equipment order and gives for each terminal the final destination of the wire or of the 2 wires leaving it, plus the color if a harness wire is involved.

This list and the one mentioned in Section 5.6 are the fundamental documents that contain all logic wiring information; they are the instruments that are used for tests, maintenance, and for any work within the frame after manufacture.

5.8

The last part of the program applies to the preparation and printing of the following lists used for factory manufacture.

(A) The list of the common wires for each platine, sorted in the sequence that allows speediest wire placement. This sequence is decided on by the engineering department and is programed into the computer.

(B) The list of harness wires and their colors. It has not been possible to obtain automatically a sequence of operations that fully optimizes the harness-manufacturing process, since this would take up too much computer time. A compromise has therefore been adopted, and the computer produces a list in which the sequence of operations is partly optimized. Consequently, for frames that are to be manufactured in large quantities, this list must be improved manually.

(C) The list of harness connections gives the terminal for each end of the harness wires. This list has been drawn up in a way that permits checking the harness and thus makes unnecessary one of the lists normally required.

To recapitulate, the program goes through three distinct stages.

(A) Entry information is read and checked; this information is supplemented by data from a general catalog, and a specific catalog is created that facilitates the translation of schematic names into wiring names.

(B) Schematic lines are separated into elementary connections and distributed by category; their length is calculated and the paths of common wires are determined. In addition, the harness wires are color-coded. This part requires the computer's capabilities of logic and decision.

(C) Several ways of processing wire data permit issuance of the various lists. Each process has its own method of assembly, provides the presentation that is most suitable for specialized users, and controls typography print-out by the line printer.

It should be noted that the documents produced by the computer are similar in form and presentation to those formerly drawn by hand. The paper used by the computer has been chosen after many inquiries and tests. It is preprinted and provides an original on standard white paper and a copy on tracing paper. The tracing paper can be reproduced by the usual heliographic processes. Moreover, it is convenient

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to file and may be corrected in the same way as documents drawn up by hand.

6. Conclusion

The automatic system was first used in April 1963. Between then and June 1964, approximately 200 frames were processed by this method. It is estimated that by 1965 some 400 frames will be treated per year. At this rate the study and development costs of the program should be redeemed within 2 years.

The main advantage of this automated method is that it frees the technicians from a great number of operations that are repetitious, tedious, and subject to error. Moreover, the total time necessary to produce a complete set of wiring documents has been very-considerably reduced. Finally, it must be noted that the computer produces sets of documents that are certain to be mutually coherent and conform more systematically to wiring rules than those drawn up by hand (which may vary according to the whims of the writer).

It must be said that this method of automated wiring design may be extended to other electrical or electronic equipment, provided that a suitable program is written in each case. The possibilities offered by "thinking" computers, beginning with the earliest stages of new equipment design, should not be neglected; the programing of any such undertaking would then be much simpler and more complete than the one described here, the aim of which was more limited—to reduce the time required to prepare wiring documents.

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Power-Frequency Induction on Coaxial Cables With Application to Transistorized Systems *

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1. Introduction

For the transmission of power to dependent repeaters of coaxial cable systems, it is customary to employ the inner conductors of the coaxial tubes themselves using either direct current or alternating current of industrial frequency. The use of direct current has the great advantage of requiring less-involved and consequently less-costly equipment. On the other hand it implies metallic continuity throughout the power-feeding circuit. This circuit may extend over the full distance of perhaps 120 kilometres (75 miles) or more between power-feeding stations, but is reduced to 60 kilometres (37 miles) or so if the circuit is looped back midway between feeding stations. Circuits of such length need special consideration if the cable is exposed to induction from electric traction or power lines, especially where transistors and other miniature components are included in the repeaters.

The inductive effects are largely dominated by the earthing arrangement of the coaxial conductors. With coaxials insulated from earth, the effects are considerably smaller than with coaxials earthed at one or at several points. This study is mainly concerned, therefore, with the insulated arrangement. On cable routes where induction from external power sources in insignificant, the earthing arrangements do not affect the operation of the system.

Throughout this study, "earth" refers to a remote ground in cases where there is no overall metallic cable sheath and to the metallic sheath where such is provided. The expression "power-feeding section" means a section of line the repeaters of which are fed by the same power-feeding circuit.

In the first part of this study, the general case is considered of a coaxial cable exposed to induction from an external power source. A method is developed of estimating the currents that appear in such circumstances on the coaxial inner conductors carrying the power-feeding current. The method also shows how the voltages that arise between coaxial inner and outer conductors can be calculated.

In the second part of the study, the method is applied to a case of particular interest.

2. Theoretical Treatment

2.1 Coaxials Insulated Throughout Power-Feeding Section

The type of cable envisaged consists of coaxial tubes arranged in concentric layers. If there is more than one layer, no consideration will be given to the tubes in the inner layers, for these being shielded to some extent by the outer layer are less susceptible to induction from external sources. For the same reason, cables are excluded in which the coaxial tubes are surrounded by one or more layers of balanced pair circuits. The worst condition in practice is that of a small cable consisting of four, six, or eight coaxial tubes with their complement of service pairs, and it is this condition which forms the principal object of the present study.

The tubes may be individually insulated or in metallic contact with each other and may be enclosed in a plastic or metallic sheath with or without armour.

In Figure 1 the length of the coaxial power-feeding section is denoted by l, the length exposed to induction from an external source by $\alpha_2 l$, and the lengths of the unexposed sections on either side by $\alpha_1 l$ and $\alpha_3 l$, $(\alpha_1 + \alpha_2 + \alpha_3 = 1)$. α_2 varies between the limits of 0 and 1. When

* Abridgment of a study submitted to the Comité Consultatif International Télégraphique et Téléphonique on 7 June 1962.

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$\alpha_2 = 0$, induction is assumed to be concentrated at the junction of the unexposed sections; when $\alpha_2 = 1$, induction is distributed over the whole feeding section. A typical point of the power-feeding section is denoted by xl $(0 \leqslant x \leqslant 1)$.

The longitudinal electromotive force induced along the exposed section $\alpha_2 l$ is assumed to be uniform and to have a total root-mean-square value of E volts. If the coaxial tubes are enclosed in a metallic sheath with or without steel-tape armour, E is to be interpreted as the potential drop along the inside surface of the metallic sheath. For long exposures this potential drop is substantially uniform and the end effects may be ignored. For short exposures the end effects may be taken into account by the adoption of an effective exposure length over which the potential drop is assumed to be uniform.

The two types of circuit subjected to the influence of E are:

(**A**) The circuit formed by the outer conductors of all coaxial tubes in the cable in parallel and earth or the metallic sheath.

(**B**) The circuit formed by all coaxial outer conductors in parallel and the inner conductors of all tubes effectively in parallel.

By reason of symmetry, the current in (**A**) will be equally divided among all the outer conductors, as will also be the total current in (**B**). To derive the transmission equations it is therefore sufficient to confine attention to a single coaxial tube, provided that in the case of (**A**) the value adopted for the shunt admittance is that of the direct admittance of the tube to earth or sheath in the particular cable make-up under consideration.

By combining the electrical properties of the coaxial tube with those of the power-feeding circuit equipment, two equivalent uniform transmission lines are obtained (Figure 1). These are coupled together by the transfer impedance of the coaxial outer conductor and thus form a system amenable to treatment by transmission-line analysis. It is understood that the analysis is concerned only with low frequencies of the order of 50 hertz.

The following plain and barred symbols are used, the plain symbols referring to the equivalent circuit formed by the coaxial outer conductor and earth or metallic sheath, and the barred symbols to the equivalent circuit formed by the coaxial inner and outer conductors:

$Z, \bar{Z}$ = series impedance in ohms per kilometre

$Y, \bar{Y}$ = shunt admittance in mhos per kilometre

$\Gamma, \bar{\Gamma}$ = propagation constant per kilometre

Z_t = transfer impedance of coaxial outer conductor with external return in ohms per kilometre

$I_n(xl), \bar{I}_n(xl)$ = root-mean-square current in amperes

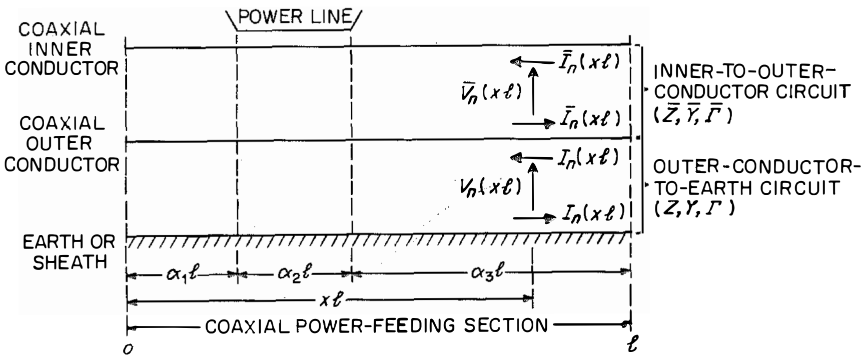

Figure 1—General case of partial exposure showing equivalent uniform transmission lines.

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$V_n(xl)$, $\bar{V}_n(xl)$ = root-mean-square voltage in volts

where n assumes the values 1, 2, or 3 depending on whether the symbols refer to the unexposed section to the left of the exposed section, the exposed section, or the unexposed section to the right of the exposed section.

The driving force per unit length at any point xl on the inner-outer circuit is taken as $Z_t I_n(xl)$.

The equations for the currents and voltages in each of the three sections have been established in the usual way by writing down for each of the three sections the general equations with two arbitrary constants, and by determining the six constants from the boundary conditions. These require that the currents at the two ends of the power-feeding section vanish in both circuits (the direct-current power source as well as the inner and outer coaxial conductors being insulated from earth) and that the currents and voltages are continuous at the two ends of the exposed section. An involved but otherwise straightforward calculation gives the following set of equations:

$$\left.\begin{aligned} I_n(xl) &= \frac{E}{Zl} G_n(\Gamma l, \alpha_1, \alpha_2, \alpha_3, x) \\ V_n(xl) &= -\frac{E}{\Gamma^2 l^2}\frac{\partial}{\partial x} G_n(\Gamma l, \alpha_1, \alpha_2, \alpha_3, x) \\ &= -\frac{1}{Yl}\frac{\partial}{\partial x} I_n(xl) \end{aligned}\right\} \quad (1)$$

$$\left.\begin{aligned} \bar{I}_n(xl) &= \frac{EZ_t}{Z\bar{Z}l}\frac{\bar{\Gamma}^2}{\bar{\Gamma}^2 - \Gamma^2}\left\{G_n(\Gamma l, \alpha_1, \alpha_2, \alpha_3, x) - \frac{\Gamma^2}{\bar{\Gamma}^2} G_n(\bar{\Gamma} l, \alpha_1, \alpha_2, \alpha_3, x)\right\} \\ \bar{V}_n(xl) &= -\frac{EZ_t}{Zl^2}\frac{1}{\bar{\Gamma}^2 - \Gamma^2}\left\{\frac{\partial}{\partial x} G_n(\Gamma l, \alpha_1, \alpha_2, \alpha_3, x) - \frac{\Gamma^2}{\bar{\Gamma}^2}\frac{\partial}{\partial x} G_n(\bar{\Gamma} l, \alpha_1, \alpha_2, \alpha_3, x)\right\} \\ &= -\frac{1}{\bar{Y}l}\frac{\partial}{\partial x}\bar{I}_n(xl) \end{aligned}\right\} \quad (2)$$

where the functions denoted by G_n have the following forms depending on whether α_2 differs from or is equal to zero, and whether n, being 1, 2, or 3, refers to the unexposed section to the left of the exposed section, the exposed section, or the unexposed section to the right of the exposed section:

For $\alpha_2 \neq 0$

$$\left.\begin{aligned} G_1(\Gamma l, \alpha_1, \alpha_2, \alpha_3, x) &= \frac{\cosh \Gamma l(1-\alpha_1) - \cosh \Gamma l\alpha_3}{\alpha_2 \sinh \Gamma l}\sinh \Gamma lx \\ G_2(\Gamma l, \alpha_1, \alpha_2, \alpha_3, x) &= \frac{\cosh \Gamma l(1-x) - \cosh \Gamma l\alpha_3}{\alpha_2 \sinh \Gamma l}\sinh \Gamma lx \\ &\quad + \frac{\cosh \Gamma lx - \cosh \Gamma l\alpha_1}{\alpha_2 \sinh \Gamma l}\sinh \Gamma l(1-x) \\ G_3(\Gamma l, \alpha_1, \alpha_2, \alpha_3, x) &= \frac{\cosh \Gamma l(1-\alpha_3) - \cosh \Gamma l\alpha_1}{\alpha_2 \sinh \Gamma l}\sinh \Gamma l(1-x). \end{aligned}\right\} \quad (3A)$$

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For $\alpha_2 = 0$

$$\left.\begin{aligned} G_1(\Gamma l, \alpha_1, 0, \alpha_3, x) &= \frac{\Gamma l \sinh \Gamma l \alpha_3}{\sinh \Gamma l} \sinh \Gamma l x \\ G_2(\Gamma l, \alpha_1, 0, \alpha_3, x) &= \frac{\Gamma l \sinh \Gamma l \alpha_1 \sinh \Gamma l \alpha_3}{\sinh \Gamma l} \\ G_3(\Gamma l, \alpha_1, 0, \alpha_3, x) &= \frac{\Gamma l \sinh \Gamma l \alpha_1}{\sinh \Gamma l} \sinh \Gamma l(1 - x). \end{aligned}\right\} \quad (3B)$$

These expressions are complex functions of as many as four independent non-dimensional variables and in consequence they are not amenable to convenient graphical representation. To make them useful for practical purposes generally, they may be simplified at the cost of restricting their field of application to symmetrical exposures.

In most practical cases the propagation constants have angles of about 45 degrees, so that no significant error is introduced by assuming that their real and their imaginary parts have the same value.

2.2 Coaxials Earthed at One End of Power-Feeding Section

Since the derivation of general transmission equations for any scheme involving an exposed section of arbitrary length and position between two unexposed sections is a laborious task and, since in this instance the results would be of little practical interest, no attempt has been made to repeat the process here. It was considered that in these circumstances it would be sufficient to examine the situation from a less-general point of view, by assuming that there is no or only one unexposed section, that is, either that induction is distributed over the whole of the power-feeding section or that it extends from one end to some arbitrary point along the section. With this restriction it is found that the equations developed for the insulated scheme are also applicable, with slight modifications, to the earthed scheme and, although the validity of the conclusions so reached is limited, these conclusions are sufficiently definite to leave little doubt as to the general situation.

Since V and $\bar{V}$ are always zero at $l/2$ in an insulated scheme with symmetrical exposure, outer and inner conductors may be earthed at that point without disturbing the current or voltage distributions on either of the circuits; in fact, half of the feeding section could be removed without affecting the interference phenomena in the other half. Each half represents precisely the earthed scheme under consideration, and hence the equations established for the insulated symmetrical scheme apply rigorously provided that l be replaced by $2l$ and E by $2E$.

Making the appropriate substitutions in the equations for I, it is found that the maximum value of I in the earthed scheme compares with that in the insulated scheme as shown in Figure 2. Thus, whenever the real part of Γl is less than about $1 \cdot 5$ (this covers the range of practical interest), the maximum current I in an earthed scheme is more than twice that encountered in an insulated scheme. For the typical system considered in Section 3, with a power-feeding section of 64 kilometres (40 miles) (real part of $\Gamma l = 0 \cdot 7$), the effect of earthing would be to increase the maximum current by a factor of nearly 4, while for a power-feeding section of 128 kilometres (80 miles) (real part of $\Gamma l = 1 \cdot 4$) the factor would still be over 2.

Similarly for V, it may be shown that for the case of $\alpha_2 = 1$ the ratio of the maximum voltage V in an earthed scheme to that in an

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insulated scheme is equal to half the ratio plotted in the upper curve ($\alpha_2 = 0$) on Figure 2; it thus remains greater than unity over the range of practical interest. For the typical system the effect of earthing would be almost to double the maximum voltage, while for a power-feeding section of 128 instead of 64 kilometres, it would still be increased by a factor of about $1 \cdot 2$.

The corresponding ratios of $\bar{I}$ and $\bar{V}$ for earthed and insulated schemes are rather involved for general analytical treatment; it would strictly be necessary to represent them by families of curves for each value of the factor α_2 with the ratio $\Gamma/\bar{\Gamma}$ as parameter. However, it is found in the case of $\bar{I}$ that, if the ratios are plotted against $\bar{\Gamma}l$ as independent variable, the curves do not differ greatly over the range of practical interest for different values of α_2 or $\Gamma/\bar{\Gamma}$ and have a general form similar to that of the Figure 2 curves. Thus, for small values of $\bar{\Gamma}l$ the maximum current $\bar{I}$ in an earthed scheme in all cases approaches 16 times that in an insulated scheme; it remains more than twice that in an insulated scheme for values of the real part of $\bar{\Gamma}l$ up to about 3 and values of $\Gamma/\bar{\Gamma}$ up to about $0 \cdot 5$, that is, over the range of practical interest.

Similar considerations apply in the case of $\bar{V}$, except that the ratios are about half the corresponding $\bar{I}$ ratios. Thus, for small values of $\bar{\Gamma}l$ the maximum voltage $\bar{V}$ in an earthed scheme approaches 8 times that in an insulated scheme and it remains greater than that in an insulated scheme over the range of practical interest.

Investigation shows that for a power-feeding section of 64 kilometres (real part of $\bar{\Gamma}l = 1 \cdot 75$) the effect of earthing would be to increase the maximum current $\bar{I}$ by a factor of nearly 7, while for a power-feeding section of 128 kilometres (real part of $\bar{\Gamma}l = 3 \cdot 5$) the factor would still be over 2. The corresponding factors for the maximum voltage $\bar{V}$ would be over 3 and about $1 \cdot 2$ respectively.

2.3 Coaxial Outer Conductors Earthed Throughout Power-Feeding Section

In this case there is only one circuit to be considered, namely the circuit formed by the outer conductors of all coaxial tubes in the cable in parallel with the metallic sheath, if any, and the inner conductors of all tubes effectively in parallel. As before, it is sufficient to confine attention to a single coaxial tube provided that, in determining the effective series impedance due to the outer conductor, allowance is made for the impedance of the sheath in relation to the number of tubes. The current on any coaxial inner conductor and the voltage between any coaxial inner conductor and its outer conductor are then given by (1).

The above procedure refers to the case in which the inner conductors are insulated throughout. The case in which the inner conductors are earthed at one end of the power-feeding section may be treated on an analogous basis to that used in Section 2.2.

In both these cases it may be shown that the magnitudes of current and voltage are substantially greater than for the coaxial inner-to-outer-conductor circuit in an insulated scheme.

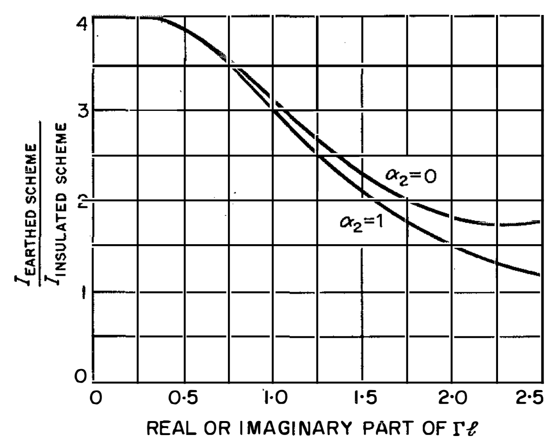

Figure 2—Ratio in which maximum current in the coaxial outer-to-earth circuit is increased when coaxial conductors are earthed at one end of the power-feeding section, instead of being insulated throughout.

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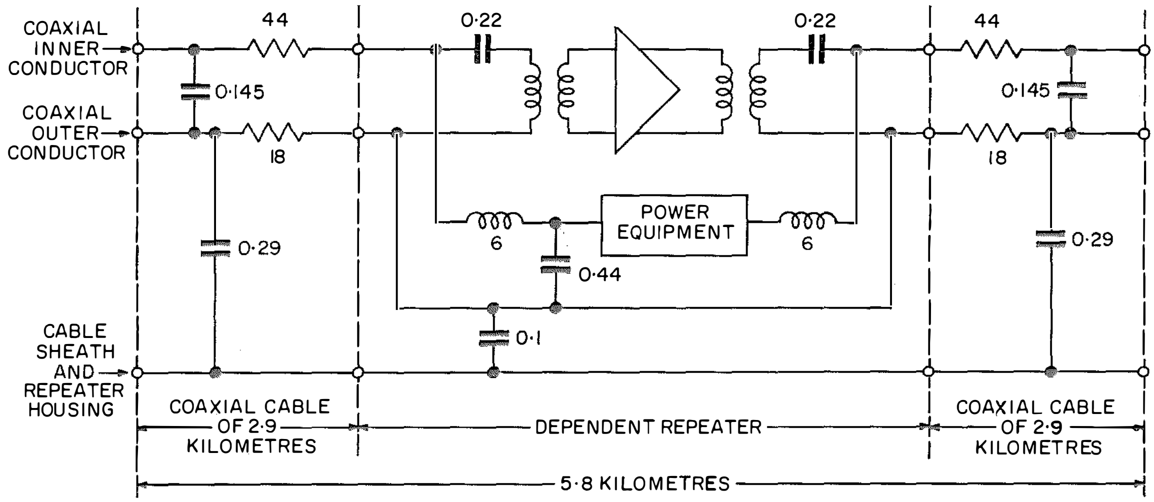

Figure 3—Repeater section of small-diameter coaxial cable system with direct-current power feeding. Capacitance values are in microfarads, and all other values are resistances in ohms.

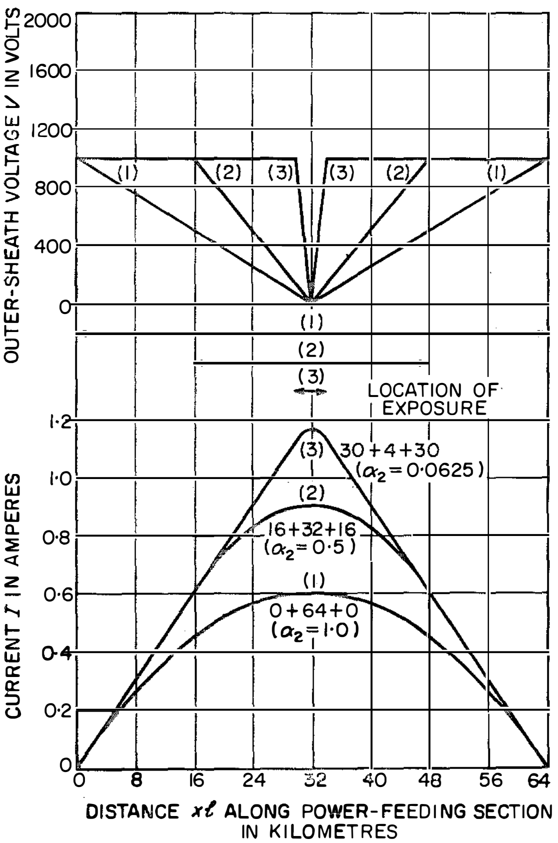

Figure 4—Voltages and currents in outer-to-sheath circuit for symmetrical exposures. Unexposed and exposed lengths are in kilometres.

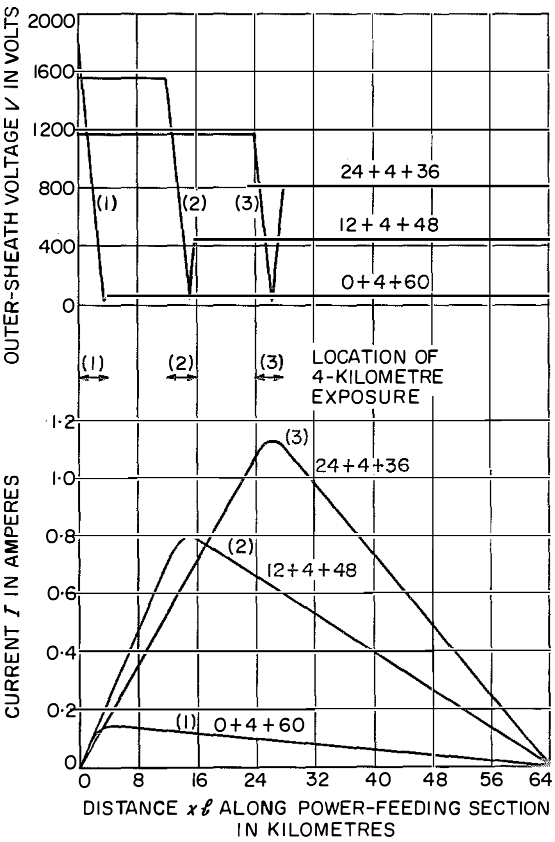

Figure 5—Voltages and currents in outer-to-sheath circuit for asymmetrical exposures. $\alpha_2 = 0 \cdot 0625$. Unexposed and exposed lengths are in kilometres.

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3. Application of General Equations to a Typical Insulated System

This system conforms with recommendations of the Comité Consultatif International Télégraphique et Téléphonique for 300-channel systems on small-diameter coaxial cable. The nominal spacing of repeaters is 5·8 kilometres (3·6 miles) and the length l of the direct-current power-feeding section is about 64 kilometres (corresponding to a spacing of about 133·5 kilometres (83 miles) between power-feeding points). Figure 3 is a diagram of the dependent repeater and the two adjoining half repeater sections of coaxial cable (4·42 millimetres (0·174 inch) diameter).

The repeater power equipment indicated presents a relatively small non-linear complex impedance to alternating current, and preliminary calculations showed that the effect of neglecting this impedance in comparison with the series impedance of the coaxial inner-to-outer-conductor circuit would be to increase only slightly the computed values of induced current $\bar{I}$ and voltage $\bar{V}$ in that circuit. Detailed computations have accordingly been carried out on such a basis, thereby giving maximum values of $\bar{I}$ and $\bar{V}$ that are independent of the power-equipment impedance.

In the following example the basic cable parameters were measured on a lead-sheathed cable containing four coaxial tubes. No significant deviations from these values would be expected for a cable consisting of six or eight tubes as long as the tubes are placed symmetrically with respect to the sheath.

The transfer impedance Z_t has been taken as equal to the series impedance Z of the coaxial outer conductor. The value of E has been taken arbitrarily as 2000 volts at 50 hertz, regardless of the length of the exposed section. In the case of induction distributed over the whole coaxial power-feeding section ($\alpha_2 = 1$), this corresponds to 2000/64 = 31·25 volts per kilometre (50 volts per mile) and, in the case of induction concentrated over a short section of say 4 kilometres ($\alpha_2 = 0{\cdot}0625$), to 2000/4 = 500 volts per kilometre (800 volts per mile).

If E differs from 2000 volts, all values of I, V, $\bar{I}$, and $\bar{V}$ quoted hereafter or indicated on the accompanying curves should be multiplied by the factor $E/2000$.

The range of inducing fields encountered in practice is between a few volts per kilometre and several hundred volts per kilometre. The highest values occur in the immediate vicinity of an electric railway. If a high inducing field persists in length over more than a few kilometres, for instance the length of a tunnel, shielding measures may be needed in the design of the cable.

On the basis of the parameter values indicated in Figure 3, the general equations (1) and (2) have been used to calculate the values of I, V, $\bar{I}$, and $\bar{V}$ for a comprehensive range of exposure lengths and positions. An electronic computer was employed for this purpose, the expressions being evaluated at sufficiently close intervals to determine accurately the distribution of current and voltage along the coaxial power-feeding section for each set of conditions.

Typical curves based on the results obtained are shown in Figures 4 to 9, and these are discussed below. In all cases the curves give the modulus of the current or voltage (root-mean-square value). To indicate the length and position of the exposed section and the lengths of the unexposed sections on either side of it, three numbers denoting in kilometres the lengths $\alpha_1 l + \alpha_2 l + \alpha_3 l$ (see Figure 1) are affixed to each curve on Figures 4 to 7. The value of the factor α_2 for the exposed section is also indicated.

Figure 4 shows the distribution along the coaxial power-feeding section of the current I and voltage V in the coaxial outer-to-sheath circuit for symmetrical exposures of 4, 32, and 64 kilometres ($\alpha_2 = 0{\cdot}0625$, $0{\cdot}5$, and 1), while Figure 5 shows typical distributions for an asymmetrical exposure of 4 kilometres ($\alpha_2 = 0{\cdot}0625$). It will be seen that I is greatest

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within the exposed section, increasing or decreasing substantially linearly over the adjoining unexposed sections, and reaches a maximum when the induction is concentrated at the centre point of the power-feeding section; V is greatest at the ends of the exposed section, remaining substantially unchanged over the unexposed sections and varying linearly within the exposed section, and reaches a maximum when the induction is concentrated at either end of the power-feeding section.

Figures 6 and 7 show the corresponding distributions of the current $\bar{I}$ and voltage $\bar{V}$ in the coaxial inner-to-outer-conductor circuit. Here it will be seen that $\bar{I}$ is greatest near the centre of the coaxial power-feeding section and reaches a maximum (as for I) when the induction is concentrated at the centre point; $\bar{V}$ is greatest at the ends of the coaxial power-feeding section and reaches a maximum when the induction is concentrated at a point slightly to either side of the centre of the section.

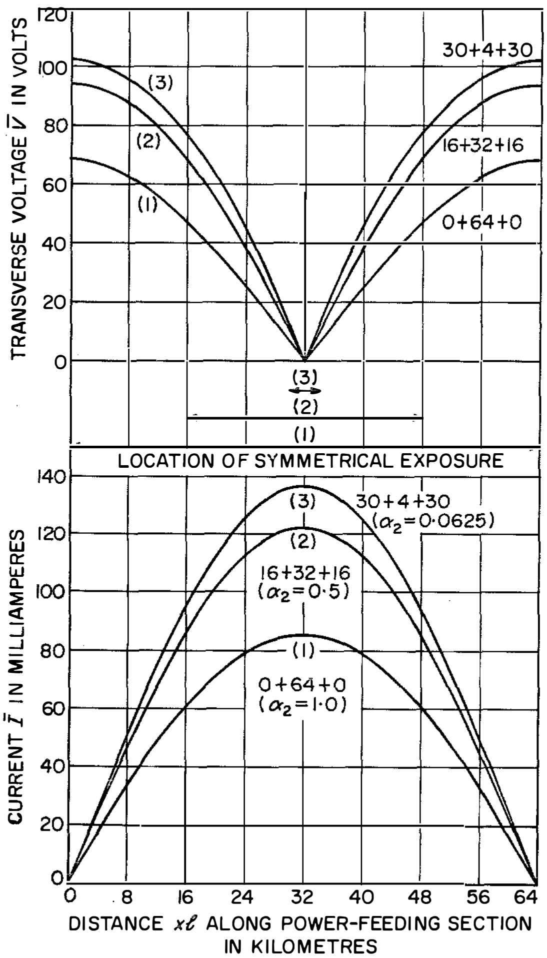

Figure 6—Voltages and currents in inner-to-outer-conductor circuit for symmetrical exposures. Unexposed and exposed lengths are in kilometres.

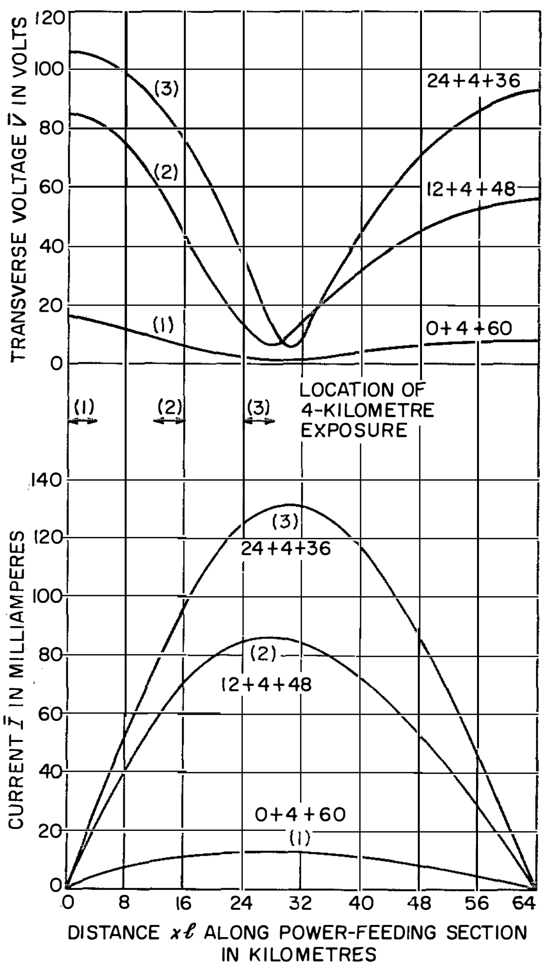

Figure 7—Voltages and currents in inner-to-outer-conductor circuit for asymmetrical exposures. $\alpha_2 = 0 \cdot 0625$. Unexposed and exposed lengths are in kilometres.

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The important role played by the position of the exposure is well illustrated in Figure 7. As the exposure of 4 kilometres moves from one end towards the centre of the power-feeding section, the maximum of $\bar{I}$ increases from 13 to over 130 milliamperes while the maximum of $\bar{V}$ increases from 16 to 105 volts. Thus if the power-feeding section includes for instance a 4-kilometre tunnel of an electric railway, but can otherwise be routed so as to be free from induction, it is desirable that one of the ends of the power-feeding section should coincide with the entry to the tunnel.

This striking effect of the position of the exposure on the maximum values of $\bar{I}$ and $\bar{V}$ is not obvious and has only been revealed by mathematical analysis.

Figures 8 and 9 summarize the computed results for the inner-to-outer-conductor circuit in a form suitable for estimating the maximum current $\bar{I}$ and the terminal voltage $\bar{V}$ for exposures of any length and position within a coaxial power-feeding section. In Figure 9 the same curves may be used for determining the

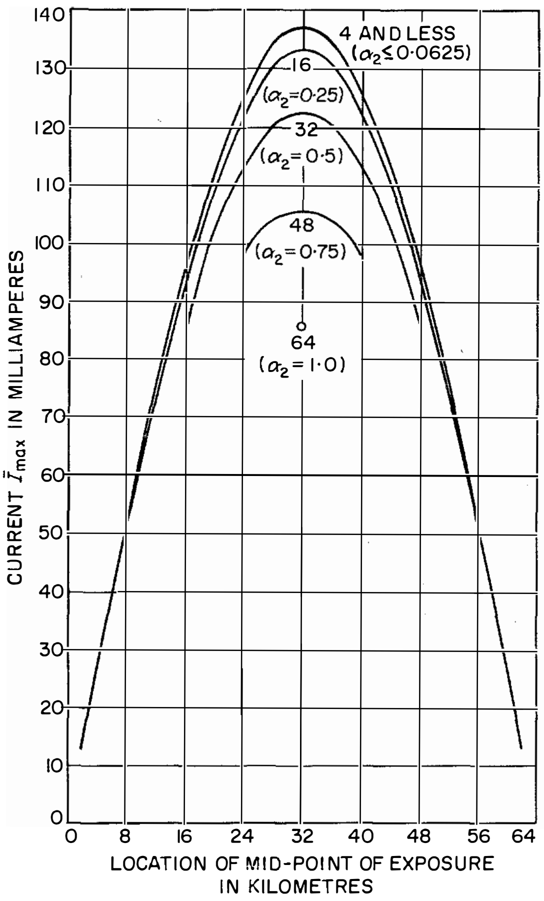

Figure 8—Variation of current $\bar{I}_{max}$ in inner-to-outer-conductor circuit with change of location and length of exposure. The length of exposure in kilometres for which each curve applies is indicated beside the curve. For all locations and lengths of the exposure, $\bar{I}_{max}$ occurs within approximately 5 kilometres of the centre of the power-feeding section.

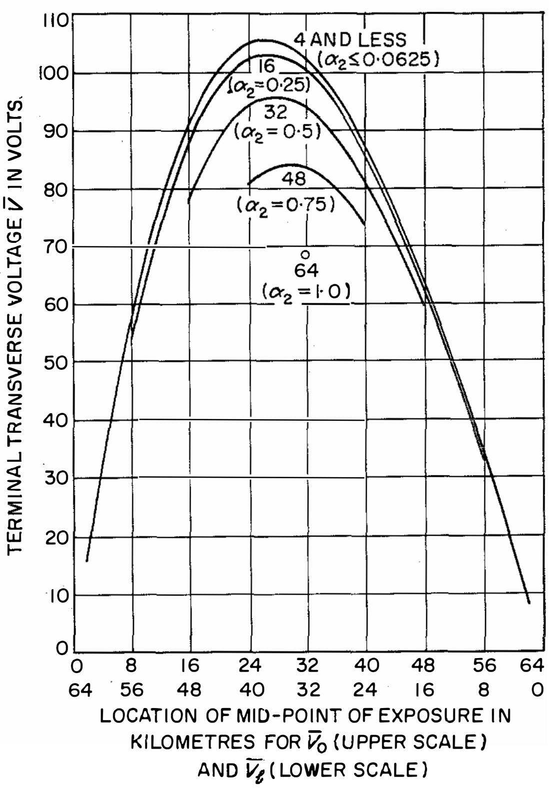

Figure 9—Variation of terminal voltage $\bar{V}$ in inner-to-outer-conductor circuit with change of location and length of exposure. The length of exposure in kilometres for which each curve applies is indicated beside the curve. For all locations with the mid-point of exposure at $<l/2$ (that is, 32 kilometres), $\bar{V}_0$ exceeds $\bar{V}_l$; for all locations with the mid-point of exposure at $>l/2$, $\bar{V}_l$ exceeds $\bar{V}_0$.

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voltages at the two ends of the section (designated by $\bar{V}_0$ and $\bar{V}_l$) by employing the appropriate distance scale, the figures on both scales representing the distance $\{\alpha_1 + (\alpha_2/2)\}l$.

It should be noted that, with the particular basic parameter values assumed, the maximum current $\bar{I}$ always occurs within about 5 kilometres of the centre of the power-feeding section for any position or length of the exposure, while the maximum terminal voltage $\bar{V}$ occurs when the mid-point of the exposure lies within a similar distance of the centre of the power-feeding section. If different basic parameter values were employed, these limits would in general be modified, although the fundamental form of the curves would remain unaltered.

4. Conclusions

The arrangement in which coaxial conductors and associated equipment are insulated throughout the power-feeding section is superior to that in which the coaxial outer and inner conductors are earthed at one end of the power-feeding section, since it gives considerably smaller effects arising from currents in the dielectric. The following conclusions apply only to this insulated arrangement:

(**A**) For a given total induced longitudinal electromotive force, the current in the circuit formed by the coaxial outer conductor and earth or sheath is always greatest within the exposed section, increasing or decreasing substantially linearly over the adjoining unexposed sections, and reaches a maximum when the induction is concentrated at the centre of the power-feeding section.

The voltage between the coaxial outer conductor and earth or sheath is always greatest at the ends of the exposed section, remaining substantially unchanged over the unexposed sections, and reaches a maximum when the induction is concentrated at either end of the power-feeding section.

(**B**) For a given total induced longitudinal electromotive force, the current in the circuit formed by the coaxial inner and outer conductors is always greatest near the centre of the coaxial power-feeding section, reaching a maximum when the induction is concentrated at the centre and a minimum when it is concentrated at either end of the section.

The voltage between coaxial inner and outer conductors is always greatest at the ends of the power-feeding section, reaching a maximum when the induction is concentrated slightly to either side of the centre of the section and a minimum when it is concentrated at either end of the section.

(**C**) For any given length and position of exposure, the currents and voltages in both circuits are directly proportional to the induced longitudinal electromotive force.

(**D**) For a given total induced longitudinal electromotive force, increasing the length of the coaxial power-feeding section increases the maximum current in the circuit formed by the coaxial outer conductor and earth or sheath, as well as the maximum current and voltage in the circuit formed by the coaxial inner and outer conductors.

The proportional increase is greatest in the case of the current in the coaxial inner-to-outer-conductor circuit.

John Kemp received a diploma in electrical engineering in 1921 from the Imperial College of Science and Technology in London. He then joined the International Western Electric Company and, after training in the United States in the Bell System cable factories and laboratories, he assisted in introducing into Europe the long-distance telephone cable developed in the United States.

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Since 1946 Mr. Kemp has been a member of the senior technical staff of Standard Telecommunication Laboratories engaged chiefly with inductive coordination of communication and power lines and with the protection of communication cables against lightning.

He is a member of the Institution of Electrical Engineers. He has published several papers on waveguides. Since 1946 he has served on inductive coordination and protection study groups of the Comité Consultatif International Télégraphique et Téléphonique.

Howard W. Silcock was born in Maidstone in 1916. After studying at University College, Southampton, he received a B.Sc. (London) general degree in 1936, and a B.Sc. (London) special degree in physics in 1937.

He then joined the Systems Planning Division of Standard Telephones and Cables and worked on carrier cable and open-wire projects. He later transferred to the newly formed Standard Telecommunication Laboratories, where he has been concerned with crosstalk and other aspects of line transmission, including carrier cable balancing techniques and open-wire transposition schemes.

C. J. Steward was born in 1908, was educated at Southampton, and received a B.Sc. (London) general degree. He joined Standard Telephones and Cables at North Woolwich in 1936, when carrier telephony was becoming established on cables, and dealt with crosstalk problems.

After spending two years with the Valve Division, he returned to North Woolwich and worked on television transmission. He transferred with other members of the Systems Planning Division to Standard Telecommunication Laboratories, where he has been studying interference on cables near lines carrying heavy 50-hertz currents.

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Photo-Etching for Reliability in Electron-Tube Manufacture

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The reliability of an electron tube can be no greater than the reliability of its components. It follows that improving the quality and reliability of any component is a step toward achieving a reliable tube. The purpose of this article is to draw attention to photo-etching as a tool for improved quality.

The general principles of the photo-etching process as applied to the making of sheet-metal components are simple. The metal sheet from which the component is to be made is coated with a photo-sensitive material that is exposed to light through a mask made by photographic processes to exactly the pattern of the desired design. This exposure stabilizes the coating and makes it impervious to chemical attack; unexposed areas of the coating are then removed to leave bare metal. The bare metal is etched away chemically, the stabilized coating defining precisely the area to be retained.

In various forms the general process has been practised for many years—as photogravure in the printing trade and as "contour milling" or "chemical milling" in the aircraft industry. In electronics it has been used very extensively in recent years in the manufacture of etched-foil printed circuits, and there are references in the literature and trade publications to the making of precision electron-tube parts. The process can be applied to all the metals commonly used in tube construction. Parts of any shape, simple or complex, can be made; parts intricate in profile or intricate in the pattern of holes or lines present no difficulty; indeed, it has been said that if a drawing can be made of a part, then that part could be made by photo-etching. For components within the normal range of metal thicknesses met in electron-tube manufacture, say 0·05 to 0·25 millimetre (0·002 to 0·01 inch), dimensional tolerances of ± 10 microns are readily maintained and can be bettered if necessary.

The purpose of this paper is not to describe in detail the photo-etching process but rather to stress certain features that make it particularly interesting with relation to reliability.

The prime virtue of the process is that it provides a method for making components of good and reproducible quality in prototype quantities very rapidly and very cheaply. The time required for a component of simple design may be only a few hours. The time may be increased to 20 or 30 hours for a complicated design requiring a good deal of detailed drawing work in making the mask. These are perhaps one-

Figure 1—Components for high-figure-of-merit pentode.

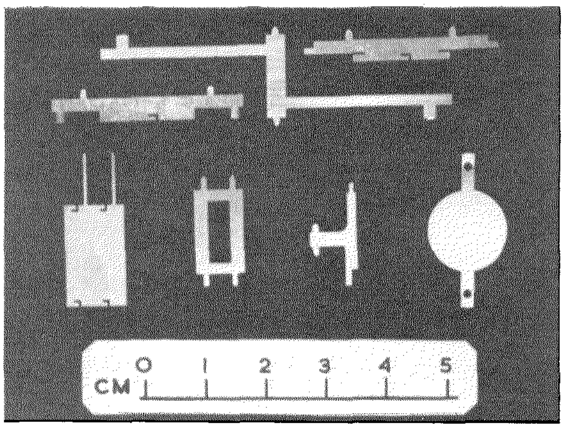
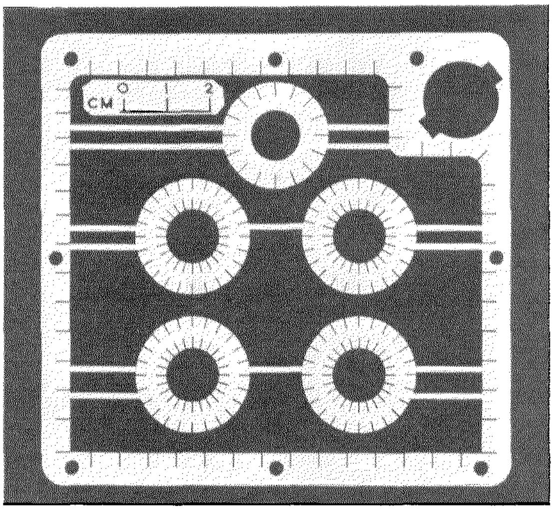


Figure 2—Contacts for waveguide and focus-coil assembly.



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tenth of the times needed for similar parts made by conventional methods, which would usually be press-tooling; indeed, in many cases the saving would be even greater.

Figure 1 shows a set of components for a high-figure-of-merit pentode made from 0·2-millimetre (0·008-inch) nickel. Experimental quantities of each of the components were produced by photo-etching in about 90 hours; this included all the drawing, photographic, and chemical processes. An estimated 1500 hours would be required just to construct the press tools for making the components.

The photo-etching process is equally applicable to making components for ancillary equipment. Figure 2 illustrates a group of contacts for the waveguide and focus-coil assembly of a travelling-wave-tube circuit. The contacts were produced by etching 0·15-millimetre (0·006-inch) beryllium copper; prototype quantities were made in a total of less than 30 hours.

This feature of photo-etching is of greater significance if changes in design are involved. Minor design modifications can be handled with sometimes no more than a few minutes of work; even major modifications amounting to a virtual re-design of a component can be dealt with at no greater cost than a few hours of drawing and photographic work. With press-tooling the changes would involve making new press-tools. This happens very often and the consequent delays impose a severe practical limit on the number of design changes that can be tolerated. The freedom from this limitation provided by photo-etching is a contribution to reliability in that it enables the engineer to test tentative designs of a component in the most-effective way (that is, operationally) and later to verify a design on a pilot run. This is a facility that is beyond the capability of other methods of component manufacture.

Some of the ways in which this facility has been exploited are illustrated below. Figure 3 shows the inter-electrode shield from the tube previously mentioned made to (A) the initial design, (B) an interim design, and (C) the final design. The changes from one design to another were considerable, but photo-etching enabled parts to the three designs to be made and tested at a cost one-tenth of that estimated for the construction of a single press-tool.

Figure 4 shows the final version of a ladder line for the slow-wave structure of a backward-wave-oscillator tube produced by etching 0·08-millimetre (0·003-inch) molybdenum. Many different designs were tested that required making a small number of ladder lines. Changes

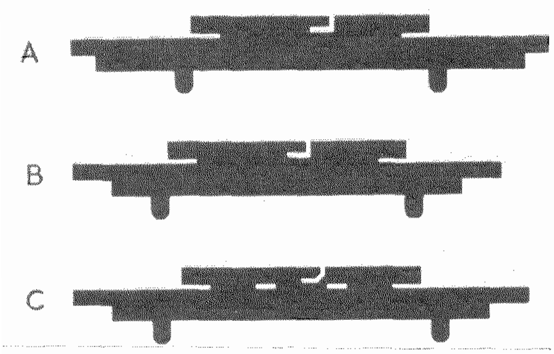

Figure 3—Different designs of inter-electrode shield.

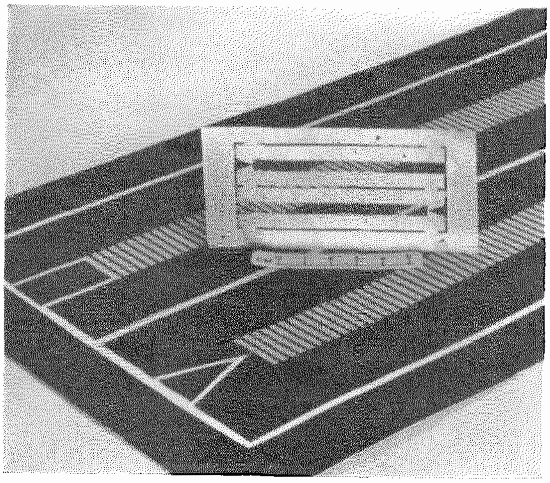

Figure 4—Ladder line for slow-wave structure. The ladder is shown against the background of a master pattern.

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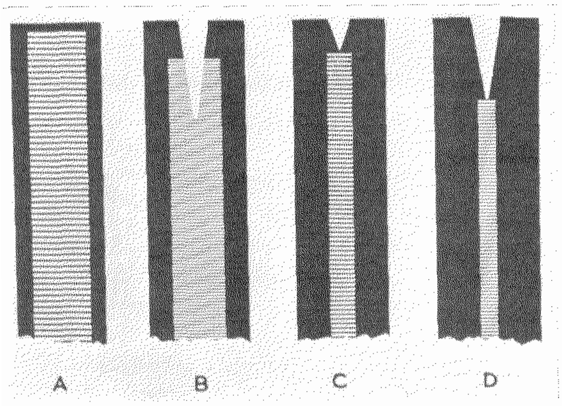

Figure 5—Alternative designs of ladder line.

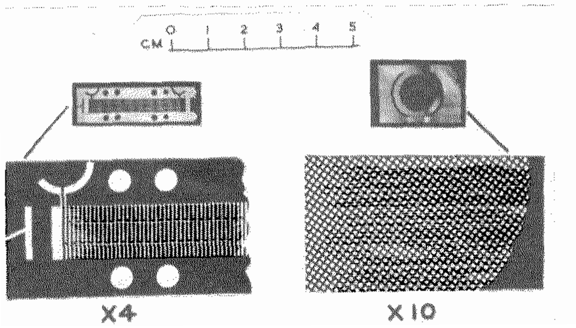

Figure 6—The component at left is a delay line for a backward-wave oscillator and is etched from 0·07-millimetre (0·003-inch) molybdenum. The component at right is a planar grid for an ultra-high-frequency triode and is etched from 0·05-millimetre (0·002-inch) molybdenum. Both components are shown in two sizes for comparison.

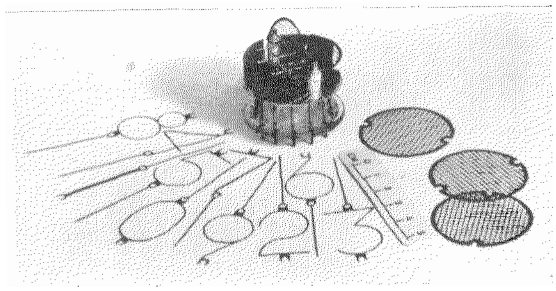

Figure 7—Number electrodes and shields for character display tubes. They are produced by etching 0·1-millimetre (0·004-inch) stainless steel.

were made to the pitch, length, and width of the rungs, and to the shape and dimensions of the terminations. Alternative designs are illustrated in Figure 5.

Certain designs that would be impracticable to realize by press-tooling can be met without difficulty by photo-etching. Such designs usually are those with intricate patterns of fine perforations or projections. In the etching process the width of the perforations and of the metal left between perforations (called "land") can be as little as about 0·05 millimetre (0·002 inch), although these dimensions generally cannot be less than the thickness of the metal. Meshes of up to about 200 per linear inch (8 per linear millimetre) can be made; so too can projections 50 or 80 microns in width and of any length. The latter point is significant in considering reliability.

Projections such as cathode tails, connecting legs for numerals, and support tags in general are difficult for press-tools to make integrally with the component. They are usually made as separate parts and spot-welded to the main bodies. These welds, however carefully made and even if reinforced, cannot have the same reliability as the projections made as an integral part of the component.

Figures 6 and 7 illustrate the diversity of design realized by photo-etching.

There are other ways in which photo-etching contributes to reliability. Once the detailed design of a component has been established experimentally, the required quantity—either a single one or hundreds of thousands—can be produced from the same master mask. All the parts will be uniform; whatever the production quantities, there will be no degradation of quality due to tool wear or to differences in the temper of the metal stock.

Burrs on components are a great hazard to reliability. They distort dimensions; they make assembly difficult and prevent the mating of contiguous parts; they fray micas and abrade glass and coatings. If they become detached

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they form conducting particles within the tube. Parts made by press-tools sooner or later tend to suffer from burrs. There are numerous chemical and mechanical methods of deburring sheet-metal parts but success is gained only by sacrificing some aspect of quality. At best, complete uniformity of dimension or of edge profile cannot be achieved.

There are no burrs on photo-etched parts. Correctly done, the process leaves clean smooth edges completely free from burrs and raggedness. The parts are free from variable stress patterns and are dimensionally stable, there is no work-hardening, and the need for stress-relief annealing is avoided.

Complete cleanliness of the components within an electron tube is essential to reliability. Initially, and throughout life, the reliability of a tube is greatly affected by the presence even of trace quantities of contaminants. For this reason, materials of high purity are chosen for components and these are inspected to rigorous standards before use.

Nevertheless, if press-tooled parts are used contamination cannot be avoided. The sheet metal, even though intrinsically of high purity, carries a high concentration of contaminants on its surface, and these are added to by the mechanical working of the material during the pressing operation. The contamination is of indefinite composition, being partly metallic and partly derived from lubricant residues. Experience has shown that the bulk of this contamination is contained in a very-thin outer layer. In a cathode sleeve, for example, the sulphur contamination has been halved by removing a skin of metal 2 microns thick. These contaminants are not just superficial. They are intimately worked into the surface and are not removed by the de-greasing and washing processes normally used in tube manufacture (more-drastic cleaning treatments are usually impracticable for precision parts).

In photo-etching, the first process, before the metal sheet is coated with photographic resist, is to etch away the top skin of metal; this is done primarily to help the coating adhere but it is also very effective in removing the surface contamination. Thereafter, the coating of resist protects the metal from contamination until it is removed immediately before the components are used.

In describing some of the ways that photo-etching can help to improve reliability, it has been difficult to avoid seeming to over-stress the demerits of press-tooling. The merits of press-tooling are well known, but it happens that for electron tubes with the special emphasis on reliability, photo-etching has certain advantages. These may be summarized as follows.

(A) The photo-etching process makes possible the use of novel design features of significance to reliability.

(B) By providing prototype parts very quickly, and especially by the ease with which changes in design can be made, photo-etching enables a reliable component to be developed within a practicable period.

(C) Photo-etching provides components of superior quality in matters of importance to reliability, notably uniformity, freedom from burrs, and cleanliness.

Bernard Thomson was born in Preston, Lancashire, in December 1914, and was educated at the school of St. Ignatius Loyola and the Harris Technical Institute. He served an apprenticeship in pharmaceutical chemistry, then worked for two years in the cathode-ray-tube laboratory of Electrical Musical Industries.

Mr. Thomson joined Standard Telephones and Cables at North Woolwich in 1938. Most of his time has since been spent in the process-chemistry laboratory, of which he has been in charge for the past 10 years.

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Selenium Rectifiers in 1965

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1. Introduction

Remarkable progress has been made in recent years in our knowledge of the physics of selenium and of the selenium rectifier. Even today there is still no complete theory of electrical conduction in selenium, but the many individual findings from experiments carried out by scientific institutions and industrial laboratories enable us to see certain features that will be included in future theory. The scientific research on selenium has also been put to practical use—to improve the electrical components made from this semiconductor, particularly the selenium rectifier.

This paper reports on work being done on selenium and selenium rectifiers, from both the scientific and engineering aspects. Particular attention is paid to the extent to which selenium differs in its electrical properties from the generally better known semiconductors germanium and silicon, permitting a better understanding of the special properties of the selenium rectifier compared with germanium and silicon rectifiers. It also shows how the operating characteristics of the selenium rectifier, particularly its load capacity and life, have been improved as a result of the increased knowledge of its physical principles.

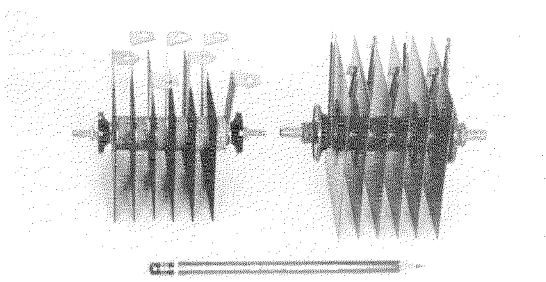

Figure 1—Selenium (right) and Silring silicon (left) rectifiers for 3-phase-current bridge circuit. The selenium rectifier delivers 27 amperes at 25 volts input, while the Silring rectifier delivers 30 amperes at 380 volts.

2. Monocrystalline and Polycrystalline Semiconductor Rectifiers

Semiconductor rectifiers are divided into monocrystalline and polycrystalline groups according to the structure of the semiconductor layers. The most-important example of the first group is the silicon rectifier, and by far the most-important of the second group is the selenium rectifier. The rectifier effect in both cases occurs because two semiconductor layers of different types of conduction are contiguous. In the silicon rectifier the transition from *n*- to *p*-type conduction takes place in a uniform monocrystalline basic material, for which reason we refer to a *p-n* junction. In the selenium rectifier, on the other hand, two different polycrystalline semiconductor materials are adjacent to each other, *p*-conducting selenium and *n*-conducting cadmium selenide. It is therefore customary to call this a *p-n* heterojunction. There are many common features in the electrical properties and in the construction of both types of rectifier, but there are also certain differences, particularly with respect to current densities and inverse voltages.

Silicon rectifier layers tolerate without difficulty current densities up to 200 amperes per square centimeter (1290 amperes per square inch). It is therefore possible to make the rectifier cells very small. On the other hand, the power loss is dissipated through a very-small surface and a special cooling member must be provided. In the selenium rectifier current densities of not more than 0.3 ampere per square centimeter (1.9 amperes per square inch) are allowed. Therefore relatively large-surface elements are needed but no special cooling is necessary.

Figure 1 shows the great resemblance in the external structures of selenium and silicon rectifiers. In this figure a Silring silicon rectifier and a selenium rectifier, both dimensioned for about the same current rating, are compared. While the semiconductor layers of the silicon rectifier are contained in the spacer rings and the plates merely act as cooling members, the semiconductor layers of the selenium rectifier

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are uniformly distributed over one surface of each plate and the spacer rings are of solid metal. Since both rectifiers must dissipate about the same amount of heat, approximately the same volume is obtained.

The inverse voltages differ as much as the current densities do. While silicon rectifier cells are made for 1000 volts and higher, the peak value of permissible inverse voltages for selenium rectifier plates at the present time is at most 60 volts. This automatically restricts the fields of use, particularly in power engineering [1].

The large effective surfaces of selenium rectifiers render them extraordinarily insensitive to extreme current loading (short-circuiting) for brief periods. This is due partly to the high thermal capacity of the semiconductor layer, which has the effect that the energy released in the case of very-brief overloading does not result in a harmful increase of temperature. Also, the heat resistance between the semiconductor layer and cooling agent is very low, so that for a longer overload the heat due to energy losses can easily dissipate. This is particularly important if heavy currents are used, for example in electroplating and welding, in which short-circuits or current peaks frequently occur during operation.

The mode of operation of the monocrystalline semiconductor rectifiers and their physical laws have been explained in all essential features. Even the electrical engineer who is not directly concerned in their development and production knows the most-important mechanisms. In contrast, the scientific principles of the selenium rectifier have hitherto been surrounded by something akin to secrecy and savor somewhat of empiricism. There are, of course, good reasons for this.

For one thing, in the 1920's when the then Süddeutsche Apparatefabrik, today one of the Standard Elektrik Lorenz factories, put the first selenium rectifiers on the market, the physics of semiconductors was still in its early stages. For another thing, the physical properties of selenium are far-more complicated and even today have not been so thoroughly investigated and understood as those of silicon or germanium. Therefore development and production had to be carried out mainly on an empirical basis until the 1950's.

3. Electrical Conductivity of Selenium

Electrical conductivity of selenium is not a basic property, but the product of two variable parameters that are to a great extent independent of each other. These are the concentration of the charge carriers (designated by n and p) and their mobility (μ_n and μ_p, respectively). It is assumed that one type of charge carrier largely predominates. If this were not so, the relation becomes more complicated, but this condition does not occur with selenium. With the electron charge equal to e, the complete equation for the conductivity of an n- or p-conductor is

$$\sigma = e\mu_n n \quad ; \quad \sigma = e\mu_p p. \qquad (1)$$

If any change occurs in the test conditions (for example, the temperature) it must be established to what extent this affects the mobility and/or the concentration of the charge carrier. An effect is used for this purpose that depends on only one of the two parameters (for example, the Hall effect or the Seebeck effect, both dependent only on the carrier concentration).

Measurements had already been made on trigonal-crystalline selenium at the beginning of this century and these showed, remarkably enough, that the conductivity may increase, decrease, or even remain constant with temperature according to the doping and pretreatment of the sample. Occasionally a maximum or minimum occurs; Ries [2] assumed in 1908 that two different modifications of trigonal selenium exist, one metallic with positive temperature coefficient and one semiconductive with negative temperature coefficient.

However, the works of Lehovec [3], Borelius and Gullberg [4], Plessner [5], Henkels [6], and Nijland [7] have produced proof that as

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the temperature rises the carrier concentration decreases and mobility increases (Figure 2). Depending on the course of both relationships, a negative or positive temperature coefficient is obtained for the product of the two magnitudes, that is, the conductivity.

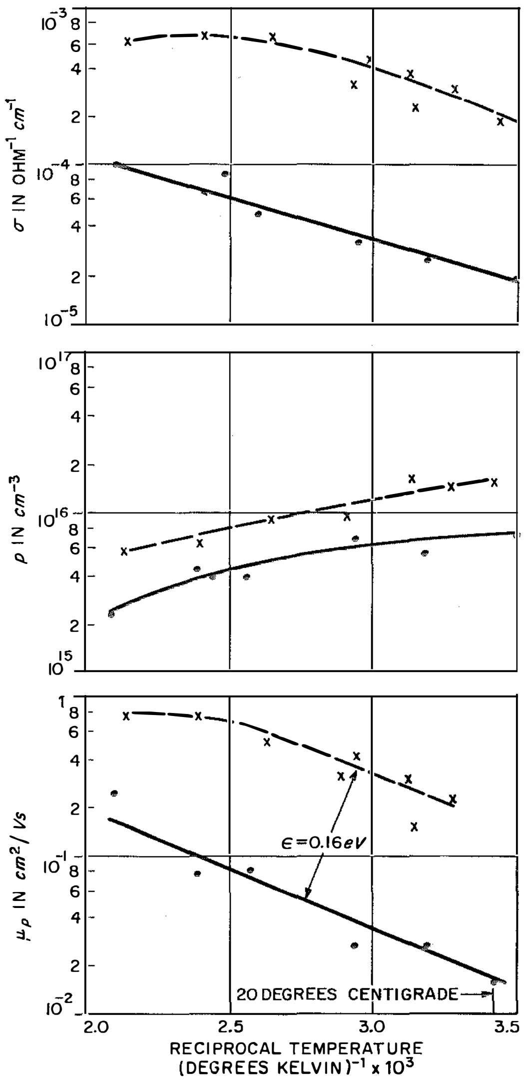

Figure 2—Conductivity σ, charge-carrier concentration p, and carrier mobility μ_p in polycrystalline selenium layers, as a function of temperature (according to Plessner). The solid curves are for iodine doping 3×10^{-6} molar (practically pure); the broken curves are for iodine doping 3×10^{-4} molar (rectifier selenium).

This is a very-remarkable finding. For a "normal" semiconductor the carrier concentration increases with temperature and the mobility decreases because the frequency of impact increases with the oscillating lattice atoms and with lattice defects.

Only for pure scattering by ionized impurity centers should the mobility increase with temperature (not exponentially, but only in proportion to $T^{3/2}$). As demonstrated in Section 4, in selenium the carrier mobility, however, does not decrease but increases with the impurity-center concentration (Figure 3). This contradicts the assumption that ionized impurity centers predominate.

4. Determination of Carrier Mobility in Selenium

The numerical values ascertained for carrier mobility in selenium do not behave according to the accustomed model of a semiconductor. The mobility of holes in pure selenium amounts to less than 0.1 cm^2/Vs. Up to now accurate values for mobility do not exist, because it is difficult with such low mobility to determine the charge-carrier concentration from the Hall effect with sufficient reliability. The measured Hall voltage

$$U_{\mathrm{Hall}} = R_H \cdot \frac{I \cdot B}{d} = \frac{3\pi}{8ep} \cdot \frac{I \cdot B}{d} \qquad (2)$$

in the case of a given Hall constant R_H depends practically only on the ratio I/d (I = current through the sample, d = thickness thereof), since the induction B of the magnet is determined by the saturation of the iron. The current I in turn is limited by the power

$$\frac{N}{F} = \frac{I^2 \cdot R}{2 \cdot l \cdot b} = \frac{I^2}{2 \cdot l \cdot b} \cdot \frac{l}{\sigma \cdot b \cdot d} \qquad (3)$$

that can be drawn off from the surface $F = 2 \cdot l \cdot b$ of the sample in the given cooling conditions. If we insert

$$I = \left(2 \cdot b^2 \cdot d \cdot \sigma \cdot \frac{N}{F}\right)^{1/2} \qquad (4)$$

in (2) and then calculate the Hall power

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$P_{\text{Hall}} = U^2_{\text{Hall}} \cdot G_{\text{Hall}}$, we will obtain with $G_{\text{Hall}} = \frac{d \cdot l'}{b} \cdot e\mu_p p$

$$P_{\text{Hall}} = \left(\frac{3\pi}{8}\right)^2 \cdot \mu_p{}^2 \cdot 2 \cdot b \cdot l' \cdot \frac{N}{F} \cdot B^2 \qquad (5)$$

(l' being the width of the Hall contacts).

Thus it is the mobility that—in addition to the size of the sample—determines the power available for measurement.

The evaluation of the Seebeck effect presumes a knowledge of the effective mass and of the dispersion mechanism, neither of which is known. It also gives only relatively inaccurate values, because the measured thermoelectric force enters the calculation as an exponent.

A third effect, which in principle makes it possible to determine carrier concentration independently of mobility, is the field effect. The semiconductor layer to be investigated is used as the plate of a capacitor and the conductivity is varied measurably by introducing known charge quantities; this assumes that the original carrier concentration is sufficiently small and the introduced carriers remain freely mobile. Apparently this is the case with selenium.

Tausend [8] succeeded in determining the mobility in selenium of the highest degree of purity at 0.02 cm^2/Vs at room temperature. For comparison, in germanium $\mu_n = 3900\ cm^2/Vs$ and $\mu_p = 1700\ cm^2/Vs$.

The fact that such low mobilities are incompatible with conventional ideas of charge transport in metals and semiconductors can be recognized, for example, by working out the mean free-drift-path length. Assuming that dispersion prevails on lattice oscillations, it amounts to

$$\lambda_p = \frac{3(2\pi m^* kT)^{1/2}}{4e} \cdot \mu_p. \qquad (6)$$

At room temperature, therefore

$$\lambda_p = 10^{-8}\left(\frac{m^*}{m}\right)^{1/2} \cdot \mu_p \text{ (centimeters)}.$$

If we insert $m^*/m = 10$ and $\mu_p = 0.02\ cm^2/Vs$, we obtain

$$\lambda_p \approx 6 \cdot 10^{-10} \text{ centimeters}.$$

Compared with this the atomic interval in the selenium lattice is about $5 \cdot 10^{-8}$ centimeters.

Both the very-small value of mobility and its anomalous temperature-dependence show that in electron conduction in selenium there can be no question of the movement of free

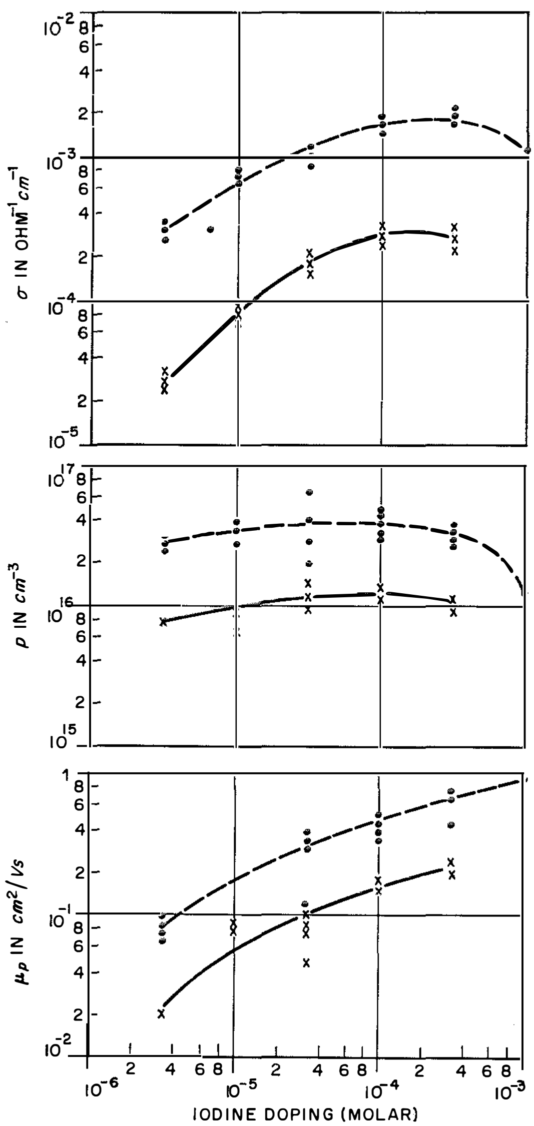

Figure 3—Conductivity, charge-carrier concentration, and carrier mobility, as a function of the iodine doping (according to Plessner). The solid curves are for coarse crystalline (crystallized at 175 degrees centigrade); the broken curves are for fine crystalline (crystallized at 110 degrees centigrade).

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holes in an incompletely occupied valence band in the usual meaning of the term. This anomaly of selenium is in no way restricted to polycrystalline layers, but is also to be found in single crystals. Figure 4 shows the temperature-dependences of σ, p, and μ_p measured by

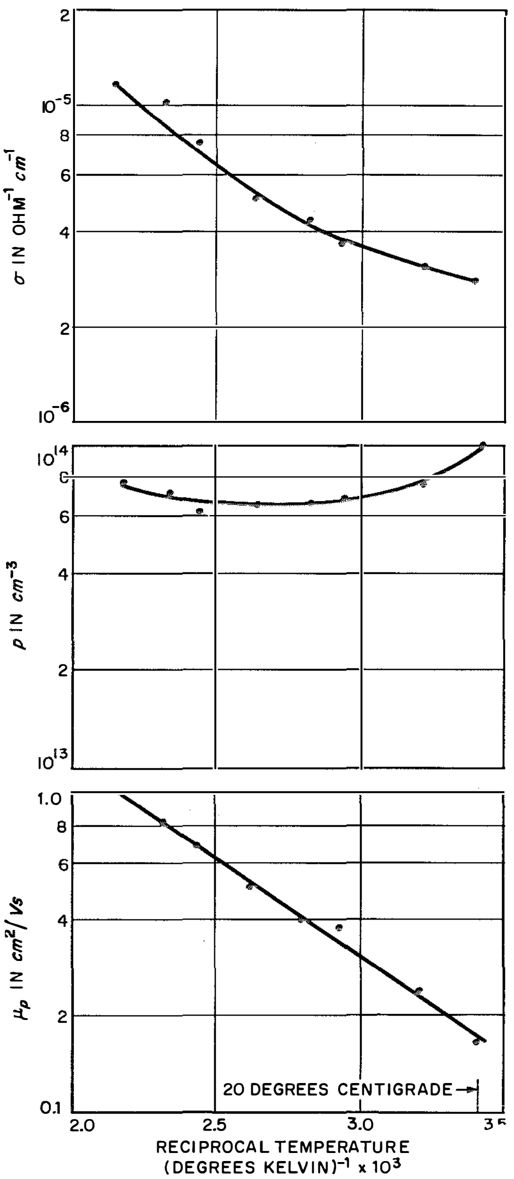

Figure 4—Conductivity, charge-carrier concentration, and carrier mobility in a selenium single crystal (according to Plessner).

Plessner [9] on a single crystal. Here μ_p also varies exponentially with $\epsilon = 0.13\ eV$.

At room temperature $\mu_p = 0.14\ cm^2/Vs$.

Spear [10] investigated the drift mobility of the charge carriers in amorphous selenium and found for this parameter also an exponential rise with temperature. At room temperature μ_p is equal to 0.1 to 0.2 cm^2/Vs, corresponding approximately to the values measured on trigonal monocrystals. Similar investigations by Hartke [11] led to the same result.

The question arises of what a transport of charge carriers would look like with mobilities lower than 1 cm^2/Vs and showing an exponential rise with temperature. First of all, one naturally thinks of a kind of diffusion process, since diffusion processes always obey the law

$$D = D_0 \cdot \exp\,(-\,\epsilon/kT) \qquad (7)$$

where D is the diffusion constant. D is linked to the mobility μ by the Nernst-Townsend-Einstein relation

$$D = \frac{kT}{e} \cdot \mu. \qquad (8)$$

Based on thermodynamic considerations, Heikes and Johnston [12] found the following expression for D_0.

$$D_0 = L^2 \gamma \nu_0 \qquad (9)$$

where γ is a geometric factor of the order of magnitude 1, L the diffusion length, and ν_0 the frequency of the lattice vibrations (order of magnitude $10^{13} s^{-1}$). They started from the supposition that each charge carrier polarizes the surrounding lattice atoms, thus creating a potential trough in which the carrier traps itself. As a consequence of thermal excitation the charge carriers diffuse through the crystal lattice by jumping from one trough to the other. With the aid of this model Heikes and Johnston interpret conductivity measurements made on transition metal oxides.

Applying Heikes' and Johnston's equation

$$\mu = \frac{e\gamma L^2 \nu_0}{kT}\, e^{-(\epsilon/kT)} \qquad (10)$$

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tentatively to measurements on selenium, the diffusion length or mean jump distance L in pure selenium is found to be

$$L = 9 \cdot 10^{-8} \text{ (centimeters)}$$

for $\epsilon = 0.13\ eV$, $\mu = 0.02\ cm^2/Vs$, $\gamma = 1$, and $\nu_0 = 10^{13}\ s^{-1}$.

This is about two times the value of the lattice constant and is consequently a rather-plausible value. Applying the same equation to the measurements made by Spear on amorphous selenium, we obtain

$$L_p = 4.4 \cdot 10^{-7} \text{ (centimeters)}$$

for $\mu_p = 0.15\ cm^2/Vs$ and $\epsilon_p = 0.16\ eV$;

$$L_n = 4.5 \cdot 10^{-7} \text{ (centimeters)}$$

for $\mu_n = 5 \cdot 10^{-3}\ cm^2/Vs$ and $\epsilon_n = 0.25\ eV$. (In all cases $T = 300$ degrees Kelvin.)

Surprisingly, the same jump distance, or—more generally expressed—the same value for D_0 is obtained for both types of charge carriers.

Unfortunately, the application of (10) to selenium involves a number of problems, since self-trapping of the charge carriers should actually be possible only in ionic crystals, and the various forms of selenium certainly cannot be numbered among these. In addition, unreasonably large jump distances will result from (10) if activation energies higher than approximately 0.5 eV are involved. Lately, however, activation energies even as high as 1.8 eV have been measured on thallium-doped solid selenium and on liquid selenium [8]. Thus it seems that another way must be found to disclose the meaning of D_0 in (7) with selenium.

5. Influence of Doping on Conductivity

The influence of doping is also unusual. As in other semiconductors, it is possible in selenium to vary the conductivity by adding small quantities of impurities. However, selenium always remains p-conductive. No one has succeeded in generating n-type conductivity in selenium. Mainly those additions that increase conductivity are interesting from a technical standpoint. These are solely the halogens: chlorine, bromine, and iodine (probably also fluorine, but no investigation of this is known). All the other elements, particularly the metals, reduce the conductivity to a greater or lesser extent.

In some circumstances, however, it is possible to obtain greatly increased conductivity by combined halogen-metal doping.

Plessner [5] investigated selenium layers with iodine doping and found that the carrier concentration was changed only negligibly by increased doping. The increased conductivity observed is based more on a considerable increase in the mobility of the holes. In very-heavy doping, the concentration of the charge carriers decreases appreciably (Figure 3). Plessner also indicates the temperature-dependence of the conductivity, carrier concentration, and mobility of samples having $3 \cdot 10^{-4}$ molar iodine concentration (Figure 2). The mobility varies exponentially with $\epsilon = 0.16\ eV$ up to a temperature that is not too high.

In addition to doping, thermal treatment of the selenium layers also greatly influences their conductivity. Our own measurements on layers and also on rectifier plates with combined halogen-metal doping show that both the carrier concentration and the mobility decrease during the thermal treatment after some 10 minutes, when all the selenium has passed into the crystalline form. Carrier concentrations in rectifier plates were determined from capacitance tests by a method indicated by Schottky [13] and expanded by Dolega [14].

A comparison of Figure 5 with Figure 2 shows that the much-higher conductivity of the layers with combined doping is caused by mobility that is 5 to 10 times greater. The carrier concentrations in both cases are about the same. (The molecular halogen concentration in the samples investigated by us amounted to about $2 \cdot 10^{-4}$.)

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6. Conduction Processes in Selenium

If we summarize these experimental and theoretical results, the following picture emerges of the conduction mechanism of selenium.

Both in amorphous and in trigonal crystalline selenium, charge carriers move in a kind of diffusion under the influence of an applied electric field. This movement is characterized by a constant D_0 that is related to the diffusion length L and by an activation energy ϵ. This energy is 0.13 to 0.16 eV for holes both in amorphous and in mono- or polycrystalline selenium, and 0.25 eV for electrons in amorphous selenium. Under certain conditions, however, values of up to 1.8 eV may be obtained.

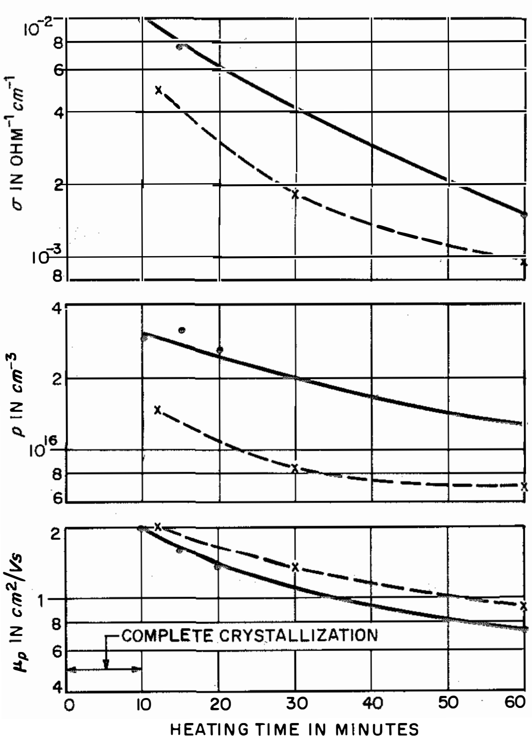

Figure 5—Conductivity, charge-carrier concentration, and carrier mobility in selenium layers with combined halogen-metal doping, as a function of length of heat treatment at 217 degrees centigrade. The solid curves are for layers measured parallel to the surface (carrier concentration from the Hall effect); the broken curves are for layers of rectifier plates, measured perpendicular to the surface (carrier concentration from capacitance tests).

The diffusion length and therefore D_0 can be considerably increased by halogen or combined halogen-metal doping. The activation energy remains practically unchanged.

The hole concentration in pure selenium is much higher than in the single crystal. The carrier concentration is irreversibly reduced by heat treatment of layers slightly below the melting temperature of the selenium. It is not appreciably affected by halogen or combined doping. It is also practically independent of temperature. (At high temperature, a temperature response can be simulated by irreversible changes!) On the other hand, conductivity can be reduced by pure metal doping, whereupon the size of the crystal grain increases in turn. The relation found by Gobrecht, Tausend, and Plümecke [15] between conductivity and total surface of the crystal grains therefore is probably due chiefly to the carrier concentration.

In amorphous selenium the carrier concentration is much less than in the monocrystal ($\approx 10^8\ cm^{-3}$ at room temperature) but mobility is about the same.

It may be possible within the forseeable future to combine all this new knowledge, at present only fragmentary, into one complete theory of electrical conduction. In particular it is probable that there is a close relationship between the diffusion-like movement of the charge carriers and the structure of the amorphous and of the trigonal crystalline selenium containing spiral chains of atoms.

However, this new knowledge has already produced useful results, since it has contributed to a physical consolidation of the hitherto predominantly empirical work on the future development of the selenium rectifier and other technical products, such as selenium photocells and xerographic plates. This particularly applies to the methods of manufacturing selenium layers of very-high or, even, very-low conductivity by doping and heat treatment.

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7. Rectifier Effect

In the past few years some gratifying advances have also been made with respect to the theory of the rectifier effect in selenium rectifiers.

The first practicable theory of the semiconductor rectifier (at that time chiefly copper-oxide and selenium rectifiers) was that developed by Schottky [13] in 1938–1942 on the metal-to-semiconductor contact. However, it could only explain the current-versus-voltage characteristic of the selenium rectifier approximately and at voltages that were not excessive. Later, Poganski [16] proved that the rectification process did not occur between the selenium and the metal of the "upper" electrode, as assumed in the theory of the metal-to-semiconductor contact, but that there is a selenide layer between, so that the rectification takes place between the semiconductors selenium and cadmium selenide.

Attempts were then made to apply the theories of rectification in *p-n* junctions of monocrystalline semiconductors (which arose in the meantime) to the selenium rectifier. Here again, however, considerable differences were found. This was not surprising, since two important prerequisites of these theories: (A) that all the processes occur in a uniform crystal lattice and (B) that the recombination of the charge carriers be distributed over a relatively large volume, are not satisfied in the selenium rectifier.

In 1962 Dolega [17] produced a theory of the *p-n* heterojunction between semiconductors with different crystal lattices. By means of this theory it is possible to describe quantitatively the whole course of the current-versus-voltage characteristic of the selenium rectifier. At the same time it contains the Schottky theory of the metal-to-semiconductor contact as a special case.

Dolega's theory, above all, can predict the effect of certain changes in the physical properties of either of the two semiconductors (for example, the charge-carrier concentration or the mobility) on the rectifier characteristic. This is obviously very important for the further technical development of the selenium rectifier.

8. Advances in Selenium Rectifiers

It is possible to summarize the progress of selenium-rectifier technology, which has been greatly affected by increasing knowledge of the physical principles, by charting the power rating of a typical rectifier of certain effective size year by year (Figure 6). Apart from an obvious stagnation during the war years 1939–1945, the figure shows a uniform rise of about two powers of ten.

The most-striking progress is that made in reducing the forward loss. At the outset an undoped selenium was used, which was only relatively roughly purified and which had a specific resistance of about 5000 ohm-centimeters. Today specific resistances are obtained down to 50 ohm-centimeters (at room temperature) by means of combined halogen-metal doping of a practically spectrally pure raw material. Figure 7 compares the characteristics of a rectifier made in 1953 with one made in 1963.

On the other hand, as is recognized today, the possibilities of improving the reverse-blocking properties are fundamentally limited.

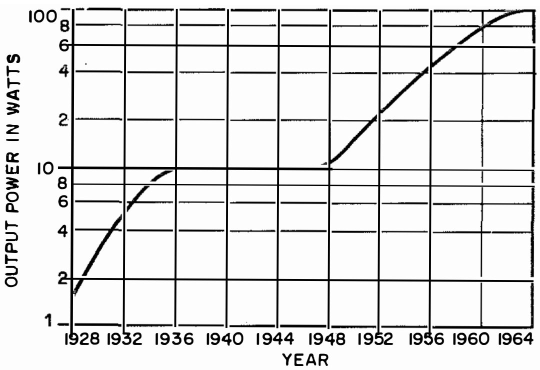

Figure 6—Increase in power output (according to catalog) of a selenium rectifier unit of 4 plates, each having 20 square centimeters (3 square inches) of effective surface, since manufacture of selenium rectifiers started.

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In both silicon and selenium rectifiers it is a question of creating, in the vicinity of the barrier layer, a region of low conductivity into which a space charge can expand when the inverse voltage is applied. This region of poor conductivity, on the other hand, naturally increases the forward resistance of the rectifier and therefore the losses. In the selenium rectifier there can be no conductivity modulation as exists in silicon and germanium rectifiers, and for this reason the high-resistivity region may not be made too thick.

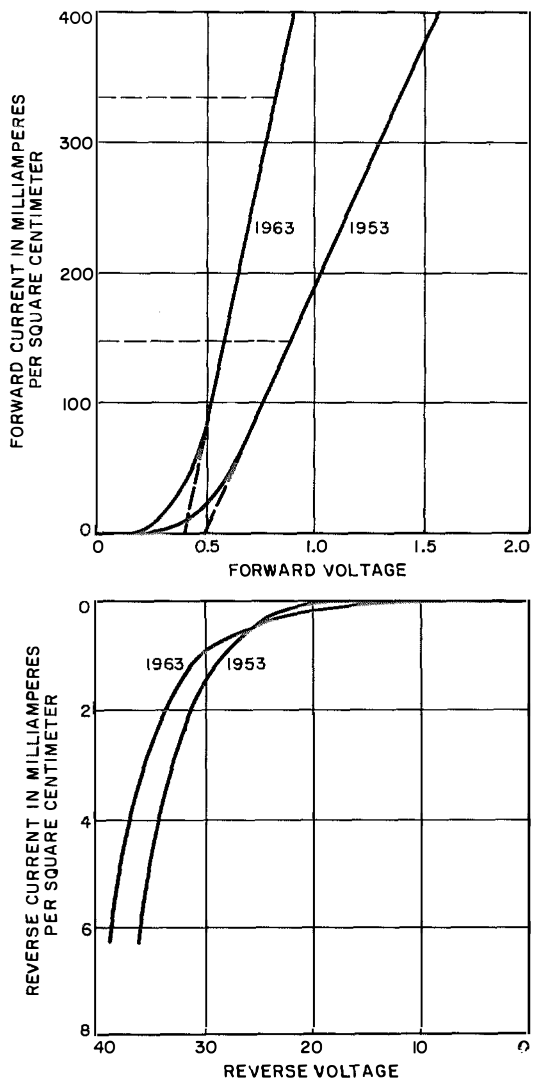

Figure 7—Forward and reverse characteristics of selenium rectifiers in 1953 and 1963. The broken lines indicate the peak values of the rated current densities.

Also, as pointed out by Schottky [13] and Dolega [17], there is no saturation current with the *p-n* heterojunction, but the inverse current increases exponentially from low voltages onward. There is therefore an upper limit to the peak reverse voltage obtainable with a selenium rectifier element. This is situated at about 100 volts as proved also by investigations into the breakdown strength [18].

According to the purpose for which the rectifier is to be used, the aim is to obtain either optimum forward quality with relatively low reverse-blocking properties, or high inverse voltage. In power rectifiers the forward loss should be kept as low as possible to obtain high efficiency. In many cases inverse voltages of 25 to 30 volts are sufficient for this. Figure 8 compares the efficiencies of power rectifiers made in 1953–1963 having 25-volt inverse voltage. The efficiency in the newest type is higher than in the 10-year-old type despite current densities that are more than twice as high as in the old.

On the other hand, for high-voltage rectifiers it is not so much a question of high current density as of high inverse voltage with the minimum possible loss, so that the rectifier unit may be kept small. In any case, however, a sufficiently large safety margin must be maintained between the peak value of the reverse voltage and the limiting breakdown voltage. For this reason, an inverse voltage of 45 volts (peak value 63 volts) for the individual cells generally may not be exceeded.

In judging a selenium rectifier, besides considering electrical values at the time of manufacture, its behavior in continuous operation is particularly important. This is another area in which gratifying progress has been made in recent years.

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9. Behavior in Continuous Operation

The phenomenon of aging of the selenium rectifier has long been well known. It is manifested by a progressive increase in the forward loss, with the result that the direct voltage delivered by the rectifier unit progressively decreases.

Therefore, in the earlier types of rectifiers, the transformer in unregulated units had to be provided with additional windings; from time to time the applied alternating voltage could then be increased to compensate for the aging of the rectifier. In addition, the increased loss raised the operating temperature of the rectifier, which in turn intensified the aging process. This could make the rectifier unit entirely useless after a time.

Although, with normal loading, aging only becomes noticeable after some years in the majority of rectifiers manufactured today, manufacturers have tried to reduce this phenomenon and eliminate it if possible. In the past few years we have carried out extensive research [19–21] to learn about the aging mechanism in earlier types of rectifiers. It was found that the diffusion of doping substances was chiefly responsible for this aging.

As already stated, a thin high-resistance layer is made immediately at the *p-n* heterojunction to improve the reverse properties of the rectifier. Some of the doping substances hitherto employed have the undesirable property of migrating after a time along the grain boundaries of the polycrystalline selenium layer and gradually producing a high resistance in the deeper layers. Once this was recognized, it was possible to set about eliminating it. In 1956 a great step forward was taken that resulted in a rectifier in which the aging is practically negligible at normal operating temperature of 70 degrees centigrade. It has since been found possible to eliminate aging almost completely.

Since aging is a diffusion process, the aging behavior of a rectifier depends strongly on the temperature at which it is operated and stored. It is unimportant in this respect whether a certain operating temperature is brought about by electrical loading or by high ambient temperature.

Since the cooling conditions in a particular rectifier are established in advance in the majority of cases, the aging and power rating of a rectifier are mutually dependent. For example,

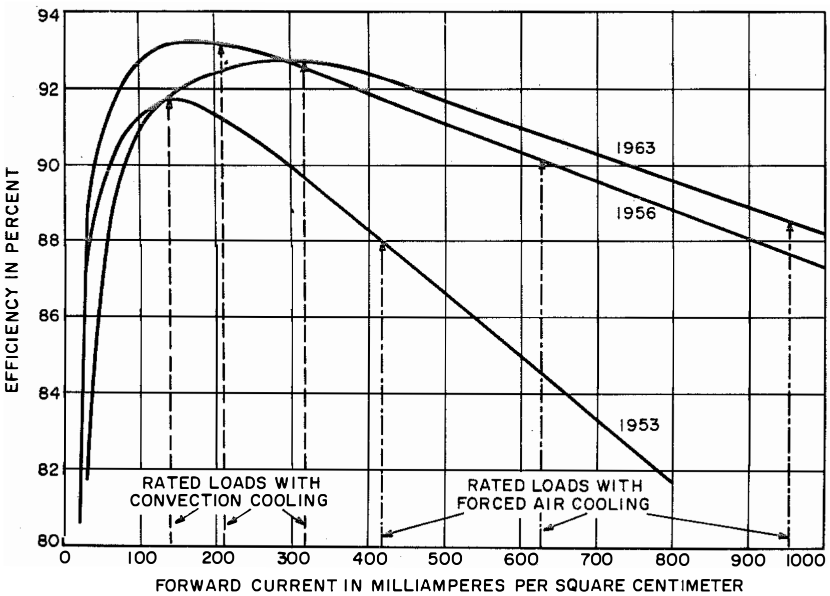

Figure 8—Efficiency of various types of selenium rectifiers as a function of current load for reverse ratings of 25 volts.

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aging can be considerably reduced by operating the rectifier far below its permissible load. On the other hand, the load can be raised above the highest permissible value if only a relatively short life is acceptable. The extent that this can be done depends in the long run on the temperature dependence of the reverse current, which differs for different manufacturers. In any case the input voltage must be reduced to such an extent at high operating temperatures that the reverse losses remain approximately constant.

The increase in aging with increased operating temperature is used in making accelerated life tests on selenium rectifiers [21]. Figure 9 shows the results of two such investigations on our selenium rectifiers made in 1956 and 1963, both carried out at operating temperature of 110 degrees centigrade. This temperature was obtained by operating the rectifiers with the same specific load at an ambient temperature of 75 degrees centigrade. It can be seen that the 1956 type had aged to such an extent after only 10 000 hours that it no longer delivered the direct voltage specified in the catalog. The voltage drop across the rectifier had just doubled in that period. On the other hand, the 1963 rectifier shows no change whatever in its forward properties under the same operating conditions. (Of course the reverse currents were also measured; they remained constant during operation.) Thus it was found possible for the first time to produce a rectifier that does not age even under extreme loading conditions. This permitted raising the operating temperature to 85 degrees centigrade, so that the rectifier can be given a much-higher load under constant cooling conditions.

As the life test has proved, the rectifier can work at even-higher operating temperatures without any appreciable aging. Of course, the input voltage must be reduced. The absolute limit is 120 degrees centigrade. At this temperature only half the rated input voltage is permissible.

10. Practical Examples

Progress in the technology of the selenium rectifier has resulted in the dimensions of the conventional power-rectifier stacks being considerably reduced, while at the same time life is lengthened.

In addition, it became possible to develop very-small efficient units that have many uses, not only as low-power rectifiers for the alternating-current supply, but also as diodes, voltage stabilizers, suppressors, and generally as non-linear resistors [22]. To develop these miniature and subminiature rectifiers, rectifier plates were necessary that in particular satisfied two conditions.

(A) Low specific losses, that is, the best for-

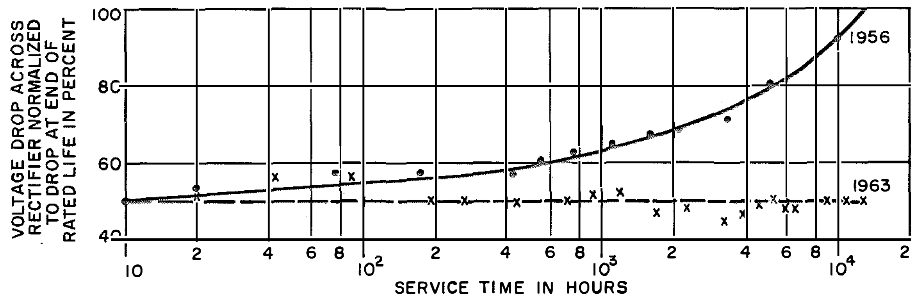

Figure 9—Continuous operation behavior of our selenium rectifiers manufactured in 1956 and 1963. The forward voltage drop across the rectifier is plotted in relation to its maximum value at the end of life. The measured points are averages in each case of about 30 cells. Ambient temperature = 75 degrees centigrade; operating temperature = 110 degrees centigrade; and load = 25 volts at 0.17 ampere per square centimeter.

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ward characteristics, so that excessive heat is not developed despite the small dimensions.

(B) High breakdown strength, because the rectifier plates are stacked on each other without interstices and hence there can be no burnout through voltage puncture.

Figure 10 shows a number of selenium rectifiers. The compact bridge rectifiers are chiefly used for the alternating-current supply of transistor television receivers and audio amplifiers. They are much smaller than conventional selenium-rectifier stacks of the same power, but nevertheless have about the same effective surface. This rectifier circuit has very-low internal resistance and the direct-current voltage regulation is very good. The latter characteristic has great importance for apparatus with power transistors.

The tubular rectifiers shown are used in television receivers to rectify the high voltage for the picture tubes. They are superior to the electron tubes hitherto used for this purpose, being smaller, having longer life, and needing no specially derived filament voltage. Finally, miniature rectifiers of the type at the lower right have many different uses in electronic equipment with printed circuits.

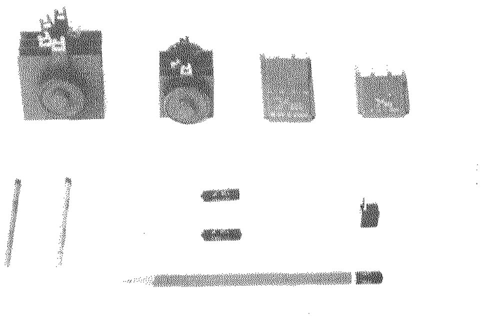

Figure 10—Conventional selenium rectifiers and modern miniature rectifiers. Top row from left to right: conventional 25-volt 2.5-ampere selenium rectifier (1956); 25-volt 2.2-ampere rectifier (1963); two compact 30-volt rectifiers for 2.5 and 1.2 amperes, respectively. Bottom row: Two high-voltage tubular rectifiers, two 500-volt 7.5-milliampere tubular rectifiers; plus one rectifier in a plastic housing filled with synthetic resin, for printed circuits.

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Hans-Peter Hempel was born in Halle-on-Saale, Germany, on 3 October 1926. In 1952, he received a degree in physics from the University of Leipzig.

From 1953 to 1957 Mr. Hempel was with the Institute of Physics in Leipzig where he worked on photoconductors and ultra-high-vacuum applications.

In 1958 he joined the Components Division of Standard Elektrik Lorenz. He has been head of the laboratory for selenium rectifiers since 1959.

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Etching of Valdemar Poulsen

Valdemar Poulsen (1869–1942), Danish engineer, is the subject of the latest in a series of etchings of eminent figures published by the International Telecommunication Union. A member of the technical staff of the Copenhagen telephone company, he is best known for two inventions.

In 1898 he invented the telegraphone, in which the current from a microphone sets up a magnetic field in a recording head. A long thin steel wire moving through the head is magnetized in spots corresponding to the instantaneous strength of the microphone signal. If the wire is then drawn through the head, its motion and retained magnetism will induce currents in the head that will operate a telephone receiver, thus reproducing the signals.

Although the telegraphone was produced commercially as an office dictating machine with the additional capability of answering the telephone and recording a message during the absence of the subscriber, it needed the vacuum-tube amplifier that did not then exist. The later invention of amplifiers, of magnetic particles that could be glued to paper tape and give improved recording at greatly reduced speed of tape travel, and of broadcasting to develop a desire in large numbers of people to record speech and music, brought the telegraphone back as the modern magnetic tape recorder.

In 1903, Poulsen showed that if the Duddell singing arc were operated in a hydrocarbon atmosphere it would produce radio-frequency currents suitable for radio transmission. The Poulsen arc was brought to the United States by C. F. Elwell about 6 years later and the companies he set up in California are linear ancestors of the present International Telephone and Telegraph Corporation. The United States Navy adopted the arc for its long-distance radio stations, and during the first world war the Federal Telegraph Company built for the Navy the 1000-kilowatt arc station near Bordeaux, France. This was the most-powerful radio transmitter in the world for about a quarter of a century.

Thus, Poulsen saw his telegraphone fade and then flower while his arc flowered and then faded. Both contributed substantially to the work and pleasure of the world.

The etching of Poulsen is the thirtieth in a series started in 1935. They are on a good grade of paper and measure 23 by 17 centimeters (9 by 6.6 inches) including margins. Copies of all these etchings are still available at 3 Swiss francs each, postpaid, from the General Secretariat, Union Internationale des Télécommunications, Place des Nations, 1200 Geneva, Switzerland.

The eminent persons honored by these etchings include: Ampère, Armstrong, Baudot, Bell, de Forest, Edison, Erlang, Faraday, Ferrié, Fresnel, Gauss and Weber, Heaviside, Hertz, Hughes, Kelvin, Kirchhoff, Lodge, Lorentz, Marconi, Maxwell, Morse, Popov, Poulsen, Pupin, Rayleigh, Siemens, Sommerfield, Tesla, Van der Pol, and Von Karman.

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World's Telephones—1964 *

There was a net increase of 9.9 million telephones in the world in 1963, bringing the world total to 171,000,000 on January 1, 1964. Numerically, this was the heaviest gain ever reported, and the increase amounted to 6.1%.

Japan now ranks second to the United States in the number of telephones. Included for the first time in this year's figures for Japan, as well as in the data for the past years, is a class of telephone that previously has not been reported. These telephones are located in rural areas of Japan and belong to nearly 2700 small municipal and cooperative systems which are joined together into an association known as the National Wired Broadcasting Telephone Association. Such telephones have loudspeakers and are used part of the time to receive broadcasts on weather, crop, and fishing conditions, as well as musical programs. Between broadcast periods the instruments are used as telephones. Until 1 January 1964 none of them could be connected with the facilities of the Nippon Telegraph and Telephone Public Corporation, which operates the general telephone system of Japan and which reports data for the country as a whole. Such connections are now being made, and this year for the first time the Nippon Telegraph and Telephone Public Corporation has reported these wired broadcasting instruments as telephones.

* Abridgement from the 1964 issue of a booklet, "The World's Telephones," published yearly by the chief statistician's office of the American Telephone and Telegraph Company, New York, New York.

Hungary now raises to 28 the list of countries having as many as half a million telephones. These 28 countries have nearly doubled the combined count of their telephones in the past decade. On a comparable basis, Japan has more than quadrupled, and India more than trebled, their counts of telephones in this period. Countries that have fewer than 500 000 telephones, not included in this list, have in the aggregate more than quadrupled the number of their telephones during the past ten years.

A new telephone cable system, laid by the Bell System between Hawaii and Japan, via Midway, Wake, and Guam, was interconnected at

TABLE 1

TELEPHONES IN CONTINENTAL AREAS—1 JANUARY 1964

Area	Number in Service			Privately Operated (1)		Automatic	
	Number 1964	Percent of World	Per 100 Population	Number 1964	Percent of Total	Number 1964	Percent of Total
North America	90 831 400	53.1	43.4	89 668 000	98.7	89 442 100	98.5
Middle America	1 389 000	0.8	1.9	990 100	71.3	1 226 500	88.3
South America	3 872 800	2.3	2.5	1 880 900	48.6	3 400 100	87.8
Europe	53 377 600	31.2	8.7	9 322 400	17.5	46 764 200	87.6
Africa	2 241 500	1.3	0.8	29 100	1.3	1 708 000	76.2
Asia (2)	15 475 400	9.1	0.8	10 236 300	66.1	9 949 300	64.3
Oceania	3 812 300	2.2	21.2	285,500	7.5	3 113 600	81.7
World	171 000 000	100.0	5.3	112 412 300	65.7	155 603 800	91.0

(1) Necessarily the distinction in this classification is with respect to operation rather than ownership. In particular it is to be noted that systems that are government owned in whole or in part may be privately operated, as in Italy and Japan. The word "government" refers to nations, states, or municipalities.

(2) These data include allowances for the Asiatic parts of Turkey and the Union of Soviet Socialist Republics.

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Hawaii in June 1964 with the existing cable system from California to Hawaii and with that from Canada to Australia. A second cable was placed in service in July between the United States mainland and Hawaii. By the end of 1964 a cable system between Guam and the Philippines, and one from Florida to St. Thomas, Virgin Islands, will be in service. Wake and Midway are new points reached by telephone cable.

Operator distance dialing (already in effect between the United States and Australia, Bermuda, France, Jamaica, Puerto Rico, United Kingdom, Virgin Islands, and West Germany) was extended to Belgium, Italy, Japan, and The Netherlands during 1964.

TABLE 2
TOTAL NUMBER OF TELEPHONES IN SERVICE

Area	1963	1962	1961	1960	1959	1954
North America	87 029 400	83 186 400	79 830 600	76 036 400	71 799 300	54 003 100
Middle America	1 275 800	1 167 300	1 075 900	1 008 000	910 800	670 000
South America	3 732 600	3 475 500	3 337 600	3 145 900	2 999 600	2 245 500
Europe	49 734 800	46 377 000	43 172 700	40 340 900	37 598 100	25 979 000
Africa	2 155 100	2 081 800	2 005 300	1 904 500	1 768 600	1 181 200
Asia (1)	13 577 200	11 903 300	10 353 400	9 110 000	7 555 700	3 881 400
Oceania	3 595 100	3 408 700	3 224 500	3 054 300	2 867 900	2 039 800
World	161 100 000	151 600 000	143 000 000	134 600 000	125 500 000	90 000 000

(1) These data include allowances for the Asiatic parts of Turkey and the Union of Soviet Socialist Republics.

TABLE 3
TELEPHONES BY COUNTRIES AS OF 1 JANUARY 1964

Area		Number of Telephones	Per 100 Population	Percent Automatic	Telephones by Type of Operation (2)	
					Private	Government
NORTH AMERICA						
Canada		6 664 000	34.89	93.5	5 570 000	1 094 000
Greenland		0	—	—	—	—
St. Pierre and Miquelon		400	8.00	0.0	0	400
United States	(1, 2)	84 167 000	44.26	98.9	84 098 000	69 000
MIDDLE AMERICA						
Bahama Islands		16 325	12.05	97.8	1 103	15 222
Barbados		12 412	5.22	100.0	12 412	0
Bermuda	(3)	18 300	38.94	100.0	18 300	0
British Honduras		992	0.97	79.1	794	198
Canal Zone	(3, 4)	8 943	23.17	100.0	0	8 943
Cayman Islands		32	0.40	0.0	0	32
Costa Rica*		20 400	1.49	84.0	19 400	1 000
Cuba		223 745	3.08	92.7	0	223 745
Dominican Republic		27 514	0.81	93.8	27 084	430
El Salvador*		20 600	0.74	50.0	0	20 600

* Estimated.

(1) Data for the State of Alaska are included. Data for the State of Hawaii are included under Oceania, rather than here under North America.
(2) The number shown as governmentally operated is estimated. More than half of such telephones are in the State of Alaska.
(3) Data exclude telephone systems of the armed forces.
(4) Data are as of 30 June 1963.
(5) Data are as of 31 March 1964.
(6) Data for China (Mainland) are as of 1 January 1948. Those parts of the telephone system shown in the table as privately operated continued under such operation until 1949, when they came under government operation.
(7) Data are as of 30 September 1963.

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TABLE 3—*Continued*

TELEPHONES BY COUNTRIES AS OF 1 JANUARY 1964

Area	Number of Telephones	Per 100 Population	Percent Automatic	Telephones by Type of Operation (2)	
				Private	Government
Guadeloupe	5 330	1.78	0.0	0	5 330
Guatemala*	20 000	0.48	92.0	0	20 000
Haiti	4 400	0.10	86.0	0	4 400
Honduras	9 266	0.45	93.0	0	9 266
Jamaica	43 041	2.52	98.8	43 041	0
Leeward Islands:					
Antigua	1 148	1.91	0.0	0	1 148
Montserrat*	220	1.69	0.0	0	220
St. Kitts	645	1.06	72.9	0	645
Total	2 013	1.50	23.3	0	2 013
Martinique	7 703	2.53	74.0	0	7 703
México	659 785	1.69	84.6	658 569	1 216
Netherlands Antilles	17 120	8.39	100.0	0	17 120
Nicaragua*	13 900	0.89	82.0	0	13 900
Panamá	39 086	3.28	98.5	38 486	600
Puerto Rico	172 009	6.77	96.3	162 624	9 385
Trinidad and Tobago	35 060	3.76	86.2	0	35 060
Turks and Caicos Islands	130	2.17	65.4	85	45
Virgin Islands (United Kingdom)	74	0.93	0.0	0	74
Virgin Islands (United States)	6 415	18.33	100.0	6 415	0
Windward Islands:					
Dominica	826	1.31	0.0	0	826
Grenada	1 801	1.94	100.0	1 801	0
St. Lucia	972	1.03	65.8	0	972
St. Vincent	550	0.65	0.0	0	550
Total	4 149	1.24	58.8	1 801	2 348
SOUTH AMERICA					
Argentina	1 425 002	6.50	88.9	101 255	1 323 747
Bolivia*	19 500	0.54	92.0	17 745	1 755
Brazil	1 207 566	1.57	83.0	1 144 773	62 793
British Guiana*	10 100	1.64	92.0	0	10 100
Chile	235 046	2.83	83.0	233 471	1 575
Colombia	372 217	2.44	97.5	15 077	357 140
Ecuador	44 087	0.92	97.4	0	44 087
Falkland Islands and Dependencies	447	22.35	0.0	0	447
French Guiana	1 234	3.43	0.0	0	1 234
Paraguay	12 939	0.66	92.5	0	12 939
Peru	126 325	1.05	87.9	126 325	0
Surinam	6 576	1.97	98.2	0	6 576
Uruguay	169 463	6.60	82.1	0	169 463
Venezuela	242 264	2.93	96.4	242 264	0
EUROPE					
Albania*	6 000	0.34	50.0	0	6 000
Andorra	700	6.36	0.0	0	700
Austria	866 275	12.04	94.3	0	866 275
Belgium	1 370 848	14.70	93.7	0	1 370 848
Bulgaria	225 400	2.78	57.0	0	225 400
Channel Islands:					
Guernsey and Dependencies	14 666	31.14	53.0	0	14 666
Jersey	20 977	33.30	63.5	0	20 977
Total	35 643	32.37	59.2	0	35 643
Czechoslovakia	1 298 766	9.27	85.7	0	1 298 766
Denmark	1 247 958	26.34	62.0	1 104 488	143 470
Finland	728 785	15.98	86.4	513 517	215 268
France	5 336 374	11.09	83.8	0	5 336 374
Germany, Eastern	1 515 271	8.82	97.8	0	1 515 271
Germany, Federal Republic	7 599 571	13.12	99.9	0	7 599 571
Gibraltar (3)	3 497	14.57	100.0	0	3 497
Greece	356 378	4.19	94.6	0	356 378
Hungary	505 198	5.00	72.8	0	505 198
Iceland	47 938	25.64	75.2	0	47 938
Ireland	192 558	6.78	75.0	0	192 558
Italy	5 056 947	9.99	97.3	5 056 947	0
Liechtenstein	5 718	31.77	100.0	0	5 718
Luxembourg	67 899	20.76	100.0	0	67 899
Malta (5)	20 834	6.41	95.6	0	20 834

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TABLE 3—*Continued*

TELEPHONES BY COUNTRIES AS OF 1 JANUARY 1964

Area		Number of Telephones	Per 100 Population	Percent Automatic	Telephones by Type of Operation (2)	
					Private	Government
Monaco*		10 600	48.18	100.0	0	10 600
Netherlands		2 023 258	16.80	100.0	0	2 023 258
Norway		838 223	22.77	75.6	48 207	790 016
Poland		1 088 686	3.52	80.1	0	1 088 686
Portugal		485 177	5.35	76.7	337 925	147 252
Rumania*		376 700	2.00	76.0	0	376 700
San Marino		1 115	6.56	100.0	1 115	0
Spain		2 283 465	7.32	78.5	2 260 180	23 285
Sweden		3 222 699	42.25	95.0	0	3 222 699
Switzerland		1 997 957	33.95	100.0	0	1 997 957
Turkey		286 450	0.95	79.2	0	286 450
U.S.S.R.*		6 502 000	2.87	60.0	0	6 502 000
United Kingdom	(5)	9 345 000	17.41	89.0	0	9 345 000
Yugoslavia*		330 400	1.72	84.0	0	330 400
AFRICA						
Algeria		159 479	1.52	70.7	0	159 479
Angola		12 443	0.25	68.5	0	12 443
Ascension Island	(5)	62	17.82	91.9	62	0
Basutoland		599	0.08	52.3	0	599
Bechuanaland		1 372	0.25	0.0	0	1 372
Burundi*		2 300	0.09	94.0	0	2 300
Cameroons		4 087	0.09	74.0	0	4 087
Cape Verde Islands		310	0.14	80.6	0	310
Central African Republic*		2 300	0.18	90.0	0	2 300
Chad		2 472	0.09	87.4	0	2 472
Comoro Islands		245	0.13	0.0	0	245
Congo (Brazzaville)		7 186	0.85	92.5	0	7 186
Congo (Léopoldville)*		30 000	0.20	85.0	0	30 000
Dahomey		3 129	0.14	86.5	0	3 129
Ethiopia		17 865	0.08	83.4	0	17 865
Gabon		2 817	0.61	94.5	0	2 817
Gambia		921	0.27	93.9	0	921
Ghana		30 741	0.41	72.7	0	30 741
Guinea*		4 200	0.12	72.0	0	4 200
Ifni		244	0.49	0.0	0	244
Ivory Coast*		12 400	0.33	80.0	0	12 400
Kenya		47 706	0.53	81.9	0	47 706
Liberia*		2 800	0.27	100.0	800	2 000
Libya		12 128	0.95	90.7	0	12 128
Madagascar		17 631	0.29	62.6	0	17 631
Malawi		6 533	0.21	89.3	0	6 533
Mali		4 010	0.09	67.5	0	4 010
Mauritania		802	0.08	68.6	0	802
Mauritius and Dependencies		11 185	1.57	68.4	0	11 185
Morocco		141 335	1.10	89.0	18 765	122 570
Mozambique		17 319	0.25	77.3	0	17 319
Niger*		1 700	0.05	80.0	0	1 700
Nigeria		58 658	0.11	79.6	0	58 658
Portuguese Guinea		885	0.17	55.8	0	885
Réunion		7 815	2.07	0.0	0	7 815
Rhodesia		91 992	2.26	89.6	0	91 992
Rwanda		712	0.02	0.0	0	712
Sahara, Spanish		540	2.16	0.0	0	540
St. Helena		138	2.76	0.0	0	138
São Tomé and Principe		444	0.79	67.8	0	444
Sénégal		24 255	0.71	82.4	0	24 255
Seychelles		277	0.62	100.0	277	0
Sierra Leone*		5 500	0.25	80.0	100	5 400
Somalia*		2 500	0.11	0.0	0	2 500
Somaliland, French		1 492	2.13	100.0	0	1 492
South Africa	(5)	1 069 612	6.16	72.4	0	1 069 612
South West Africa		19 901	3.57	41.1	0	19 901
Spanish Equatorial Region		1 156	0.45	91.3	1 156	0
Spanish North Africa		7 973	5.08	100.0	7 973	0
Sudan		32 746	0.25	82.3	0	32 746

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TABLE 3—*Continued*

TELEPHONES BY COUNTRIES AS OF 1 JANUARY 1964

Area		Number of Telephones	Per 100 Population	Percent Automatic	Telephones by Type of Operation (2)	
					Private	Government
Swaziland*		2 900	1.01	60.0	0	2 900
Tanganyika and Zanzibar						
Tanganyika		18 878	0.19	65.9	0	18 878
Zanzibar		1 890	0.58	74.1	0	1 890
Total		20 768	0.20	66.6	0	20 768
Togo		2 610	0.16	74.5	0	2 610
Tunisia		31 840	0.69	57.2	0	31 840
Uganda		16 138	0.23	79.4	0	16 138
United Arab Republic*		264 400	0.93	85.0	0	264 400
Upper Volta*		1 900	0.04	30.0	0	1 900
Zambia		29 001	0.82	96.0	0	29 001
ASIA						
Aden*		6 700	2.95	100.0	0	6 700
Afghanistan*		9 200	0.06	60.0	0	9 200
Bahrein		5 016	3.10	100.0	5 016	0
Bhutan		0	—	—	—	—
Brunei*		1 200	1.28	95.0	0	1 200
Burma		20 828	0.09	74.0	0	20 828
Cambodia		3 875	0.06	78.6	0	3 875
Ceylon*		41 000	0.38	96.0	0	41 000
China, Mainland	(6)	244 028	0.05	72.9	94 945	149 083
China, Taiwan		132 524	1.12	65.9	0	132 524
Cyprus		24 820	4.18	99.3	0	24 820
Hong Kong		178 285	4.89	100.0	178 285	0
India	(5)	684 284	0.15	57.6	3 105	681 179
Indonesia		149 090	0.15	19.2	0	149 090
Iran*		160 000	0.71	75.0	0	160 000
Iraq	(5)	59 831	0.86	79.9	0	59 831
Israel		185 358	7.63	100.0	0	185 358
Japan	(5)	10 682 492	11.06	62.3	10 232 053	450 439
Jordan*		25 000	1.33	70.0	0	25 000
Korea, North		n.a.	—	—	—	—
Korea, Republic of		170 765	0.62	59.1	0	170 765
Kuwait*		18 400	5.11	100.0	2 700	15 700
Laos		1 031	0.05	74.6	0	1 031
Lebanon*		95 000	5.16	86.0	0	95 000
Macao		2 626	1.52	100.0	0	2 626
Malaysia:						
Malaya		98 471	1.28	84.6	0	98 471
Sabah		4 991	1.00	99.6	0	4 991
Sarawak		6 451	0.81	79.8	0	6 451
Singapore		73 879	4.11	100.0	0	73 879
Total		183 792	1.70	91.1	0	183 792
Maldive Islands		0	—	—	—	—
Mongolia		12 234	1.19	69.0	0	12 234
Muscat and Oman		296	0.05	100.0	296	0
Nepál*		1 500	0.02	97.0	0	1 500
Pakistan		107 334	0.11	69.6	0	107 334
Philippine Republic		146 663	0.48	80.1	130 372	16 291
Portuguese Timor		538	0.10	0.0	0	538
Qatar		4 105	7.33	100.0	4 105	0
Ryukyu Islands	(3)	21 000	2.30	81.0	0	21 000
Saudi Arabia*	(4)	26 500	0.40	40.0	0	26 500
Sikkim		249	0.15	0.0	0	249
South Arabia, Protectorate of		0	—	—	—	—
Syria		67 718	1.27	89.2	0	67 718
Thailand	(7)	55 219	0.19	90.0	0	55 219
Trucial Oman		775	0.70	100.0	775	0
Viet-Nam, North		n.a.	—	—	—	—
Viet-Nam, Republic of*		19 300	0.13	85.0	0	19 300
West Irian		2 581	0.34	13.7	0	2 581
Yemen		900	0.02	100.0	0	900
OCEANIA						
Australia	(4)	2 522 522	23.11	80.9	0	2 522 522
British Solomon Islands		490	0.38	98.2	0	490
Canton Island		60	20.00	100.0	0	60
Caroline Islands		750	1.39	0.0	0	750
Christmas Island	(4)	99	3.00	100.0	99	0

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TABLE 3—*Continued*

TELEPHONES BY COUNTRIES AS OF 1 JANUARY 1964

Area	Number of Telephones	Per 100 Population	Percent Automatic	Telephones by Type of Operation (2)	
				Private	Government
Cocos (Keeling) Islands	63	6.30	100.0	0	63
Cook Islands	373	1.96	0.0	0	373
Fiji Islands	9 123	2.06	59.7	0	9 123
Gilbert and Ellice Islands	0	—	—	—	—
Guam	13 867	20.70	99.7	0	13 867
Mariana Islands (less Guam)	350	3.50	100.0	0	350
Marshall Islands	1 055	5.96	97.2	0	1 055
Midway Island	1 230	41.00	100.0	0	1 230
Nauru	0	—	—	—	—
New Caledonia	3 993	4.54	76.4	0	3 993
New Hebrides Condominium	510	0.77	100.0	0	510
New Zealand (5)	901 955	35.00	78.4	0	901 955
Niue Island	119	2.38	0.0	0	119
Norfolk Island	38	3.80	0.0	0	38
Papua and New Guinea	8 371	0.40	80.9	0	8 371
Pitcairn Island	0	—	—	—	—
Polynesia, French	2 107	2.48	0.0	0	2 107
Samoa, American	650	3.10	100.0	0	650
Samoa, Western	1 222	1.02	0.0	0	1 222
Tokelau Islands	0	—	—	—	—
Tonga (Friendly) Islands	630	0.90	0.0	0	630
United States: Hawaii	285 427	39.21	100.0	285 427	0
Wake Island	180	12.00	100.0	0	180

TABLE 4

TELEPHONE CONVERSATIONS DURING 1963

Country or Area	Thousands of Conversations			Average Conversations per Person
	Local	Long Distance	Total	
Algeria	68 050	6 207	74 257	7.1
Angola	21 172	557	21 729	4.3
Argentina	4 029 135	49 793	4 078 928	187.8
Australia (1)	1 809 000	84 500	1 893 500	175.1
Bahama Islands	38 943	351	39 294	304.6
Barbados, West Indies	24 500	20	24 520	103.9
Belgium	740 657	167 115	907 772	97.7
Bermuda	17 043	89	17 132	364.5
Brazil	6 491 887	106 499	6 598 386	86.6
Cambodia	4 329	262	4 591	0.8
Cameroons	3 618	581	4 199	0.9
Canada	11 054 000	259 259	11 313 259	597.7
Chad	2 119	28 (4)	2 147	76.7
Channel Islands	19 414	1 052	20 466	185.9
Chile	594 773	20 872	615 645	74.9
China, Taiwan	460 668	14 388	475 056	40.6
Colombia	1 142 571	17 194	1 159 765	76.8
Congo (Brazzaville)	5 590	576 (4)	6 526	7.8
Cuba	1 202 655	12 977	1 215 632	168.8
Cyprus	26 897	2 802	29 699	50.4
Czechoslovakia	854 313	109 283	963 596	69.1
Dahomey	3 040	438	3 478	1.5
Denmark	1 276 377	314 470	1 590 847	337.0
Ethiopia	27 038	1 107	28 145	1.3
Fiji Islands	9 000	1 001	10 001	23.0

(1) Data are for the year ended 30 June 1963.
(2) Data are for the year ended 31 March 1963.
(3) Data are for the year ended 30 September 1963.
(4) Three-minute units.

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TABLE 4—*Continued*

TELEPHONE CONVERSATIONS DURING 1963

Country or Area		Thousands of Conversations			Average Conversations per Person
		Local	Long Distance	Total	
France		1 184 353	774 406	1 958 759	40.9
Gabon		2 367	74 (4)	2 441	5.3
Germany, Eastern		839 281	202 089	1 041 370	60.7
Germany, Federal Republic		4 132 706	1 528 389	5 661 095	98.3
Ghana		25 957	2 683	28 640	3.9
Gibraltar		9 695	100	9 795	408.1
Greece		773 664	20 130	793 794	93.6
Guadeloupe		2 438	437	2 875	9.7
Hungary		561 211	34 897	596 108	59.1
Iceland		103 172	3 180	106 352	574.9
Ireland		161 231	18 195	179 426	63.2
Italy		6 185 541	729 218 (4)	6 914 759	137.0
Jamaica, West Indies		70 596	1 042	71 638	42.5
Korea, Republic of		1 427 319	18 283	1 445 602	53.0
Libya		17 186	581	17 767	14.0
Liechtenstein		2 122	2 110 (4)	4 232	235.1
Madagascar		14 480	944	15 424	2.6
Mali		2 960	163	3 123	0.7
México		1 431 548	26 229	1 457 777	37.9
Morocco		128 337	10 851	139 188	11.0
Mozambique		18 791	1 726	20 517	3.0
Netherlands		1 255 755	613 887	1 869 642	156.2
Netherlands Antilles		35 000	32	35 032	173.4
New Caledonia		2 446	235	2 681	31.5
Nigeria		70 225	4 148	74 373	1.3
Norway		605 781	64 907	670 688	182.9
Papua and New Guinea		6 630	186	6 816	3.3
Peru		616 807	8 170	624 977	52.7
Philippine Republic		1 078 800	1 543	1 080 343	35.7
Polynesia, French		1 250	115	1 365	16.9
Portugal		449 935	53 563	503 498	55.7
Puerto Rico		311 311	6 283	317 594	126.4
Réunion		3 559	336	3 895	10.5
Sarawak		14 862	933	15 795	19.8
Senegal		15 000	839	15 839	4.7
Singapore		234 790	1 581	236 371	133.2
South Africa	(2)	1 377 304	94 760	1 472 064	85.8
South West Africa		16 783	2 626	19 409	35.0
Sweden	(1)	2 962 000	446 500	3 408 500	446.9
Switzerland		768 986	760 201 (4)	1 529 187	263.2
Thailand	(3)	110 170	427	110 597	3.9
Togo		1 916	362	2 278	1.5
Trinidad and Tobago, West Indies		108 054	8 855	116 909	127.1
Tunisia		21 317	8 200	29 517	6.4
Turkey		281 338	18 527	299 865	10.1
United Kingdom	(2)	5 150 000	628 800	5 778 800	107.9
United States		103 941 000	3 984 000	107 925 000	570.0
Virgin Islands (United States)		19 065	92	19 157	547.3

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International Telephone and Telegraph Corporation
Principal Divisions and Subsidiaries

NORTH AMERICA

MANUFACTURING—SALES—SERVICE

Canada

ITT Canada Limited, Montreal
Royal Electric Company (Quebec) Ltd., Pointe Claire, P.Q.

Jamaica

ITT Standard Electric of Jamaica Ltd., Kingston

Mexico

Industria de Telecomunicación, S. A. de C. V., Mexico City
Materiales de Telecomunicación, S. A., Toluca
McClellan, S. A., Mexico City
Standard Eléctrica de México, S. A., Mexico City

Panama

ITT Standard Electric of Panama, S. A., Panama City

Puerto Rico

ITT Caribbean Manufacturing, Inc., Rio Piedras
ITT Caribbean Sales and Service, Inc., Rio Piedras

United States of America

Airmatic Systems Corporation, Saddle Brook, N. J.
Barton Instrument Corporation, Monterey Park, Calif.
Federal Electric Corporation, Paramus, N. J.
Industrial Products Division, San Fernando, Calif.
Intelex Systems Incorporated, Paramus, N. J.
International Standard Electric Corporation, New York, N. Y.
International Telephone and Telegraph Corporation, Sud America, New York, N. Y.
ITT Arkansas Division, Camden, Ark.
ITT Bell & Gossett Hydronics Division, Morton Grove, Ill.
Stover branch, Freeport, Ill.
ITT Cannon Electric Division, Los Angeles, Calif.
ITT Data and Information Systems Division, Paramus, N. J.
ITT Direct Fired Equipment Division, Columbus, Ohio; Mercer, Pa.; Torrance, Calif.
ITT Electron Tube Division, Easton, Pa. and Roanoke, Va.
ITT Export Corporation, New York, N. Y.
ITT Federal Laboratories Division, Nutley, N. J.
ITT Financial Services, Inc., New York, N. Y.
ITT Aetna Finance Company, St. Louis, Mo.
Kellogg Credit Corporation, New York, N. Y.
International Telephone and Telegraph Credit Corporation, New York, N. Y.
Great International Life Insurance Company (50% interest), Atlanta, Ga.
ITT General Controls Division, Glendale, Calif.
ITT Gilfillan Inc., Los Angeles, Calif.
ITT Hammel-Dahl Division, Warwick, R. I.
ITT Industrial Laboratories Division, Fort Wayne, Ind.
ITT Kellogg Communications Systems Division, Chicago, Ill.
ITT Kellogg Telecommunication Division, New York, N. Y.; Corinth, Miss.; Milan, Tenn.; Raleigh, N. C.
ITT Mackay Marine Division, Clark, N. J.
ITT Marlow Division, Midland Park, N. J.
ITT Mobile Telephone, Inc., San Fernando, Calif.
ITT Nesbitt Division, Philadelphia, Pa.
ITT Process Systems Division, Lawrence, Mass.
ITT Semiconductors Division, West Palm Beach, Fla. and Lawrence, Mass.
ITT Terryphone Corporation, Harrisburg, Pa.
ITT Wire and Cable Division, Pawtucket, R. I.; Woonsocket, R. I.; Clinton, Mass.
Jennings Radio Manufacturing Corporation, San Jose, Calif.

TELEPHONE OPERATIONS

Puerto Rico

Puerto Rico Telephone Company, San Juan

Virgin Islands

Virgin Islands Telephone Corporation, Charlotte Amalie

SOUTH AMERICA

MANUFACTURING—SALES—SERVICE

Argentina

Compañía Standard Electric Argentina, S.A.I.C., Buenos Aires

Brazil

Standard Eléctrica, S. A., Rio de Janeiro
Electrônica Industrial S. A., Sao Paulo

Chile

Compañía Standard Electric, S.A.C., Santiago

Colombia

ITT Standard Electric de Colombia, S. A., Bogotá

Venezuela

Standard Telecommunications C.A., Caracas

TELEPHONE OPERATIONS

Brazil

Companhia Telefônica Nacional, Curitiba

Chile

Compañía de Teléfonos de Chile, Santiago

Peru

Compañía Peruana de Teléfonos Limitada, Lima

EUROPE, MIDDLE EAST, AFRICA

MANUFACTURING—SALES—SERVICE

Algeria

Société Algérienne de Constructions Téléphoniques, Algiers

Austria

Standard Telephon und Telegraphen Aktiengesellschaft, Czeija, Nissl & Co., Vienna

Belgium

Bell Telephone Manufacturing Company, Antwerp
ITT Europe, Inc. (branch), Brussels
ITT Industries, Europe Inc. (branch), Brussels

Denmark

Standard Electric Aktieselskab, Copenhagen

Finland

Standard Electric Puhelinteollisuus Oy, Helsinki

France

Centre Français de Recherche Opérationelle, Paris
Compagnie Générale de Constructions Téléphoniques, Paris
Les Téléimprimeurs, Paris
Compagnie Générale de Métrologie, Annecy
Laboratoire Central de Télécommunications, Paris
Le Matériel Technique Industriel, Paris
Le Matériel Téléphonique, Paris
Société Industrielle de Composants pour l'Electronique, Courbevoie

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Germany (West)

Standard Elektrik Lorenz Aktiengesellschaft, Stuttgart, and subsidiaries

Iran

Standard Electric Iran AG, Tehran

Italy

Fabbrica Apparecchiature per Comunicazioni Elettriche Standard S.p.A., Milan
Società Impianti Elettrici Telefonici Telegrafici e Costruzioni Edili S.p.A., Florence
ITT Domel Italiana S.p.A., Milan

Netherlands

Internationale Gas Apparaten N.V., The Hague (joint venture)
Nederlandsche Standard Electric Maatschappij N.V., The Hague

Nigeria

Kollerich (Nigeria) Limited, Lagos

Norway

Standard Telefon og Kabelfabrik A/S, Oslo

Portugal

Standard Eléctrica, S.A.R.L., Lisbon

Republic of South Africa

Standard Telephones and Cables (South Africa) (Proprietary) Limited, Boksburg East, Transvaal
Supersonic Africa (Pty.) Ltd., Johannesburg, Transvaal

Spain

Compañía Internacional de Telecomunición y Electrónica, S.A., Madrid
Compañía Radio Aérea Marítima Española, S.A., Madrid
Standard Eléctrica, S.A., Madrid

Sweden

ITT Norden AB, Barkarby
Standard Radio & Telefon AB, Barkarby

Switzerland

Intel S.A., Basle
ITT Standard S.A., Basle
Standard Téléphone et Radio S.A., Zurich
Steiner S.A., Berne

Turkey

Standard Elektrik ve Telekomünikasyon Limited Şirketi, Ankara

United Kingdom

Creed and Company Limited, Brighton
ITT Industries Limited, London, and subsidiaries
Standard Telephones and Cables Limited, London
Standard Telecommunication Laboratories Limited, London, and other subsidiaries

FAR EAST AND PACIFIC

MANUFACTURING—SALES —SERVICE

Australia

Standard Telephones and Cables Pty. Limited, Sydney

Hong Kong

ITT Far East and Pacific, Inc. (branch), Hong Kong
ITT Far East Ltd., Hong Kong

Japan

ITT Far East and Pacific, Inc. (branch), Tokyo

New Zealand

Standard Telephones and Cables Pty. Limited (branch), Upper Hutt, Wellington

Philippines

ITT Philippines, Incorporated, Makati, Rizal

INTERNATIONAL COMMUNICATIONS OPERATIONS

American Cable & Radio Corporation, New York
All America Cables and Radio, Inc.
Commercial Cable Company, The
ITT Cable and Radio, Inc.—Puerto Rico
ITT Central America Cables & Radio, Inc.
ITT Communications, Inc.—Virgin Islands
ITT World Communications Inc.
Mackay Radio and Globe Wireless of the Philippines
Companhia Rádio Internacional do Brasil, Rio de Janeiro
Compañía Internacional de Radio Boliviana, La Paz
Compañía Internacional de Radio, S.A., Buenos Aires
Compañía Internacional de Radio, S.A., Santiago
Cuban American Telephone and Telegraph Company (50% interest), Havana
Radio Corporation of Cuba, Havana

ASSOCIATE LICENSEES FOR MANUFACTURING

(MINORITY INTEREST)

Australia

Austral Standard Cables Pty. Limited, Melbourne

France

Lignes Télégraphiques et Téléphoniques, Paris

Italy

Società Italiana Reti Telefoniche Interurbane, Milan

Japan

Nippon Electric Company, Limited, Tokyo
Sumitomo Electric Industries, Limited, Osaka

Spain

Marconi Española, S.A., Madrid

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