# ELECTRICAL COMMUNICATION

ITT

VOLUME 38 • NUMBER 4 • 1963

# ELECTRICAL COMMUNICATION

Technical Journal Published Quarterly by
INTERNATIONAL TELEPHONE and TELEGRAPH CORPORATION
320 Park Avenue, New York 22, New York
President: H. S. Geneen
Secretary: J. J. Navin

## CONTENTS

**Volume 38** **1963** **Number 4**

Notes

EDITOR, H. P. Westman
ASSISTANT EDITOR, M. Karsh

Subscription: $2.00 per year
50¢ per copy

# This Issue in Brief

**Analog Signal Multiplier**—A cascade-bridge type of analog signal multiplier forms its product in stable parametric fashion, using an uncomplicated structure. The signal voltage required may be of the order of 0.1 volt, or less.

The use of voltage-adjustable capacitors, or parametric diodes, permits the desired product to be effected over a very-wide frequency range. Products of complicated time functions can be obtained within the present bandwidth capabilities of approximately 400 kilocycles per second in the developmental model. Extension to megacycle-per-second bandwidths is entirely feasible.

Products of various types of analog signal voltages are illustrated. A number of applications for the multiplier are also suggested, ranging from the practically instantaneous phase reversal of waveforms to the more-sophisticated application of cross correlation to improve signal-to-noise ratios.

**Transmission-Loss Measurements and Diversity Tests Made on Two Knife-Edge Diffraction Paths in Europe**—Results are given of radio propagation measurements made at 1965 megacycles per second on two knife-edge diffraction paths. The measured transmission loss indicated that the diffraction gain was not the same on the two paths. On each path, the measured loss was less than was estimated for normal beyond-the-horizon paths having the same angular distance, but was greater than the values calculated from knife-edge diffraction theory.

Diversity tests made with receiving antennas spaced 47 feet (14.3 meters) horizontally and 10 feet (3 meters) vertically showed that there was little correlation on either path between the carrier-level variations at the two diversity receiving antennas. A reason for this low correlation may be that in each case the knife-edge obstruction was not smooth but was a ridge varying in elevation across the path.

**Marine Transmitter with Single-Sideband Facility**—A marine transmitter for the 400-to-535-kilohertz and 1.6-to-25-megahertz ranges produces a peak power of 1200 watts. For telegraph operation, the carrier may be keyed or the carrier and one sideband of audio-frequency modulation may be keyed. For telephony, single sideband may be transmitted with either suppressed or full-amplitude carrier.

A linear power amplifier using four *4X250B* tubes in parallel is driven by a preamplifier, which is operated from a frequency-determining and modulating exciter. The transmitter uses 94 quartz crystals, without temperature control, to produce the multiplicity of channels in the various bands. A frequency-drift-cancelling circuit is used.

The carrier and single-sideband signal components are generated independently of each other and may be combined in any proportion as required.

A $\pi$-$L$ network permits proper loading of the many types of antennas found on shipboard without modification of either the equipment or antennas.

**Submarine-Coaxial-Cable Manufacture at Southampton Factory 2**—The construction of the second submarine-cable factory at Southampton was initiated in 1961. It is used for the manufacture of a new design of armourless deep-sea cable developed by Bell Telephone Laboratories for their *SD* long-distance submarine-cable telephone systems. The new cable is of coaxial construction and has the strength member, consisting of 41 high-tensile-steel wires, inside the inner-conductor copper tube. These two components are produced from one machine in which a strander, a tube-forming mill, an argon-arc seam welder, and a tube-drawing mill are operated in tandem. A complex electronic controller maintains correctly the relative speeds of the four units of this machine during starting, running, and stopping. A current proportioner varies the arc current as a preset function of the line speed.

The core insulant is a high-molecular-weight low-density polythene, and after extrusion the external diameter is shaved at 20 degrees centigrade to within a tolerance of 0·1 per cent of the specified diameter of 1·000 inch (25·4 millimetres). The shaving operation also corrects for extruder failure to keep the inner conductor concentrically within the polythene insulant.

A single copper tape is applied longitudinally with overlapping edges over the shaved core to form the outer conductor. In these operations, the tubing mill is driven in tandem with the sheathing extruder. A sheath of high-molecular-weight high-density black polythene plastic compound completes the new deep-sea cable, for which the manufacturing unit is a 20-nautical-mile (37-kilometre) repeater-section length.

**Telepuls and FMX Modular Electromechanical and Electronic Remote Control**—A group of standard units permit the needs of various types of users of remote indication and control systems to be met without special designs. Telepuls 10 and 12 are pulse-code binary systems using electromechanical and electronic means, respectively, and are so fully compatible that their units may be intermixed for certain purposes.

A mark and space, each of 40 milliseconds duration, represent a binary 0 and a similar 160-millisecond pair indicate binary 1. By reference to two 60-millisecond pulses that are initiated by the start and the end of each received signal pulse, the shorter pulses cancel each other and do not affect a flip-flop circuit but the longer signals turn the flip-flop on and then off, thus selecting only the 1's for decoding the signals.

Memories supply both the addresses of the equipment to be queried or controlled and the proper form of message for the type of equipment. The usual mimic diagram shows the system conditions by lighted lamps. Analog information is converted into decimal and then binary code for transmission.

The *FMX1* system uses 15 tone frequencies for identifying equipment and is limited to party-line control of substations having only 1 or a few controlled units. Presence of a tone indicates "good" or "on" and its absence the alternate condition.

**Digitel 108, A Digital Telemetering System**—All data are transmitted in binary digits, suitable transducers being employed to convert analog measurements to this form. A basic group of 8 inputs has been selected and multiple groups may be used for larger installations. A telemetering precision within ±0.5 percent is obtained. Transmission is within a 120-hertz voice-frequency telegraph channel, which may be derived from a telephone channel.

Binary coding uses a frequency deviation of −30 hertz for 0 and +30 hertz for 1. Address and message codes are taken from memories and a parity pulse is added if necessary to make an even number of pulses for each complete telegram. Converters permit parallel inputs and outputs to be transmitted serially over the telegraph channel.

Plug-in printed circuits and transistors permit great flexibility in the design and assembly of systems. Convenient test points, built-in light indicators, reference oscillograms in the instruction manual, programs for routine tests, and rapid replacement of units minimize out-of-service time.

**Transradar—A Narrow-Band Radar Relay Equipment Using Storage Capacitors**—For air-traffic control, it is helpful to use several surveillance radars whose coverages overlap or supplement each other. For economy and flexibility, the megahertz video band from each radar may be greatly compressed to permit remote plan-position indicators to be operated over telephone channels.

Bandwidth compression is obtained by sequentially charging a bank of capacitors, each corresponding to a different increment of range from the radar antenna. A writing timer under control of the radar trigger determines the sequence in which the capacitors are charged from the radar video signal and the charges are read at a slower rate. This derived narrow-band video signal, supplemented by trigger and azimuth information in digital form, is encoded. It is then transmitted for long distances, using vestigial-sideband modulation, over conventional radio or telephone lines. The transmitting unit may be installed at unattended radar stations, which may then be operated by remote control.

The received signal may operate a slow-scan plan-position indicator directly. However, if the conventional high-speed indicator is used, another bank of storage capacitors will be required at the receiver to permit bandwidth expansion corresponding to the compression in the transmitter. By time sharing, signals from more than one radar may be reproduced on a single indicator.

**Optimization of Network Configurations Through Queueing Theory**—A major problem in system communication is the optimum location of manned and/or machine information-processing stations. The problem is further complicated if the information is processed on a priority basis.

A queueing-theory model that simplifies the process of obtaining an optimum network configuration and a computer flow-chart of the model are presented. The output of this model is a generalized set of curves showing traffic density versus queue length and traffic density versus waiting time, both as functions of processing rates and priority.

A numerical example is included to illustrate the method by which the expected queue lengths and waiting times can be found for a station in the iterative process of determining an optimum network configuration for a semiautomatic communication system.

**Traffic Simulation and Its Application in Telephony**—Approximate computation of the traffic-handling capacity of telephone switching networks may be verified by measurements based on artificially generated traffic. Electronic computers may replace the traffic machine to generate artificial traffic and at the same time simulate the switching network to be tested.

In a simple example, the number of rejected or lost calls from an infinite number of inlets to 3 outlets for a traffic of 2 erlangs is to be found. A series of random numbers from 1 through 6 is produced by repeated casting of a 6-sided die. Numbers 1, 2, and 3 indicate release of any call that may be present on the correspondingly numbered outlet. Numbers 4 and 5 initiate a call on the lowest-numbered free outlet. No action is initiated by 6. Although the successive steps do not correspond to constant time intervals, either real or artificial, this method produces calls having exponentially distributed holding times. The test should encompass several thousand numbers in the random series to ensure a balance of statistical variations.

In a practical test, the number of outlets and of calls would be much larger than in the example above. The paper is concerned with the application of a computer to the essential elements of developing the required random numbers, their use as artificial traffic, and the simulation of the switching network in handling the traffic offered to it, to find the quantity of calls rejected because outlets were not immediately available to them. Lost calls were not stored for later attempts for completion.

**Automatic Switching for Telex**—The telex offices of the American Cable and Radio Corporation in San Francisco and in New York, with extension to Washington, District of Columbia, now employ *7E* rotary switching equipment to expand facilities and increase automation. Compatibility of operation must be maintained with the domestic telegraph and telephone telex networks and with corresponding overseas networks to provide subscriber-to-subscriber service with minimum operator assistance.

A 3-digit prefix that identifies the geographical area in which the telex subscriber's number applies permits worldwide calling on an automatic basis. Operator assistance may be required on calls switched through the domestic telephone network for certain tone signals that are not automatically translated into teleprinter signals.

Each telex operator position can handle 2 calls and make 6 connections simultaneously. Overflow positions are provided for peak-traffic intervals. Billing is automatic.

The telex subscriber may receive traffic and also send messages at the conventional per-word rate.

Telex subscribers of the domestic telegraph and telephone systems may gain access to the overseas network of American Cable and Radio Corporation.

The *7E* teleprinter switching system uses single-motion nonhoming 7-brush rotary switches having 100 outlets. Switches are positioned under the direct control of a register by matching 12 phases and 2 voltages of a 450-hertz supply to provide 24 distinguishable positions.

Wire-spring relays provide long life and fast operation with low actuating power. Each line circuit has 2 relays and 2 resistors that are adjusted initially to compensate for the line characteristics and to set the bias current of the repeater in the connecting circuit.

**High-Power Precision-Stabilized Rectifiers—** Three-phase full-wave rectifiers and filters are described for producing direct currents as large as 450 amperes at powers above 100 kilowatts. Regulation maintains the load current and ripple within 5 parts in $10^5$ for an 8-hour period and within 5 parts in $10^6$ for approximately 5 minutes.

The voltage drop across a precision resistor carrying the load current is compared with a reference voltage derived from 3 cascaded stages of zener diodes. Any difference voltage goes to a direct-current amplifier and the greatly amplified output is connected to the base of one or more control transistors that also carry the full load current. The voltage drop across the emitter and collector of this transistor stage controls a 3-phase motor-driven regulator transformer. An auxiliary chopper and amplifier are provided to compensate for drift in the direct-current amplifier.

# Recent Achievements

**Laval Factory Opens**—Mr. Jacques Marette, Minister of the French Posts and Telecommunications, presided over the opening on 5 July 1963 of the new Sept-Fontaines plant in Laval, about 290 kilometers (180 miles) west of Paris. The 300 guests were welcomed by Dr. Maurice Deloraine, president, and Mr. Philippe Lizon, managing director, of Le Matériel Téléphonique.

Covering an area of 15 000 square meters (161 460 square feet), the plant will assemble Pentaconta crossbar telephone equipment, its capacity being about 250 000 lines per year. In addition, manufacture of capacitors will soon be initiated. The present staff of 650 will be increased to 1000. Figure 1 shows a partially assembled Pentaconta switch being inspected.

*Le Matériel Téléphonique*
*France*

**Quasi-Electronic Telephone Exchange**—Richard Stücklen, Minister of the German Posts, Telegraphs, and Telephones Administration, may be seen in Figure 2 as he opened a quasi-electronic telephone exchange in Stuttgart on 12 July 1963 by placing a call from a push-button subscriber's set.

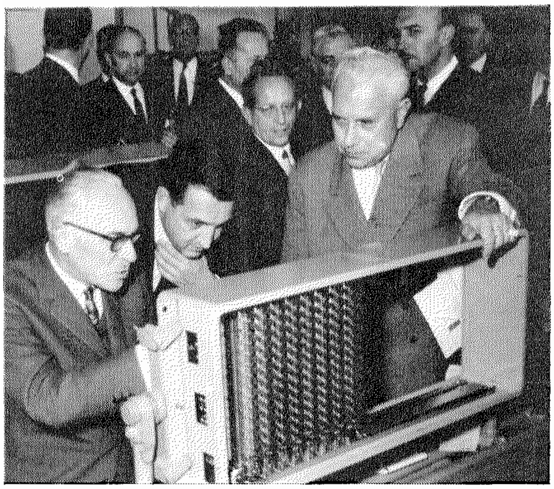


Figure 1—Examining a Pentaconta switch at the opening of the new Laval factory are, from left to right, Dr. Maurice Deloraine, president of Le Matériel Téléphonique; Mr. Jacques Marette, minister of French Posts and Telecommunications; and Mr. Philippe Lizon, managing director of Le Matériel Téléphonique.

The 2000-line Blumenstrasse office shown in Figure 3 is part of the Stuttgart local area. In addition to the standard German facilities of subscriber dialling of long-distance and international calls and of identified out-dialling and in-dialling with private automatic branch exchanges, the new office provides push-button calling and alternative routing within the local area.

This trial installation will provide practical experience with the functioning of the equipment and system as well as the acceptance by subscribers of the new features that it makes available to them.

The *HE-60* quasi-electronic switching system uses hermetically sealed dry-reed switches and electronic components exclusively, combining the speed of electronic action with the reliability of mechanical contacts. A multistage link arrangement uses 4-contact crosspoints with individual reed switches and coils for each contact.

The desired connection is established in one marker operation by the register. While the

Figure 2—Richard Stücklen opening Blumenstrasse quasi-electronic office in Stuttgart using a push-button subscriber's set.

central translator is determining the optimum route, the path through the network is found with the help of a guide wire. Subscriber lines are in groups of 2000 that are served on a one-at-a-time basis by a central control. The common units are duplicated and normally operate alternately but one unit is capable of carrying the entire load in case of trouble. Punched-tape reports are made automatically of all faults. Standardized plug-in units are used.

*Standard Elektric Lorenz*
*Germany*

**Non-Linear Speech Amplitude Quantizer**—The standard telephone channel bandwidth from 300 to 3400 cycles per second passes more information than is needed for correct recognition of human speech. To investigate this field, a non-linear amplitude quantizer in which the number of steps and their grading could be adjusted was used to find the minimum number of levels necessary for accurate reception of English language speech.

With 4 amplitude levels for each half wave having relative magnitudes of 1, 4·6, 22, and 100, corresponding to 13·4-decibel steps, the received speech was above minimum acceptable telephone quality. Optimum performance was obtained with peak clipping of about 8 decibels at the highest level. Speech became slightly harsh from harmonics of the vowel sounds when the signal was 20 decibels above this optimum.

Although the output waveform was a very crude copy of the input analogue waveform, it contained the elements necessary for understandability and for conveying the natural variations in loudness of the original speech. The understandability was maintained over a 30-decibel variation in the mean level of speech input.

*Standard Telecommunication Laboratories*
*United Kingdom*

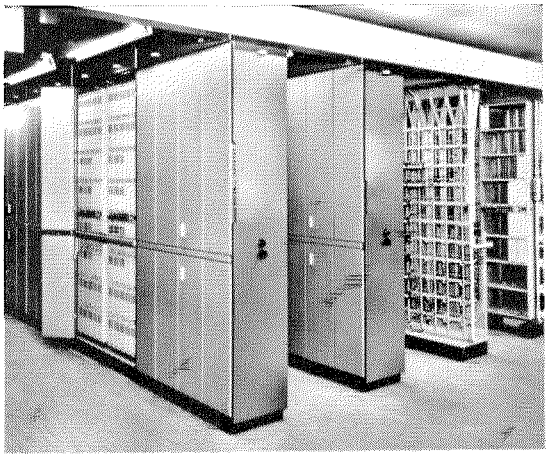


Figure 3—Quasi-electronic 2000-line exchange based on electronic switching of dry-reed contacts. It is part of the Stuttgart local-area telephone system.

**Accelerometer Calibrator**—The equipment shown in Figure 4 is used for the comparative calibration of accelerometers. When used with a reference accelerometer calibrated to within

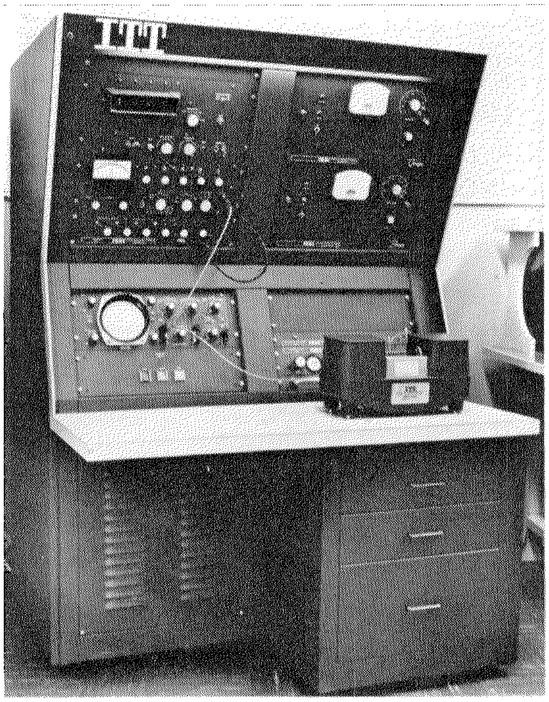


Figure 4—Equipment for comparative calibration of accelerometers.

±1 percent by the National Bureau of Standards, the system will have a total possible error of less than 2.4 percent and a probable error of only 1.2 percent.

A series of seminars on the techniques for calibrating vibration pickups conducted at our San Fernando, California, plant permitted about 50 specialists from the leading industrial companies active in this field to contribute to the design criteria for this apparatus.

*ITT Industrial Products Division*
*United States of America*

**Pentomat Pentaconta Private Branch Exchanges**—A new series of Pentaconta private automatic branch exchanges has been developed. Three sizes are available for 40 extensions and 5 trunks, 10 extensions and 3 trunks, and 6 extensions and 2 trunks, being designated Pentomat *40, 10,* and *6,* respectively.

The entire equipment, including the mains-operated power supply, is housed in a metal cabinet. The Pentaconta relays are mounted on hinged frames for easy and rapid accessibility. Operation of this all-relay system is quiet and free of maintenance.

The Pentomat *40* is operated with an attendant's set; if traffic is heavy, the Pentomat *10* will be similarly equipped. The smallest unit does not normally need this facility. The attendant's set shown in Figure 5 is a compact desk unit with provision for both dial and push-button selection.

Figure 5—Attendant's set for Pentomat *10* private branch exchange.

*Bell Telephone Manufacturing Company*
*Belgium*

**Transatlantic Telephone Cable**—Now being laid over a distance of 3600 nautical miles (6667 kilometres), *TAT-3* will be the longest single submarine cable link in the world.

In Figure 6, *HMTS Alert* is taking on a supply of the new lightweight cable from the world's largest submarine-cable factory built by Standard Telephones and Cables at Southampton Docks in England. A joint project of the British Post Office and the American Telephone and Telegraph Company, this will be the first repeatered telephone cable to link directly the United Kingdom and the United States.

*Standard Telephones and Cables*
*United Kingdom*

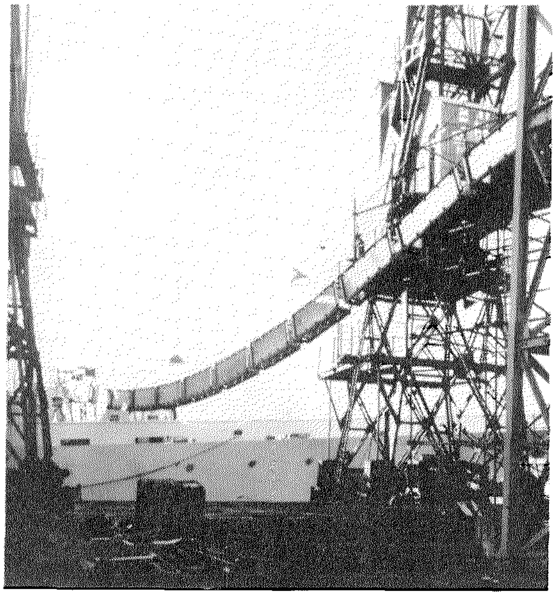


Figure 6—*HMTS Alert* taking on a cargo of the new lightweight coaxial cable from the largest submarine cable plant in the world at Southampton Docks, England.

**Coherent Light Pierces Steel**—Enormous energy may be concentrated into a very small area by focusing coherent light as shown in Figure 7.

A laser cooled by liquid nitrogen produces a coherent light that is focused by a photographic lens onto a steel plate. A single pulse lasting only 200 microseconds pierces a hole having a diameter of about 50 microns in the 1.5-millimeter (0.06-inch) plate.

*Laboratoire Central de Télécommunications*
*France*

**Netherlands Automatic Telex Exchanges**—The Leeuwarden telex exchange in The Netherlands was cut over on 18 April 1963. The network uses the indirect rotary system developed by Bell Telephone Manufacturing Company.

Separate telex exchanges have been built in Amsterdam, The Hague, Rotterdam, Bois-le-Duc, Groningen, Utrecht, and Arnhem. Additional exchanges are under construction for Venlo, Deventer, Hengelo, Eindhoven, and Haarlem.

There are more than 6000 private telex subscribers in the country. All national and 95 percent of the international connections are made automatically.

*Nederlandsche Standard Electric Maatschappij*
*The Netherlands*

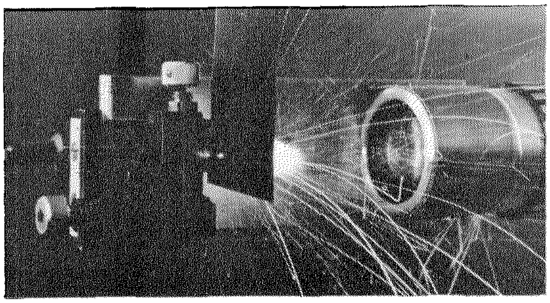


Figure 7—Coherent light from laser pierces hole in steel plate.

**Digital Recording of Wind-Tunnel Tests**—A characteristic of wind-tunnel testing is the large number of measurements that are taken in a short time.

These measurements may now be recorded by a page teleprinter that also punches a paper tape to be used as input to a Zebra computer for evaluation of the test results.

A second-generation design shown in Figure 8 makes increased use of electronic circuits for control of the teleprinter.

*Standard Telephone & Radio*
*Switzerland*

**Poly Para Xylylene Capacitor**—A very-thin film of poly para xylylene, having a permittivity of 1.7 and a breakdown strength of 2 megavolts per centimetre, is the dielectric of a new capacitor. The material may be deposited from the gas phase on to a suitable substrate such as aluminium foil, glass, et cetera, by passing para xylene vapour through a tube at 900 degrees centigrade and at a pressure of a few millimetres.

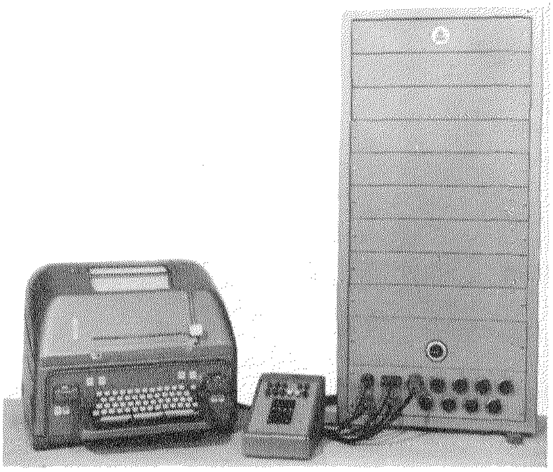


Figure 8—Combined page teleprinter and tape perforator with input equipment for recording wind-tunnel test data. The tape becomes input to a Zebra computer.

The poly para xylylene may exist in three forms, amorphous, $\alpha$, and $\beta$, depending on the pressure, temperature of the gas, and temperature of the substrate. The amorphous form converts to $\alpha$ at about 50 to 140 degrees centigrade and then to the $\beta$ form at 200 to 300 degrees centigrade. The $\beta$ form thus produced is brittle in contrast to that obtained directly from the gas phase. The amorphous and $\alpha$ forms are both mouldable but change very quickly to the $\beta$ form at moulding temperatures. Ultra-high pressures will reconvert $\beta$ to $\alpha$.

Typical capacitors having 1-square-centimetre electrodes, a calculated film thickness of 16 microns, and a weight of 0.01 gramme, have a capacitance of 130 picofarads, loss angle of 0.0012, and resistance of 1 megohm. After irradiation with 2-million-electron-volt electrons at doses up to 50 000 megarads, their electrical properties were changed to a capacitance of 122 picofarads, loss angle of 0.0026, and resistance of 500 kilohms.

*Standard Telecommunication Laboratories*
*United Kingdom*

**New Factory in Eggenberg, Austria**—On 21 June 1963, a new factory was inaugurated in Eggenberg, Austria. With a floor area of 3000 square meters (32 280 square feet), the plant will accommodate up to 350 workers. It is shown in Figure 9.

This factory will be used for assembling telephone bays, relay strips, Villaphone home telephones, and Dirigent intercommunication sets including loudspeaker and door-answering designs.

*Standard Telephon und Telegraphen*
*Austria*

**Radio Altimeter**—A new design of radio altimeter, coded *STR 52,* has simultaneous outputs of 0–2500 feet (0–762 metres) for terrain clearance in low-strike fighter operation and of 0–500 feet (0–152 metres) for use in automatic landing. The aerials may be as far as 50 feet (15 metres) from the equipment.

*Standard Telephones and Cables*
*United Kingdom*

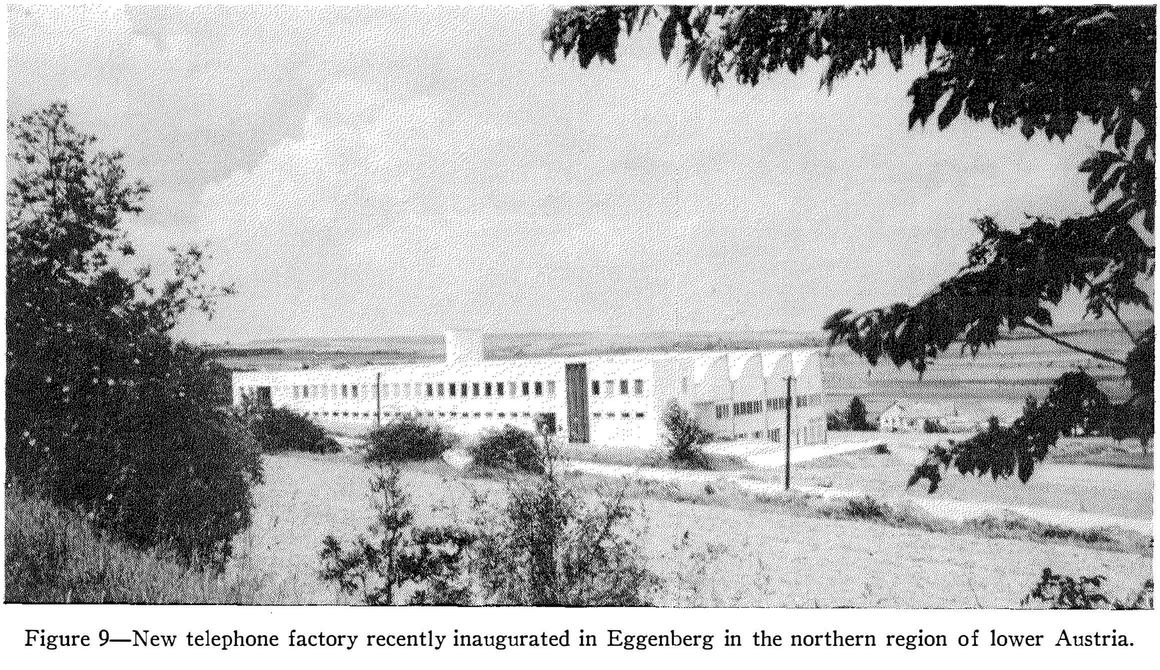


Figure 9—New telephone factory recently inaugurated in Eggenberg in the northern region of lower Austria.

**Standard Elektrik Lorenz Installs 10 000th Crossbar Private Automatic Branch Exchange**—In March 1963, Standard Elektrik Lorenz put in service its 10 000th crossbar private automatic branch exchange, which is in the new building of the central fire station of Ludwigshafen, Germany. This is the 11th installation of its kind for that city administration.

All private automatic branch exchanges having more than 10 lines are now supplied with crossbar switches by Standard Elektrik Lorenz, which has supplied crossbar equipment having a capacity of 650 000 lines to customers all over the world. Precious-metal contacts provide good noisefree connections; short travel of contacts permits high-speed, vibrationless, quiet operation; and the simple relay type of design ensures reliability, long life, and minimum maintenance.

*Standard Elektrik Lorenz*
*Germany*

**Primrose Telephone Subscriber Set**—Available in several colors to harmonize with any surroundings, the Primrose telephone subscriber set, with its illuminated dial, oblong shape, and depth of only 95 millimeters (3.7 inches), is particularly suited for bedrooms and hotels. It is shown in Figure 10.

The sound volume of the built-in ringer is adjustable. The lightweight handset is equipped with a high-quality microphone and receiver and can be furnished with a transmission equalizing circuit. The Primrose set has been approved by the Belgian telephone administration.

Figure 10—Primrose subscriber set with illuminated dial.

*Bell Telephone Manufacturing Company*
*Belgium*

**Gallium-Arsenide Crystals and Lasers**—Gallium-arsenide is a second-generation semiconductor suitable for solid-state electronic applications. Compared to silicon and germanium, its greater carrier mobility promises response at higher frequencies or operation at higher powers. Its wider energy gap should permit operation at increased temperatures and reduced leakage at reverse-biased junctions. Its recent use as a laser indicates new properties not exhibited by silicon and germanium.

For laser operation, gallium-arsenide must contain large and carefully controlled amounts of impurities such as selenium and silicon. By the beginning of 1963, single crystals having the required level of impurity and unusually high crystal perfection had been grown. Dislocation density was below 1000 per square centimetre and carrier mobility was particularly high, being above 1000 square centimetres per volt-second at $10^{19}$ carriers per cubic centimetre and 2500 square centimetres per volt-second at $10^{18}$ carriers per cubic centimetre. This material is particularly well suited for laser preparation because of its great homogeneity. A further finding was a method of growing from the melt $p$-type gallium-arsenide of the extremely high resistivity of $10^8$ to $10^9$ ohm-centimetres.

*Standard Telephones and Cables*
*United Kingdom*

**Blind Telephone Operators**—In Western Germany, about 2000 blind persons are working as telephone operators. Despite the complexity of the modern telephone exchange, they perform their duties effectively with the aid of finger guide bars, a few indications in Braille characters, and audible rather than visual signals. A typical switchboard is shown in Figure 11.

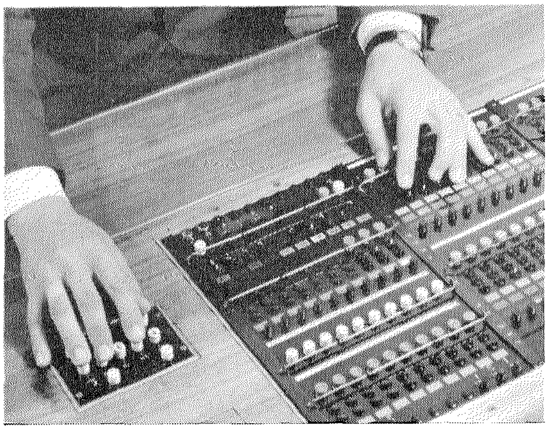


Figure 11—Telephone switchboard for the blind.

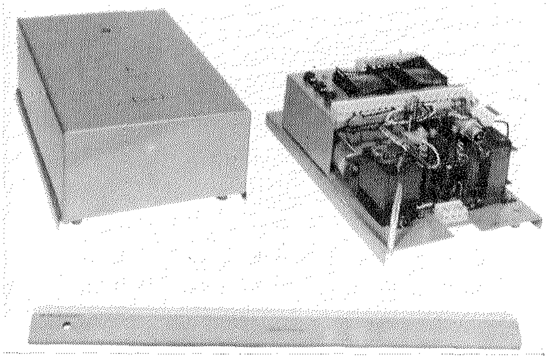


Figure 12—Magnetically stabilized rectifier-filter power supply for telephone subscriber sets.

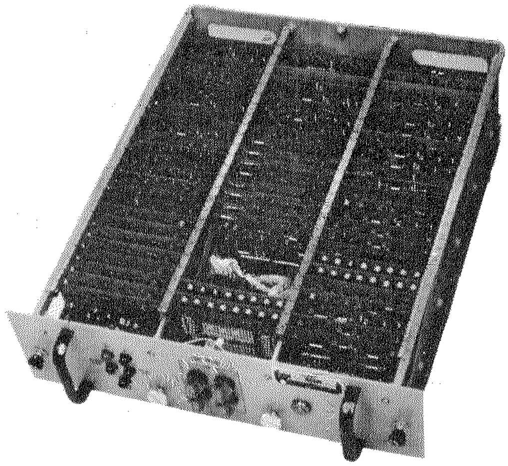


Figure 13—Line store for 256 characters operating at 10 000 bits per second.

A report based on 30 years of experience in designing telephone equipment for the blind is now available. Requests for brochure 0130/364 should be addressed to Standard Elektrik Lorenz AG, 42 Hellmuth-Hirth-Strasse, Stuttgart Zuffenhausen, Germany.

*Standard Elektrik Lorenz*
*Germany*

**Hospital Intercommunication Systems**—The modernization of hospital intercommunication systems has greatly expanded the services they provide. The early push button to call an attendant has been replaced by a microphone-loudspeaker permitting the patient to converse with the nurse or with an operator during the night when calls are concentrated. The same loudspeaker may be used for music.

Recent improvements include the use of unbreakable housings for indicator lamps, improved acoustic properties of the microphone-loudspeaker units, printed wiring on individual boards for fewer interruptions in service, control switchboards using printed-circuit panels of washable materials, and standard equipment racks with plug-in drawers for transistor amplifiers and oscillators, groups of relays, and common circuits.

*Bell Telephone Manufacturing Company*
*Belgium*

**Battery Substitute for Telephone Operation**—Figure 12 shows a battery substitute operating from the 220-volt 50-cycle-per-second power mains that produces 25 volts at 0.7 ampere of direct current for speech and 75 alternating volts for ringing of telephone subscriber sets. Magnetic stabilization makes it insensitive to mains variations of ±10 percent and to load variations. Its dimensions are 375 by 240 by 142 millimeters (14.8 by 9.4 by 5.6 inches) and a transistor version reduces this volume to less than half.

*Nederlandsche Standard Electric Maatschappij*
*The Netherlands*

**Solid Tantalum Capacitors**—The British Ministry of Aviation has issued its first such certificate for a design of sintered-anode tantalum capacitors. They have high capacitance per unit volume and may be used over a temperature range from − 55 to + 125 degrees centigrade. They are the only type of electrolytic capacitor capable of being hermetically sealed.

*Standard Telephones and Cables*
*United Kingdom*

**Magnetic-Core Line Store**—The line store shown in Figure 13 was developed for command and control systems. Up to 256 characters may be stored on magnetic cores at a rate of 10 000 bits per second with character-by-character output. It may be used as a random-input device and as a real-time store.

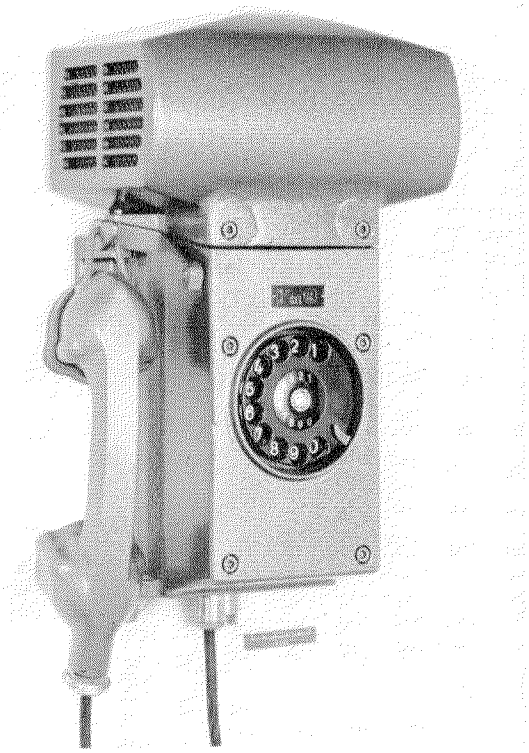


Figure 14—Telephone set for installation in mines and other places where explosive gases may be present.

Unusual features include visual operating and fault indicators, message sequence counting, and nondestructive readout. Internal circuits detect errors in data information and malfunctioning of the logic circuits.

*ITT Kellogg Communications Systems*
*United States of America*

**Safe Telephone Set**—A special telephone set has been developed for use in coal mines and other hazardous places where combustible gases can be ignited by electric sparks. The energy flow, either in normal or fault conditions, is held to so low a value that the opening and closing of electric circuits will not ignite explosive gases. No further safety provisions are required.

Operation in a noisy environment is aided by a shortened handset, which is equipped with a high-quality dynamic microphone and receiver, and a transistor amplifier that accentuates the high-frequency response. The set is shown in Figure 14.

*Standard Elektrik Lorenz*
*Germany*

**Laser and Infrared Scope**—The two boxes shown in Figure 15 house a gallium-arsenide laser and an image tube to convert the infrared output of the laser to visible radiation. This

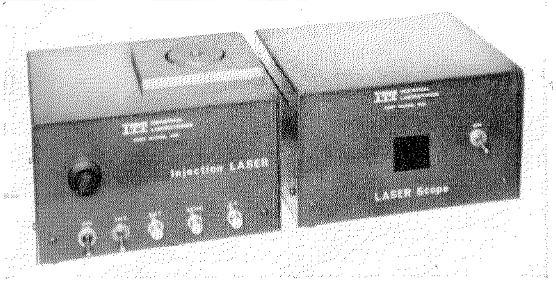


Figure 15—Gallium-arsenide laser and image-tube scope to produce visible light from the infrared laser beam.

compact self-contained battery-operated equipment is suitable for laboratory experimentation.

The laser is of the cleaved gallium-arsenide type and has a very-low operating threshold of 4 amperes. It radiates an average output power of 90 milliwatts at a wavelength of 0.874 micron. The beam is pulsed at an adjustable rate between 100 and 1000 cycles per second. A liquid-nitrogen cooling head extends from the top of the box and a thermocouple is provided to monitor the temperature of the laser.

The scope is an infrared image tube that permits visual and photographic inspection of the laser beam. It can be equipped with either objective or eyepiece lenses or with both for the study of enlarged images.

*ITT Industrial Laboratories*
*United States of America*

**Avalanche Rectifiers**—Avalanche rectifiers are now available that are capable of absorbing reverse voltage transients. A unit rated for 1.25 amperes is able to dissipate reverse transient energy at a 4-kilowatt rate without damage. Previously, silicon rectifiers have required special networks or selenium rectifiers as surge absorbers.

*Standard Telephones and Cables*
*United Kingdom*

## Transistors

Designated as "A Self-Instructional Programed Manual," this book was prepared under the editorship of I. L. Kosow by the training branch of Federal Electric Corporation, a subsidiary of International Telephone and Telegraph Corporation. The material is presented in the form of statements and questions, the answers to which follow immediately. Each answer is kept covered by the student until he has answered the associated question. If he is incorrect, the student should review enough to discover his error before going to the next question. The immediate testing of the understanding of the student is an important element in this new written teaching method of programed self-instruction.

The book is divided into the following 7 parts that are in turn subdivided into a suitable number of sets, each containing one major idea or circuit description.

1. Transistor Theory
2. Transistor Symbols
3. Rectification and Amplification
4. Oscillation
5. Modulation-Demodulation
6. Transistor Pulse and Switching Circuits
7. Logic and Gating Circuits

The page size is 9¼ by 6½ inches (23.5 by 16.5 centimeters) and the book contains 430 pages. It is published by Prentice-Hall, Englewood Cliffs, New Jersey, at $12.00 per copy.

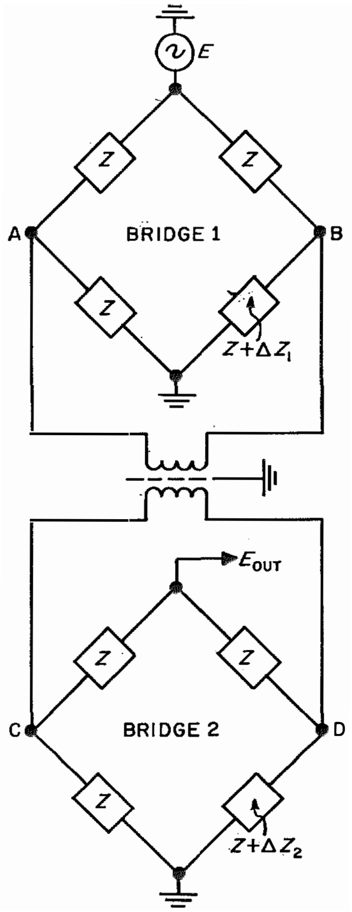
# Analog Signal Multiplier

O. S. MEIXELL

**ITT Federal Laboratories, A Division of International Telephone and Telegraph Corporation; Nutley, New Jersey**

## 1. Introduction

Communications and electronics are challenging fields for the application of a real-time analog signal multiplier. This is particularly true if the multiplier can furnish the true product of time-varying signals simultaneously with their variations. The multiplier described in this paper has been designed to achieve these objectives as closely as possible.



$$Z + \Delta Z_1 = Z(1 + \Delta Z_1/Z) = Z(1 + \mathcal{E}_1)$$
$$Z + \Delta Z_2 = Z(1 + \Delta Z_2/Z) = Z(1 + \mathcal{E}_2).$$

Thus, $\mathcal{E}_1$ and $\mathcal{E}_2$ are fractional unbalances of their respective bridges.

Figure 1—Basic multiplier. $E$ is the alternating-current carrier excitation.

## 2. Basic Principles

The basic multiplier comprises two equiarm impedance bridges in cascade, with the unbalance or detector output of the first bridge applied as excitation for the second. This is illustrated in Figure 1.

The output $E_{AB}$ of the first bridge is given closely by $k_1\mathcal{E}_1E$, where $k_1$ is a proportionality constant. Assuming an ideal unity-ratio transformer, this becomes the excitation for the second bridge, $E_{CD} = k_1\mathcal{E}_1E$. Correspondingly, the output of the second bridge is defined by $k_2\mathcal{E}_2E_{CD}$. Thus, $E_{\text{out}} = k_1k_2\mathcal{E}_1\mathcal{E}_2E$.

This relation shows that, for a given alternating-current voltage applied to the first bridge, the output of the second bridge is proportional to the product of the two fractional unbalances, $\mathcal{E}_1$ and $\mathcal{E}_2$. If these unbalances are caused by electrical or mechanical deviations, then the output of the second bridge is also proportional to the product of these deviations. For each bridge, the departure from linearity is only about half the fractional impedance unbalance.

A voltage-adjustable capacitor, or parametric diode, is an excellent means to implement the foregoing principles. The desired change of impedance $\Delta Z$ can be obtained merely by applying a signal voltage. This is accomplished over a very-wide frequency range at minimum cost of signal energy.

### 2.1 FUNDAMENTAL CIRCUIT

Two impedance bridges are used. They should be highly sensitive, with practically dissipationless elements that exhibit a conjugate-reactance relation at the bridge diagonal points. Figure 2 shows the fundamental circuit [1]. Obviously, there is no output voltage $E_{\text{out}}$ when $E_1$ and $E_2$ are both zero. This is also true if either $E_1$ or $E_2$ is zero. However, the simultaneous presence of both voltages causes two incremental capacitances $\Delta C_1$ and $\Delta C_2$ to appear. Thus

$$\begin{aligned} C_v &= C_{vo} + \Delta C_1 \\ &= C_{vo}(1 + \Delta C_1/C_{vo}) \\ &= C_{vo}(1 + \mathcal{E}_1) \end{aligned}$$

for bridge 1, and

$$\begin{aligned} C_v &= C_{vo} + \Delta C_2 \\ &= C_{vo}(1 + \Delta C_2/C_{vo}) \\ &= C_{vo}(1 + \mathcal{E}_2) \end{aligned}$$

for bridge 2.

For peak analog voltages not exceeding about 10 percent of the direct-current voltage used to bias the voltage-adjustable capacitors, the fractional unbalances $\mathcal{E}_1$ and $\mathcal{E}_2$ vary closely with $E_1$ and $E_2$, respectively. As will be shown, $E_{\text{out}}$ is therefore proportional to $Q^2E_1E_2E_c$, where $E_1$ and $E_2$ are the analog voltages, $E_c$ is the carrier voltage, and $Q = \omega_o L/R_s$ with $\omega_o = 2\pi f_o$.

Thus, for a constant carrier voltage, the output is proportional to the product of $E_1$ and $E_2$.

In practice, a portion of the carrier voltage is applied, with proper phase, in series with $E_{\text{out}}$. This becomes an amplitude-modulated wave with a carrier frequency $f_o$ and an envelope that is the product of $E_1$ and $E_2$. By demodulating the wave, the true instantaneous product of $E_1$ and $E_2$ is obtained. $E_1$ and $E_2$ may be any type of signal within the bandwidth capabilities of the two bridges. Thus, the products of sine waves, nonsinusoidal waves with or without direct-current components, square waves, pulses, noise, et cetera, can be formed.

## 2.2 COMPARISON WITH OTHER MULTIPLIERS

The product of two analog voltages is formed in essentially linear fashion. This is significant because the respective bridge unbalances caused by these voltages result from differential excursions over the voltage-adjustable capacitor's characteristic. In fact, the smaller the analog voltages are, the more linear is the formation of their product. As in other analog devices, there is a limit to the dynamic range that can be accommodated. The present model exhibits a 40-decibel range in its product. By refinement of the technique for obtaining quiescent (no-signal) bridge balance and by other means, this range can be extended materially. In contrast, the quarter-square technique [2] requires diodes or vacuum tubes that exhibit a square-law characteristic as closely as possible. Also, undistorted multiplication can take place only when the instantaneous sum or difference of the two analog signals is within the square-law region.

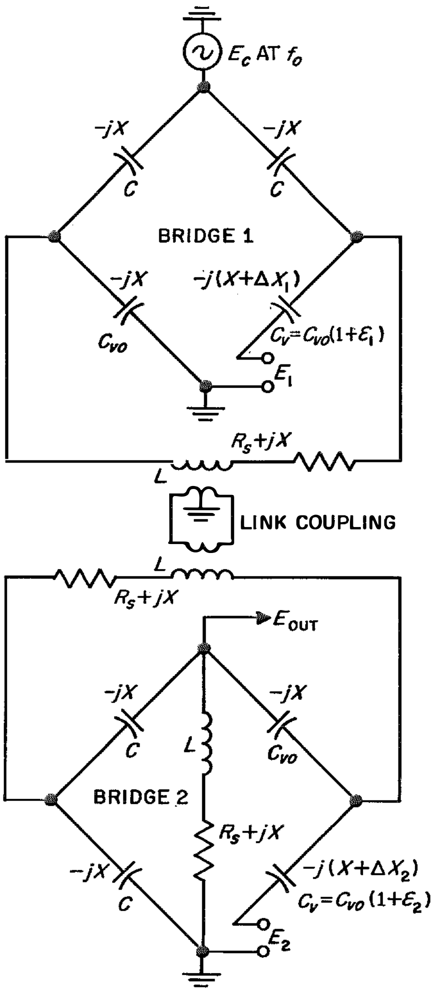


Figure 2—Multiplier with low-loss elements. $C_v$ is a voltage-adjustable capacitor whose no-signal value $= C_{vo}$. $C$ is the arm capacitance $= C_{vo}$. Therefore, both bridges are balanced when $E_1 = E_2 = 0$.

Multipliers using the Hall effect [3], [5], [7] require conversion into a high current for one analog voltage and a high magnetic field for the other. The latter conversion imposes a severe bandwidth limitation on the formation of the desired product.

Mechanical and electronic computers are available that perform multiplication and other mathematical operations accurately. The office calculating machine provides one-quadrant multiplication but not on a real-time or high-bandwidth basis. Electronic computers can perform four-quadrant multiplication. However, they are complex and expensive, and their signal bandwidth is limited because of the use of time-sharing and storage techniques, or for other reasons.

The analog signal multiplier presently does not have computer-like dynamic range and precision. However, it has unique advantages in its simplicity and ability to effect a four-quadrant low-noise product of complicated time functions over a large bandwidth.

### 2.3 ANALYTICAL CONSIDERATIONS

It might be very hard to develop and appraise a quantitative analysis of the multiplier output for generalized forms of $E_1$ and $E_2$. However, a set of conditions can be postulated that greatly simplifies the task and yet provides a representative picture of quantitative relations. Refer to the Appendix for an analysis based on the simplified assumption that the two analog voltages $E_1$ and $E_2$ are both sinusoidal, are at the same frequency, and differ only in amplitude and relative phase. Thus

$$e_1 = E_{1m} \sin \omega_s t$$

$$e_2 = E_{2m} \sin (\omega_s t + \theta_s)$$

where

$\omega_s = 2\pi f_s$

$f_s$ = signal frequency

$\theta_s$ = phase angle between $E_1$ and $E_2$.

When a carrier voltage $e_c = E_{cm} \sin \omega_o t$ is applied as excitation for the first bridge, the instantaneous output of the second bridge is described by

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} [2 \sin (\omega_o t) \cos \theta_s - \sin (\omega_o t - 2\omega_s t - \theta_s) - \sin (\omega_o t + 2\omega_s t + \theta_s)] \quad (1)$$

where $k_o$ is a proportionality constant.

Two special conditions are instructive. The first is when $E_1$ and $E_2$ are in phase, and the second is when they are in quadrature.

For the in-phase condition $\theta_s = 0$, (1) becomes

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} [2 \sin \omega_o t - \sin (\omega_o t - 2\omega_s t) - \sin (\omega_o t + 2\omega_s t)]. \quad (2)$$

A 100-percent amplitude-modulated wave is described in (2). Its modulation envelope represents the instantaneous product of the two analog voltages, which can be restored by demodulation of the wave. Note that the frequency of this product is twice that of the analog voltages.

For the quadrature condition $\theta_s = -90$ degrees, (1) becomes

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} \times [- \sin (\omega_o t - 2\omega_s t + \pi/2) - \sin (\omega_o t + 2\omega_s t - \pi/2)]. \quad (3)$$

Equation (3) shows that only the sideband components of the wave exist. The envelope no longer describes the instantaneous product of the two analog voltages. To do so, a sufficiently strong fixed carrier component must be added to $e_{\text{out}}$ before demodulation. If this fixed carrier component is in proper phase (90 degrees with respect to the impressed carrier), the envelope then becomes the true instantaneous product of $E_1$ and $E_2$ regardless of their relative phase.

The foregoing quantitative relations are illustrated in Figure 3 by waveforms derived from the multiplication of two sinusoidal analog voltages having the same amplitude and frequency but differing in relative phase (0, 90, and 180 degrees). Row *A* shows the analog voltages and row *B* the corresponding outputs of the multiplier. These outputs are pictorial representations of (1). However, the output shown for the in-phase condition of $e_1$ and $e_2$ is described by (2) and for the 90-degree condition by (3).

If the multiplier outputs are rectified at this point, the resultant waveforms in row *C* would be false products except for the fortuitous case where $e_1$ and $e_2$ are in phase. To prevent this, a fixed carrier component must be added before demodulation. This results in the waveforms in row *D*. The carrier level shown is minimum; a higher level is advisable in practice, to improve rectification efficiency.

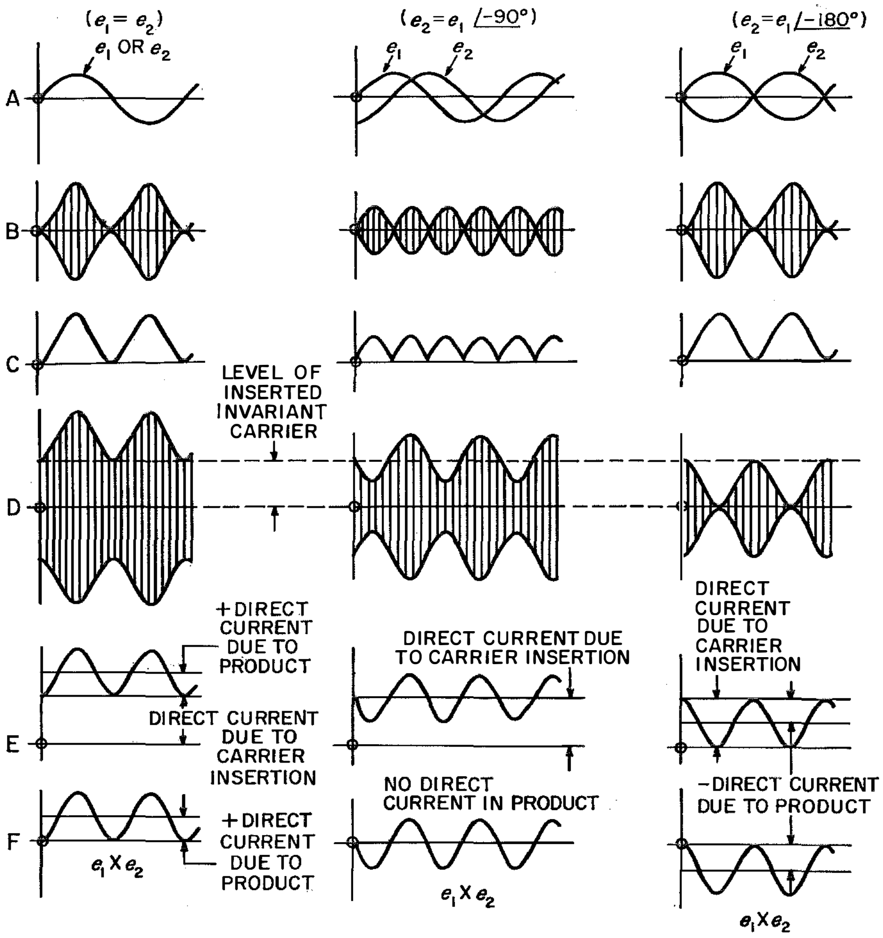


Figure 3—Analog signal multiplier waveforms.

*A* = analog voltages for 3 different phase relations.

*B* = products before insertion of invariant carrier.

*C* = demodulated false products.

*D* = products with inserted invariant carrier.

*E* = demodulated products.

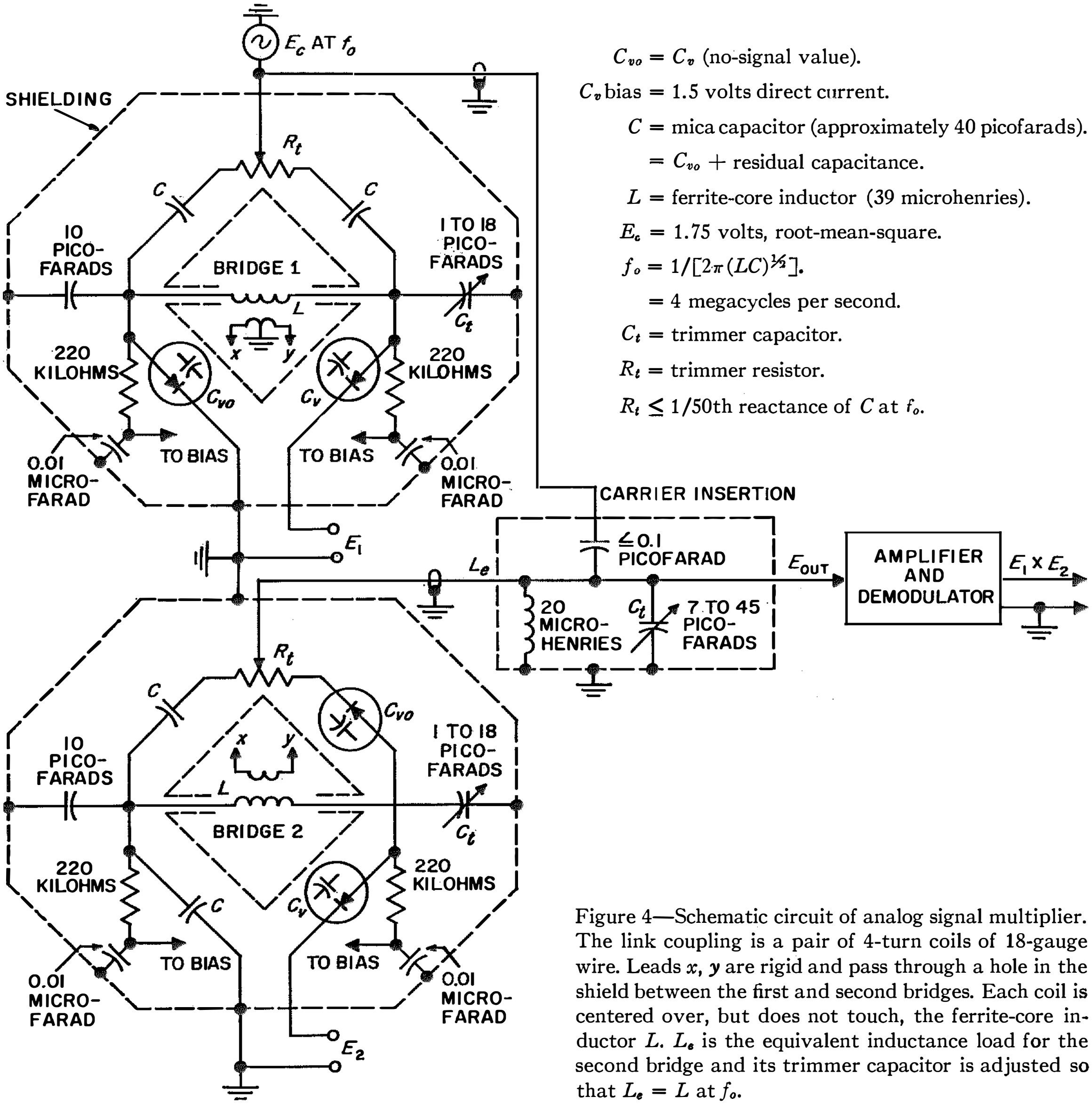
*F* = true products with direct-current voltage due to invariant carrier removed.

Row *E* shows that the demodulation now provides the product of the original analog voltages raised above a fixed direct-current level. Row *F* shows the true instantaneous products after the fixed level has been removed.

Note that row *F* represents the instantaneous power function when $e_1$ is proportional to the voltage applied to a load and $e_2$ is proportional to the current.

### 3. Design of Developmental Model

Figures 4 and 5 show, respectively, the schematic diagram and a photograph of the developmental model of the analog signal multiplier.



$C_{vo} = C_v$ (no-signal value).
$C_v$ bias = 1.5 volts direct current.
$C$ = mica capacitor (approximately 40 picofarads).
$= C_{vo}$ + residual capacitance.
$L$ = ferrite-core inductor (39 microhenries).
$E_c$ = 1.75 volts, root-mean-square.
$f_o = 1/[2\pi(LC)^{1/2}]$.
= 4 megacycles per second.
$C_t$ = trimmer capacitor.
$R_t$ = trimmer resistor.
$R_t \leq 1/50$th reactance of $C$ at $f_o$.

Figure 4—Schematic circuit of analog signal multiplier. The link coupling is a pair of 4-turn coils of 18-gauge wire. Leads *x*, *y* are rigid and pass through a hole in the shield between the first and second bridges. Each coil is centered over, but does not touch, the ferrite-core inductor *L*. $L_e$ is the equivalent inductance load for the second bridge and its trimmer capacitor is adjusted so that $L_e = L$ at $f_o$.

It has four major connections. One is for application of the carrier voltage $E_c$ to the first reactance bridge, located in the lower half of the structure. The analog voltage $E_1$ is also applied to this bridge. Through the action of the voltage-adjustable capacitor* in series with $E_1$, a suppressed-carrier modulation appears across the primary winding $L$ of a special transformer which link-couples this output to the input of the second reactance bridge. The over-all coupling is made low and thus the inductance of the primary winding is virtually unchanged relative to its unloaded value. Furthermore, the carrier frequency $f_o$ is such that the reactance of this high-$Q$ winding is the conjugate of the reactance presented to it by the four capacitance arms of the bridge. Thus, under the quiescent and balanced condition $E_1 = 0$, if $+X$ is the reactance of $L$, each of the four capacitance arms has a reactance of $-X$. In simpler terms, this corresponds to a resonant condition of $L$ and $C$ at the carrier frequency.

The other analog voltage $E_2$ is correspondingly applied to the second of the reactance bridges, located in the upper half of the structure. Through its action on the associated Varicap, the suppressed-carrier modulated voltage coupled to the input of this bridge is again modulated to become $E_{out}$.

For sinusoidal analog voltages at the same frequency, instantaneous $E_{out}$ appears as shown in row $B$ of Figure 3. Adding the original invariant carrier voltage at the proper 90-degree phase results in the waveform of row $D$. This envelope displays the true instantaneous product of $E_1$ and $E_2$ for the conditions postulated.

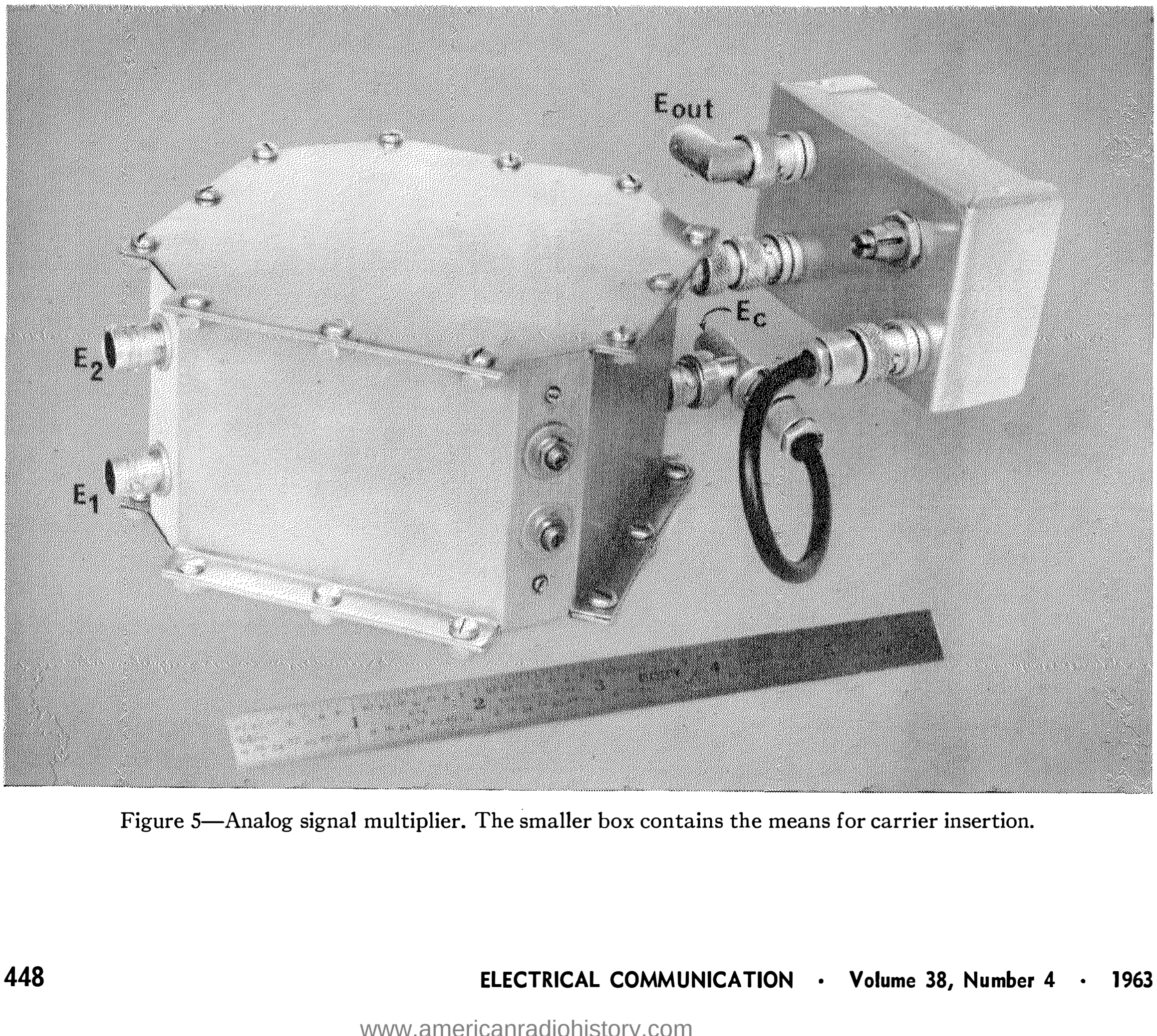


Figure 5—Analog signal multiplier. The smaller box contains the means for carrier insertion.

---

* Varicap, type *PC-116-22*, a product of Pacific Semiconductors, Incorporated.

The method of coupling the output of the first bridge into the second is especially important. The over-all coupling must be strictly inductive and on a balanced-to-ground basis. There can be no capacitive or longitudinal coupling because a capacitance change caused by the analog signal of one bridge would also appear as a change in the other. The link coupling solves this problem since its impedance is so low that the induced voltage gradient across the winding is practically zero. When center tapped and grounded, as shown in Figure 4, the link-coupled windings constitute a Faraday shield that effectively prevents any capacitive coupling between the output of the first bridge and the input of the second.

An important design feature is the means used for dynamic compensation to improve linearity and stability of performance. Figure 4 shows an additional Varicap in the arm contiguous with that containing each signal-controlled Varicap. Note that in the first bridge half of the impressed carrier voltage $E_c$ is impressed cophasally on each Varicap. Since these are identically poled, the variations in their capacitances caused by the carrier voltage are essentially equal and of like sign during its positive and negative excursions. Thus, these capacitance variations cause no unbalance of the bridge. Unbalance can occur only when $E_1$ is present. The second bridge works similarly and can only be unbalanced when $E_2$ is present.

Stability is also improved by the use of these additional Varicaps. Since they are all of the same type and characteristics, capacitance variations caused by temperature changes are of like sign and magnitude, and quiescent balance is maintained.

The source impedances of $E_1$ and $E_2$ should be low at $f_o$ compared to the reactance of $C$ at $f_o$ and should offer a direct-current return path. If these requirements are not met, suitable shunt capacitance and resistance must be applied at the analog-signal input terminals. When a common analog source is used, as for obtaining the square of a signal, two aperiodic voltage dividers should be used between the source and the analog-signal input terminals for decoupling purposes.

The multiplier equivalent circuit in the Appendix shows that the permissible bandwidth of the analog signals is a function of the equivalent $Q$ value of the inductors used in the link coupling and the output. It is also proportional to the carrier frequency. Since the intrinsic $Q$ value of the ferrite-core inductors is high, suitable series or shunt resistance can be used to lower it, although at the expense of sensitivity. For the present model, the $Q$ value of the inductive elements has been reduced to about 10 at the 4-megacycle-per-second carrier frequency. A product bandwidth has thus been observed from direct current to 400 kilocycles per second, which is 10 percent of the carrier frequency. It appears possible to modify the existing structure for operation with a 40-megacycle-per-second carrier and achieve a tenfold extension of the product bandwidth, to 4 megacycles per second.

Several design principles are useful regardless of the carrier frequency chosen. First, choose a voltage-adjustable capacitor whose $Q$ value is high at the carrier frequency $f_o$. Use the lowest possible negative bias to obtain the largest usable quiescent capacitance $C_{vo}$, remembering that this bias should not be less than about half the carrier peak maximum $E_c/2$ plus the peak analog signal. The largest $C_{vo}$ provides the highest possible sensitivity in the presence of unavoidable residual capacitances to ground. The arm capacitance $C$ is the sum of the quiescent and residual arm capacitances. This establishes the inductance $L$ in the output arm of the first bridge and the input arm of the second in accordance with $L = 1/\omega_o^2 C$.

Second, the mechanical layout of each of the constituent bridges should be symmetrical. All arms must be electrically isolated from each other by shielding. Also, the detector arm of the first bridge and the input arm of the second should be isolated carefully from the bridge arms containing the fixed and voltage-adjustable capacitors. The second bridge must be isolated from the first except for the balanced link coupling which should be mechanically and electrically symmetrical.

Third, the output of the multiplier should be amplified 20 to 40 decibels before demodulation. This can be accomplished best by a band-pass amplifier with a center frequency at $f_o$ and a bandwidth sufficient to accommodate the sidebands in the product of $E_1$ and $E_2$.

Most signal sources are single-ended. Also, a usable direct-current component may reside in the analog signal. The present model is adapted to these conditions. However, if balanced-to-ground or push-pull analog signals are used, distortion can be reduced by applying $E_1$ and $E_2$ suitably to both Varicaps in their respective bridges. Of course, if transformers are used for this purpose, any direct-current signal components are lost and the bandwidth is limited to that of the transformer.

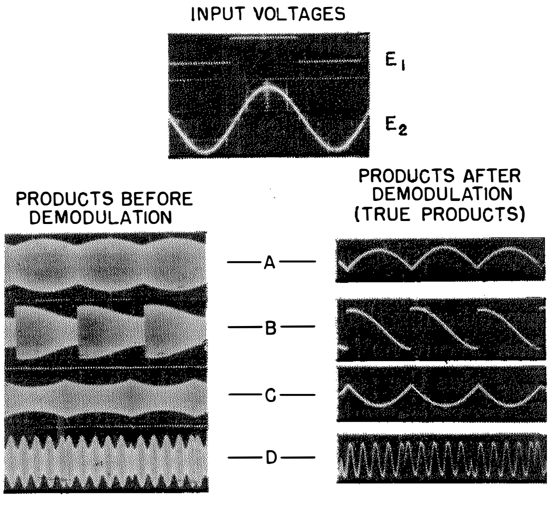


Figure 6—Products of rectangular and sine waves. $E_1$ and $E_2$ are the input signals at 300 cycles per second. Their amplitudes are 0.16 and 0.30 volt, respectively, peak to peak. The carrier is 1.75 volts (root-mean-square) at 4.1 megacycles per second. Oscilloscope sweep = 500 microseconds per centimeter.

$A$ = $E_1$ and $E_2$ in phase.

$B$ = $E_2$ advanced 90 degrees.

$C$ = $E_2$ advanced 180 degrees.

$D$ = $E_2$ increased from 300 to 3000 cycles per second

## 4. Multiplier Performance

Figures 6 through 9 illustrate typical multiplier output waveforms for various analog signal voltages. Figure 6 shows the products of a rectangular wave and sinusoids. Note that discontinuities are cleanly formed and little distorted, confirming the high bandwidth capability and good dynamic range.

Figure 7 shows the products of sinusoidal analog voltages. The first group is a superimposed display of the products of two 300-cycle-per-second sine waves at relative phase displacements of 0, 90, and 180 degrees. Observe that these are replicas of the waveforms shown in row $F$ of Figure 3. Further evidence of bandwidth and dynamic range capability is provided by the two remaining displays of Figure 7, where the frequency of the second sinusoid is increased 10 times and then 100 times above the first, with no change in amplitude.

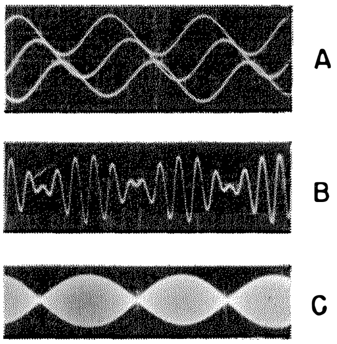


Figure 7—Products of sine waves. The amplitude of each voltage is 0.30 volt peak to peak. Oscilloscope sweep = 500 microseconds per centimeter. At $A$, the top curve is the product for $E_1$ and $E_2$ in phase. The middle and bottom curves of $A$ are for $E_2$ shifted 90 and 180 degrees, respectively. $E_1$ and $E_2$ are each at 300 cycles per second and the displays are superimposed.

At $B$ and $C$, are the products of $E_1$ at 300 cycles per second and $E_2$ at 3000 and 30 000 cycles per second, respectively.

The square of a triangular wave, shown in Figure 8, consists of contiguous parabolas. Also, the bottom waveform in this figure demonstrates the capability to multiply two signals each having a direct-current component (in this case, equal to half the peak-to-peak value of the analog).

Figure 9 demonstrates the capability to obtain the product of two identical noise functions; first as instantaneously cophasal and, second, in an instantaneous antiphasal relationship. Observe that the product for the cophasal condition contains a positive direct-current component, while the product for the antiphasal condition contains a negative one.

## 5. Multiplier Applications

One of the most important applications of an analog multiplier is as a correlator to improve the signal-to-noise ratio of various types of signals. For example, when a noisy signal, $f(t)$ + noise, is multiplied by a noise-free replica that periodically passes through time coincidence, sufficient integration can reveal the presence of the signal even when it is deeply immersed in noise. By using this cross-correlation technique [4], the analog signal multiplier has detected sinusoidal signals having a signal-to-noise ratio of 1/30. Of course, the bandwidth of the signal information that can be restored by this process suffers with increased integration time. The signal-to-noise ratio can be improved materially for noisy pulse-like signals when they

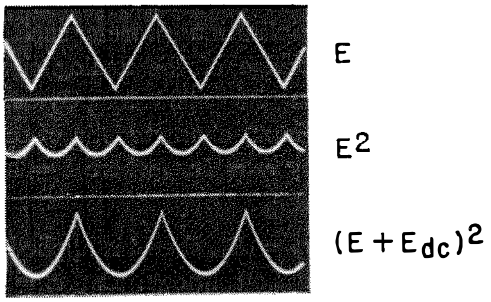


Figure 8—Square of a triangular wave. $E$ = 920 cycles per second at 0.07 volt peak. $E_{dc}$ = peak value of $E$.

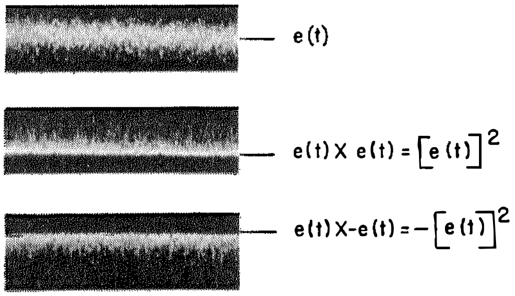


Figure 9—Products of white noise over a band from zero to 20 kilocycles per second.

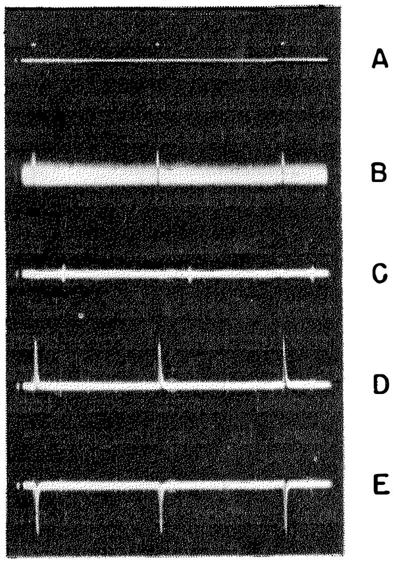


Figure 10—Products of noisy and noise-free pulses. Pulses are of 10-microsecond duration with a repetition frequency of 1000 per second. Oscilloscope sweep = 200 microseconds per centimeter. The light output of the oscilloscope was integrated over a 0.1-second exposure for each display on Pola Pan 200 (Type 42) film. The lens was adjusted to f5.6. Electrical time constant is about 22 microseconds.

$A$ = signal pulses $E_1$ at 0.03 volt. (The correlating pulses $E_2$ at 0.3 volt are not shown.)

$B = E_1 + 0.01$ volt root-mean-square white noise from zero to 500 kilocycles per second.

$C = E_1 + \text{noise} \times E_2$, not correlated.

$D = E_1 + \text{noise} \times E_2$, correlated.

$E = -E_1 + \text{noise} \times E_2$, correlated.

are multiplied by time-coincident noise-free pulses. This is illustrated in Figure 10. Here the signal-to-noise improvement is obtained with only a moderate amount of integration. Additional improvement should result from cross-correlating the product in a second multiplier.

An analog multiplier can be employed as a cross-correlator for signal recognition or analysis. The use for signal recognition contemplates the formation of the product of an incoming signal with voltages having various normalized trial waveforms. On time-sweeping with these waveforms and integrating the various products, the largest of the maxima identifies the waveform that is closest to the unknown. Williams and Denker [5] describe the application of a Hall-effect multiplier for analysis of a nonsinusoidal wave. It appears that the bridge-type multiplier can implement this analysis technique more readily and, certainly, over a greater bandwidth.

Each bridge of the analog signal multiplier constitutes a low-level linear modulator. The device (in toto) offers a means for a wide variety of modulations. Before delineating these in detail, it will be helpful to compare more-conventional modulation techniques with those used in the bridge-type analog multiplier.

It is common practice to obtain amplitude modulation of a carrier by applying the carrier and signal voltages as inputs to a circuit that has a nonlinear output-input characteristic. For various types of vacuum-tube, transistor, or diode modulators, this relation may be represented by the series expansion

$$e_o = k_1 e + k_2 e^2 + k_3 e^3 + \cdots + k_n e^n$$

where $e_o$ is the output, $e = e_c + e_s$ (the carrier-plus-signal input), and $k_1$ through $k_n$ are amplitude coefficients.

Let the first three terms suffice to describe a typical nonlinear characteristic. Substituting $e = e_c + e_s$, we obtain

$$\begin{aligned} e_o = k_1 e_c + k_1 e_s + k_2 e_c^2 + 2k_2 e_c e_s \\ + k_2 e_s^2 + k_3 e_c^3 + 3k_3 e_c^2 e_s \\ + 3k_3 e_c e_s^2 + k_3 e_s^3. \quad (4) \end{aligned}$$

The terms $k_1 e_c + 2k_2 e_c e_s$ in (4) afford undistorted amplitude modulation for a modulation index of less than unity. The other terms are not wanted since they include modulation of the carrier by a square of the signal, modulated and unmodulated second and third harmonics of the carrier, et cetera. Normally, most of these are removed by the use of a band-pass filter that accommodates only the desired carrier frequency and sideband components.

Assuming $e_c = E_{cm} \sin \omega_c t$ and $e_s = E_{sm} \sin \omega_s t$, then

$$\begin{aligned} e_o &= k_1 e_c + 2k_2 e_c e_s \\ &= k_1 E_{cm} \sin \omega_c t + 2k_2 E_{cm} \sin \omega_c t E_{sm} \sin \omega_s t. \quad (5) \end{aligned}$$

This result can also be achieved more directly by a linear multiplication process such as occurs in either bridge of the analog signal multiplier. Within its allowable dynamic range, the output-input characteristic of each bridge is defined by $e_o = k e_c e_s$ where $k$ = amplitude coefficient.

Assume a carrier $e_c = E_{cm} \sin \omega_c t$ is applied as excitation for the first bridge. Let $e_s = e_{dc} + E_{sm} \sin \omega_s t$ be applied to the analog signal port of this bridge (see Figure 4). Thus

$$\begin{aligned} e_o &= k E_{cm} \sin \omega_c t (e_{dc} + E_{sm} \sin \omega_s t) \\ &= k e_{dc} E_{cm} \sin \omega_c t \\ &\quad + k E_{cm} \sin \omega_c t E_{sm} \sin \omega_s t. \quad (6) \end{aligned}$$

Except for the constant coefficients, (6) is the same as (5). Accordingly, undistorted modulation can be realized directly without the unwanted products that characterize usual practice. This can be accomplished at signal levels as low as a fraction of a volt. In fact, the lower the levels of carrier and signal, the more linear is the formation of the modulation. Any small extraneous products are removed by the band-pass action of the bridge network.

Table 1 summarizes the modulation and related capabilities of the analog multiplier when energized by a carrier voltage $E_c$.

This versatile modulation capability permits additional uses. For example, a version of the multiplier can serve as an electronic phase shifter of the carrier. Two bridges, identical in form with the first bridge, would be excited in parallel; the first by $E_c$ and the second by $E_c$ shifted 90 degrees. If $E \sin \alpha$ is applied to the $E_1$ port and $E \cos \alpha$ to the $E_2$ port, the vector sum of the outputs of the two bridges, taken in series, is the original carrier with a phase shift of $\alpha$. When $\alpha = \omega_s t$, it can be shown that the vector sum of the outputs consists solely of the lower-sideband modulation of $E_c$ and the vector difference consists solely of the upper sideband. This single-sideband modulation technique has been anticipated [6] using conventional types of modulators but use of the bridge structures for this purpose deserves consideration because they constitute low-level linear modulators having a minimum of complexity.

Waveform $A$ of Figure 6 obviously duplicates the full-wave rectification of a sinusoid by conventional means. The upper group of waveforms in Figure 7 suggests use of the multiplier either as a synchronous detector [7] or for the detector function in a synchrodyne type of receiver. The latter is treated in the literature under other designations such as homodyne and product-detector types of receivers.

## 6. Extension of Design

Size can be reduced advantageously in the present model and extension to higher carrier frequencies appears feasible, with further increase of allowable bandwidth in the product. For designs at carrier frequencies higher than 50 megacycles per second, the present Varicap is not suitable at the low bias employed. Other types of parametric diodes may have to be used singly or in parallel groups. The use of ultra-high-frequency and microwave carriers

TABLE 1

MULTIPLIER MODULATION CAPABILITIES

| Modulation or Process | Signals to be Applied | | Desired Modulation or Process Appearance |
|---|---|---|---|
| | at $E_1$ Port | at $E_2$ Port | |
| Modulation of $E_c$ by $E$: | | | |
| Suppressed carrier | $E$ | $E_{dc}$ | $E_{out}$ port with no inserted $E_c$, or by coupling to the input of the second bridge, which is not used. |
| Normal amplitude modulation | $E + E_{dc}$ | $E_{dc}$ | |
| Modulation of $E_1$ by $E_2$ or vice versa: | | | |
| Suppressed carrier | $E_1$ (or $E_2$) | $E_2$ (or $E_1$) | Output of amplifier-demodulator whose input is $E_{out}$ + inserted $E_c$. |
| Amplitude modulation of $E_1$ by $E_2$ | $E_1$ | $E_2 + E_{dc}$ | |
| Amplitude modulation of $E_2$ by $E_1$ | $E_1 + E_{dc}$ | $E_2$ | |
| Commutation or phase reversal of: | | | |
| $E_c$ (method 1) | ± $E_{dc}$ steps | $E_{dc}$ | Same as for modulation of $E_c$ by $E$. $E_{out}$ port with no inserted $E_c$. |
| $E_c$ (method 2) | $E_{dc}$ | ± $E_{dc}$ steps | |
| $E$ (method 1) | $E$ | ± $E_{dc}$ steps | Same as for modulation of $E_1$ by $E_2$. |
| $E$ (method 2) | ± $E_{dc}$ steps | $E$ | |
| On-off switching of: | | | |
| $E_c$<br>$E$ | Same as for commutation of $E_c$ and $E$, respectively, but with ± $E_{dc}$ step = 0 for off condition. | | |

will no doubt require distributed-parameter bridge elements. Extension of designs into these areas will lead to many more-challenging applications.

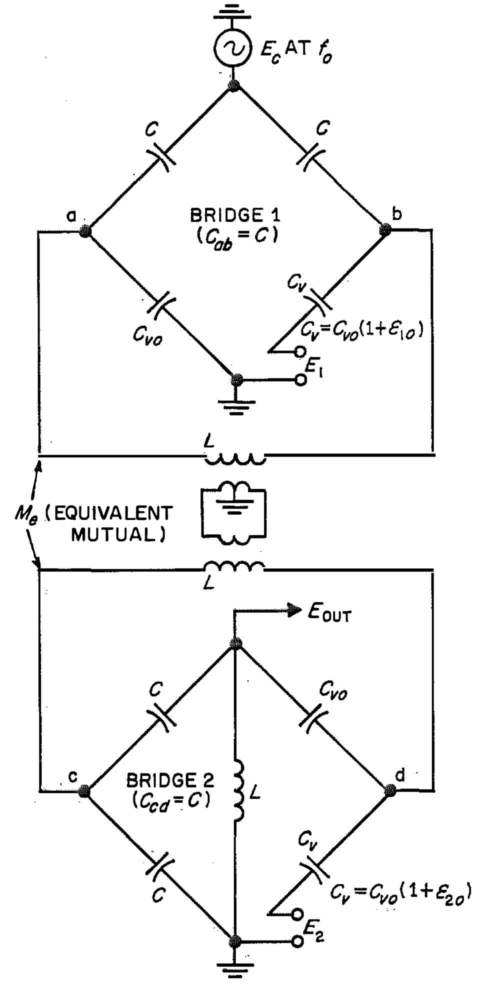


Figure 11—Multiplier equivalent circuit.

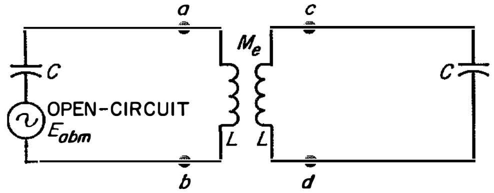


Figure 12—Thévenin's equivalent circuit.

## 7. Acknowledgements

The author greatly appreciates the encouragement and support given by Mr. H. N. Capen, Director, and Mr. W. A. Miner, Jr., Associate Director, of the Electronic Defense Laboratory. He is also indebted to Mr. G. Hulst and Mr. F. A. Modavis for several important applications whose classified nature precludes disclosure in this paper. The fundamental circuit of the multiplier was ably implemented by the design of Mr. W. E. Behrendt and by the construction of Mr. W. Hoag.

## 8. References

1. O. S. Meixell, "Analog Signal Multiplier," ITT Federal Laboratories report 62-327; September 1962.

2. S. Seely, "Electron-Tube Circuits," McGraw-Hill Book Company, New York, 1950; pages 152–153.

3. W. A. Scanga, A. R. Hilbinger, and C. M. Barrack, "Hall-Effect Multipliers," *Electronics*, volume 33, pages 64–67; 15 July 1960.

4. Y. W. Lee, "Statistical Theory of Communications," John Wiley & Sons, New York, 1960; pages 17–22 and 288–291.

5. J. D. Williams and S. P. Denker, "Fourier Analyzer Uses the Hall Effect," *Electronic Industries*, volume 20, pages 108–110; September 1961.

6. R. B. Dome, "Wideband Phase Shift Networks," *Electronics*, volume 19, pages 112–115; December 1946.

7. E. D. Sisson, "HALL-PAK Magnetic Circuit Applications," *Electrical Design News*, pages 38–50; February 1963.

## 9. Appendix

The multiplier equivalent circuit is shown in Figure 11.

$C_v$ = voltage-adjustable capacitance (no-signal value = $C_{vo}$)

$C$ = total arm capacitance
($C = C_{vo}$ + fixed residual capacitance)

$L$ = inductance arm

Carrier frequency $(f_o) = 1/[2\pi(LC)^{1/2}]$.

Let $\mathcal{E}_1$ be fractional unbalance of first bridge, then

$$\mathcal{E}_1 = \frac{\Delta C_1}{C} = \frac{\Delta C_1}{C_{vo}}\frac{C_{vo}}{C} = \mathcal{E}_{1o}\frac{C_{vo}}{C}.$$

$\Delta C_1$ is caused by $E_1$ and $C_{vo}/C$ takes residual capacitance into account.

For $\mathcal{E}_1 \leq 0.05$

$$\text{open-circuit } E_{ab} = \frac{\mathcal{E}_1 E_c}{4} \text{ (very closely)}$$

$$\text{open-circuit } E_{ab} = \frac{\mathcal{E}_{1o}}{4}\frac{C_{vo}}{C}E_c$$

$$\text{open-circuit } E_{abm} = \frac{\mathcal{E}_{1o}}{4}\frac{C_{vo}}{C}E_{cm}$$

where subscript $m$ = maximum value.

Characteristics of a Varicap are such that

$$\frac{\Delta C_1}{C_{vo}} = \mathcal{E}_{1o} = \frac{E_{1m}}{2E_{bo}} \text{ (closely for } E_{1m} \leq 0.1\ E_{bo})$$

where

$E_{1m}$ = maximum value of $E_1$

$E_{bo}$ = applied direct-current bias.

Accordingly,

$$\begin{aligned}\text{open-circuit } E_{abm} &= \frac{1}{8}\frac{E_{1m}}{E_{bo}}\frac{C_{vo}}{C}E_{cm}\\ &= K_1 E_{1m} E_{cm} \qquad (7)\end{aligned}$$

where

$$K_1 = \frac{1}{8}\frac{C_{vo}}{CE_{bo}}.$$

This leads to the Thévenin's equivalent circuit shown in Figure 12. For critical coupling $(1/Q)$ or a coupling up to twice this value (or $2/Q$)

$$E_{cdm} \doteq \frac{Q}{2}E_{abm}$$

where $Q$ is the $Q$ value of the identical primary and secondary inductors $L$ in the presence of coupling.

From (7)

$$\text{open-circuit } E_{abm} = K_1 E_{1m} E_{cm}$$

$$\therefore E_{cdm} = \frac{Q}{2}E_{abm} = K_1\frac{Q}{2}E_{1m}E_{cm}. \qquad (8)$$

With further application of Thévenin's theorem, it can be shown readily that the output of the second bridge is

$$E_{\text{out } m} = j\frac{Q}{8}\frac{E_{2m}}{E_{bo}}\frac{C_{vo}}{C}E_{cdm} = jQK_2E_{2m}E_{cdm} \qquad (9)$$

where

$Q$ = $Q$ value of load $L$ on second bridge

$$K_2 = \frac{1}{8}\frac{C_{vo}}{CE_{bo}}.$$

When the bridges are identical, with the same $Q$ values and with a common bias $E_{bo}$ applied to all Varicaps, $K_1 = K_2 = K$. Substituting (8) in (9), we obtain

$$E_{\text{out } m} = jK^2(Q^2/2)E_{1m}E_{2m}E_{cm}. \qquad (10)$$

Equation (10) shows that, for a constant $E_c$, the output of the second bridge is proportional to the product of $E_1$ and $E_2$.

Rewriting (10) in terms of instantaneous sinusoidal voltages

$$e_{\text{out}} = k_o e_1 e_2 e_c$$

where

$k_o = jK^2(Q^2/2)$

$e_1 = E_{1m} \sin \omega_s t$

$e_2 = E_{2m} \sin (\omega_s t + \theta_s)$

$e_c = E_{cm} \sin \omega_o t$

$\omega_o = 2\pi f_o$

$f_o$ = carrier frequency.

Thus

$$\begin{aligned}e_{\text{out}} = k_o E_{1m}E_{2m}E_{cm}&[\sin \omega_s t]\\ &\times [\sin(\omega_s t + \theta_s)][\sin \omega_o t].\end{aligned}$$

By straightforward trigonometric development

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} [2 \sin (\omega_o t) \cos \theta_s - \sin (\omega_o t - 2\omega_s t - \theta_s) - \sin (\omega_o t + 2\omega_s t + \theta_s)]. \quad (11)$$

When $e_1$ and $e_2$ are in phase ($\theta_s = 0$), (11) becomes

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} \times [2 \sin (\omega_o t) - \sin (\omega_o t - 2\omega_s t) - \sin (\omega_o t + 2\omega_s t)]. \quad (12)$$

Equation (12) is for a 100-percent-modulated carrier whose envelope is the true product of $E_1$ and $E_2$.

When $e_1$ and $e_2$ are in quadrature ($\theta_s = 90$ degrees), (11) becomes

$$e_{\text{out}} = \frac{k_o E_{1m} E_{2m} E_{cm}}{4} \times [- \sin (\omega_o t - 2\omega_s t - \pi/2) - \sin (\omega_o t + 2\omega_s t + \pi/2)]. \quad (13)$$

Only the upper and lower sidebands exist in this case. The envelope is not the true product. It becomes so upon insertion of a carrier component of proper phase and having sufficient magnitude.

**Oliver S. Meixell** was born in Reading, Pennsylvania, on 23 March 1904. In 1927, he received a BA degree from Yale University and thereafter studied electrical engineering at Massachusetts Institute of Technology.

He spent five years as an engineer with the Western Electric Company. From 1936 to 1943, he was an instructor in communication and electronics at RCA Institutes in New York, becoming a department head in 1938. During 1943 and 1944, he was a senior instruction-book writer for the Hazeltine Electronics Corporation.

In 1944, he joined ITT Federal Laboratories and is presently technical assistant to the director of the Electronic Defense Laboratory.

Mr. Meixell has been granted four patents. He is a Senior Member of the Institute of Electrical and Electronics Engineers.

# Transmission-Loss Measurements and Diversity Tests Made on Two Knife-Edge Diffraction Paths in Europe*

**R. E. GRAY**

**ITT Communication Systems Inc., on loan from ITT Federal Laboratories, a Division of International Telephone and Telegraph Corporation; Nutley, New Jersey**

## 1. Introduction

As part of a program of transmission-loss measurements on beyond-the-horizon paths in the Mediterranean area, tests were made on two knife-edge diffraction paths having a common receiving terminal near Genoa in northern Italy. The tests were made at a frequency of 1965 megacycles per second in August 1961.

The path-loss measurements indicated that the loss on one path was about 27 decibels less, and on the other 8 decibels less, than would be expected for normal scatter paths having the same angular distances.

Diversity tests, using receiving antennas spaced 47 feet (14.3 meters) horizontally and 10 feet (3 meters) vertically, showed that there was little correlation on either path in the received carrier-level variations at the two antennas.

From the transmission-loss measurements and diversity tests, it is estimated that high-quality 72-voice-channel systems could be established on these paths, using 10-kilowatt transmitters and 60-foot (18.3-meter) antennas in the 2000-megacycle-per-second band.

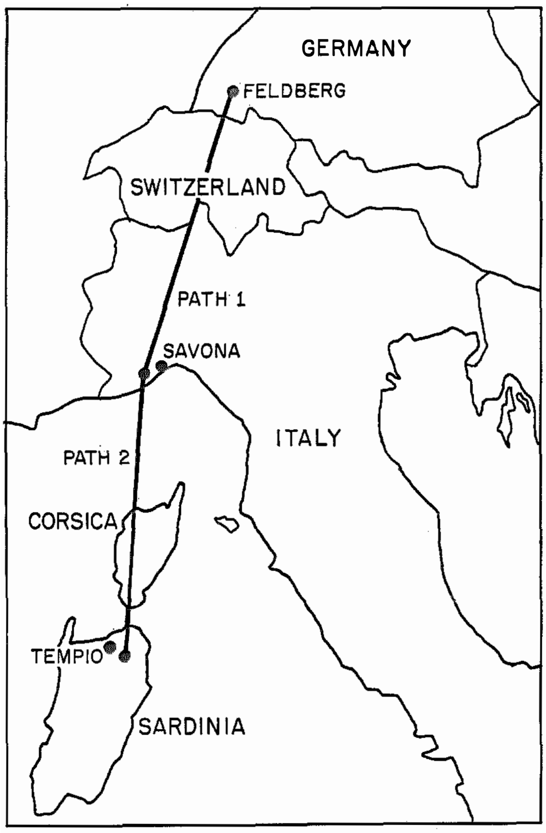


Figure 1—Transmission paths.

## 2. Transmission Paths

Path 1 extended from Feldberg, in southern Germany, across the Alps to a site near Savona in Italy; and path 2 from a point near Tempio, in northern Sardinia, across Corsica to the same receiving site. Figure 1 shows the location of the two paths.

The characteristics of the two transmission paths were as follows:

| | *Savona–Feldberg* | *Savona–Tempio* |
|---|---|---|
| Height of transmitting terminal above sea level in feet (meters) | 4900 (1493) | 3970 (1210) |
| Height of receiving terminal above sea level in feet (meters) | 3370 (1029) | 3370 (1029) |
| Height of common obstacle above sea level in feet (meters) | 13 200 (4030) | 5400 (1650) |
| Length of path in statute miles (kilometers) | 251 (404) | 240 (385) |
| Distance from receiving terminal to common obstacle in statute miles (kilometers) | 159 (256) | 134 (215) |
| Distance from transmitting terminal to common obstacle in statute miles (kilometers) | 92 (148) | 106 (170) |
| Angular distance in milliradians (assuming $K = 4/3$) | 53 | 28.2 |

---

* The tests described were part of a program of measurements made by ITT Communication Systems, Inc., under Air Force Contract AF 19(626)-4, "Eur-Med Tropo Scatter Communication System."

To allow for normal refraction, the effective radius of the earth is about one-third longer than the true radius, and $K$ is defined as the ratio of the effective radius to the true radius. Figures 2 and 3 show the profiles of the two paths, assuming a 4/3 earth radius.

## 3. Equipment

### 3.1 TRANSMITTING INSTALLATIONS

### 3.1.1 *Feldberg Site*

A 10-watt crystal-controlled transmitter operating on 1965 megacycles per second and a 28-foot (8.5-meter) Kennedy-type transportable antenna were installed at this terminal.

### 3.1.2 *Tempio Site*

Two 10-watt crystal-controlled transmitters, one operating on 1965 and the other on 1965.3 megacycles per second, and two 10-foot (3-meter) antennas were installed at this site. The centers of the antennas were spaced 55 feet (16.8 meters) horizontally and 15 feet (4.6 meters) vertically at right angles to the transmission path.

### 3.2 RECEIVING INSTALLATION

At the receiving site near Savona, two crystal-controlled narrow-band (500-cycle-per-second) path-loss-measuring receivers and two antennas, one 28 feet (8.5 meters) and the other 10 feet (3 meters) in diameter, were employed. The receiving antennas (see Figure 4) could be directed toward either Feldberg or Tempio and were spaced 47 feet (14.3 meters) horizontally and 10 feet (3 meters) vertically at right angles to the transmission path. The output of each receiver went to a separate recorder, scaled linearly in decibels. The time constants of the recorders could be varied from

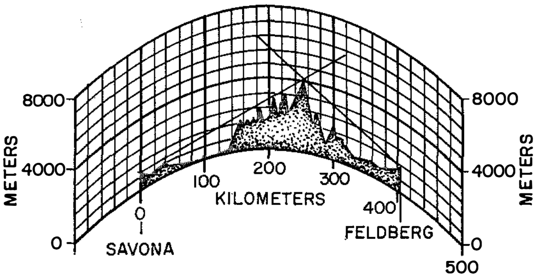


Figure 2—Profile of Savona-Feldberg path for $K = 4/3$.

TABLE 1

VERTICAL SPACING BETWEEN ANTENNA CENTERS

| Path | Calculated Antenna Spacing in Feet (Meters) for $K = 4/3$ | |
|---|---|---|
| | Receiver | Transmitter |
| Feldberg–Savona | 11.4 (3.5) | 6.0 (1.8) |
| Tempio–Savona | 27.9 (8.5) | 15.4 (4.7) |

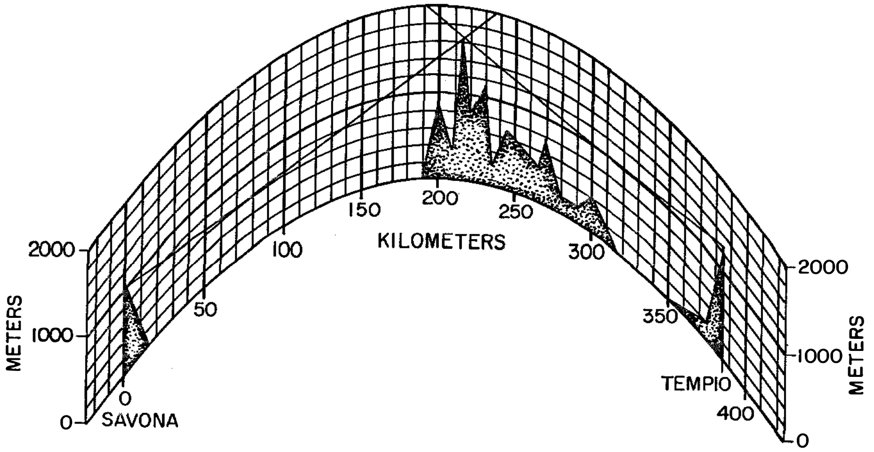


Figure 3 — Profile of Savona-Tempio path for $K = 4/3$.

3 to 50 seconds, and the speed of the recorder charts could be varied from 1 inch (25 millimeters) per minute to 1 inch per 15 minutes. Each receiver had a microammeter indicating the rectified receiver output; the over-all time constant of these meter indications was less than 1 second.

### 3.3 Antenna Spacing

In a space-diversity system, the antennas are spaced so that there is little correlation between the received level variations at the two antennas. On a normal scatter path, a horizontal antenna spacing of about 100 wavelengths is found to be adequate to reduce the correlation coefficient to a value of less than +0.4. On a knife-edge diffraction path, however, interference may exist between a direct ray and one or more rays reflected from points between the terminals and the knife-edge. In this case, vertical antenna spacing is required, and the correct distance between antennas may be calculated from the path profile. Assuming a single reflection at each side of the knife-edge, Table 1 shows the calculated vertical spacing between antenna centers necessary to ensure that when one antenna is at a minimum, the other is at or near a maximum of the interference pattern formed by the direct and indirect rays.

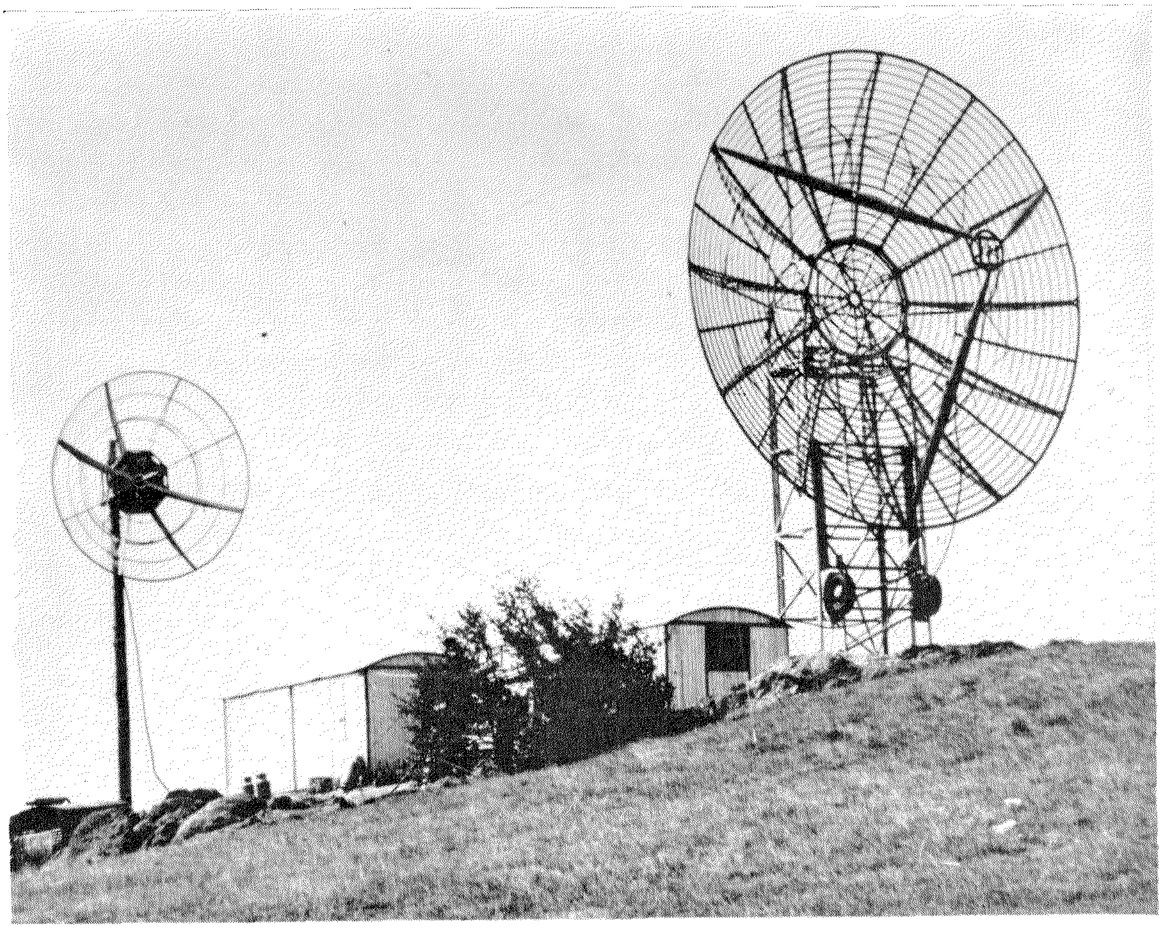


Figure 4—Receiving antennas at Savona site.

## 4. Transmission-Loss Measurements

Path-loss measurements were made at 1965 megacycles per second, first on the Feldberg-Savona path for about 600 hours, and then on the Tempio-Savona path for about 500 hours. Figures 5 and 6 show the distribution of the 15-minute medians on the two paths.

Table 2 shows the measured median loss, the calculated loss assuming knife-edge diffraction, and the estimated annual loss for paths having the same angular distances as those measured.

The calculated loss is based on the method shown in National Bureau of Standards Report 6767, Part 5, Sections 6-13, June 1961, and the estimated annual median loss is taken from an empirical curve of the average of a large number of transmission-loss measurements made on tropospheric-scatter paths.

The Feldberg-Savona path showed a higher diffraction gain than did the Tempio-Savona path. This may be due to the fact that the knife-edge on the Tempio-Savona path was the rounded side of a mountain on Corsica, whereas the knife-edge on the Feldberg-Savona path was a relatively sharp ridge in the Alps. The fading recorded on the two paths was very similar to that normally associated with tropospheric-scatter paths.

It has been estimated from the test results that the median loss for the worst hour of the worst month at 2600 megacycles per second for the Feldberg-Savona link would be 234.6 decibels. Using this value, the performance in

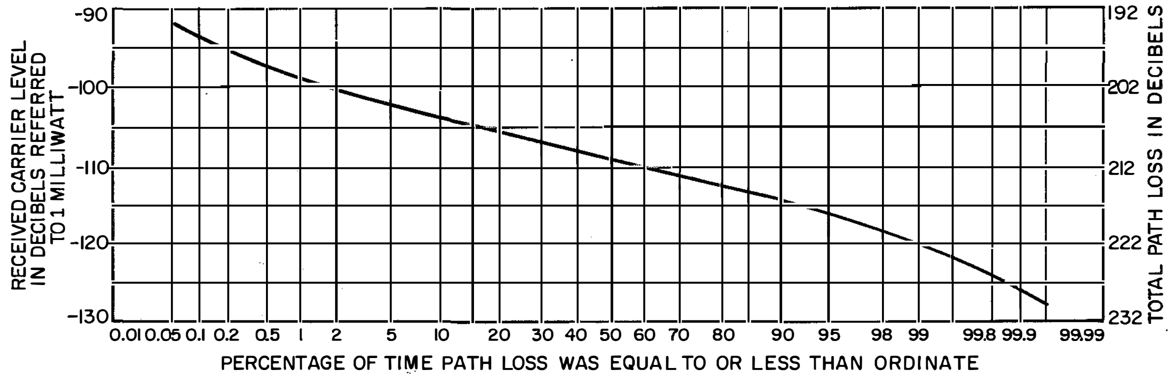


Figure 5—Path-loss results, Feldberg-Savona, for 2439 compilations of 15-minute medians.

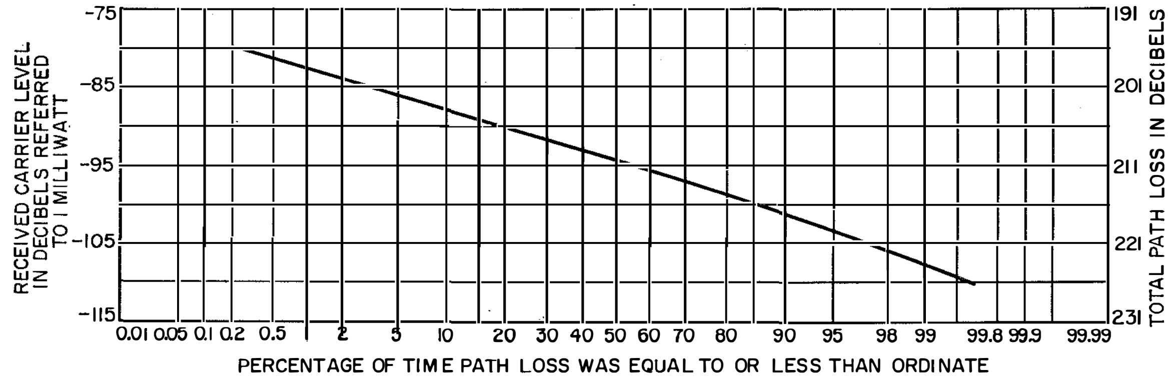


Figure 6—Path-loss results, Tempio-Savona, for 2029 compilations of 15-minute medians.

the worst channel of a 72-voice-channel system during the worst hour of propagation is outlined in Table 3.

It is re-emphasized that Table 3 represents the performance for the median of the worst hour of the year.

Since the measured diffraction gain on the Tempio-Savona path was only 8 decibels compared with 27 decibels on the Feldberg-Savona path, it is estimated that the seasonal change in loss would be about 4 decibels greater on the Tempio-Savona path.

Thus, using the same power and antenna size, the performance during the worst month on the Tempio-Savona path is likely to be about 4 decibels poorer than on the Feldberg-Savona path.

TABLE 2
MEDIAN LOSSES IN DECIBELS

| Path | Measured Median Loss | Calculated Median Loss | Estimated Median Loss |
|---|---|---|---|
| Feldberg–Savona | 210 | 198 | 237 |
| Tempio–Savona | 211 | 188 | 219 |

TABLE 3
WORST PERFORMANCE CHARACTERISTICS

| | |
|---|---|
| Transmitter power in kilowatts | 10 |
| Antenna diameter in feet (meters) | 60 (18.3) |
| Antenna gain in decibels | 86 |
| Transmission loss in decibels | 234.6 |
| Transmission-line loss in decibels | 3 |
| Received carrier level in decibels referred to 1 milliwatt | −81.6 |
| Carrier-to-noise ratio in decibels* | 25 |
| Bandwidth improvement in decibels | 27.4 |
| Frequency-modulation improvement in decibels (100-kilocycle-per-second root-mean-square deviation per channel) | −8.1 |
| Weighting, pre-emphasis, and 4-fold diversity improvements in decibels | 13.5 |
| Channel signal-to-noise ratio in decibels | 57.8 |

* Assuming 2-decibel noise-figure receivers with parametric amplifiers.

## 5. Diversity Measurements

Dual space-diversity tests were made by recording simultaneously the received carrier level on two receivers at the Savona terminal, and also by noting the reading on the two receiver output meters.

### 5.1 FELDBERG-SAVONA PATH

For diversity tests on this path, a transmitter and a single antenna were used at Feldberg, and two receivers with spaced antennas were used at Savona. The received carrier level was recorded on each receiver using a recorder with a time constant of about 3 seconds and a chart speed of 1 inch (25 millimeters) per minute.

### 5.1.1 *Diversity Test Results*

The results obtained are given in Figures 7 through 9, which show sample values taken from the recorded carrier levels. These were obtained at different times during a 4-day period in August 1961. The difference in median received level on the two antennas is a function of the difference in gain of the antennas and the difference in loss of the two antenna transmission lines; 10 divisions on the recorded scales represent a 5-decibel difference in receiver level.

### 5.1.2 *Correlation Coefficient*

The calculated correlation coefficients between the recorded carrier levels in decibels as shown in Figures 7 through 9 are given in Table 4.

### 5.2 TEMPIO-SAVONA PATH

For diversity tests on this path, two transmitters and two spaced antennas were available at Tempio, and two receivers with spaced antennas were used at Savona. The Tempio transmitters operated at slightly different frequencies so that the carriers could be separated at the Savona receivers. During diversity

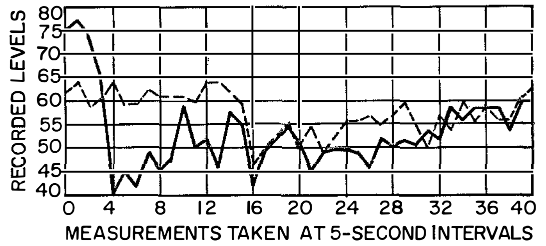


Figure 7—Recorded levels of measurements taken at 5-second intervals, Feldberg-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was +0.22.

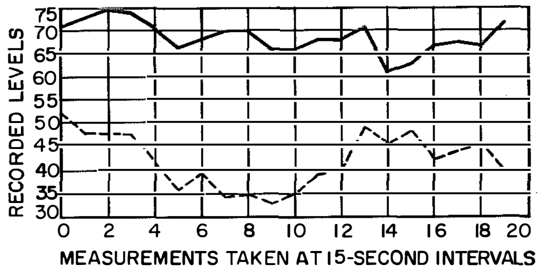


Figure 8—Recorded levels of measurements taken at 15-second intervals, Feldberg-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was +0.26.

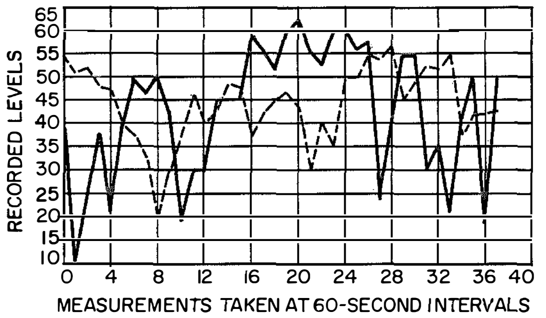


Figure 9—Recorded levels of measurements taken at 60-second intervals, Feldberg-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was −0.26.

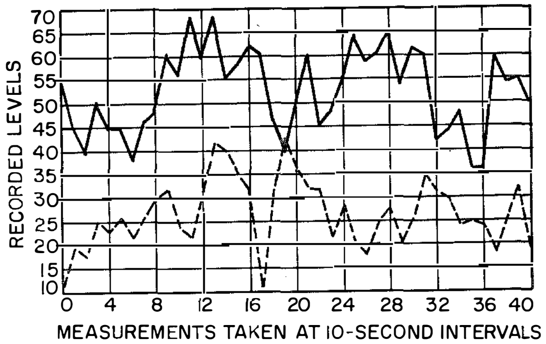


Figure 10—Recorded levels of measurements taken at 10-second intervals for one transmitting and two receiving antennas, Tempio-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was +0.05.

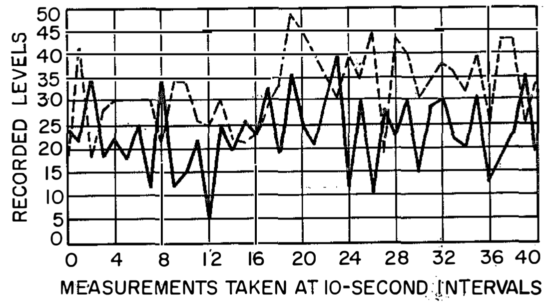


Figure 11—Recorded levels of measurements taken at 10-second intervals for two transmitting and two receiving antennas, Tempio-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was −0.08.

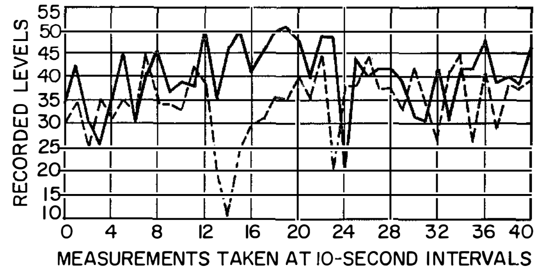


Figure 12—Recorded levels of measurements taken at 10-second intervals for two transmitting and one receiving antennas, Tempio-Savona. The broken line is for a 10-foot (3-meter) antenna and the solid line is for a 28-foot (8.5-meter) antenna. The correlation coefficient was −0.15.

TABLE 4
CALCULATED CORRELATION COEFFICIENTS, FELDBERG–SAVONA

| Time Interval in Seconds | Correlation Coefficient |
|---|---|
| 5 | +0.22 |
| 15 | +0.26 |
| 60 | −0.26 |

TABLE 5
CALCULATED CORRELATION COEFFICIENTS, TEMPIO-SAVONA

| Antenna Combination | Time Interval in Seconds | Correlation Coefficient |
|---|---|---|
| 1 transmitting and 2 receiving | 10 | +0.05 |
| | 60 | −0.27 |
| | 900 | +0.28 |
| 2 transmitting and 2 receiving | 10 | −0.08 |
| | 15 | −0.14 |
| | 60 | +0.10 |
| 2 transmitting and 1 receiving | 10 | −0.15 |
| | 15 | −0.0046 |
| | 60 | −0.16 |

tests on this path, the following antenna combinations were used.

**(A)** One transmitting and two receiving.
**(B)** Two transmitting and two receiving.
**(C)** Two transmitting and one receiving.

### 5.2.1 *Diversity Test Results*

Figures 10 through 12, respectively, show samples of the results obtained with each of the 3 combinations of antennas.

### 5.2.2 *Correlation Coefficient*

The calculated correlation coefficients between the recorded carrier levels in decibels for the various antenna combinations are given in Table 5.

## 6. Conclusion

Path-loss measurements and space-diversity tests were made at 1965 megacycles per second on two knife-edge diffraction paths in Europe. On each path, the measured loss was less than was estimated for normal beyond-the-horizon paths having the same angular distance, but was greater than the values calculated from knife-edge diffraction theory.

Diversity tests were made with antennas spaced 47 feet (14.3 meters) horizontally and 10 feet (3 meters) vertically. Correlation coefficients calculated from the recorded carrier level variations in decibels were in all cases less than +0.30.

A suggested cause of this low correlation is that in each case the obstruction was not a smooth horizontal knife-edge, but a ridge varying in elevation across the path. Such an obstruction would result in multipath transmission with an expected low correlation of amplitude variations on spaced receiving antennas and with the fading characteristics associated with multipath transmission. The recorded fading was similar to that observed on normal scatter paths.

It is estimated that high-quality 72-voice-channel systems could be established over these paths using 10-kilowatt transmitters and 60-foot (18.3-meter) antennas in the 2000-megacycle-per-second band. A Feldberg-Savona multichannel system would be of particular interest since it would form a direct communication link between Italy and Germany.

---

**R. E. Gray** was born in Wynford, England, on 15 February 1902. He received a First Class Diploma, equivalent to a bachelor's degree, from Faraday House, London, in 1924.

From 1924 to 1927, he was with Standard Telephones and Cables, London, and from 1927 through 1939 was with Laboratoire Central de Télécommunications, Paris. During the second World War, he was on loan to the Royal Aircraft Establishment and to the Telecommunication Research Establishment in England.

He participated in the first microwave communication link between France and England, in the introduction of the ground-controlled-approach aircraft landing system in England, and in the original commercial transatlantic radiotelephone service. He was also active in the installation of the first Spain-Argentina radiotelephone link, as well as the multichannel radio link joining Scotland and Ireland.

In 1945, he joined ITT Federal Laboratories, where he is presently a senior scientist. He continues his radio wave propagation studies with particular attention to forward-scatter systems.

He was on loan to ITT Communication Systems for the survey discussed in this issue.

## Eaglesfield Receives Institution of Electrical Engineers Premium

Charles Cecil Eaglesfield has received the Premium of the Institution of Electrical Engineers for his paper on "Optical Pipelines—A Tentative Assessment," which was published in the Proceedings of that society, volume 109, part 13, pages 26–32; January 1962.

Mr. Eaglesfield joined Standard Telephones and Cables in 1947 and was transferred to Standard Telecommunication Laboratories ten years later. He has been associated with the application of waveguides to long-distance communication and is presently concerned with the feasibility of communication at optical wavelengths.

# Marine Transmitter with Single-Sideband Facility

F. HEDSTRÖM

Standard Radio & Telefon AB; Bromma, Sweden

## 1. Introduction

Efforts have been concentrated in recent years on decreasing the emitted bandwidth of high-frequency radio transmitters to make best use of the very-crowded bands. Some of the methods used to reach this goal have been careful pulse shaping in radiotelegraphy and advanced single-sideband techniques in radiotelephony. Both techniques require the use of linear power amplifiers, since the amplification that follows modulation or keying must be linear all the way to the antenna. The input to the linear amplifier must therefore have the correct frequency and the required kind and degree of modulation; frequency multiplication may not be used after modulation.

In marine radio communication, telephony is increasing in importance. Distances are often very long and transmission quality must be good to allow interconnection with the public telephone network. Since directive antennas are not practicable on shipboard, the high efficiency of single-sideband operation is a great advantage.

A modern marine transmitter must provide for telegraphy of both unmodulated (*A1*) and

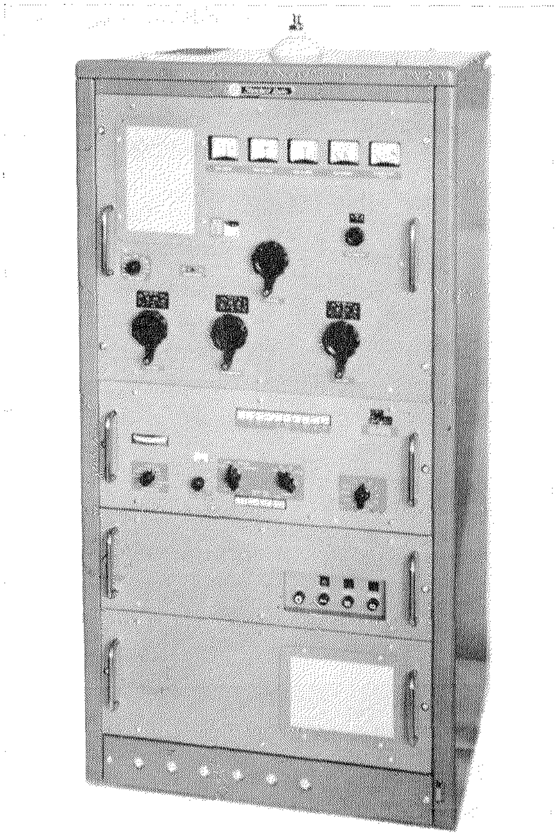


Figure 1—*ST 1200* transmitter. From top to bottom are shown the linear power amplifier, exciter unit, control unit, power supply, and fuses and connection unit.

audio-frequency modulated (*A2*) types, telephony with single-sideband and reduced carrier (*A3A*), and, to comply with international regulations, telephony with full carrier (*A3H*).

The lowest limit of the transmitter's output power (250 watts for some countries) is also based on full-carrier double-sideband operation. This corresponds to a peak effective power of 1 kilowatt from a linear power amplifier. Thus the power output of the *ST 1200* transmitter was designed to be 1.2 kilowatts peak effective power to include a margin for aging of tubes and other losses.

The main parts of the transmitter, which is shown in Figure 1, are:

**(A)** A linear power amplifier covering the frequency ranges from 400 to 535 kilohertz and from 1.6 to 25.5 megahertz. The linear power amplifier includes a matching network for the antenna impedances specified by different national rules to permit proper loading of the various types of shipboard antennas.

**(B)** A power supply with control devices for the linear power amplifier.

**(C)** An exciter unit that produces a signal of the required frequency and modulation at a power level suitable for the input of the linear power amplifier.

## 2. Linear Power Amplifier

### 2.1 Design Problems

To assure linear operation and also to reduce the load on the penultimate stage, the output tubes should give reasonable efficiency even when not driven into the grid-current region. The efficiency of such tubes is about 64 percent when driven to the grid-current point. Thus a peak effective power of 1.2 kilowatts gives an anode dissipation of 675 watts. The coupling network between the output stage and the antenna introduces losses of about 15 percent, so the output power from the tubes must be increased by about this amount. The efficiency of a linear stage decreases with decreasing output power. When the output power is zero, all the power delivered to the anodes is dissipated as heat. If this minimum level of plate input is lowered, the linearity of the stage deteriorates. Thus, as a function of linearity, this minimum level determines the required anode dissipation. As a rule of thumb, the obtainable peak effective power for modern tubes not driven into the grid-current region is about 1.2 times the allowable anode dissipation. Accordingly, the required anode dissipation in this case is about 1 kilowatt. The factor of 1.2 applies to tubes with a relatively low anode voltage of about 2000 volts; the factor increases for tubes with higher anode voltage.

Eimac tube type *4X250B* was chosen for the transmitter. To get the required power output, four of these tubes were connected in parallel. A single tube, such as Eimac type *4CX1000A*, could be employed but the type *4X250B* is more widely available. Also its anode voltage is only 2000 volts compared with 3000 volts for the *4CX1000A*. At the highest operating frequency of about 25 megahertz, 4 high-slope tubes can be operated in parallel by careful design of the input and output circuits.

Figure 2—Typical antenna coupling network for medium-frequency range from 415 to 535 kilohertz.

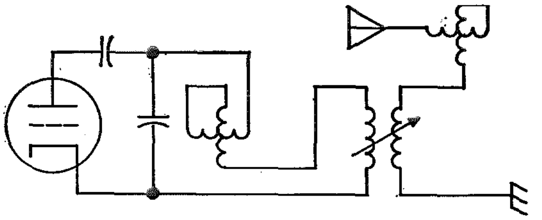
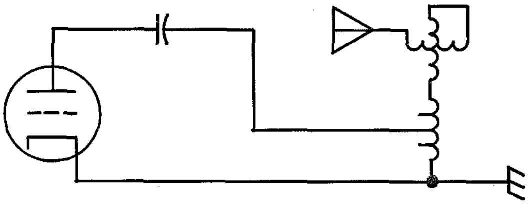


Figure 3—Simplified antenna coupling network for medium-frequency range.



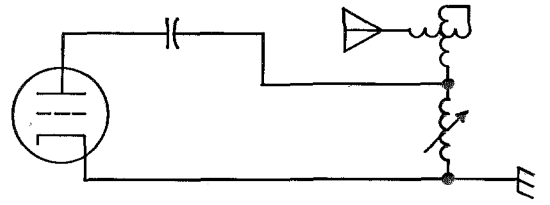
Another problem was the design of the matching network between the output stage and the antenna. Two different types of networks were required, one for the 400-to-535-kilohertz range and one for the 1.6-to-25.5-megahertz range. In the medium-frequency range, the $Q$ and thus the selectivity of the antenna is always very high. The required selectivity for suppressing harmonics can easily be obtained in the antenna circuit without the need for extra tuned circuits or filters. The high $Q$ of the antenna causes the antenna current and voltage to be high. Because of this, antenna circuit components tend to be rather bulky and expensive.

### 2.2 ANTENNA COUPLING NETWORKS

A commonly used medium-frequency antenna coupling network is shown in Figure 2. The anode and antenna circuits are separately tuned while the coupling between them is adjustable. The circulating currents are high in the two tuned circuits, especially in the antenna circuit, which requires careful design of components.

The coupling circuit used for the transmitter is shown in Figure 3. The single tuning element is an adjustable inductor in series with the antenna to form a tuned circuit with the antenna capacitance. Taps on the inductor are used to match the antenna resistance to the optimum loading resistance of the output stage. This circuit works only with an antenna of capacitance reactance but this is always the case with antennas in the medium-frequency range aboard ship. Under special conditions, a capacitor can be connected in series with the antenna to shorten it electrically. Figure 4 shows the practical circuit; to avoid taps on the inductor, a second adjustable small inductor is connected in series with the antenna tuning inductor and forms a continuously adjustable tap.

Figure 4—Practical antenna coupling network for medium-frequency range.



For the 1.6-to-25.5-megahertz range, another type of coupling network is required. Antennas in this frequency range cannot be depended on to suppress harmonics and the selectivity of the coupling network must be high. Thus three different requirements are imposed on the network:

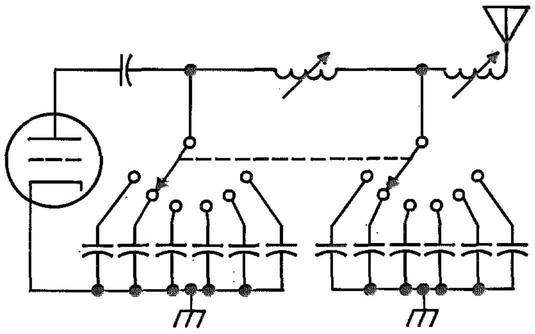
**(A)** Correcting the phase angle of the load (antenna) to zero.

**(B)** Transforming the antenna resistance to the correct value for optimal loading of the output stage.

**(C)** Providing the required selectivity to get sufficient suppression of harmonics.

To satisfy the first two requirements, at least two adjustable controls must be introduced in the network. Many different types of networks can be used and the most common one is the $\pi$ filter. However, the harmonic suppression obtained with a simple $\pi$ filter is limited and two components, one a capacitor, must be adjustable. Adjustable air dielectric capacitors are very bulky at this output power and vacuum capacitors are expensive. The problem was solved eventually by using a $\pi$-$L$ filter as

Figure 5—$\pi$-$L$ antenna coupling network used from 1.6 to 25 megahertz.



shown in Figure 5. The two adjustable components are now inductors, allowing the use of small and inexpensive fixed capacitors. It was possible to design the inductors to be usable over the entire frequency range. The fixed capacitors have to be switched in 6 bands.

In the coupling network for the medium-frequency range, it was explained that a small adjustable inductor was used in series with the antenna inductance. To minimize the number of components in the complete network, this same inductor is used as the loading coil in the $\pi$-$L$ filter. Figure 6 shows the complete tuning and coupling network for all ranges.

The amount of amplification required from the linear power amplifier depends on the input power level and the input impedance. An input power level of 0.1 watt was chosen as a compromise among such considerations as the stability of the linear power amplifier, the amount of power available from inexpensive receiving-type tubes, and the number of amplifier stages required. For standardization, an input impedance of 50 ohms was chosen.

### 2.3 Preamplifier

The preamplifier is shown in Figure 7, which is a simplified schematic diagram of the entire linear power amplifier. Two preamplifier stages are employed, one a grounded-grid and the other a conventional grounded-cathode stage. A grounded-grid circuit is used in the first stage for impedance matching and stability without neutralization. The couplings between the first and second stages and to the grids of the output stage are tuned circuits. Because of the rather high impedances in the grid circuits of the penultimate and output stages, both must be neutralized. Neutralization holds for the complete frequency range and does not require adjustment during normal operation of the transmitter. Both tuned circuits are provided with continuously adjustable inductors. In combination with the switched fixed capacitors of the tuning network, they cover the entire medium-high and high-frequency range with a single coil per circuit. The coils for both circuits are ganged. In the medium-frequency range, broad-band circuits are used and no tuning is required.

### 2.4 Tuning Indicators

The tuning of a linear power amplifier must be carried out with precision or the linearity will be adversely affected. To simplify the tuning process, the preamplifier and the output stage are provided with phase discriminators

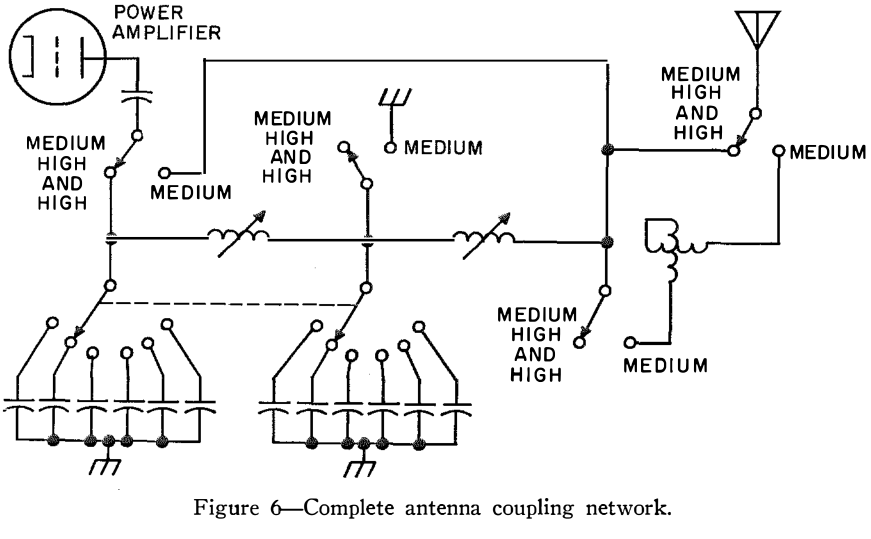


Figure 6—Complete antenna coupling network.

that indicate the resonant condition of the tuned circuits. The output stage is also provided with an amplitude-comparison circuit (loading discriminator) and all the discriminators are provided with zero-center meters.

## 3. Exciter Unit

The exciter unit provides the linear power amplifier with the proper input signal, which must have the required frequency plus the appropriate kind and degree of modulation. Therefore the exciter must contain two different circuits, a frequency-generating circuit and a modulating circuit.

### 3.1 FREQUENCY-GENERATING CIRCUIT

The number of frequencies required from a marine transmitter has increased very rapidly during the past few years and their harmonic relation, which was once a rule, does not fully hold today. This means that the number of crystals for a modern marine telegraph and telephone transmitter is very high; the *ST 1200* transmitter is equipped with 94 replaceable and switchable crystals. For the medium-high-frequency band, 1.6 to 3.8 megahertz, the crystal frequency is the same as the output frequency. For the high-frequency band, 4 to 26 megahertz, the crystal frequencies range from 2 to 3 megahertz and frequency multiplication is used to get the required output frequencies. Crystal ovens are not needed for the frequency stabilities required by international regulations.

When higher frequency stabilities are required for certain types of transmission, other methods must be employed. The *ST 1200* transmitter obtains the required output frequency by mixing frequencies from three different sources:

**(A)** A crystal oscillator in the 2-to-3-megahertz range.

**(B)** A crystal oscillator adjustable from 41.2 to 66.2 megahertz in 1-megahertz steps.

**(C)** A 9-megahertz amplitude-modulated frequency derived from a 1-megahertz crystal oscillator.

The simplified block diagram of Figure 8 illustrates the frequency-generating circuit for the

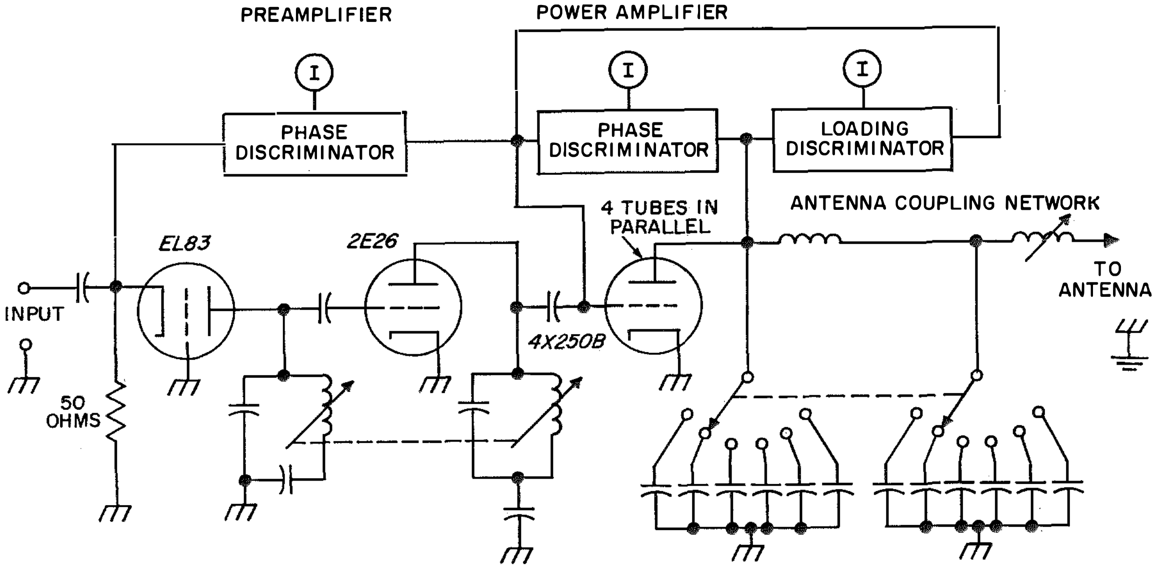


Figure 7—Diagram of the linear power amplifier.

medium-high and high frequencies. The 2-to-3-megahertz crystal oscillator is called the channel oscillator and the 41.2-to-66.2-megahertz oscillator is called the band oscillator. The reasons for these designations are that the channel oscillator generates the frequency required within a band, whereas the band oscillator determines the required frequency band.

The output from the channel oscillator is subtracted in mixer *M1* from the output of a correction oscillator that has been tripled to 34.2 megahertz. The difference frequency of approximately 31 to 32 megahertz passes through a band-pass filter to mixer *M2*, where it is added to the 9-megahertz amplitude-modulated frequency. The resulting 40-to-41-megahertz frequency and a frequency from the band oscillator are impressed on mixer *M3*. The frequencies from the band oscillator occur only at 1-megahertz intervals from 41.2 to 66.2 megahertz. The exact frequency from the band oscillator depends on the required frequency band, and the output from mixer *M3* is the required output frequency with the modulation contained in the 9-megahertz signal supplied to mixer *M2*.

If, for instance, the required frequency is 12 440 kilohertz, then the crystal frequency for the channel oscillator is 2440 kilohertz. The output frequency from mixer *M1* is 31 760 kilohertz, and the output frequency from mixer *M2* is 40 760 kilohertz, which means that the output frequency from the band oscillator must be 53.2 megahertz.

Frequency errors caused by the band oscillator are corrected as follows. The band-oscillator output is also supplied to mixer *M4*, which mixes it with sharp pulses having a repetition frequency of 1 megahertz controlled by a 1-megahertz crystal oscillator. Mixer *M4* develops a frequency spectrum containing mixer products between the band-oscillator frequency and the 1-megahertz harmonics. The 34.2-megahertz mixed product is supplied through a band-pass filter to a phase discriminator, which is also connected to the correction-oscillator circuit. The phase discriminator controls the frequency of the correction oscillator by means of a reactance element. Thus the output frequency of the tripler following the correction oscillator is exactly the same as the output frequency from mixer *M4* selected by

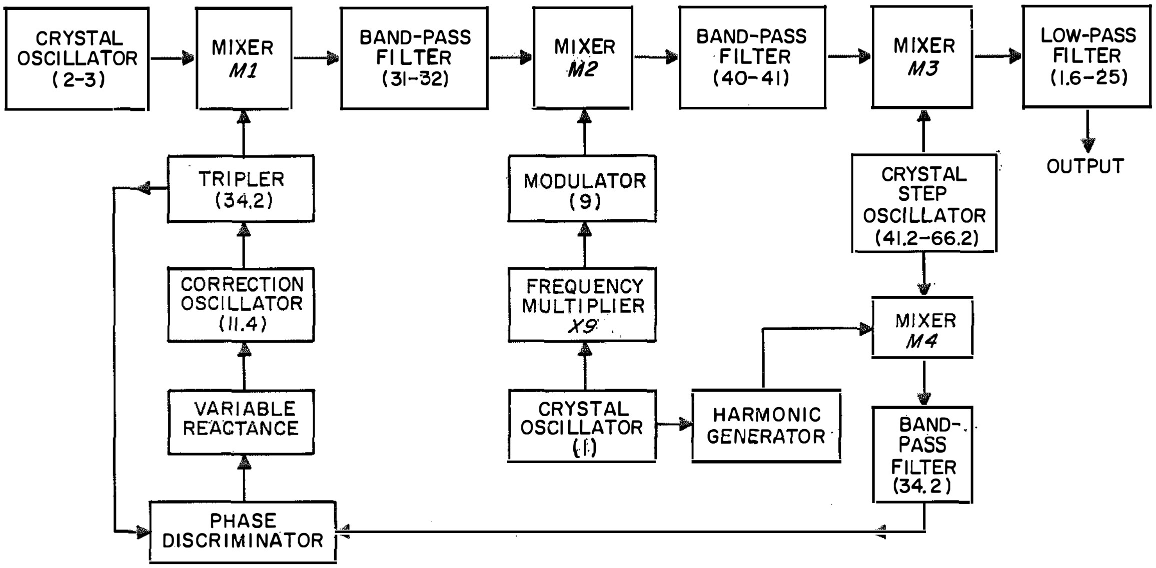


Figure 8—Diagram of frequency-generating circuit used on medium-high-frequency and high-frequency bands. The frequencies within parentheses are in megahertz.

the 34.2-megahertz band-pass filter. If the band-oscillator frequency varies by $\Delta F$, the frequency from mixer *M4* also changes by this amount causing the frequency of the oscillator and tripler to change by the same amount. Since the latter frequency is also supplied to mixer *M1*, the frequencies in the chain from mixer *M1* to mixer *M3* will change. In this way $\Delta F$ is cancelled, and the only frequency errors are the sum of those in the highly stable channel oscillator and 1-megahertz oscillator. The correction oscillator with its associated tripler should be considered only as a filter and is necessary to minimize the spurious signals supplied to mixer *M1*.

An amplifier is used to raise the output level from mixer *M3*. This amplifier uses two wideband stages, with a gain control in the first stage. The nominal output is 0.1 watt peak effective power into 50 ohms.

### 3.2 Modulating Circuit

All modulation of the transmitter, including *A1* and *A2* keying, is introduced into mixer *M2* of the frequency-generating circuit. The 9-megahertz signal supplied to this mixer must contain all the necessary modulation.

Figure 8 shows that the 1-megahertz crystal oscillator supplies a signal to a frequency multiplier with an output frequency of 9 megahertz. The 9-megahertz signal is supplied via two separate amplifiers to two modulators, one for *A1* modulation and the other for *A3* modulation, as shown in Figure 9. The *A1* modulator supplies one of the inputs to a summation amplifier and the *A3* modulator supplies the other via the single-sideband filter. The output of the summation amplifier is supplied to mixer *M2* in the frequency-generating circuit.

The modulating signals, from either a microphone or a telegraph key, are supplied to the modulators via a switching unit, controlled by push buttons and rotary switches on the front panel. For *A1* emission, the telegraph key interrupts a direct-current source to produce the pulses representing telegraph signals. The direct-current pulses are routed through the switching unit to the *A1* modulator. A low-pass filter shapes the pulses to minimize the emitted bandwidth. Since the transmitter is

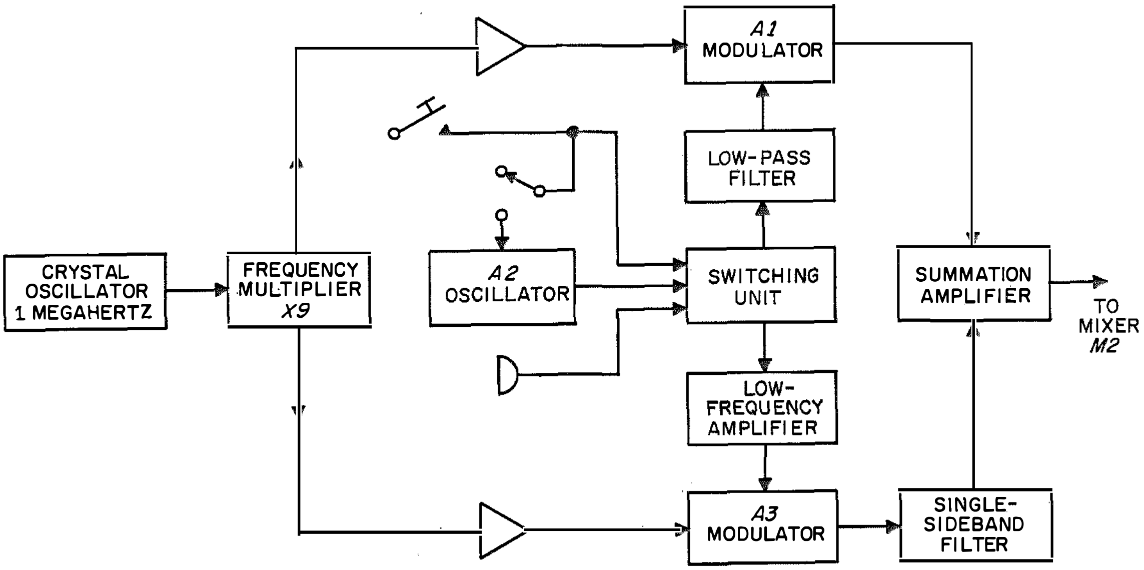


Figure 9—Diagram of modulating circuit.

linear from this point on, the pulse shape undergoes no further change. To suppress keying noises, the telegraph key operates a keying unit that provides the output amplifier with a blocking voltage.

For *A2H* modulation, in addition to the *A1* operation described above, the telegraph key also operates an *A2* oscillator whose output goes via the switching unit and the low-frequency amplifier to the *A3* modulator, which emits the sideband. Accordingly, both the *A1* and *A3* modulators are used during *A2H* operation; the *A1* modulator supplies the keyed carrier and the *A3* modulator supplies the keyed sideband.

For voice operation, *A3A* and *A3H,* the *A3* modulator also supplies the sideband and the *A1* modulator supplies the carrier. Adjustment of the *A1* modulator controls the carrier level. A switch, usually placed in the microphone handle, turns the transmitter on and off in the same way that the telegraph key does. The low-frequency amplifier that increases the output signals from the *A2* oscillator is also used to amplify the voice signals.

## 4. Power Supply and Cooling System

Separate power supplies are used for the exciter unit and the linear power amplifier. The latter supply provides 2000 volts for the output stage. This power supply is a 3-phase full-wave rectifier with the high-voltage transformer designed for either delta or star connection of the primary, to allow the transmitter to be connected to a source of either 220 or 380 volts. The rectifier is a compact unit provided with silicon elements.

The screen-grid and control-grid voltages for the output stage are developed by electronically regulated power supplies.

The transmitter is forced-air cooled by two blowers. One blower cools the output tubes and the tuning coils, especially the antenna tuning coil for the medium-frequency band. The second blower cools the silicon rectifiers and transformers. Its air stream is also ducted to the control and exciter units. Both blowers are provided with replaceable front-panel filters. The cooling system makes it possible to operate the transmitter in ambient temperatures of +55 degrees centigrade (+131 degrees fahrenheit).

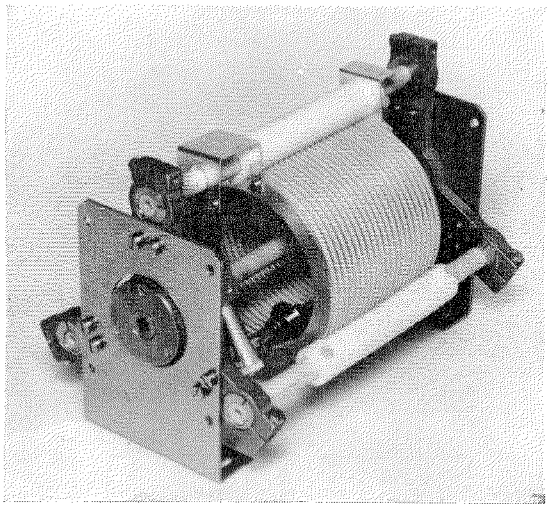


Figure 10—Inductor used in the $\pi$-$L$ filter of the antenna-coupling network.

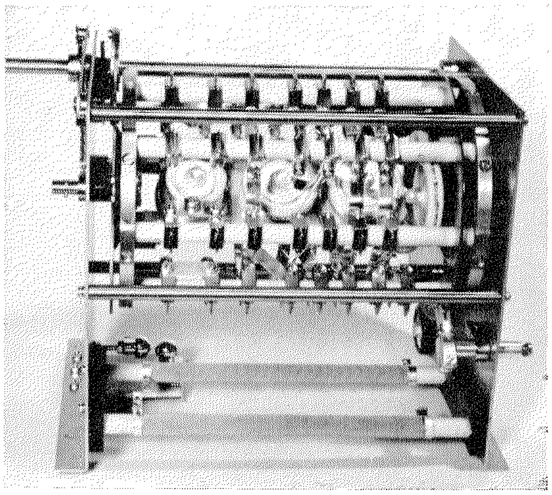


Figure 11—Main band switch.

TABLE 1

*ST 1200* TRANSMITTER PERFORMANCE

| Characteristic | Specification |
|---|---|
| Frequency coverage | Medium-frequency: 400 to 535 kilohertz.<br>Medium-high frequency: 1600 to 3800 kilohertz.<br>High frequency: 4, 6, 8, 12, 16, 22, and 25 megahertz, within marine mobile bands. |
| Number of frequency channels | Medium-frequency: 8 telegraph channels.<br>Medium-high frequency: 25 telephone channels.<br>High frequency: 3 telegraph channels in each band; 7 telephone channels in each band except 3 channels in 6-megahertz band.<br>All channels are crystal controlled. |
| Frequency tolerance attainable | Short term: ±40 hertz.<br>Long term: ±100 hertz. |
| Types of emission | Medium-frequency: *A1, A2H.*<br>Medium-high frequency: *A1, A2H, A3H, A3A.*<br>High frequency: *A1, A3H, A3A.*<br>*A1:* Telegraph-keyed carrier.<br>*A2H:* Telegraph-keyed audio-frequency-modulated carrier with carrier at full amplitude.<br>*A3A:* Telephony with single sideband and reduced carrier.<br>*A3H:* Telephony with single sideband and full carrier. |
| Harmonic suppression | Greater than 50 decibels. |
| Suppression of lower sideband | Greater than 46 decibels. |
| Suppression of carrier (*A3A*) | Adjustable from 16 to 40 decibels below the sideband power. Normally set for 26 decibels. |
| Keying speed | 50 bauds (approximately 60 words per minute). |
| Audio-frequency band | 350-2700 hertz within 6 decibels. |
| Audio input | Dynamic or carbon microphone, 40 or 200 ohms, respectively. |
| Power into antenna | Medium-frequency:<br>*A1:* 250 watts carrier.<br>*A2:* 100 watts carrier, 50 watts sideband.<br>Medium-high and high frequency:<br>*A1:* 1200 watts carrier.<br>*A3H:* 600 watts carrier and 600 watts sideband for 100-percent modulation.<br>*A3A:* 1200 watts sideband. |
| Antenna impedance | Tunable to all normal antennas on shipboard. |
| Power supply | 3-phase, 220 or 380 volts, 50 to 60 hertz. |
| Power consumption | 3 kilowatts maximum, 0.9 power factor. |

## 5. Special Components

The components used in the antenna-coupling network in the output stage of the linear power amplifier were designed especially for this transmitter. Figure 10 shows one of the continuously adjustable inductors used in the $\pi$-$L$ filter. The winding is held by three ceramic rods. Grooved teflon pads are used to fix the turns and the spacing between them. To reduce the weight of the moving parts, the coil is fixed and the shaft with the contact arm and wheel moves. To avoid deterioration of the contact surfaces, rhodium plating is used. The contact wheel has a graphite bushing bearing to avoid the need for lubrication.

A special main band switch was designed for the fixed capacitors in the output stage. Figure 11 shows its general design. The insulation is ceramic. The contact pressure is relieved before the switch is rotated, making rotation easy despite high contact pressures. The fixed tuning capacitors are mounted inside the switch turret, making very-short connections possible. The mechanism for turning the switch employs a Maltese cross to ensure very-positive positioning. The switch contacts can also be moved along the ceramic rods, permitting numerous switching combinations.

## 6. Conclusion

The performance specifications of the transmitter are shown in Table 1. Measurements are for typical production units. A number of transmitters have been installed on ships and confirm the known advantages of single-sideband operation over conventional *A3*. Installation of the transmitter on ships has proved to be easy, particularly since the antenna coupling network makes matching to the different antennas very simple.

**Fred Hedström** was born at Lysekil, Sweden, in 1914. In 1934, he graduated from the Technical Institute of Gothenburg as an engineer.

From 1939 to 1953, he was engaged in the development of receivers and amplifiers. In 1953, he joined Standard Radio & Telefon in Bromma, and now is in charge of the development of communication radio equipment.

From 1950 to 1953, Mr. Hedström lectured on radio engineering at the Norwegian Technological Institute in Oslo, and, from 1957 to 1959, he taught transmitter engineering at the Technical University in Stockholm.

# Spicer Honoured

L. R. Spicer was appointed a Member of the Order of the British Empire by Her Majesty the Queen in the June Honours.

This recognition was made for his work on submarine cables for the Admiralty. Mr. Spicer is a cable development engineer for Standard Telephones and Cables at the Newport plant.

# Submarine-Coaxial-Cable Manufacture at Southampton Factory 2

**ERIC BAGULEY**
**Standard Telephones and Cables Limited; London, England**

## 1. Introduction

An account has been given of the problems met with and the solutions achieved in setting up facilities for the large-scale production of submarine coaxial cables at Southampton [1] starting from a vacant building site in 1955 and completing the first 5000 miles of cable in 1962.

The most-significant project contributing to the rapid increase in demand for these repeatered submarine telephone cables was the 1956 transatlantic cable [2] for which flexible one-way submerged repeaters were developed by Bell Telephone Laboratories in the United States. In 1960 Bell Telephone Laboratories announced to cablemakers at a symposium in Boston, Massachusetts, their new *SD* 128-channel submarine-telephone-cable system based on two-way working with rigidly housed submerged repeaters. A new design of armourless submarine cable was specified.

The cable, shown in Figure 1, consists of a central strength member of 41 high-tensile-steel wires enclosed within a seam-welded copper tube that is the inner conductor of the coaxial core. The outer conductor is a second copper tube insulated from the inner conductor by a cylinder of polythene 1·000 inch (25·4 millimetres) in external diameter. Over the outer conductor is an extended sheath of black polythene compound 1·25 inches (32 millimetres) in diameter to afford protection to the outer conductor during handling in the factory and on board ship during laying. Thus, the cable design has only 5 components; a steel-wire strand, 2 tubed copper tapes, and 2 polythene extrusions. For these simple components precise control of dimension and of the properties of the materials used in their fabrication was specified.

After due consideration of these proposals it was decided to build a second submarine cable factory at Southampton. A vacant site was available next to the first factory, and building operations commenced with pile driving in January 1961. Production of the new design of cable was under way by the middle of 1962, and the first shipment was made on 3 January 1963.

## 2. Production Engineering

There are 5 principal processes in the manufacture of the cable, namely: centre-conductor assembly, core extrusion, core shaving, outer conductor combined with sheathing, and transmission testing. Repeater couplings are then fitted to the cable ends. Of these only the core extrusion process lacked novelty. Processes new to submarine-cable engineering, such as argon-arc seam welding and a new method of core shaving, had been tried out at a pilot plant by Bell Telephone Laboratories, and recommendations had been made as to how to implement them in commercial production. From the information received a factory layout was planned for a cable output rate of 80 to 100 nautical miles (150 to 185 kilometres) per week and consisted of 4 centre-conductor machines, 2 core extruders, 4 core-shaving machines, 2 sheathing extruders, a testing enclosure, and a jointing room for the repeater couplings.

For the *SD* system the repeater-section cable

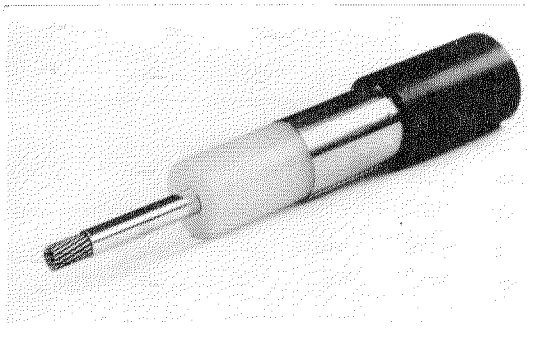


Figure 1—The new type *SD* armourless 1·000-inch (25·4-millimetre) coaxial submarine telephone cable comprising a central strength member, welded-copper-tube inner conductor, polythene dielectric, single-copper-tape outer conductor, and black polythene-compound outer sheath.

length is approximately 20 nautical miles (37 kilometres), and it was specified that this length should obtain at all stages of production, that is, 20-mile lengths of centre conductor, 20-mile lengths of core, and 20-mile lengths of completed cable, preferably without any joints in the central strength member. This manufacturing unit has to be taken into account in designing the various machines, in planning storage space for work in progress between machines, and for incoming materials. To ensure continuity of production at the stated rate it was decided to have storage space for not less than 3 weeks of stock of all materials and all work in progress. For some of the processes this means that storage facilities occupy more space than the production machines.

On completion of manufacturing, testing, and terminating, the 20-nautical-mile repeater-section lengths of cable have to be stored until required for loading into a cable ship. This is a major planning problem in a submarine-cable factory. To deal with the increasing demand for cables, new cable ships are being built capable of carrying up to 3000 nautical miles (5560 kilometres) of cable. The factory must, therefore, be able to store sufficient cable to enable ships to be loaded at a rate not less than 500 nautical miles (930 kilometres) per week.

A new method of storing completed cable was requested by the customer. Each 20-mile repeater-section length was required to be stored separately from all other lengths and to be accessible for ship loading in any sequence best suited to the transmission requirements of the system. Storage on drums was unacceptable because of the weight of cable involved, approximately 46 tons (47 tonnes). Instead, transportable cylindrical steel tanks were designed into which the 20-mile length of cable may be coiled down horizontally in the traditional submarine-cable manner. A 75-ton (76-tonne) travelling gantry crane presides over a storage yard in which up to 200 of the cable tanks may be moved around and stacked two high.

Part of the runway covered by the crane is inside the factory building, and within this enclosure the completed cable sections are coiled down into the tanks. When filled with cable the tanks are lifted and then lowered into fixed test tanks supplied with circulating water at 10 degrees centigrade, the temperature specified for the cable during transmission

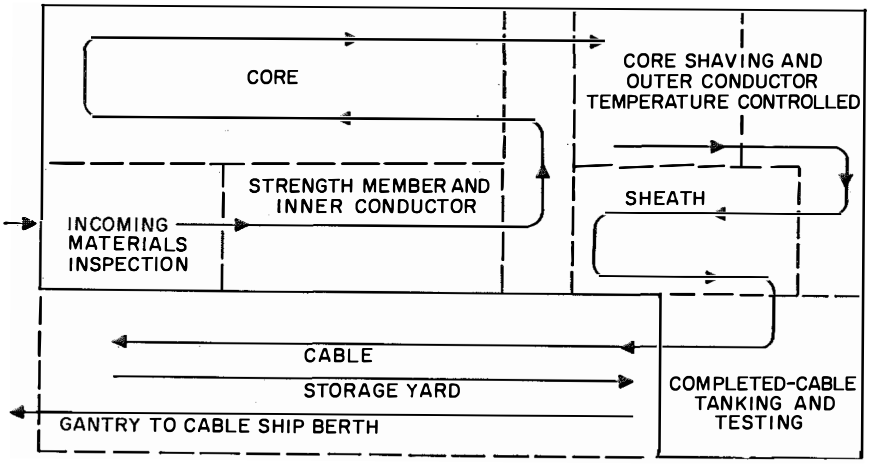


Figure 2—The layout of factory 2 in Southampton indicating diagrammatically the product flow line from raw-material intake to finished-cable delivery to ship.

tests. After testing, the cable tanks are taken out of the water tanks and placed near to the terminating room into which both ends of the cable are passed for the fitting of repeater couplings. The product line plan adopted for the new factory is shown diagrammatically in Figure 2.

## 3. Strength Member and Inner Conductor

There are 4 principal sequential operations to produce a copper tube tightly fitting round a steel-wire strand.

### 3.1 STEEL-WIRE STRANDING

A conventional tubular stranding machine was selected as best suited to deal with 40 bobbins of steel wire, each containing 21 nautical miles (39 kilometres) of wire. In this type of strander, extensively used for stranding conductors of power cables, the bobbins do not rotate in a flat plane about the axis of the strand but are mounted in line in horizontal yokes that float inside a long tubular frame. This tube rotates about the axis of the strand, and the wires from the bobbins pass through guides mounted inside the tube. All 40 of them, plus the centre wire, emerge in a circular array from the output end of the tube. Here the wires pass through a closing die that brings them closely together twisted by the rotation of the tube, all in the same direction, with the specified length of lay of 6 inches (152 millimetres). The wires are of 5 different gauges to give a better space factor (ratio of steel cross-section to total cross-section) than the more-usual single-gauge wire strands. Five of these long machines are shown side by side in the background of Figure 3.

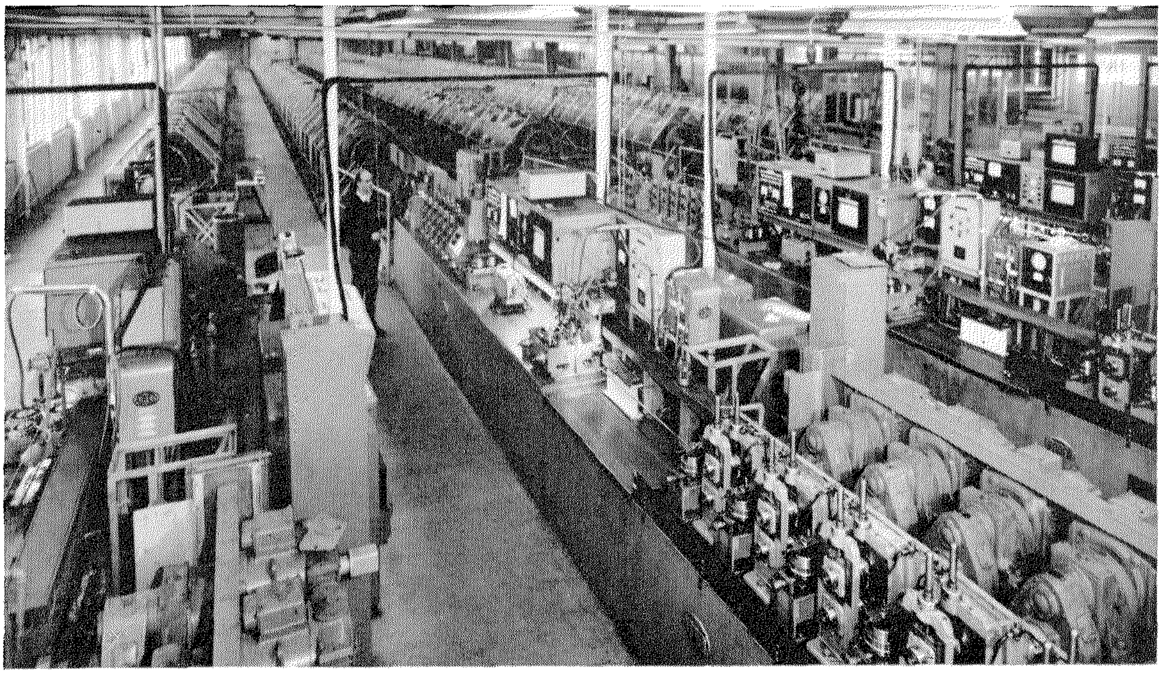


Figure 3—The inner-conductor lines. In the background are the 40-bobbin tubular stranding machines for the steel-wire strength member. In the right foreground is a tape-tubing mill, the argon-arc-welding unit, and the tube-drawing mill with completed inner-conductor assembly leaving the machine at the lower right.

### 3.2 Copper Tape Tubing

Copper tape for the inner conductor is supplied to the line immediately following completion of the steel-wire strand. The tape comes up from the floor below that on which the machines are mounted and passes along in a horizontal plane underneath the strand into a pair of slitting rolls that cut a narrow strip of copper off each edge. Both the tape and the strand then pass into a 6-stage tubing mill, which rolls the copper into a tube 0·50 inch (12·7 millimetres) in diameter enclosing the steel strand and with the butting edges of the tape along the top of the tube.

It will have been noted that the stranding machine does not have to stop to load steel wire during a 20-nautical-mile (37-kilometre) run. The same facility was required of the tubing unit, but this cannot be loaded with 20 miles of copper tape. The tape, being 1·65 inches (42 millimetres) wide, is not readily handled in pads weighing more than about 200 pounds (90 kilogrammes). This means that a new pad of tape has to be welded into the unit at intervals of about 2 hours. To avoid having to stop the machine when this occurs, sufficient copper tape for 15 minutes running is stored in a mechanical accumulator sited between the tape supply and the tubing mill.

### 3.3 Continuous Seam Welding

In Figure 4 the copper tube is shown on the left leaving the last stage of the tubing mill. It moves to the right into the welding enclosure where the butting edges of the tape pass under a direct-current arc torch with a non-consumable electrode operating in a piped supply of argon. The electrode is mounted vertically, and the top of it may be seen projecting above the enclosure. An indirect view of the arc is displayed in the dark-glass periscope unit to the right of the enclosure.

Basically the unit is a proprietary argon-arc welder, but a great deal of development effort was expended locally to obtain sufficiently precise control of the process to make possible non-stop runs of 20 nautical miles (37 kilometres) lasting 5 to 6 days. Brief reference may be made to a few of the difficulties encountered and overcome.

The copper tape is only 0·023 inch (0·6 millimetre) thick, and a sound seam weld will be obtained only when the arc provides the right amount of heat in the right place, directly over the butting edges of the tape. A round tube was found to have a tendency to rotate to and fro about its axis whilst passing through the machine, and at times the tape edges were not under the electrode. This was corrected by making the tube slightly oval. With the tube travelling at constant speed past the electrode there is an optimum value for the arc current; too little gives imperfect welding, and too much burns holes in the copper and may damage the steel strand inside the tube. The machine, however, does not run at constant speed all the time; it has to start and stop, and on occasions it has to run at less than full speed. Long studies were made of arc current as a function of tube speed, and of other variables such as electrode composition, electrode diameter, tip profile, and length of arc. The results of these studies were built into automatic arc-current–tube-speed controllers with facilities also for marginal manual adjustment by the machine driver. Improvements were found necessary to ensure more-complete shrouding in argon of the arc and of the hot copper on both sides of the tape [3].

After being welded the tube passes through a weld integrity checker shown on the extreme right of Figure 4. In this device two similar coils are coaxially mounted round the tube a few inches apart. They form part of a balanced high-frequency network whose output signal is displayed on a cathode-ray oscilloscope. When the portion of copper tube within one coil is in the same physical condition as the portion within the other coil the network is balanced and the oscilloscope trace is a flat ellipse; but when a faulty weld or defect in the copper tape moves into one coil the network balance is upset and an alarm is sounded. There is a second welding unit about 6 feet (2 metres) downstream from the checker at which repairs may be effected. Welding may continue at this torch for a short time should adjustments to the main torch be necessary.

### 3.4 Tube Drawing

From the second welding torch the copper tube passes into a tube-drawing mill, one of which may be seen in the lower right of Figure 3. Here the tube is reduced in diameter from 0·5 inch (12·7 millimetres) to 0·33 inch (8·4 millimetres). At this new diameter the copper is drawn tightly on to the steel-wire strand inside it to form a compact centre-conductor assembly which is taken up in 20-nautical-mile (37-kilometre) lengths on to large steel drums as shown in Figure 5.

The external diameter and the effective thickness of the copper after draw-down are required to be held within very close limits of stated nominal values to ensure that the copper is firmly keyed into the steel-wire strand throughout all lengths. During the drawing-down process the linear speed at which the tube is moving through the machine increases, and the factor principally controlling the

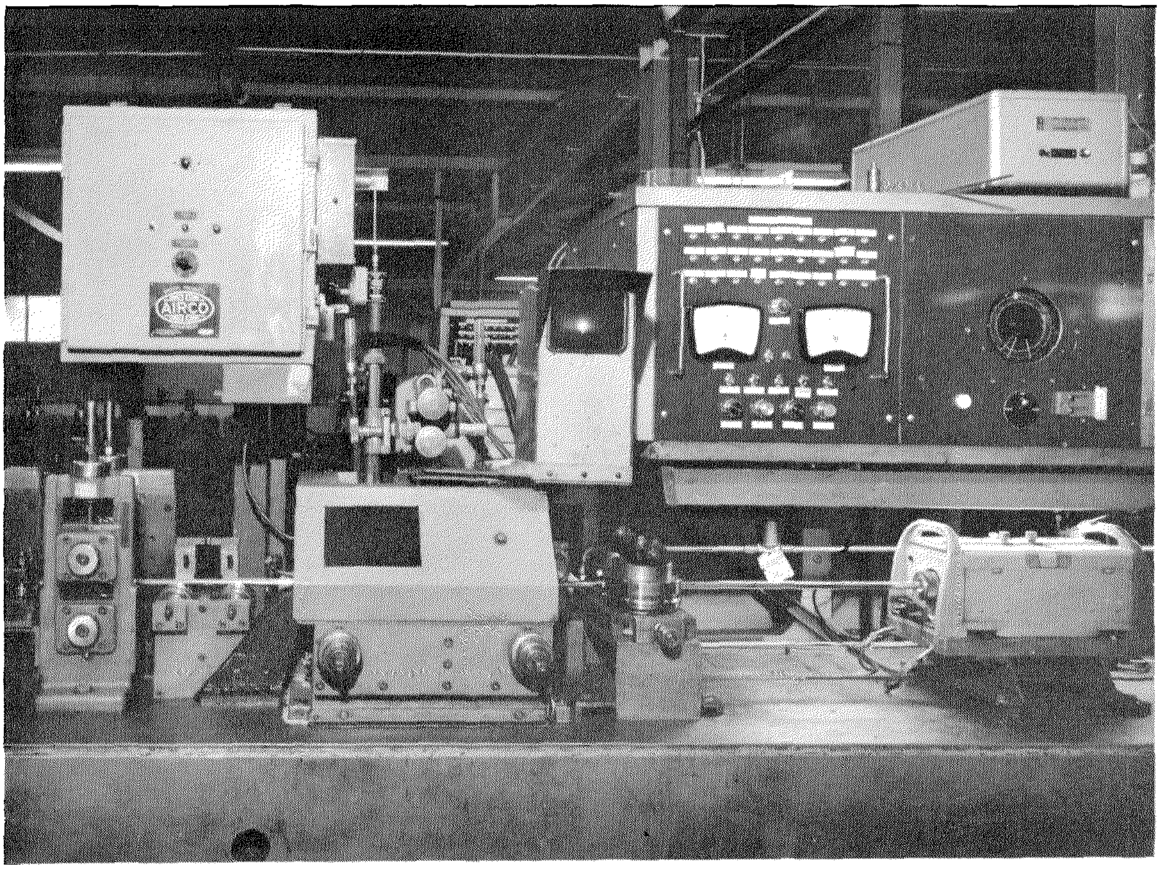


Figure 4—Inner-conductor argon-arc seam-welding unit with tube moving from the last stage of the tubing mill on the left into the welding box. Argon passes to the arc region through the electrode holder, which projects out of the top of the box. The arc is displayed in the viewer to the right. After seam welding the copper tube passes through the weld integrity checker at the extreme right.

dimensions obtained is the ratio of the speed of the finished conductor to the speed of the copper tape entering the welding unit. These two speeds are monitored by a ratiometer device, which displays the ratio on a dial and also provides a command signal to the complex control gear for the various electric motors driving the stranding machine, the 2 tubing mills, and the haul-off caterpillars. Ratio control during starting and stopping proved to be a difficult problem. The copper tube between the welding unit and the haul-off caterpillar is the mechanical link between the high-speed group of units and the low-speed group. It is a weak link, and unless the speed ratio is correctly held during speed changes the tube may either buckle or fracture.

In the tube-drawing mill it is necessary to lubricate the outside of the tube with oil, and then it is equally necessary to remove all traces of it after the tube leaves the final sizing die. This is effected in a trichloroethylene degreasing unit, which first sprays the tube with trichloroethylene and then dries it by air blast.

## 4. Core

Completed drum lengths of inner conductor are lifted by a travelling gantry crane from the take-up stands, Figure 5, and when re-

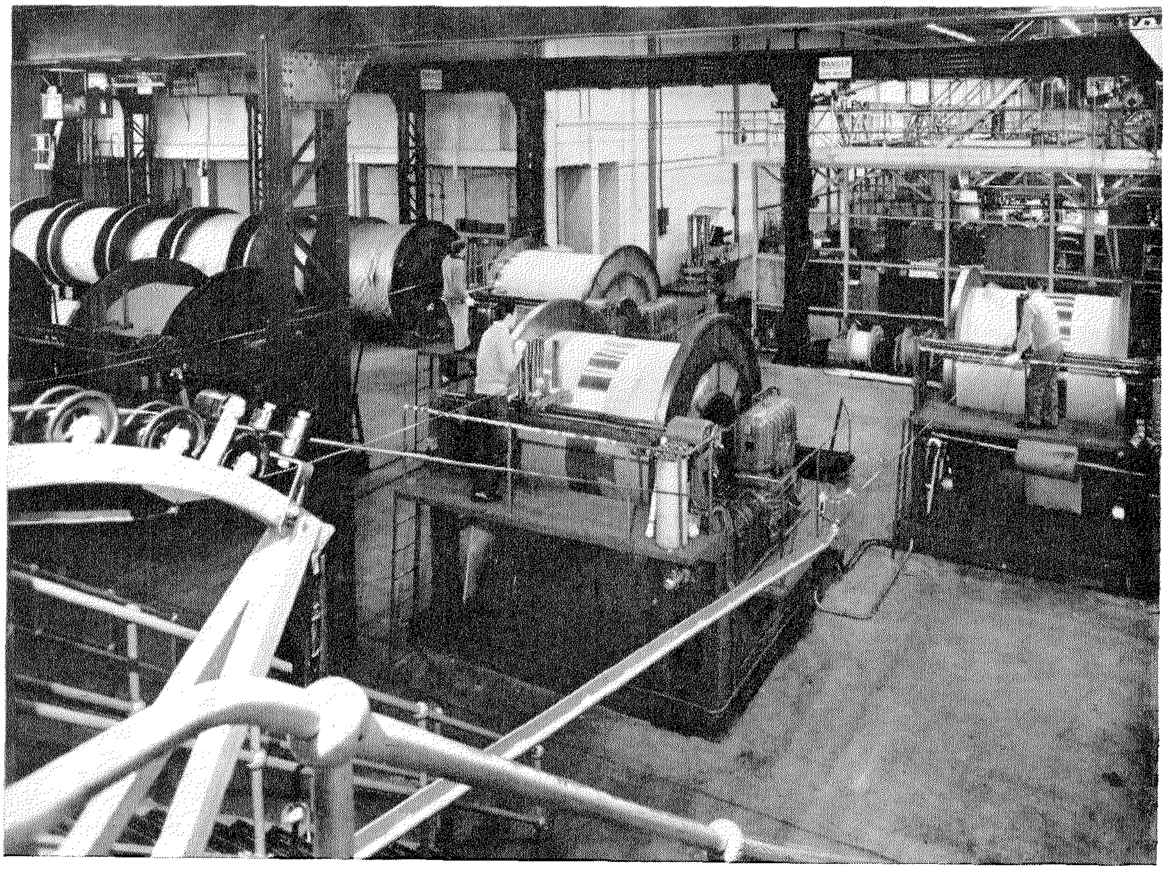


Figure 5—Completed inner-conductor assembly being taken up in 20-nautical-mile (37-kilometre) lengths on to large steel drums. Filled drums awaiting loading into extruder lines are stored in the left background. Drums are moved to left and to right on a 25-ton (25·4 tonne) travelling gantry crane.

quired are placed in the extruder supply stands at the left-hand end of the crane runway. Storage space for completed drum lengths forming the work-in-progress stock is shown in the left background of Figure 5.

The 2 core extruders are 6-inch-long (152-millimetre) screw machines designed to deal effectively with high-molecular-weight (melt index about 0·1) insulating-grade polythene. In common with all the machines in the factory, tractive effort is applied to the product by double-track caterpillar haulers equipped for servo-controlled line tension. (There is one on the right in Figure 8.) These replace the various types of single- and double-wheel capstans more-commonly used in cable making as the product is not subjected to bending and straightening stresses when passing through a caterpillar machine.

The first caterpillar in the extruder line hauls the inner conductor from its drum and pushes it into the extruder head at a linear speed controlled by a diameter measurement on the polythene-insulated core as it leaves the head. There are graded cooling troughs totalling 400 feet (122 metres) in length from the hot end, then a turn-round sheave followed by 400 feet of cold trough back towards the extruder. At the end of the first section there is a concentricity head, Figure 6, comprising two pairs of

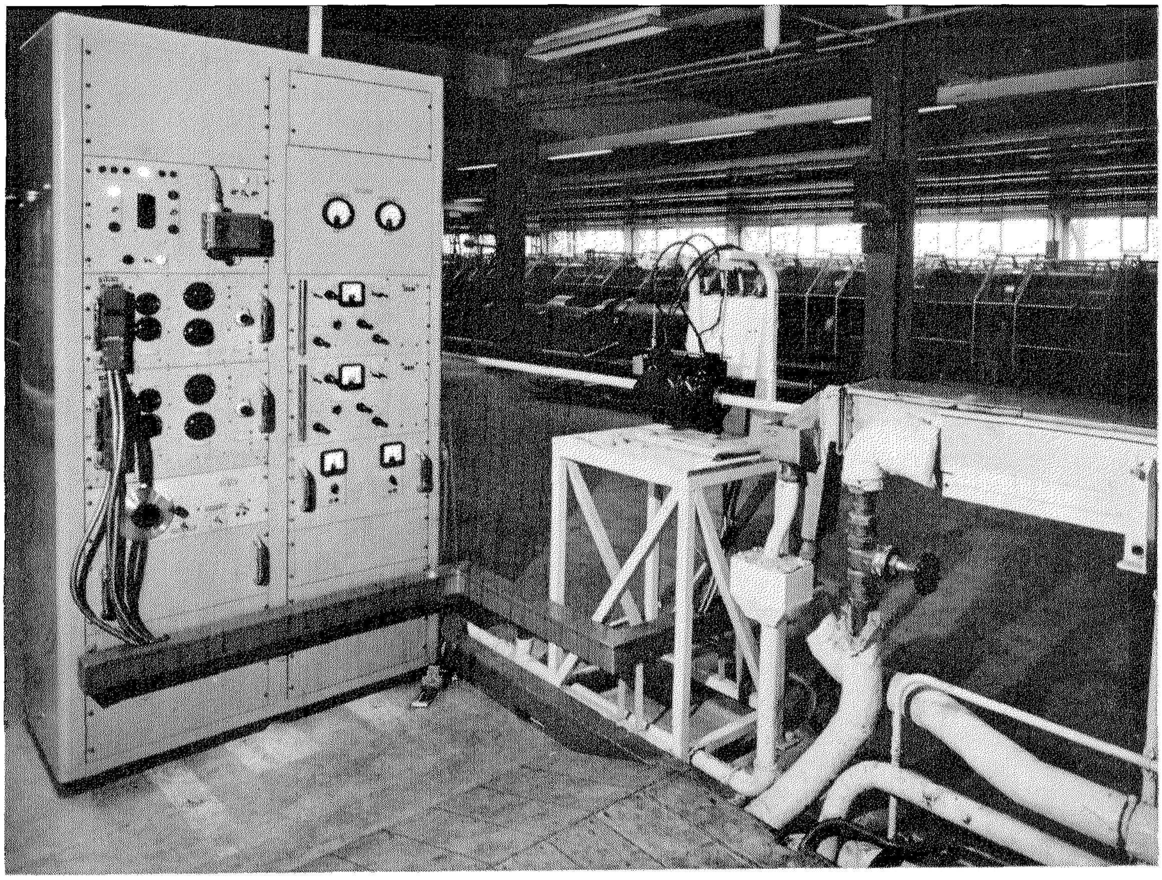


Figure 6—The eccentricity-monitoring head in the core-extruder line. The box containing two pairs of capacitance electrodes rides on the newly formed 1·000-inch-diameter (25·4-millimetre) core when it passes from one section of cooling trough into another section. Two racks of equipment at left integrate the signals from the electrodes and operate recorders (not shown) at the extruder console.

capacitance-sensing electrodes bearing on the core and aligned at right angles to each other. The capacitance measured between each electrode and the inner conductor of the core is a function of the distance between them, and by comparing the four readings the position of the inner conductor with respect to the axis of the polythene covering may be determined. When first setting up the core extruder, adjustments are made to the point and die to give an indicated conductor eccentricity not greater than 0·015 inch (0·4 millimetre) in the 1·05-inch (27-millimetre) core. Thereafter the capacitance readings are charted to provide quality assurance data.

## 5. Shaved Core

It was specified that the diameter of the core as presented to the outer-conductor tubing mill should be 1·000 ± 0·001 inch (25·4 ± 0·025 millimetres) at 20 degrees centigrade, and that the eccentricity (displacement from core axis) of the inner conductor should not exceed 0·005 inch (0·13 millimetre). Such limits are not easily met by core extruders, so a new method of core shaving developed from a process first applied by Standard Telefon og Kabelfabrik of Oslo, Norway, was introduced.

Because of the comparatively large coefficient of expansion of polythene, variations of core diameter due to ambient temperature could be

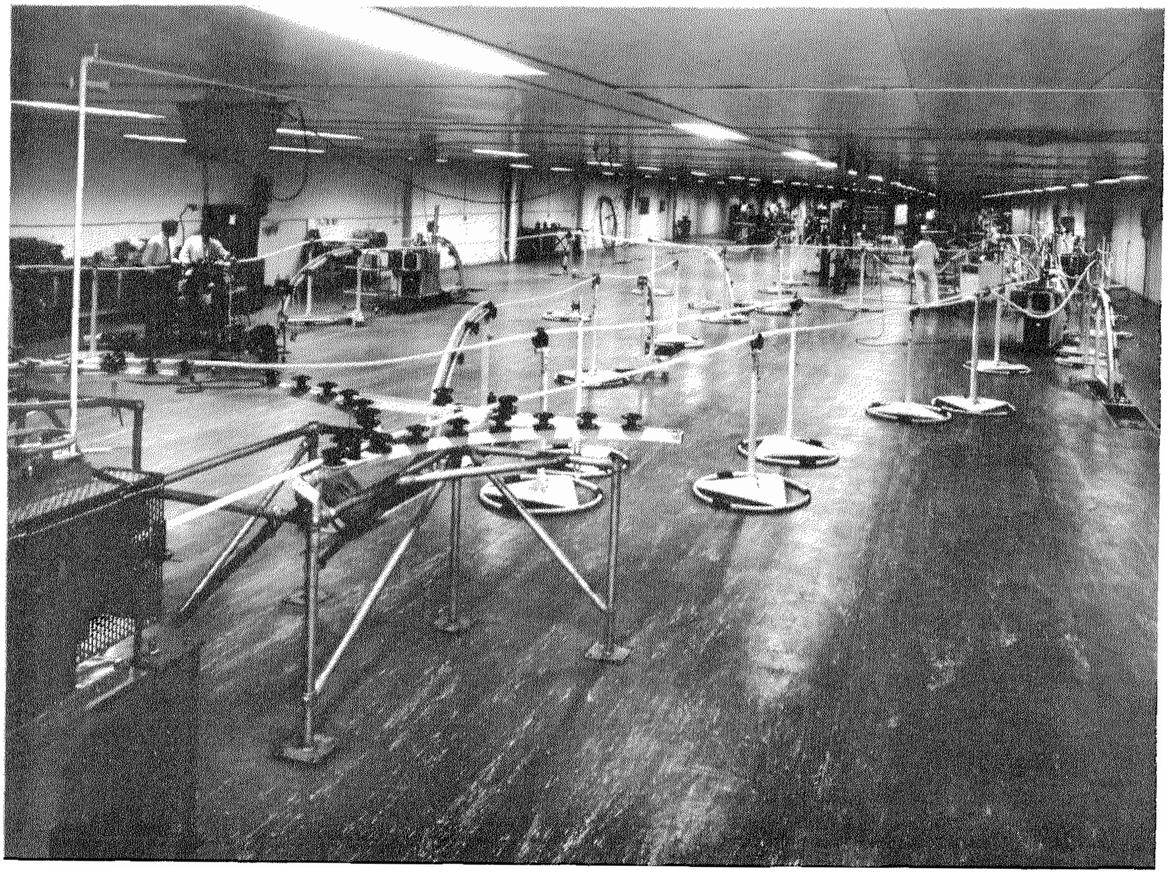


Figure 7—The mezzanine floor within the temperature-controlled core-shaving room. Newly completed cores from the extruder lines are entering the room from the left and go down through the floor into storage creels on the ground floor of the room. Also shown are core-jointing operations.

greater than 0·1 per cent. Thus the first requirement for the shaving process is a controlled working temperature, and to achieve this a large core-shaving room was built equipped with an air-conditioning plant capable of holding the temperature within it to 20 ± 1 degrees centigrade. A view of the mezzanine floor within this enclosure is given in Figure 7.

Cores from the extruders enter at this end of the room and pass down through slots in the floor to coil down in open creels on the floor below, which is within the temperature-controlled area. Each creel will hold a 20-nautical-mile (37-kilometre) length, and there is a storage accommodation for more than 300 nautical miles (560 kilometres). Each completed length may therefore be held here until it has become acclimatized to 20 degrees centigrade. The core may be run from one creel to another by way of the mezzanine floor, as shown, for quality inspection, cutting, jointing, and repairing of extrusion faults, as may be necessary.

When a length of core has passed process inspection and is acclimatized, it returns to the mezzanine floor and passes through one of the

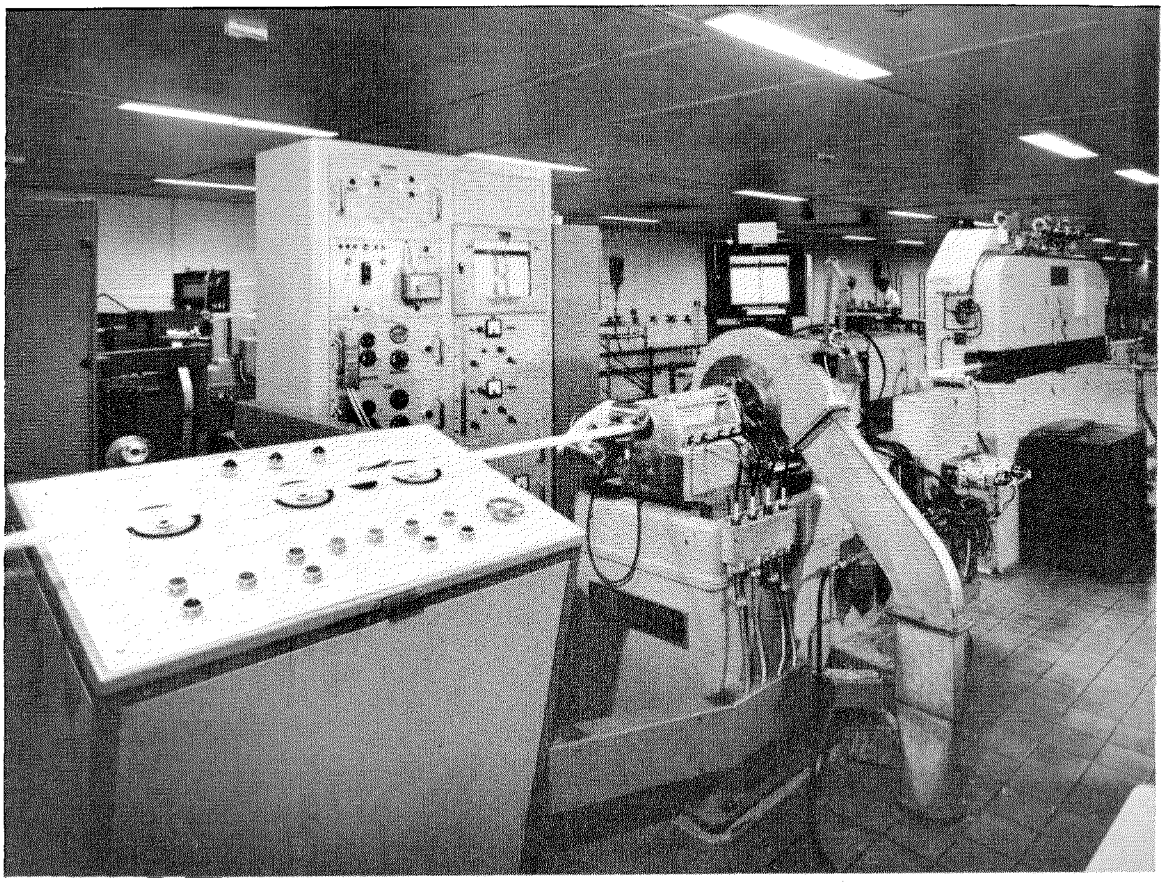


Figure 8—One of the core-shaving machines. At its extreme right is a caterpillar hauler for bringing the core to the shaving head within the circular blower casing where the polythene shavings are exhausted into the ducting going down through the floor. Up against this is the eccentricity head which, with the two racks of equipment behind the line, servo-controls the cutter axis about the conductor axis and records conductor displacement in two planes. Shaved-core-diameter record is behind the cutter head.

4 shaving machines at the balcony end of the floor in the centre of the air-conditioned room. The design of the shaving machine shown in Figure 8 stems from earlier development work at the pilot plant of Bell Telephone Laboratories and it has been further developed locally for commercial production use.

A high-speed 3-blade rotary cutter skims the polythene diameter to $1{\cdot}000 \pm 0{\cdot}001$ inch ($25{\cdot}4 \pm 0{\cdot}025$ millimetres) and a complex servo-controlled mechanism homes the axis of the cutter head around the axis of the inner conductor within the core rather than around the axis of the polythene covering, if the two differ. This takes care of the reduction of the conductor eccentricity to less than 0·005 inch (0·13 millimetre) after shaving.

The command signal for this control is obtained from a capacitance concentricity head (centre of Figure 8) of a type similar to that on the extruder lines. Both the shaved-core diameter and the eccentricity readings are recorded throughout the operation. as may be noted in Figure 8. The charts, as with other processes, form part of the quality assessment made prior to clearance for subsequent processing.

Each shaving head produces a large volume of

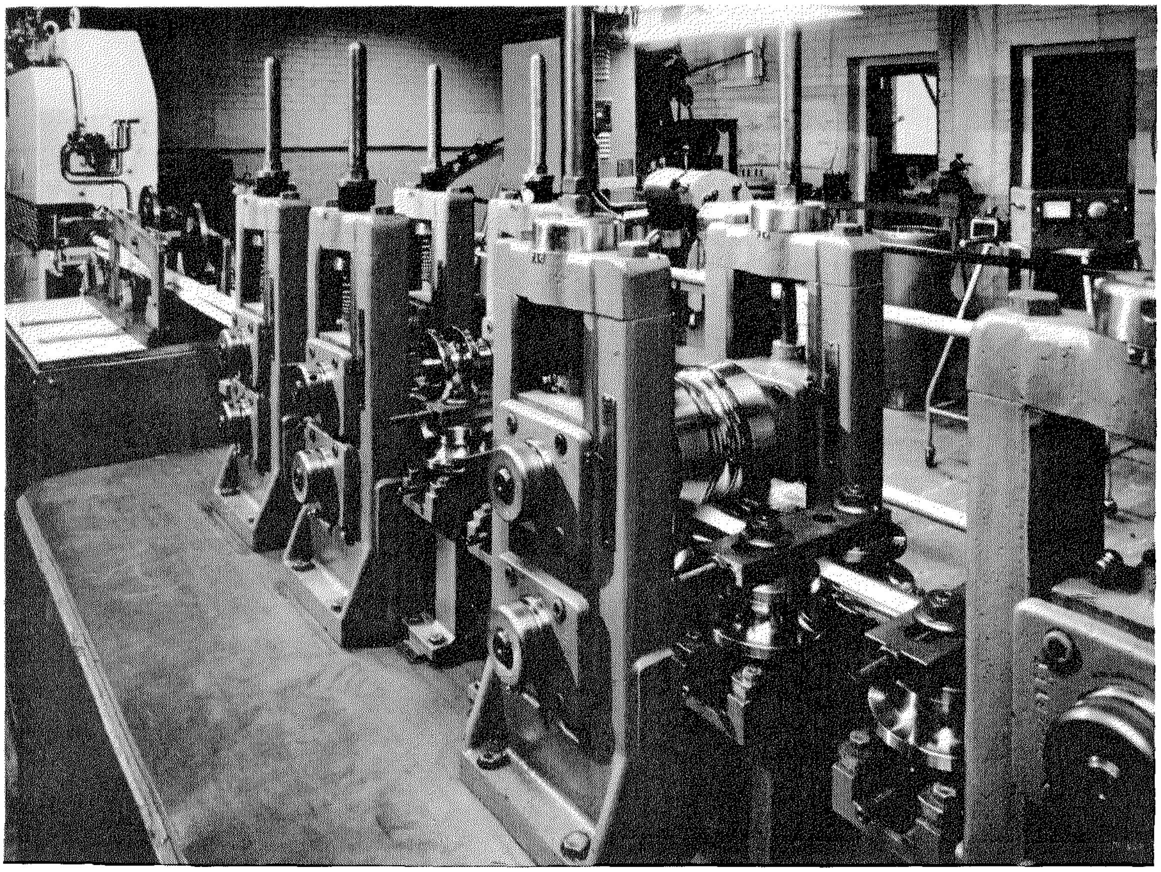


Figure 9—Tubing mill applying outer-conductor copper tape to shaved core coming into picture from caterpillar hauler at top left. Copper tape comes up from floor below where the core leaves the hauler. A ratiometer device monitors the relative speeds of the core and the copper tape as they enter the mill. By holding the ratio to nearly unity excessive stretching of the copper tape during tubing is avoided.

filamentary polythene shavings, which are sucked away from the core and blown down the trunking to the right of Figure 8 into a hopper room adjoining. There were problems here in keeping the polythene shavings moving along the ducting without stoppages caused by "bird nesting," and without acquiring excessive electrostatic charges. If the shavings are not removed, clogging may occur at the cutters, with damage to the core. Another difficulty proved to be maintenance of shaved-core diameter during starting and stopping of the machines; the cutters tended to dig into the polythene at very-low speeds. This is not a serious problem providing that not more than a few inches of core are undercut. A repair to the polythene may then be made in a core-jointing mould.

The shaved cores are coiled down into a second set of storage creels occupying the output end of the shaving room, where facilities for core examination, patching, and jointing are again provided.

## 6. Outer Conductor and Sheath

The most-efficient outer conductor for a coaxial cable is a copper tube and in the new *SD* submarine cable a very-close approximation to this is achieved by folding a single copper tape longitudinally around the shaved polythene core. The edges of the tape are overlapped but not joined together. A view of the tubing mill is given in Figure 9. Here the shaved core enters the mill from the caterpillar pusher (top left). The single copper tape comes up from the floor below and moves into the forming rolls with the core. All this equipment is inside the temperature-controlled shaving room so as to ensure that a core of constant diameter is presented to the copper tape.

When in position round the core the copper tape is not self-supporting, and in consequence the outer plastic sheath of the cable has to be applied, immediately after tubing, by a 6-inch (152-millimetre) extruder driven in tandem with the tubing mill. The completed cable leaving the extruder head is shown in Figure 10. Development problems here were concerned with keeping the copper tape in intimate contact with the polythene core, and free from ripples or waviness whilst passing from the tubing mill to the extruder. The copper tape supplied must accurately conform to specified dimensions and be free from edge bow and surface irregularities. A copper-tape accumulator similar to those on the inner-conductor lines is installed between the pad stand and the tubing mill so that the sheathing extruder need not be stopped whilst new pads of copper tape are loaded into the line. Unlike most cable-making processes, this combined operation is more-successfully carried out at high speed. A 20-nautical-mile (37-kilometre) length can be dealt with in less than 48 hours non-stop running time.

After passing through the extruder cooling troughs, the completed cable is coiled down into one of the repeater-section storage tanks, is tested, and terminated as already described.

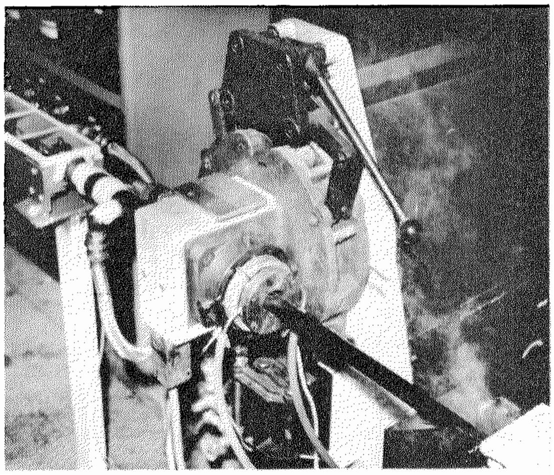


Figure 10—Completed sheathed cable leaving extruder head. This extruder is driven in tandem with the tubing mill in Figure 9. The copper-wrapped core comes from the mill in the enclosed channel way (top left), which is maintained under vacuum to remove air from underneath the copper tape to prevent extrusion blows in the hot sheath.

## 7. Conclusion

A brief account has been given of some of the development problems associated with the commissioning of the second submarine-cable plant at Southampton between 1961 and 1962. On 3 January 1963, the cable ship *H.M.T.S. Alert* left Southampton with 570 nautical miles (1056 kilometres) of the new *SD*-type cable on board. This now forms part of the submarine telephone cable between Port Everglades, Florida, and Kingston, Jamaica; it was completed on 10 February 1963 and was the first *SD* system to be installed.

## 8. Acknowledgements

The author acknowledges that his primary role has been to report on the achievements of the Submarine Cable Division of Standard Telephones and Cables in Southampton under the management of Mr. G. K. Duddridge. Also to be acknowledged is design data made available by the Bell Telephone Laboratories Ocean Cable Group under Mr. A. W. Lebert.

## 9. References

1. E. Baguley, "Submarine-Coaxial-Cable Manufacture at Southampton," *Electrical Communication,* volume 37, number 4, pages 316–335; 1962.

2. J. Kelly, Gordon Radley, G. W. Gilman, and R. J. Halsey, "Transatlantic Telephone Cable," *Proceedings of the Institution of Electrical Engineers,* volume 102, Part B, pages 117–138; March 1955.

3. British Patent Applications 20742/62 and 28060/62.

**Eric Baguley** was born in Lancashire, England, in 1905. He graduated in electrical engineering from Manchester University in 1924.

On graduation, he joined Western Electric Company in London and was soon assigned to the engineering of communication cables. Since then, he has been concerned with all types of communication cables.

From 1927 to 1932, he was with ITT Laboratories at Hendon. He was then transferred to Standard Telephones and Cables and in 1940 was placed in charge of the telephone cable development laboratories at North Woolwich. In 1955, he was transferred to the newly formed submarine cable division at Southampton, where he is now chief cable engineer.

Mr. Baguley is a Member of the Institution of Electrical Engineers.

# Telepuls and FMX Modular Electromechanical and Electronic Remote Control

**K. BARTELS**

Standard Elektrik Lorenz, AG; Pforzheim, Germany

## 1. Introduction

As customer requirements in Germany for remote control and indication systems were found to be not too divergent, systems were developed using standard components that could be combined to meet the most varied needs of such organizations as power companies, railroads, the post office, and gas, water, and pipeline operators.

Three systems were developed. Two of them operate with pulse-code transmission in binary form and are compatible with each other. The system called Telepuls 10 is electromechanical and the second system, Telepuls 12, is fully electronic. Using semiconductors throughout, Telepuls 12 operates at 1200 bauds compared with 25 bauds for Telepuls 10. Any combination of the standard units of the two systems may be used as shown in Table 1. This facilitates a progressive conversion from electromechanical to electronic operation. Furthermore, fully electronic substations are able to interwork with electromechanical central stations, which often are preferred for their particular methods of display.

The third system employs frequency multiplexing, which is advantageous if a number of substations along a party line require only one, or a few, indications each. This frequency-multiplex system, *FMX1,* may also be used for the transmission of pulse signals at speeds below 25 bauds.

## 2. Telepuls 10

In the design of the new Telepuls 10 system [1], experiments made with previous systems and an extended analysis of customer requirements were evaluated. The system must permit point-to-point, parallel, and star operation of substations, and transmission of digital and analog values must be possible over the indication path when it is not engaged with the transmission of indication signals. The code used must have a high degree of error detection but must avoid excessive redundancy. The elements for detecting changes in the positions of the indicating contacts in the substations must be very reliable. Operation must be ensured over a wide range of voltage fluctuations and with distorted pulses. The equipment must be divided into a minimum number of units that permit any combination within a system as well as interworking with the Telepuls 12 with only simple wiring changes.

TABLE 1

POSSIBLE COMBINATIONS OF TELEPULS 10 AND TELEPULS 12

| Central Station | Converter | Converter | Substation |
|---|---|---|---|
| Control Path | | | |
| electromechanical | electromechanical | electromechanical | electromechanical |
| electromechanical | electromechanical | electronic | electronic |
| electronic | electronic | electronic | electromechanical |
| electronic | electronic | electronic | electronic |
| Indication Path | | | |
| electromechanical | electromechanical | electromechanical | electromechanical |
| electromechanical | electromechanical | electronic | electronic |
| electromechanical | electronic | electronic | electronic |
| electronic | electronic | electronic | electronic |

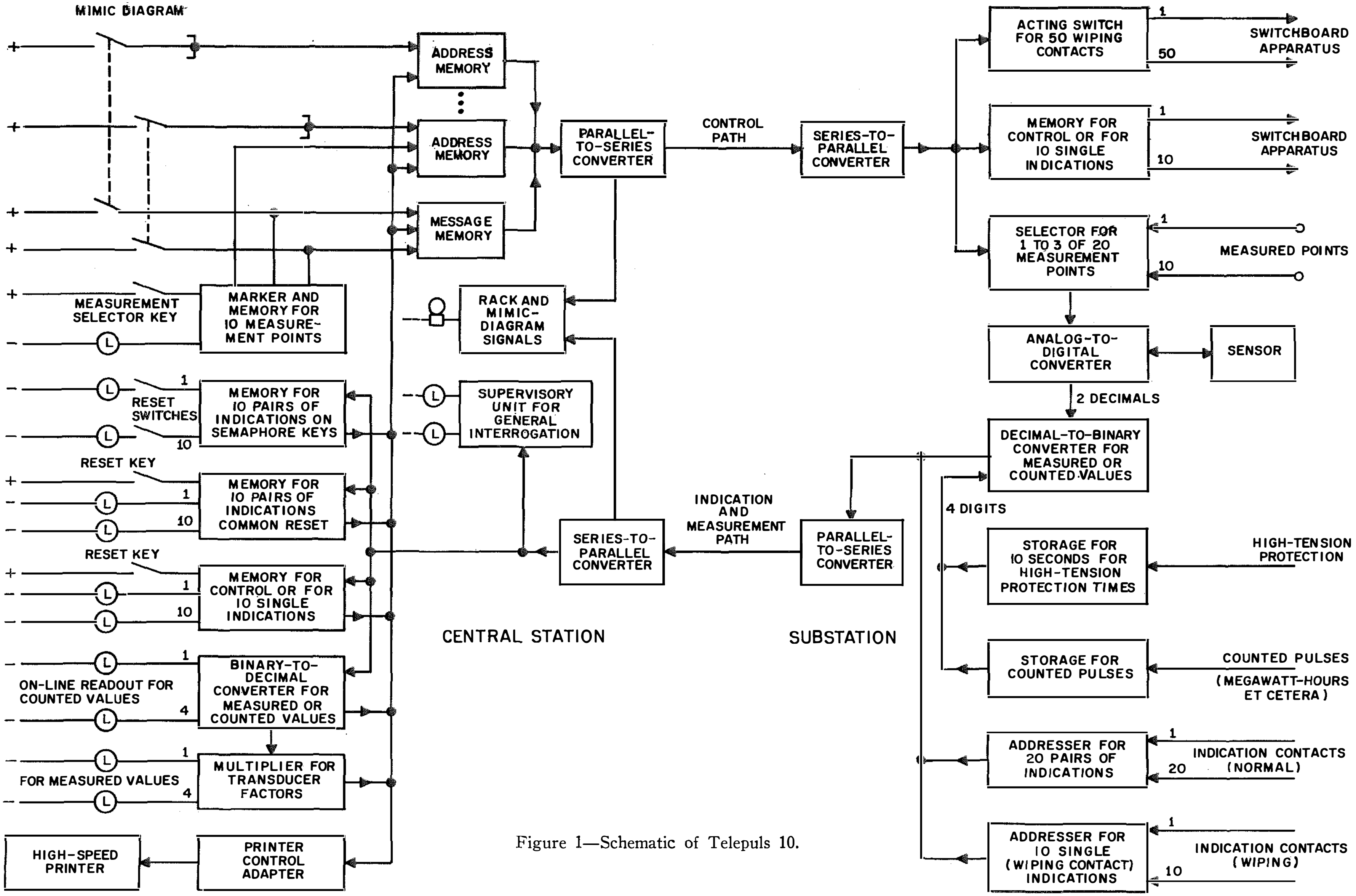


Figure 1—Schematic of Telepuls 10.

Figure 1, a schematic of the Telepuls 10 system, shows separate control and indication transmission paths.

In all pulse trains, the binary 0 is represented by a short mark and a short space each of 40 milliseconds. The binary 1 is a long mark and a long space, each of 160 milliseconds. A time detector at the receiver guarantees even greater safety against transmission errors than is achieved by the conventional method of a backward retransmission and a subsequent forward acknowledgment signal [5].

Each pulse train contains an address in the form of a 2-out-of-10 code, thus permitting 45 different combinations, and a message consisting of one or several groups of 10 pulses and spaces arranged in a code suitable for the type of information to be transmitted.

Control orders are given by operating a key or semaphore key, indications associated with the keys being shown by lighted or unlighted lamps. The address and message, the latter in a 1-out-of-10 code, are passed to a parallel-to-series converter, which transmits the two parts sequentially to the series-to-parallel converter of the receiver. Figure 2 shows an example of such a pulse telegram. After checking the pulse train for transmission errors, the decoded information is used to actuate the control equipment of the address unit.

Indications of contact position changes are reported by the control equipment to an addresser. Here a parallel-to-series converter forms the address and message, the latter containing in an *n*-out-of-10 code all indications concerned with this address. For each contact in a group of 10, the report is either a 1 for being closed or a 0 for being open. In the receiver, the message is converted into parallel output and, after a code check, is compared with the existing indication display associated with the address. Any change of information is then normally displayed by a flashing light.

In addition to control orders and indications, decimal counts and measured analog values may be transmitted from a substation to the central station. For transmission of counted values, a pulse counter in the substation stores the pulses, for example, those generated by a kilowatt-hour meter, and adds them up to a decimal figure. To handle measured analog values, the analog quantity is converted to a direct voltage between 0 and 10 volts, which is transformed by a special analog-to-digital converter into a decimal figure. For the decimal figures obtained from decimal counters and from analog values, each decimal digit is represented by marking 1 of 10 terminals. This marking is then changed by a converter into a 4-bit binary code. Thus two decimal digits are transmitted by the standard group of 10 information pulses, leaving 2 pulses for other purposes. Decimal figures with more than 2 digits are transmitted by a series of several 10-bit groups. In the central station, the receiver performs the reverse code conversion, resulting in a decimal output. It then marks 1 of 10 terminals for each decimal digit, which may be displayed by an on-line readout, a printer, or a tape perforator.

Auxiliary facilities are provided to check the proper operation of the system and to signal

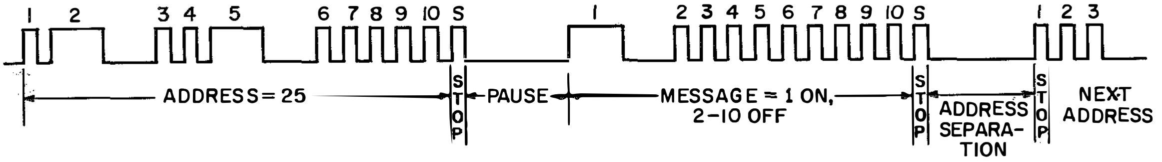


Figure 2—Example of a pulse-train. Each pulse and its following space of equal length represents a character. The short pulses and spaces are 40 milliseconds each and signify 0. The longer pulses and spaces are of 160 milliseconds and signify 1. The address-separation space is at least 320 milliseconds long.

such faults as blown fuses, transmission interruptions, and code failures. The transmission speed of 25 bauds permits operation on ordinary signalling and telegraph channels with a high safety margin; the average time for transmission of an address and message is about 3 seconds.

The mechanical design complies with German Industrial Standards *DIN 47432* and also permits installation of the units in racks conforming with International Telephone and Telegraph System practices. Figure 3 shows typical units for full supervision of an unattended 20-megavolt-ampere transformer and distribution station [4].

## 3. Telepuls 12

As mentioned previously, the electronic system, Telepuls 12 [2], is compatible with the electromechanical system, Telepuls 10. Accordingly, they use the same coding and have corresponding subunits.

Figure 4 is a schematic diagram of Telepuls 12; its similarity to Figure 1 is obvious. It is significant that in most cases there is a direct substitution of an electronic unit for a corresponding electromechanical unit. However, some of the units of Telepuls 10 are retained in Telepuls 12, notably the substation memory for the controls made by contacts and in the

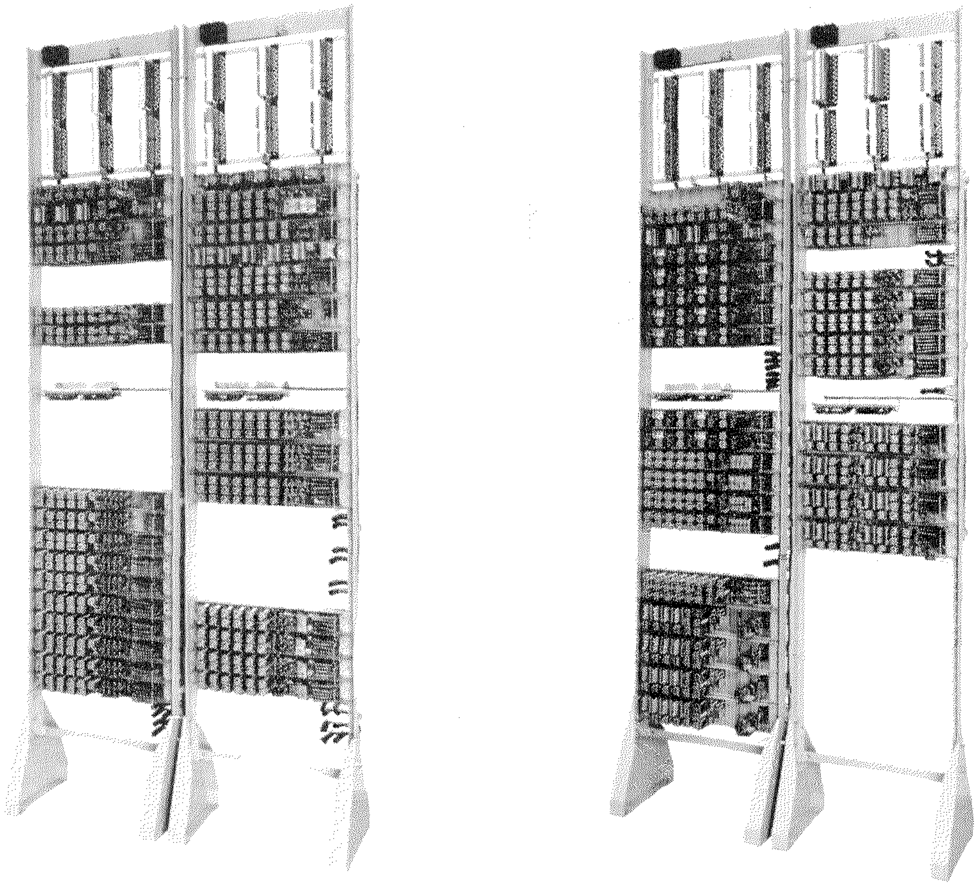


Figure 3—Racks for central and substation installations for Telepuls 10.

central station the memories for lamp indications on a board (flashing light, continuous light, acoustic signals, et cetera). It is not always economical to use transistors, which are most effective if indications are to be given by a change-over from one lamp to another.

In Telepuls 12, the magnetic counters of the series-to-parallel and the parallel-to-series converters used in the Telepuls 10 equipment are replaced by conventional shift registers. While Telepuls 10 uses relay flip-flops to generate pulses and relays with time delay to check the

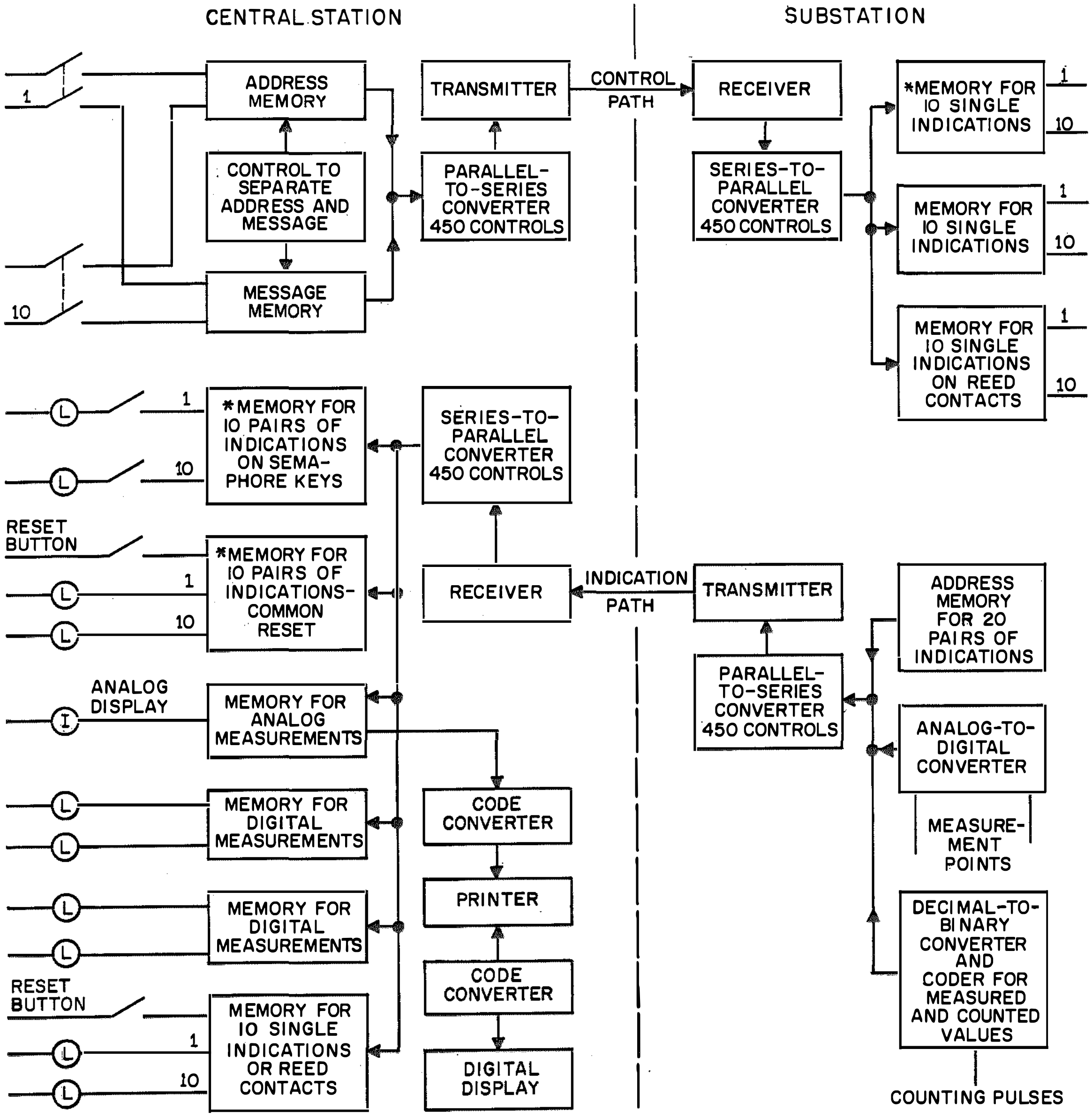


Figure 4—Schematic of Telepuls 12. All the units are electronic except those marked by an asterisk, which are electromechanical units of the Telepuls 10 System.

long and short marks, Telepuls 12 uses single-shot multivibrators, with coupled units in the master clock and single units for checking the marks.

Figure 5 shows how the pulse lengths are distinguished. At the input is shown a binary 0 of 40-millisecond mark and space followed by a binary 1 of 160-millisecond mark and space. The capacitor and rectifier in the left-hand branch permit the positive-going wave front at the start of a pulse to actuate the single-shot multivibrator to produce a 60-millisecond pulse, which is compared with the original signal by the AND gate.

In the right-hand branch, the input pulses are inverted and what was originally the negative-going wave front at the end of the pulse actuates a similar multivibrator, the output of which is also compared with its input by an AND gate. So long as there is an uninterrupted sequence of 40-millisecond marks and spaces, there will be no signal to operate the flip-flop but 160-millisecond marks and spaces will turn the flip-flop to mark and then to space. Thus the binary 1 is distinguished from the 0.

Using these electronic means, the transmission speed is limited only by the characteristics of the connecting telephone channel. At 1200 bauds, which is one of the speeds recommended by the Comité Consultatif International Télégraphique et Téléphonique for data transmission, it takes less than 0.1 second to transmit a pulse telegram.

The mechanical design complies with International Telephone and Telegraph System practices and is also compatible with German Standards *DIN 47432*. Figure 6 shows the units of a central station and of a substation, which have been installed together in one rack

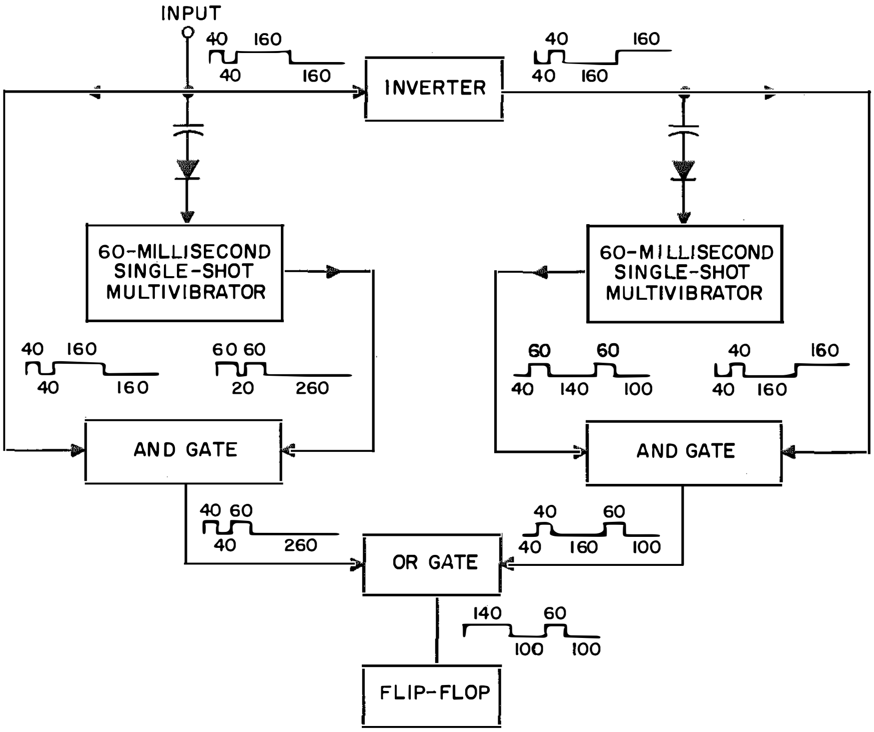


Figure 5—Distinguishing between 0 and 1 pulse lengths in Telepuls 12. The mark and space durations are given in milliseconds.

for testing purposes. The upper part comprises the equipment for the central station. These units are combined with Telepuls 10 panels, which provide for the controls and indications.

## 4. FMX1

As a reasonable relationship must exist between the cost of an entire installation to be supervised and that of the supervisory equipment, pulse-code transmission is uneconomical for very-small stations. Meshed power networks, for example, have a large number of stations that are required to transmit only 1 indication to a central station. For such installations, the *FMX1* system [3], which uses frequency separation of signals, was developed.

On a party line, each substation employs 1 or a few audio frequencies out of a total of 15 to identify the station and particular equipment being reported on. Reception of each tone at the control station indicates the state "good" or "on" and absence of the signal designates a state of "bad" or "off." Such indications are frequently stored in subcentral stations for common transmission by a pulse-code system to the dispatching center.

The *FMX1* system uses transistors exclusively. The signalling frequencies follow a geometric progression being separated by the factor of 1.224. This permits the use of very-simple filters and renders transmission distortion unimportant. Furthermore, the filter loss is almost the inverse to the cable loss for that frequency, thus permitting any required distribution along the party line. The generators and receivers have shunt-type impedance characteristics, that is, they have high input or output impedances outside their pass bands. Thus, no restrictions need be placed on their distribution along the line. Power consumption, especially of the generators, has to be very low. It was further demanded that it should be possible to use different supply voltages (6, 12, 24, or 60 volts) with a wide tolerance of ±15 percent. The system uses neither contacts exposed to the atmosphere nor components subject to wear.

Figure 7 shows the four subunits. Only the

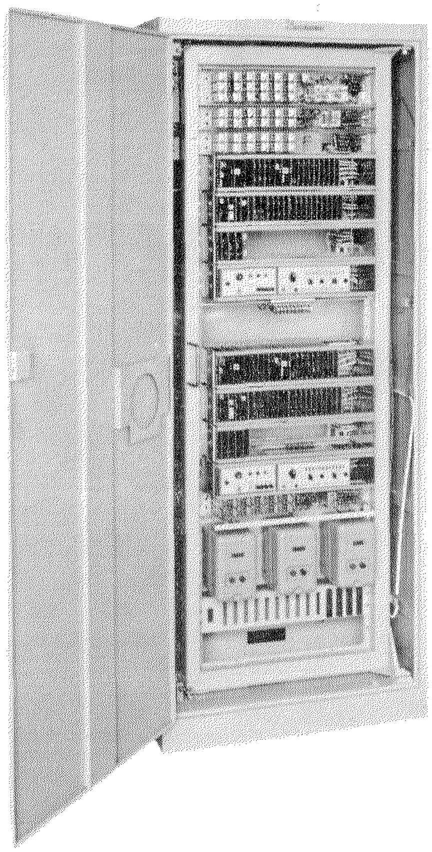


Figure 6—Telepuls 12 bay housing both the central and the substation equipments for testing purposes.

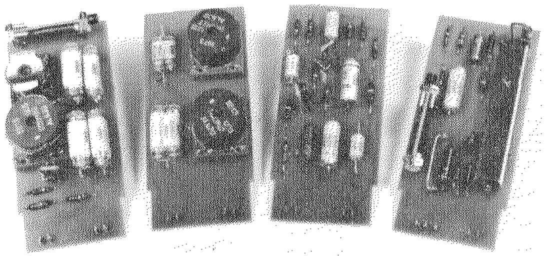


Figure 7—Subunits of *FMX1* system.

generator (left) and the band-pass filter (second from left) differ for the various frequencies. The amplifier and the alternating-to-direct-current converter (right) are identical for all 15 frequencies. Panels and housings, especially of the generators in the substations, are available in various designs to permit interworking of the receivers with the Telepuls 10 or Telepuls 12 components.

## 5. References

1. Bartels and Oklobdzija, "Telepuls 10, das Ergebnis einer Entwicklungsreihe electromechanischer Fernwirksysteme," *SEL-Nachrichten,* volume 3, pages 138–144; 1961.

2. Trautwein, "Telepuls 12, ein elektronisches Fernwirksystem," *SEL-Nachrichten,* volume 3, pages 145–153; 1961.

3. Hitzinger, "FMX1, ein Fernwirkgerät für Frequenz-Multiplex," *SEL-Nachrichten,* volume 3, pages 154–157; 1961.

4. Zube, "Fernwirkanlagen mit kombinierter Meldungs-, Mess- und Zählwertübertragung in Mittelspannungsfreileitungsnetzen," *Nachrichtentechnische Fachberichte,* volume 25, pages 20–24; 1962.

5. Marko, "Die Fehlerkorrekturverfahren für die Datenübertragung auf stark gestörten Verbindungen," *Nachrichtentechnische Fachberichte,* volume 25, pages 101–108; 1962.

**Konrad Bartels** was born in Bischofsburg, Germany, on 12 October 1919. He studied at the Technische Hochschule in Danzig.

After the second world war, he joined Standard Elektrik Lorenz in Pforzheim, where he is now chief of the laboratories for auxiliary equipment for radio links and remote-control equipment.

Mr. Bartels represents Standard Elektrik Lorenz in the study group on remote control of the Nachrichtentechnische Gesellschaft.

# Digitel 108, A Digital Telemetering System

J. LHOEST
B. MATTLET

**Bell Telephone Manufacturing Company; Antwerp, Belgium**

## 1. Introduction

In telemetering, there are substantial advantages in converting the original analog values of the quantities under consideration immediately into digital or numerical values for transmission with modulation methods that provide protection against noise, that are not affected by distortion due to variations in line characteristics, that permit simple error-correction codes to be used, and in which several channels may be multiplexed for prompt transfer of information.

## 2. Digitel 108 System

A telemetering system must be able to handle all kinds of physical quantities, such as, current, voltage, power, pressure, flow, temperature, resistance, displacement, level, counting, et cetera. Transducers convert these quantities into proportional values of direct current that then actuate analog-to-digital converters.

The transducer may be a simple directly coupled adjustable resistor, thermocouple, differential transformer, or a more-complicated electrodynamic system.

For convenience, 8 inputs have been selected as a basic group. This is a good average of the needs met in practice; for larger installations multiple groups may be employed.

A precision within ±0.5 percent has been adopted as being adequate for almost all industrial applications. Frequently, the telemetering system includes commercial recording equipment that has an accuracy within only 1.5 percent.

To make maximum use of high-cost transmission lines, time multiplexing is employed with several channels of coded information being sent in rapid predetermined sequence.

Existing frequency-modulated telegraph channels conforming with the recommendations of the Comité Consultatif International Télégraphique et Téléphonique may be used. These 120-hertz channels, which may be part of a 4000-hertz telephone channel, are sufficiently wide to accommodate several telemetering signals simultaneously.

Binary coding has been selected as having the highest information capacity for a given number of pulses. For error detection, the address, message, and such other pulses as may be required by the transmission system are treated as a complete telegram or parity block that must contain an even number of pulses, an extra parity pulse being added at the transmitter if the count would otherwise be odd. The loss of a pulse or the appearance of a parasitic pulse in the received block will cause its rejection.

The transmitted measurements are usually indicated on moving-coil instruments of various types. The trend is to present them in digital form by illuminated figures or to print them, particularly if there are large numbers of measurements and if it is desirable to maintain written records to provide for later analysis of operations. Ultimately, it should be possible to record the measurements on punched cards, on tape, and in some other similar form suited to data processing. If permanent records are not made, it is possible to store on a loop of magnetic tape sufficient transmitted information to permit an analysis to be made of any important occurrences.

## 3. Description of System

The Digitel 108 telemetering system is divided into plug-in units corresponding to functional elements.

### 3.1 TRANSMITTER

As shown in Figure 1, the elements to be measured $X1, X2, \ldots, X8$ are permanently connected to appropriate transducers that indicate their states by proportional voltages $V1, V2, \ldots, V8$.

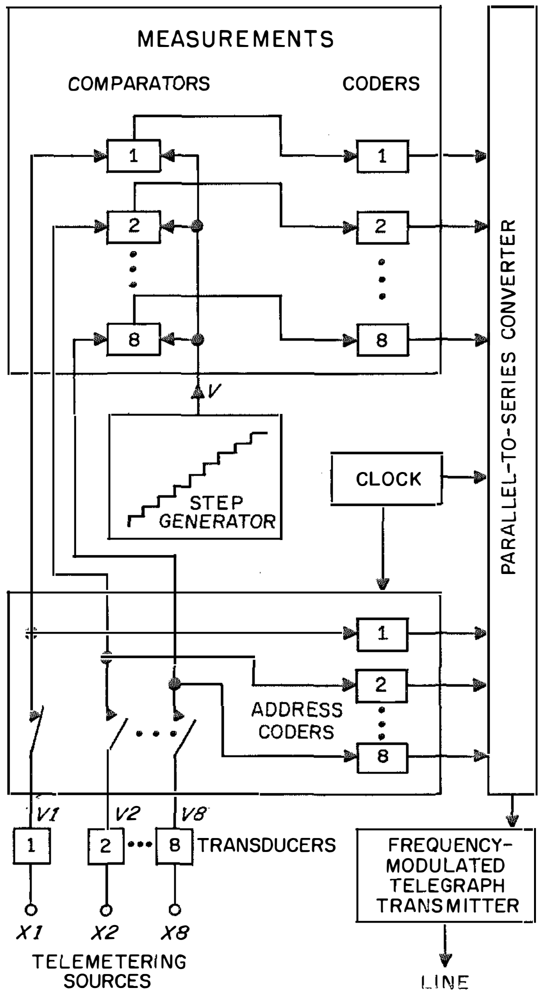


Figure 1—Transmitter. Under control of the clock, each of 8 units to be reported on is examined sequentially and its address and condition are transmitted as a frequency-modulated binary-code telegram to the receiver over a 120-hertz telegraph channel.

The address coder electronically switches each channel voltage *V1 . . . V8* sequentially to the measurement coder. As each channel is so connected, the appropriate address set up on bistable flip-flop circuits in binary code is connected to the parallel-to-series converter so that the address and message may be transmitted over the telegraph channel as a single telegram.

The measurement coder includes 8 comparators to which the proportional voltages of the channels are switched by the address coder. Each voltage is compared with *V*, a stepped voltage wave produced in the step generator. The step matching the proportional voltage to the comparator causes the appropriate output set up in binary code to be connected to the telegraph channel. The step generator is in principle a 7-bit counter controlling transistor switches that connect 7 resistors in a voltage-divider arrangement.

The parallel-to-series converter accepts both address and message and converts the binary signals into serial form for delivery to the telegraph transmitter. All the timing for the binary pulses and the switching sequence is provided by the clock, which is a crystal-controlled oscillator and frequency divider. Figure 2 shows the content of a telegram. A start pulse is followed by a group code to identify the group of channels if more than 8 inputs are involved. Following is the address code, the measurement code, and the parity pulse if needed. A synchronization pulse is then transmitted until the next start pulse. The space pulses are part of the synchronizing system.

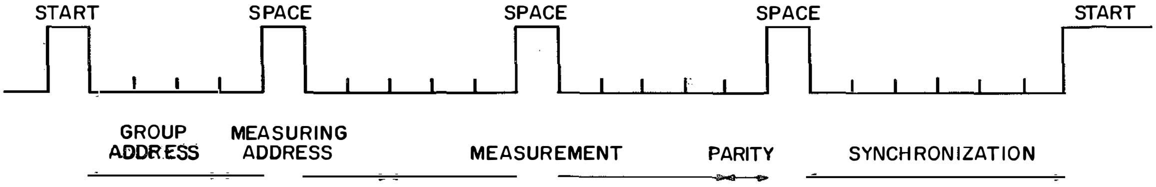


Figure 2—Typical binary-code pulse telegram.

The start and spacing pulses are transmitted at a voice frequency $f_n$, the binary 0 at $f_n-30$ hertz, and binary 1 at $f_n+30$ hertz.

### 3.2 RECEIVER

After demodulation, as shown in Figure 3, in the telegraph channel receiver, the pulses go to the distributor, where serial-to-parallel conversion takes place. The message and address are written in the bistable memories of their respective decoders, the measurement code being stored temporarily in a buffer memory while parity and synchronization are being verified.

The distributor, under control of the clock and synchronizer, connects the proper address switch in the address coder to the display converter and initiates the transmission of the corresponding measurement from its decoder to the display converter. Each display decoder receives its message in the original sequence established at the transmitter.

Each display converter can be equipped as an analog device giving a proportional output current corresponding to the encoded message value; producing an indication on a milliammeter or recorder. It may, optionally, contain an amplifier to operate relays of a binary-to-decimal converter to control luminous display units or a printing register, or both.

## 4. Construction and Maintenance

The use of plug-in printed circuits and transistors provides great flexibility in design of systems and in maintenance. Routine testing is possible with automatic program control and indicator lamps. Test points are provided for oscilloscope comparisons with reference pulse trains reproduced in an instruction manual. Thus, effective and rapid testing together with ready replacement of plug-in units minimizes out-of-service time.

Figure 4 shows a plug-in unit having 20 test points along the side and a switch for checking by the numbered light indicators on the panel the main functions and calibration of the unit.

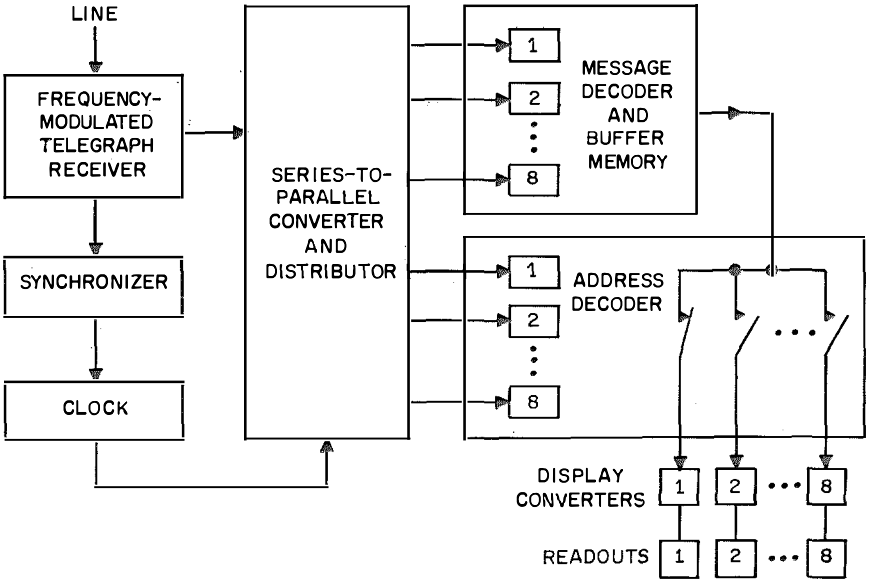


Figure 3—Receiver that reverses the processes accomplished in the transmitter.

Some typical printed-circuit boards are shown in Figure 6.

## 5. Combined Telemetering and Signalling

In many applications, remote signalling may be required in addition to remote measuring. Pipelines, for instance, may require supervision of many pumping stations and valves to control the flow. For such cases, the 7 information bits corresponding to the message in each telegram can be considered on a binary basis to indicate on-and-off conditions. Thus a group of 56 such reports can be accommodated by a full group of 8 channels. Using the full capacity of the group code, 448 on-and-off conditions can be transmitted. At the receiver, the signals are recorded on relays.

## 6. Applications

As examples, two applications are outlined, one being for a purely scientific purpose and the other is industrial.

### 6.1 RADIO ASTRONOMY

Radiation from astronomical bodies is not limited to light and heat but may be found in the radio spectrum. The radio telescope has consequently become an important astronomical tool and the Belgian Royal Observatory in Uccle has installed such equipment.

Signals are received from our sun on paraboloidal antennas tuned for operation at 153, 408, 600, and 900 megahertz. A system of 48 paraboloidal antennas in a cross arrangement operates as an interferometer at 408 megahertz.

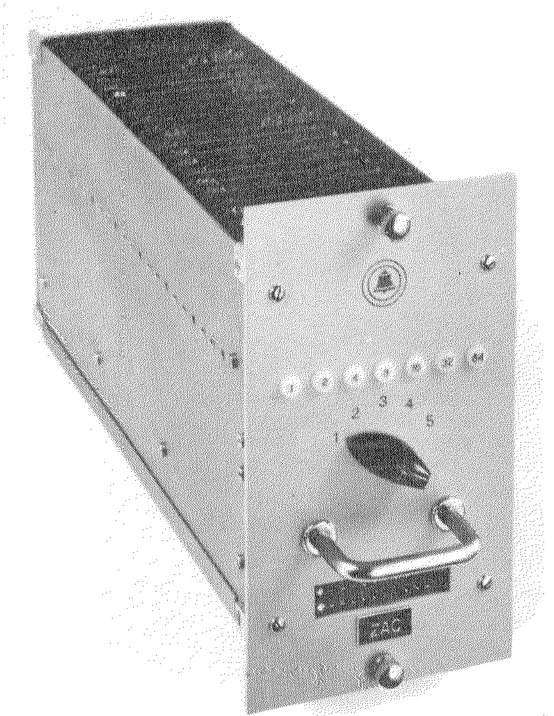


Figure 4—Typical plug-in unit with numbered lights on the panel and a row of test points along the side to facilitate testing and calibration.

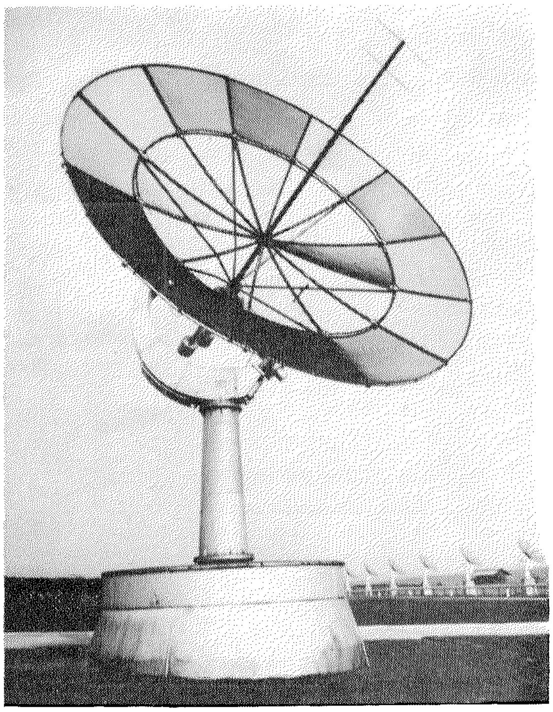


Figure 5—Paraboloidal antenna and in background part of 48-antenna array used for radio astronomy at the Humain field station of the Belgian Royal Observatory.

Another antenna working at 27 kilohertz is used for the study of signals originating in thunderstorms.

The information obtained from these antennas goes to a central control room to operate potentiometer-type recorders. The slide of an auxiliary potentiometer is coupled to certain of the recorders to provide sources of potential for telemetering transmissions.

Telemetering reports on 12 measurement systems are transmitted from the station at Humain to the observatory at Uccle over a 120-kilometer (75-mile) 2-wire telephone line. A 120-hertz band at a nominal frequency of 2460 hertz is used, the lower frequencies of the transmission path being retained for a telephone circuit with its attendant signalling frequencies.

At the receiver, the 12 telemetered reports are recorded on a 12-pen recorder. A punched-tape perforator will be added to permit simple transfer of the material to a computer.

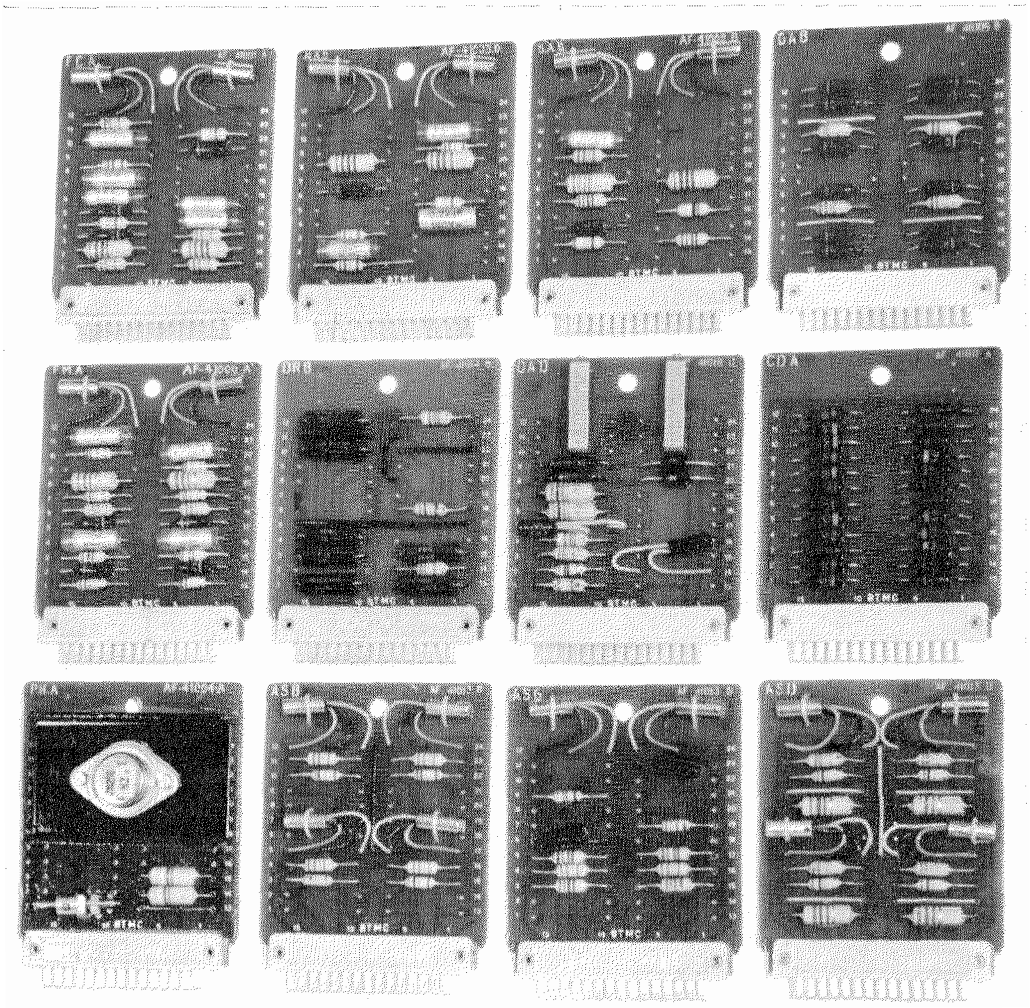


Figure 6—Assortment of printed-circuit panels.

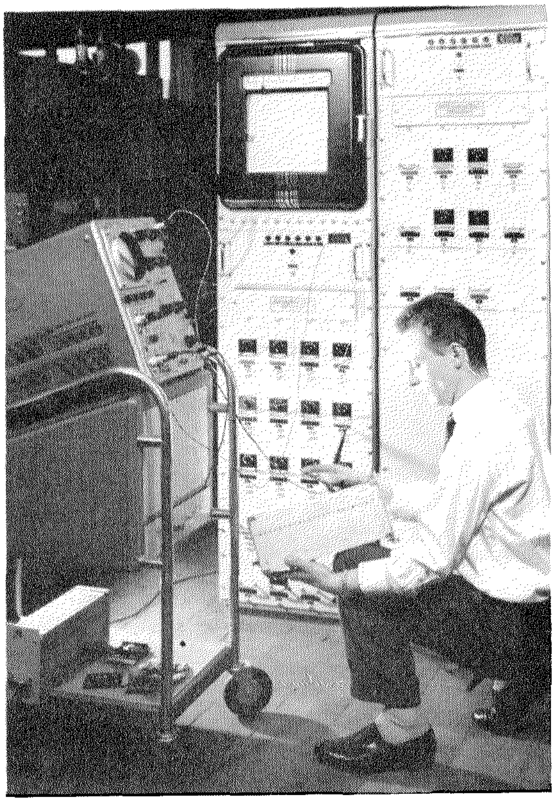


Figure 7—Transmitter and receiver installed at the Uccle observatory in Belgium.

### 6.2 Water System

In Holland, a water substation at Osdorp must keep the central station in North Amsterdam informed about two flow values derived from differential pressure-measuring systems with potentiometer indication, two pressures from manometers also having potentiometer outputs, input current to pump motors using transducers to convert alternating current into proportional direct currents between 0 and 5 milliamperes, and two integrated flows recorded on counters. An electromechanical transducer converts the count values to proportional pulse signals that are stored in bistable counting circuits for transmission.

Transmission is over a pair of telephone wires in a 120-hertz band at 2700 hertz. The transmission speed is 50 bauds. Remote control is obtained over the same channel but at a lower frequency.

At the receiver, the information is displayed on recorders for the flows and pressures, on milliammeters for the electric currents, and on electromechanical counters for the integrated flows.

**J. C. Lhoest** was born in 1931 in Liège, Belgium. He studied electrical engineering at the University of Liège and specialized in electronics at the Institut Montefiore of Liège from which he received the degree of civil engineer in 1956.

In 1958, he joined the Bell Telephone Manufacturing Company in Antwerp. Mr. Lhoest is now in charge of the development of electronic telemetering systems.

**B. Mattlet** was born in 1930 in Kitega, Ruanda. He studied electrical engineering at the University of Louvain, Belgium, from which he received the degree of civil engineer in 1954. He did postgraduate work on pulse techniques at the Massachusetts Institute of Technology in the United States of America until 1955. Two years later, he received a degree in administrative sciences from the University of Louvain.

After working at Electrorail in Brussels, Mr. Mattlet joined Bell Telephone Manufacturing Company in Antwerp in 1958, and is now head of the instrumentation and control laboratory.

# Transradar—A Narrow-Band Radar Relay Equipment Using Storage Capacitors

**K. JEKELIUS**

Standard Elektrik Lorenz AG; Stuttgart, Germany

## 1. Introduction

The effective coverage area of airspace surveillance radar equipment is limited, especially by shadowing effects such as from the horizon, mountains, and clouds. With increasing aircraft speed and traffic density, the time a target appears on a single radar screen is no longer sufficient for evaluation. Effective airspace surveillance requires the evaluation of indications obtained from a number of distant radar stations used in a meshed radar network. The information obtained from each radar station must be relayed to neighboring radar stations, evaluation centers, or other interested agencies.

Depending on the distances involved and the operational situation, one of the methods outlined in Figure 1 may be employed for transmission of the radar information.

Direct connection by means of several parallel cables is standard practice for short distances up to 5 kilometers (3.2 miles).

For longer distances up to 50 kilometers (32 miles), the signals must be suitably mixed to permit their transmission without complications. Such broad-band transmission, shown in *A* of Figure 1, is simple, but costly and inflexible [1, 2].

For distances greater than 50 kilometers (32 miles), narrow-band transmission as shown in *B* of Figure 1 is substantially superior.

The technique of automatic data transmission shown in *C* of Figure 1 requires much larger

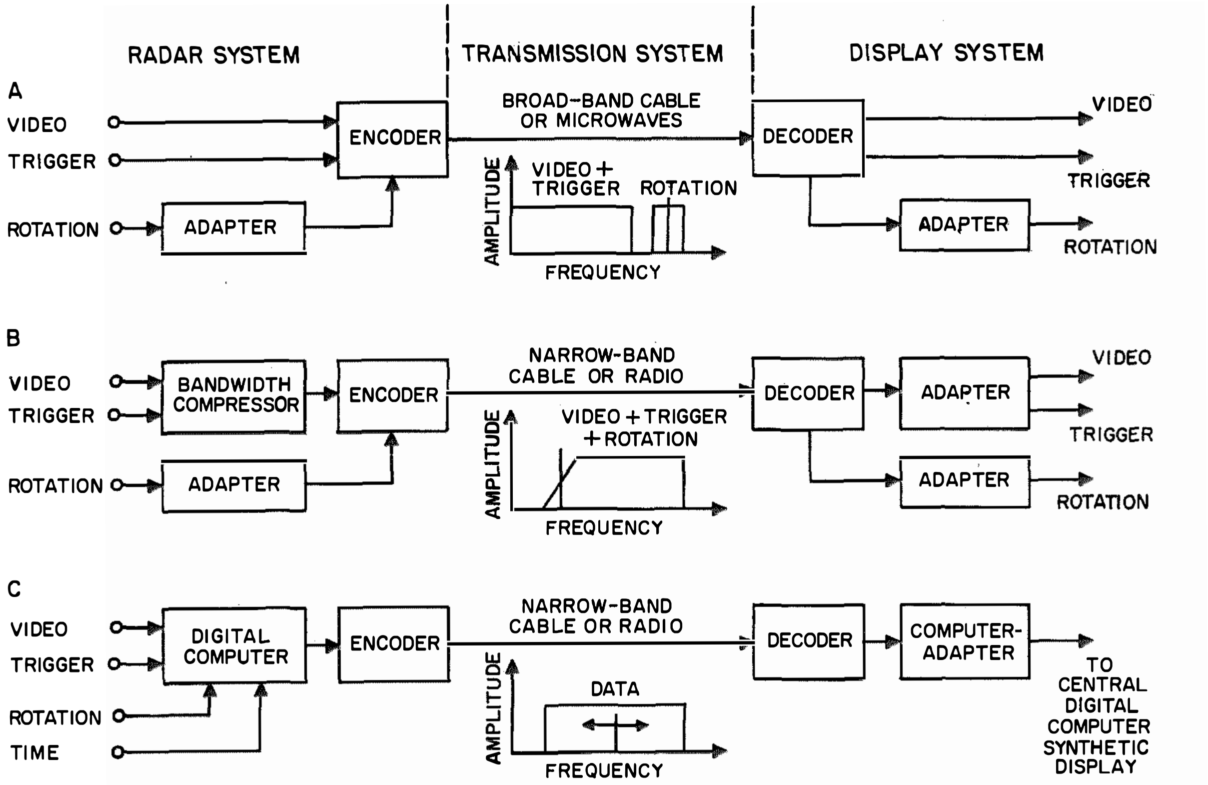


Figure 1—Relaying of radar indications over transmission systems of differing bandwidths. *A* = broad-band. *B* = narrow-band using bandwidth compression. *C* = narrow-band using digital computers.

equipment investments and is still experimental [3, 4]. Thus far, only semiautomatic data systems are being used in practical operations. These systems do not use the principal advantage of meshed radar networks, which is the correlation of overlapping information *before* detection. However, this lack of correlation is avoided by the narrow-band technique discussed below.

One of the decisive requirements for a practical meshed radar network is that the radar information must be compatible with the available transmission channels. These are:

**(A)** Telephone channels (bandwidth 3 kilohertz).

**(B)** Broadcast channels (bandwidth 8 to 15 kilohertz) or with similar data lines.

**(C)** Higher-order channels in carrier telephone systems (bandwidth 48 to 240 kilohertz).

**(D)** Television channels (5000 kilohertz).

An increasing volume of information can be transmitted with increasing channel width, but the transmission network becomes more inflexible, costly, sensitive to noise, and susceptible to failures. Satisfactory operation of a meshed radar surveillance system therefore depends primarily on the conversion of radar video, trigger, and rotation signals into a narrow-band signal. Since the signal-to-noise ratio is the cardinal criterion of radar technology, it must not be substantially deteriorated by bandwidth compression. Suitable means have to be used to ensure that a minimum of the reflected target energy is lost [5–9].

A bandwidth-compression method meeting these requirements is possible because of the redundancies contained in the radar signal [10]. For ground surveillance radar, to which this paper is limited, the redundancies consist of the following:

**(A)** Multiple scanning during one revolution of the antenna, with each target appearing in several (for example $n = 10$) successive pulse periods.

**(B)** Safety factor against overrange (for example $K = 2$), which is the ratio between the unambiguous and nominal maximum ranges, considering variations in target areas.

**(C)** Dead time when the operative range is shorter than the maximum range.

**(D)** For remote operation, possible reconsideration of the compromises among pulse length, relative resolution, and evaluation accuracy. For example, transmission accuracy can be adapted to possible location errors, depending on application and evaluating method, or adjusted to equal resolution in radial and tangential directions at half the maximum range [11].

**(E)** Low target density. Usually about 10 percent of the roughly 200 000 picture elements on the radar screen are activated at once. Of those activated, only a small number are associated with true flying objects while the rest are caused by ground clutter, clouds, and noise [4].

Elimination of redundancies *A* through *D* permits a reduction of the transmission bandwidth in most cases by 2 to 3 orders of magnitude, leading to the transmission of a corresponding time-extended radar signal. The equipments required for this bandwidth compression are relatively simple, inexpensive, and rugged, and their development is sufficiently advanced to use them for practical operation [12–15].

Elimination of redundancy *E* is more difficult. It normally requires extensive use of electronic computers [3, 4]. The transmitted signal containing the target coordinates is then a pulse train approximately 30 bits long per target. Substantially higher equipment investments are needed for such a digital data system, and stringent requirements would have to be imposed on the characteristics of the radar signal and the suppression of noise in it, since only a very-limited number of targets can be transmitted over narrow-band channels (50 targets per second over a telephone line). Such a definitive limitation, however, is highly questionable, since radar surveillance attains its greatest importance during periods of heavy disturbances produced by bad weather and jamming. In practice it is advisable to introduce a margin of 1 to 2 orders of magnitude above the theoretical limitation. This results in approximately the same bandwidth for digital transmission as for narrow-band picture transmission.

Simple picture transmission with bandwidth compression, therefore, can be used both as a first stage in establishing an integrated radar system and later for interference-insensitive parallel operation. This type of transmission is well suited to digital data systems for monitoring, emergencies, and as a service channel. It will also be able to hold its place in all future integrated surveillance systems [16, 17].

## 2. Principle of Bandwidth Compression

Bandwidth compression requires the provision of a storage system between the output of the radar receiver and the input of the transmission path to the slave display unit. This storage unit can operate on the cathode-ray-tube picture produced at the radar or on the video output signal from the radar.

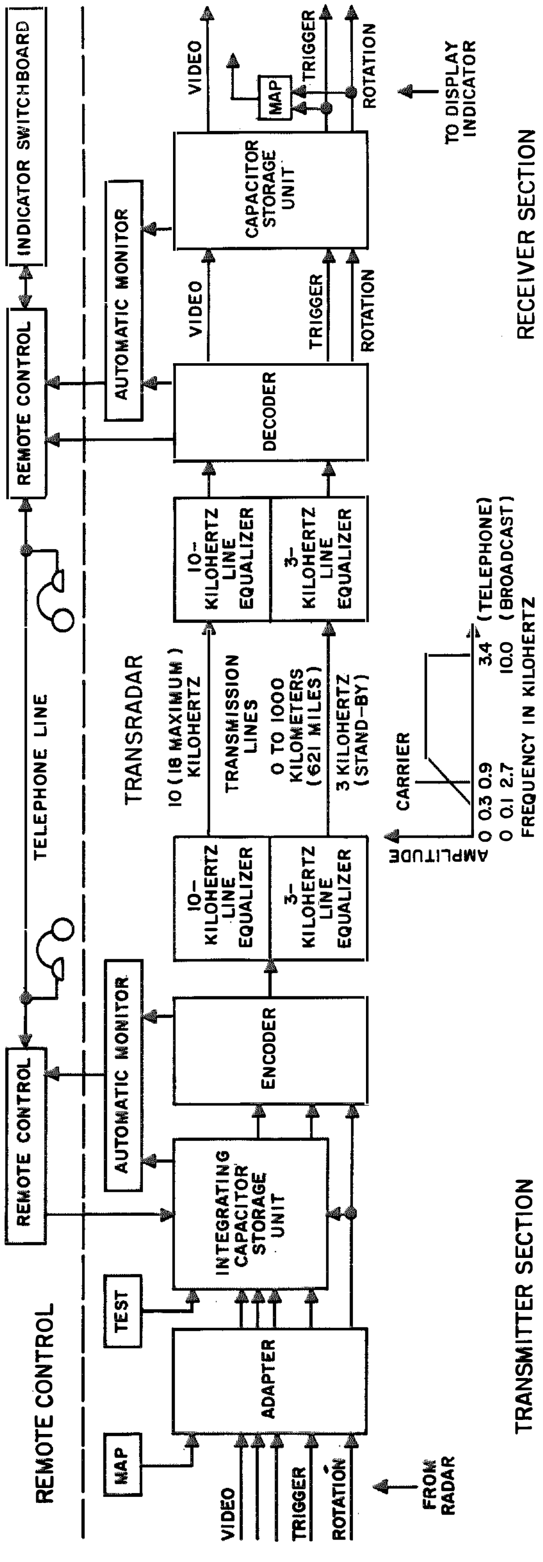
A picture storage unit has the duty of storing the complete radar signal received during one or several antenna revolutions. It therefore allows free choice of the scanning mode, which may be a conventional television scan [14, 15, 18], and under certain circumstances permits some use of the low target density for further bandwidth compression and track formation. Currently, the only available storage device for this purpose is the signal storage tube.

A video-signal storage unit, on the other hand, stores the information for only a small sector of antenna rotation. After integration and evaluation, the information is erased and replaced by that of the following sector [11–13, 17]. For airspace surveillance where the targets move rapidly, practically a full pattern of information must be transmitted during each antenna revolution. For this purpose, a rugged video-signal storage device for bandwidth compression at the origin of the transmission line, combined with a simple picture storage unit such as a storage tube or scan converter in the slave display unit for track formation, was found to be a most-suitable solution operationally and economically.

A thorough investigation of the principles that can be used for the realization of a video-signal storage device led to the choice of a bank of discrete capacitors, their quantity (250, 500, or 1000) depending on the required range resolution. This system uses pulse-operated gate circuits instead of tolerance-sensitive scanning by electron beams, and direct-current connections instead of coupling by secondary emission as is conventional for storage tubes.

The functions of this integrating storage device are thus based partly on analog and partly on digital techniques. Analog processing of the video signal achieves economy, excellent linearity of the integration characteristic, and lenient timing requirements. Performing all scanning operations with digital circuits ensures maximum reliability, high accuracy, and easy maintenance. Scanning can be started and stopped as required, using radar systems with nonconstant pulse frequency, and the scanning speed may be changed instantaneously with change of range or transmission bandwidth.

The behavior of a building block in a radar system with respect to natural and man-made interference is important from the viewpoint of information theory. The best arrangement for discerning a periodic pulse sequence in statistically distributed noise is a filter whose response-frequency curve provides periodically alternating pass and stop bands distributed so that all harmonics of the pulse frequency are within the narrow pass bands while the intervening noise power is suppressed by the stop bands. However, since the useful signal in a radar system is not absolutely periodic but



has a finite number of pulses per target, the pulse harmonics form sidebands that require wider pass bands. The practical realization of such periodic or cam filters presents certain difficulties as in obtaining very-steep filter characteristics and in the alignment procedure.

Much easier, however, is the realization of the equivalent arrangement in the time domain, this filter being a storage integrator or combining filter [5–9]. The selectivity then stems from linear addition of the coherent signals and increases with the number of identical summed pulse periods. Here also, the interference suppression is limited by the finite number of pulses per target [8, 9].

In this respect, the capacitor store shows its superiority. First, the integration characteristic of the storage elements is very linear. Second, digital switching by the pulse distributor readily accommodates aperiodic radar signals such as are encountered in phase modulation of the transmitted pulses (jitter) without loss of accuracy. To accomplish this, the digital pulse distributor is designed to stop when arriving at the last storage capacitor and to remain in start position on the first storage capacitor until the next triggering pulse arrives, thus providing dead time. The digital feature of this start-stop operation requires practically no transient time. The accuracy of distribution, and therefore both the effectiveness of interference suppression and the accuracy of distance measurement in the narrow-band signal, is independent of the duration of the variable dead time.

The writing and scanning circuits of the capacitor storage unit use only transistors as active elements. It is unsurpassed for ruggedness and reliability. The principle of operation practically excludes distortions in geometry as

← Figure 2—Diagram of transradar system, with remote control, inserted between radar and display systems.

well as voltage and temperature influences. Adjustment to the operating data of individual radar sets is possible without using special measuring instruments.

## 3. Transradar Design

A bandwidth compression system has been developed for long-range narrow-band transmission of radar signals and operates on the principles outlined above. The equipment requires the same connections to the radar system as a normal display unit, and delivers the transmitted signals in the form required by the respective remote indicators. Thus it may be considered as an extension of the connecting cable between the radar receiver and the indicators.

Figure 2 is a block diagram of the equipment, with a remote-control facility, incorporated in a radar system.

The signals arriving from the radar equipment (video, trigger, rotation, and sometimes map video) are passed through adapter networks where they are matched to the operating requirements of the bandwidth-compression device. Free selection is possible of either normal, moving-target indication, or logarithmic receiver video signals, and another smaller-amplitude signal produced in synchronism (map; identification, friend or foe; et cetera) may then be superposed on the selected video signals. The rotation signals are converted into angular pulses, used in the bandwidth-compression equipment for controlling the narrow integration sector, and added within the encoder to the narrow-band transradar video signal to be transmitted. A number of different adapter networks are available to permit connection of the equipment to different types of radar sets.

Bandwidth compression is performed by the integrating capacitor storage unit, whose functions are illustrated in Figure 3. This video-signal storage device comprises a bank of 500 capacitors with electronic write and read

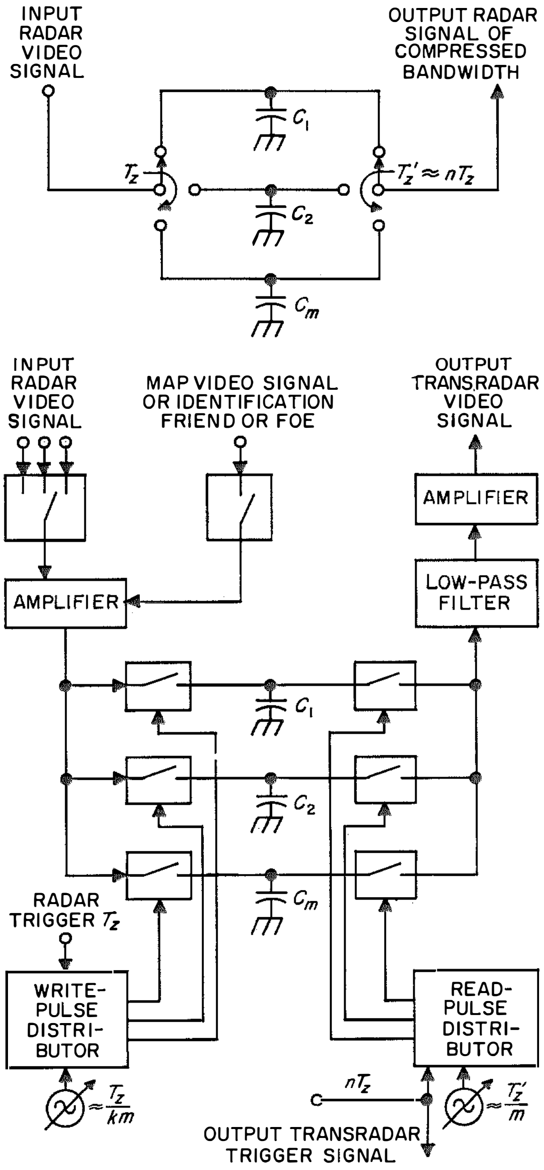


Figure 3—Diagram of the integrating capacitor storage unit. At top is the functional arrangement for storing the radar signals sequentially in capacitors, which discharge at a much slower scanning rate and thereby compress the frequency bandwidth. Below is the same functional arrangement using pulse-operated electronic gates to switch the storage capacitors. $m$ = number of range elements. $n$ = number of hits per target. $k$ = unambiguous range divided by range in use.

gates, one capacitor being provided for each distance element ($C_1$ through $C_m$, $m = 500$).

A writing timer synchronized by the radar trigger distributes the video signals of consecutive scan periods $n$ to the $m$ storage capacitors arranged as a radial line. The capacitor line is traversed once during each pulse period. The writing time may be selected as required for range resolution in the storage unit.

The read-timing depends on the integrated angular sector derived from the radar antenna pattern plus the bandwidth of the transmission path (range resolution determined by transient time of the transmission line). The charges stored and integrated in the capacitors are removed via read gates and delivered to the transmission device via a low-pass filter.

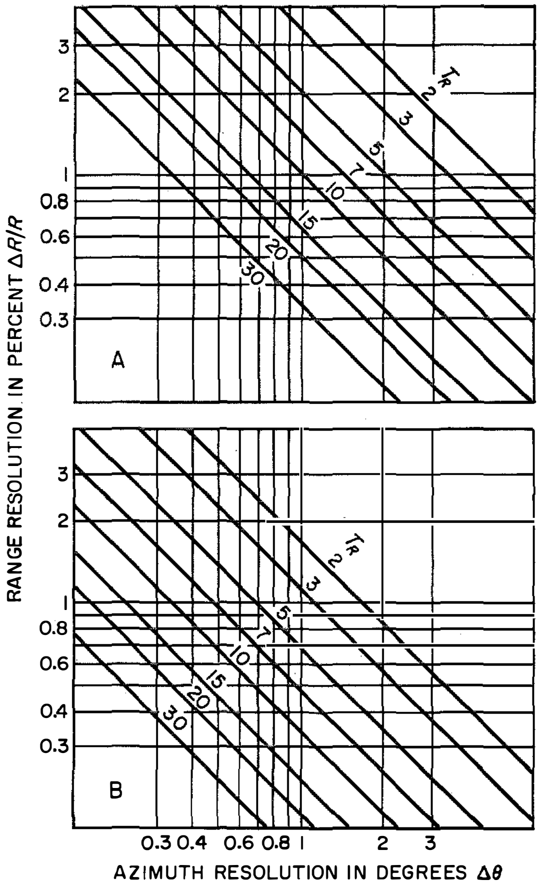


Figure 4—Resolution obtainable over 3-kilohertz telephone line *A* and 10-kilohertz line *B* for indicated values of antenna-rotation time in seconds.

The narrow-band transradar video signal, the read trigger, and the rotation pulses are first added to form a composite signal at the input of the carrier terminal equipment. This signal is then translated to the transmission frequency range by vestigial-sideband signal techniques, using double conversion and a Nyquist filter. The relative width of the vestigial sideband corresponds approximately to well-established television criteria, while the modulation depth can be held substantially lower since transradar signals are relatively insensitive to transmission interference. The carrier position at the lower end of the pass band permits both good and bad lines to be used, which is an important advantage in meshed networks made up of different line sections. The time characteristics are similar to that of a television signal with positive modulation, having large carrier amplitude for bright picture points and blanking of the carrier during synchronizing pulses. The system can work with two parallel carrier terminals to allow normal operation over a broadcast line and emergency operation over a telephone line.

The equalizers for the frequency delay characteristic are accommodated in the receiving terminal. The amplitude characteristic normally does not require equalization. The decoder is also located at the receiving terminal with the remote display unit. Here the composite signal is reconverted into video-, trigger-, and rotation-output signals in three separate channels for display.

No abnormal requirements are imposed on the transmission lines. However, the system design makes them the controlling factors for the quality of the transmitted picture.

The resolution attainable in practical operation over the most-commonly used transmission lines is shown in the diagrams of Figure 4, using the actual signal carrier frequencies and average equalization. The integrated angular sector, or scanning frequency, whose magnitude influences the angular resolution, is determined from the angular pulses derived from the antenna rotation signals. The attainable range resolution thereafter results mostly from the bandwidth (transient time) of the transmission line. By varying the scanning frequency, the relationship between angular and range resolution obtainable for a given transient time can be adjusted (see Figure 4).

The mean duration of one antenna revolution of practical radar sets is proportional to the square of maximum range. Therefore, the absolute resolution is almost independent of the type of radar set used when, as is normal in radar networks, the full range is transmitted. The absolute resolution primarily depends on the type of transmission line, while the transmission distance depends on the quality of the transmission line. Normal transmission networks may span distances of 1000 kilometers (621 miles) or more. Along the transmission line, the signal may branch off by means of standard terminal equipments to a number of remote evaluation centers.

Figure 5 shows the time diagrams for operation with various displays. Line 1 shows a radar video signal displayed on the original plan-position indicator in synchronism with the rotating directional antenna. Lines 2 and 3 show, respectively, the bandwidth-compressed video signal at the output of the integrating capacitor storage unit and the line signal. After demodulation at the receiving terminal, various forms of display are possible, depending on the type of indicator (lines 4 through 6 of Figure 5).

Video and trigger signals can be displayed directly if a special slow-scan indicator is available (line 4). As this is seldom the case, a capacitor storage device, similar to the one at the transmitting end but with inverse operating mode, is usually incorporated in the receiving terminal. The incoming video signal is written into the storage unit at slow speed and subsequently read out at high speed one or several times. This method permits the use of normal fast-scan indicators (line 5). Of special interest is the extremely long dead time that occurs when the storage is scanned only once (line 6). This dead time permits inserting additional information such as direction-finder bearings, marking symbols, or other radar signals, in indicators with interscan facilities.

Various connecting facilities are available to

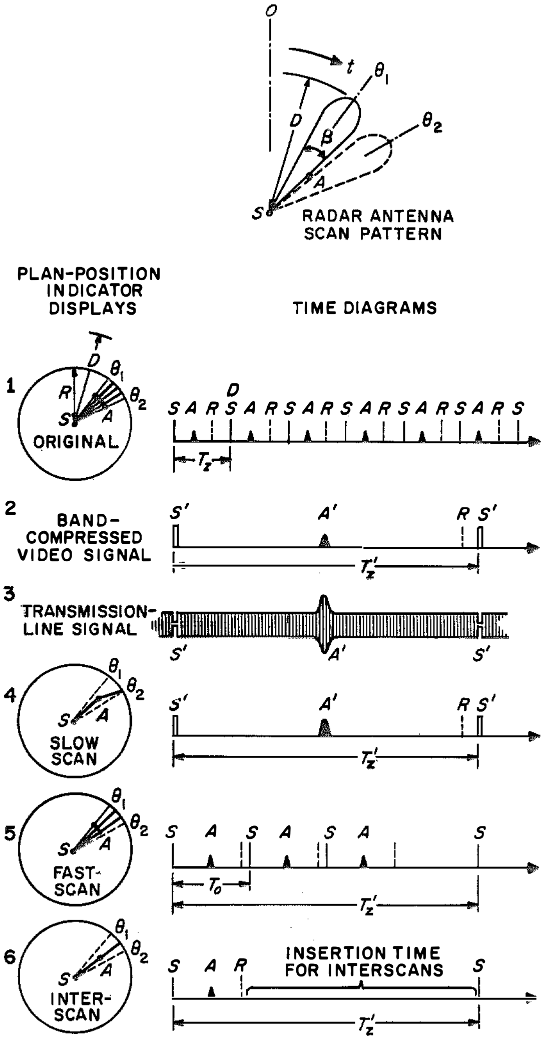


Figure 5—Plan-position indicator displays and corresponding time diagrams of signals. $A$ = target returns. $D$ = unambiguous range corresponding to pulse frequency. $R$ = range in use. $S$ = triggered pulse.

correlate the transmitted rotation signals to the deflection of the indicator systems. These facilities may be synchros with normal gear ratios or rotating shafts or they may be sine-cosine voltages varying with the azimuth. An automatic facility ensures fast reorientation to north after line interruptions or other disturbances of synchronous operation. The reacquired rotational information may also be used for synchronizing video mapping equipments or other evaluation aids.

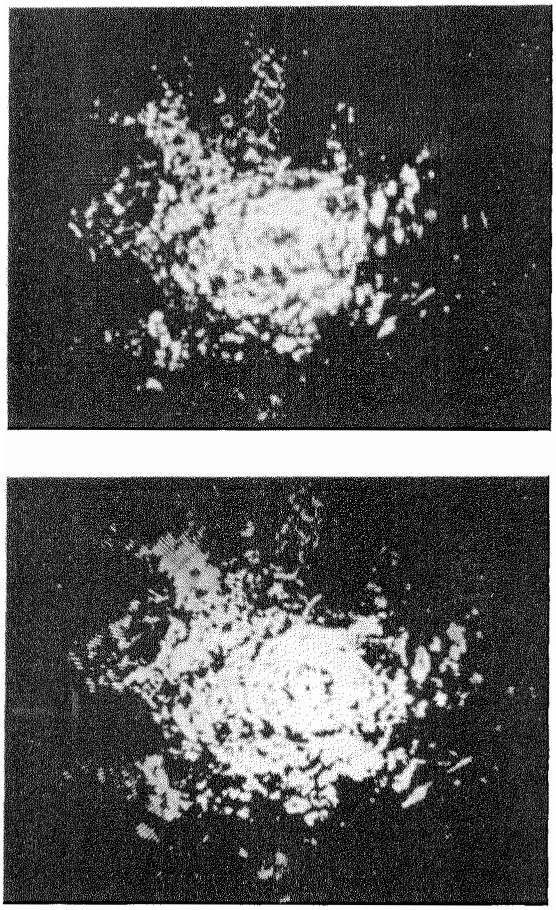


Figure 6—At top is an original radar picture, while below the same picture is shown after transmission over more than 600 kilometers (373 miles) of telephone lines. Each exposure was for exactly 1 antenna rotation. The pictures were taken on different days, which accounts for minor differences.

Figure 6 shows two photographs, one taken from an original indicator before bandwidth compression and the other from a slow-scan indicator after transmission over a normal telephone line of approximately 600 kilometers (372 miles). Both photographs were exposed for exactly one antenna revolution. Since the two pictures were taken on different days, because of the distance between the two locations, they do not correspond in all details. It is also very difficult to reproduce radar displays photographically.

Figure 7 shows the transradar transmitting and receiving equipments, which are very similar in design. They are subdivided into plug-in units, which simplify maintenance and rearrangement for special operational requirements. Transistors are the only active elements used. The equipment has been in operation for some time with very good evaluation results, and is now in production.

## 4. Applications

Narrow-band radar relay systems have the following important applications:

**(A)** Integrated radar network for airspace surveillance, permitting detection behind shadow areas, with great interscan time-sharing capacity and flexibility in emergencies.

**(B)** Relaying radar pictures from unattended or mobile gap-filled radar stations, with insertion into the long-range surveillance picture.

**(C)** Centralized air traffic control with common use of civil and military radar equipments (transmission before evaluation).

**(D)** Radar connection to positions not having their own radar sets.

**(E)** Target acquisition and coordination of short-range systems with the aid of the general display.

**(F)** Making radar surveillance immune to interference, through dense meshing of the radar sets working in space, time, and frequency diversity, by combining the output signals of several transradar systems in time multiplex.

**(G)** Combining raw radar information and synthetic information, since the extremely long dead time permits simultaneous display of both types of information on one display unit.

**(H)** Centralized weather service.

**( I )** Documentation of the air-traffic situation in a realistically reproduceable form, by using standard audio magnetic tape recording.

**( J )** Training, by using selected portions of magnetic tape records.

## 5. References

1. E. K. Peterson, H. R. Ulander, R. N. Hargis, and E. Hajic, "Microwave Remotes Aid Traffic Control," *Electronics,* volume 30, number 7, pages 144–147; July 1957.

2. J. W. Sutherland, "Microwave Links for Radar Networks," *British Communications & Electronics,* volume 5, number 9, pages 688–695; September 1958.

3. J. S. McPetrie, "Radio Aids for High-Speed Aircraft," *Engineer,* volume 204, number 5309, pages 597–598; 25 October 1957.

4. H. J. Zetzmann, "Über den Aufwand bei Radarschirmbild-Übertragungen," *Zeitschrift fuer Flugwissenschaften,* volume 6, number 9, pages 258–265; September 1958.

5. J. V. Harrington and T. F. Rogers, "Signal-to-Noise Improvement Through Integration in

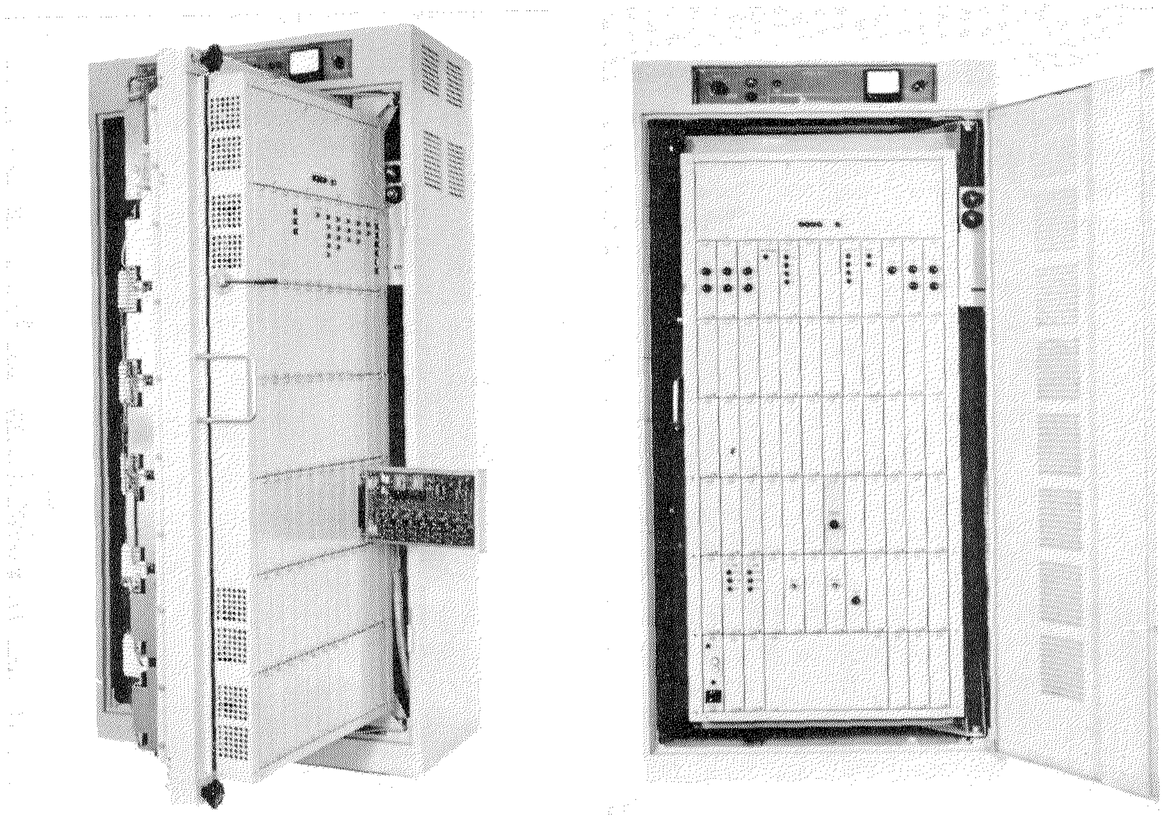


Figure 7—Transradar equipment. At left is the transmitter section for installation at the radar site. At right is the receiver section for installation at the remote indicator site. Dimensions of each unit are 1554 by 800 by 420 millimeters (61¼ by 31½ by 16½ inches).

a Storage Tube," *Proceedings of the IRE,* volume 38, number 10, pages 1197–1203; October 1950.

6. A. E. Bailey, "Integration in Pulse Radar Systems," published in "Communication Theory," Butterworth Scientific Publications, London; 1953: pages 216–230.

7. S. F. George and A. S. Zamanakos, "Comb Filters for Pulsed Radar Use," *Proceedings of the IRE,* volume 42, number 7, pages 1159–1165; July 1954.

8. J. Galejs, "Enhancement of Pulse Train Signals by Comb Filters," Institute of Radio Engineers, *Transactions on Information Theory,* volume IT-4, number 3, pages 114–125; September 1958.

9. G. Günther and G. Kraus, "Die Verbesserung des S/N-Verhältnisses durch Wiederholung von Signalen," *Frequenz,* number 10, pages 169–177; 1956.

10. A. V. Bedford, United States Patent 2 412 669; 19 June 1942.

11. H. Meinke, "Theorie und Praxis der Fernübertragung von Radarschirmbildern mit Frequenzkompression," *Bücherei der Funkortung,* volume 5, number 6, pages 11–26; 1955.

12. C. W. Doerr and J. L. McLucas, "A Narrow-Band Radar Relay System," Institute of Radio Engineers, *Convention Record,* part 8, pages 145–149; 1955.

13. F. Schröter, I. E. Otto, W. Schaffernicht, and K. Dinter, "Radarbildübertragung," *Telefunken-Zeitung,* volume 32, number 125, pages 152–182; 1959.

14. W. R. Schulz, "Teleran," *Funktechnik,* page 413; 1949.

15. H. W. Gates and A. G. Gatfield, "Scan Converter Aids Phone-Line Radar Relay," *Electronics,* volume 32, number 16, pages 48–51; 17 April 1959.

16. K. Jekelius, "Bandbreitenkompression von Radarsignalen mittels Speicherkondensatoren," *Bücherei der Funkortung,* volume 7, number 3c, pages 159–165; 1959.

17. K. Jekelius, "Entwicklung einer Anlage zur Bandbreitenkompression in Radarsignalen mittels Speicherkondensatoren und ihre Einsatzmöglichkeiten," *Nachrichten-Technische Zeitung,* volume 13, number 5, pages 225–233; 1960.

18. J. Fagot and R. Aste, "Transformateur d'image PPI-Television Type TI440," *Annalen de Radioelectronite,* volume 12, number 49, pages 183–189; July 1957.

**Kurt Jekelius** was born in Brasov, Rumania, on 28 April 1925. He studied telecommunications at the Technische Hochschule in Stuttgart, receiving the degree of Dipl.-Ing.

In September 1950, he joined Standard Elektrik Lorenz, where he was engaged in the development of the first television receivers, transmitters, and transmission lines. In 1953, he became head of the television laboratory, which included responsibility for the development of transradar equipment.

Since 1961, Mr. Jekelius has been in charge of technical projects in the sales field.

# Optimization of Network Configurations Through Queueing Theory

FRANK P. RANDAZZO

**ITT Kellogg, a Division of International Telephone and Telegraph Corporation; Chicago, Illinois**

## 1. Introduction

This paper describes a generalized queueing theory model that can be used to perform traffic studies for the optimization of network configurations with respect to time and equipment costs. The model considers parameters such as time of input arrivals, priorities, variable processing rates, and variable traffic densities, and may be used for both steady-state and dynamic conditions. The model demonstrates a way to reduce most multichannel problems to single-channel problems by analyzing a *type* of station rather than a *particular* station. That is, a multichannel processing station is considered as a single-channel processing station with a modified processing rate.

The queueing theory model is applicable only to networks having the property that from one point to another there is not more than one possible path. It follows that for a given configuration, the traffics corresponding to various junctions are known, whereas for a network not possessing the above property, most of those traffics would be additional unknowns.

A processing station is defined as a service unit that operates on inputs requiring service. It is considered in this paper as servicing only a single input at any specific point in time. If a processing station has parallel service units, it may be considered to be a single station with a processing rate equal to the sum of the individual processing rates of the parallel service units multiplied by an interaction constant. This constant corrects for the effect of interaction between the parallel service units. The interaction varies for each problem encountered and the constant, determined empirically, must be given as an input to the queueing model. Interaction may cause the processing rate of the station to be more or less than the sum of the individual processing rates. Where there is no interaction, the station processing rate is equal to the sum of the individual processing rates of the parallel service units

$$(\text{interaction constant} = 1).$$

The expected queue length at a specific processing station is the expected total number of inputs requiring service within a particular time increment and awaiting processing because of the unavailability of a station.

The queueing model assumes a Poisson input rate and exponential holding time. The holding-time distribution is defined as that distribution generated by determining the probabilities that service is continuing on the input being processed at discrete points in time. Accordingly, holding time is the amount of time that an input spends at the station.

The queueing theory model develops the following outputs for a processing station:

(**A**) A generalized set of curves showing traffic density versus queue length as a function of processing rate and priority.

(**B**) A generalized set of curves showing traffic density versus waiting time as a function of processing rate and priority.

The model is time-independent; therefore, in analyzing a particular processing station, variable time increments $\Delta t$ may be considered. From a generalized set of curves, it is possible to construct histograms for any particular station in a configuration. These histograms show queue length versus time of day and waiting time versus time of day, respectively. It is possible to determine the expected waiting time of an input if it arrives at a particular time of day and the expected number of inputs that must be processed before the input under consideration is processed. Both histograms are functions of priority.

An example is given of how this generalized set of curves may be used in determining the optimum configuration of a communication network, subject to equipment and cost constraints.

## 2. Queueing Theory Model

### 2.1 GENERAL

For the application of queueing theory, the characteristics that must be specified include:

(**A**) Input distribution—the frequency or time distribution of occurrence of inputs to each station in the system.

(**B**) Holding time distribution—the time or distribution of times required by the station to perform its functions.

(**C**) Queue discipline—a set of rules indicating what the system will do about an input if it arrives when the station is still serving a previous input; that is, whether a waiting line forms or the input is lost.

The model has been developed under the following assumptions (explained in Section 2.1.2):

(**A**) The input distribution to any type of station is Poisson.

(**B**) The holding or service time distribution is exponential.

### 2.1.1 *Notation*

The notation that will be used throughout this paper is:

$\lambda$ = total number of inputs introduced into the system during the time interval under consideration.

$P_i$ = percentage of the total inputs having the $i$th precedence introduced into the station during the time interval under consideration.

$\lambda_i = P_i\lambda$ = number of inputs having the $i$th precedence introduced into the station during the time interval under consideration.

$\lambda'_k = \sum_{i=1}^{k} \lambda_i$ = number of inputs having precedence greater than or equal to $k$ introduced during the time interval under consideration.

$N$ = total number of precedences.

$\mu$ = number of inputs that the station can process one at a time during the time interval under consideration.

$\rho'_k = \lambda'_k/\mu$ = traffic intensity or utilization factor for a service facility.

$Lq'_k$ = queue length composed of inputs introduced during the time interval under consideration and having precedence greater than or equal to $k$.

$Lq_k$ = queue length composed of inputs of the $k$th precedence introduced during the time interval considered.

$W_k$ = mean waiting time for an input having $k$th precedence introduced during the time interval under consideration.

It can be seen that if $\rho_k = (\lambda_k/\mu) > 1$, the queue must grow longer without limit. The probability of any specified queue length must then tend toward zero as time increases, as must the probability of any specified finite waiting time. However, it is clear that for finite intervals of time, $\rho$ may be greater than 1, although the queue length need not increase without limit over a longer period of time. Where $\rho_k < 1$, the condition is termed steady state. Where $\rho_k > 1$, it is termed dynamic or nonsteady state.

### 2.1.2 *Assumptions*

The model is developed using the following assumptions:

(**A**) Poisson input distribution.
(**B**) Exponential holding-time distribution.
(**C**) Priority queue discipline.

(**A**) According to Saaty [1], a Poisson input exists when:

(1) The total number of arrivals during any given time interval is independent of the number of arrivals before the beginning of the interval.

(2) For any interval $(t, t + dt)$, the probability that exactly one arrival will occur is $\lambda dt + f(dt^2)$, where $\lambda$ is a constant. The probability that more than one arrival will occur is of the order of magnitude of $dt^2$ and may be neglected. The input rate $\lambda$ may, of course, vary for $\Delta t$ time increments chosen at different times.

**(B)** Similarly, an exponential holding-time distribution exists when the following condition holds:

If a station is occupied at time $t$, the probability that it will become free during the following time increment is $\mu dt$, where $\mu$ is a constant. It follows that the frequency function of the service times is $\mu e^{-\mu t}$ while the mean duration of service is $1/\mu$, since the expected value of $t$ is

$$E(t) = \mu \int_o^\infty te^{-\mu t}\, dt = \frac{1}{\mu}. \qquad (1)$$

**(C)** Inputs of different types enter a station with assigned priorities for service. Whenever the station is free to accept an input, it selects the one of highest priority within each precedence class on a first-come first-served basis. However, if an input of higher priority enters the system while one of lower priority is in service, the service is not pre-empted; that is, the item being processed is not sent back to the waiting line. The model requires that provisions be made for storing queued items on a higher-to-lower-priority basis. It is understood that items of highest priority are also queued as required. Therefore, the expected length of the line is

$$Lq = \sum_{i=1}^{N} \lambda_i W_i. \qquad (2)$$

### 2.2 QUEUE LENGTHS

#### 2.2.1 *Steady-State Condition* $\rho < 1$

It has been shown [2] that the expected number of units in the queue is given by

$$Lq = \frac{\rho}{1-\rho} - \rho = \frac{\rho^2}{1-\rho}. \qquad (3)$$

This presupposes that there are times when the channel is not occupied because of the nature of the Poisson distribution.

#### 2.2.2 *Nonsteady-State Condition* $\rho > 1$

Clarke [3, 4] details a proof for the nonsteady-state condition when he derives the following expression.

$$Lq = (\rho - 1)\mu\tau + \frac{1}{\rho^m(\rho - 1)} + m + 0(\tau) \quad (4)$$

where $m$ is the queue length at the beginning of the time interval under consideration, $\tau$ is the length of the time interval, and $0(\tau) \rightarrow 0$ as $\tau \rightarrow \infty$. Since $\tau$ is considered large when it is much larger than $1/\mu$ ($\tau$ being considered a unit of time length, $\tau = 1$) $0(\tau)$ may be considered 0.

Therefore

$$Lq = (\rho - 1)\mu + \frac{1}{\rho^m(\rho - 1)} + m. \quad (5)$$

However, since we are not considering the previous time interval in the generation of a model to derive a generalized set of curves, the leftover queue length can be set equal to 0, resulting in

$$Lq = (\rho - 1)\mu + \frac{1}{\rho - 1}. \qquad (6)$$

The above analysis assumes that the system has settled down. The time required for this is called the relaxation time. For the system under consideration (single-channel Poisson input and exponential holding-time distribution) this is approximated by

Relaxation time

$$= \frac{1}{(\mu^{1/2} - \lambda^{1/2})^2} = \frac{(\mu^{1/2} + \lambda^{1/2})^2}{(\mu - \lambda)^2}. \quad (7)$$

For $\rho = 1$

$$Lq = 2\left(\frac{\mu\tau}{\pi}\right)^{1/2} + 0(\tau) \qquad (8)$$

where $0(\tau)$ is a bounded function for which $0(1)$ can be considered 0.

Therefore, for $\rho = 1$

$$Lq = 2\left(\frac{\mu}{\pi}\right)^{1/2}. \qquad (9)$$

Considering priority, modify the preceding theory in the following manner.

Let $\lambda'_k = \sum_{i=1}^{k} \lambda_i$ = the number of units having priority greater than or equal to $k$ introduced during the time interval considered.

$\rho'_k = \dfrac{\lambda'_k}{\mu}$ = the traffic intensity or utilization factor.

$Lq'_k$ = the queue length composed of inputs, having priority greater than or equal to $k$, introduced during a specific time interval.

$Lq_k$ = the queue length composed of inputs of the $k$th priority introduced during the time interval considered.

### 2.2.3 *Summary*

Since queue length is independent of priority within any group of inputs to a processing station, we conclude

$$\text{For } \rho'_k < 1,\ Lq'_k = \frac{(\rho'_k)^2}{1-\rho'_k}, \qquad k = 1, 2, \cdots, N. \qquad (10)$$

$$\text{For } \rho'_k = 1,\ Lq'_k = 2\left(\frac{\mu}{\pi}\right)^{1/2}, \qquad k = 1, 2, \cdots, N. \qquad (11)$$

$$\text{For } \rho'_k > 1,\ Lq'_k = (\rho'_k - 1)\mu + \frac{1}{\rho'_k - 1}, \quad k = 1, 2, \cdots, N. \qquad (12)$$

Once $Lq'_k$ has been computed for all $k$ values, then $Lq_k$ may be simply calculated by

$$Lq_k = Lq'_k - Lq'_{k-1}. \qquad (13)$$

### 2.3 WAITING TIMES

Kesten and Runnenburg [5] show that for the steady-state condition $\rho_k < 1$, assuming an arbitrary holding time and Poisson input distribution, the expected waiting time for $k$th priority is given by

$$W_k = \frac{\sum_{i=1}^{N} \lambda_i \mu_i^{(2)}}{2\left(1 - \sum_{i=1}^{k-1} \lambda_i \mu_i^{(1)}\right)\left(1 - \sum_{i=1}^{k} \lambda_i \mu_i^{(1)}\right)}, \qquad k = 1, 2, \cdots, N \qquad (14)$$

where

$$\mu_i^{(p)} = \int t^p \, dF_i(t) \qquad (15)$$

and where $F_i(t)$ is the cumulative holding-time distribution for the $i$th priority units in a single-channel queueing operation.

In the saturation (nonsteady-state) condition $\sum_{i=1}^{N} \lambda_i \mu_i^{(1)} \geq 1$, we can find an integer $s$ $(0 \leq s \leq N)$ such that

$$\sum_{i=1}^{s} \lambda_i \mu_i^{(1)} < 1 \quad \text{and} \quad \sum_{i=1}^{s+1} \lambda_i \mu_i^{(1)} \geq 1. \qquad (16)$$

Testing whether precedence $k$ is less than $s$

$$W_k = \frac{\sum_{i=1}^{s} \lambda_i \mu_i^{(2)} + \left(\frac{\mu^{(2)}_{s+1}}{\mu^{(1)}_{s+1}}\right)\left(1 - \sum_{i=1}^{s} \lambda_i \mu_i^{(1)}\right)}{2\left(1 - \sum_{i=1}^{k-1} \lambda_i \mu_i^{(1)}\right)\left(1 - \sum_{i=1}^{k} \lambda_i \mu_i^{(1)}\right)}, \qquad k = 1, 2, \cdots, (s-1). \qquad (17)$$

For $k \geq s$, we modify (2) to derive the expected waiting time.

$$Lq = \sum_{i=1}^{N} \lambda_i W_i.$$

Therefore

$$Lq'_k = \sum_{i=1}^{k} \lambda_i W_i \qquad (18)$$

and

$$W_k = \frac{Lq'_k - \sum_{i=1}^{k-1} \lambda_i W_i}{\lambda_k}. \qquad (19)$$

An approximation that can be used instead of (19) is

$$W_k = \frac{(\rho'_k - 1) + W_{k-1}\left(1 + \frac{\lambda_k}{\lambda_{k-1}}\right)}{2} . \qquad (20)$$

2.3.1 *Summary—Waiting Times*

2.3.1.1 *Steady-State Condition*

Using a Poisson input distribution and exponential holding-time distribution, (14) becomes

$$W_k = \frac{\rho'_N}{\mu(1 - \rho'_{k-1})(1 - \rho'_k)} , \qquad \rho'_k < 1, \rho'_o = 0. \qquad (21)$$

2.3.1.2 *Nonsteady-State Condition*

Finding the first and second moment of the exponential holding-time distribution and substituting these expressions in (17) we obtain

$$W_k = \frac{1}{\mu(1 - \rho'_{k-1})(1 - \rho'_k)} , \qquad k < s, \rho'_o = 0. \qquad (22)$$

When $k \geq s$

$$W_k = \frac{Lq'_k - \sum_{i=1}^{k-1} \lambda_i W_i}{\lambda_k} \qquad (23)$$

or

$$W_k = \frac{(\rho'_k - 1) + W_{k-1}\left(1 + \frac{\lambda_k}{\lambda_{k-1}}\right)}{2} , \qquad \rho'_o = 0. \qquad (24)$$

## 3. Generalized Application Procedure

Figure 1 gives a generalized procedure to determine an optimum configuration for any given network.

(**A**) Through a traffic analysis, determine the set or sets of message priority percentages to be considered.

(**B**) Using the queueing model, develop a set of generalized curves for the set of percentages desired.

(**C**) Considering operation functions and other appropriate factors, determine the processing rates for each station as a function of the length of the chosen time intervals.

(**D**) Hypothesize a station configuration for the specific application.

(**E**) Through a traffic analysis, determine

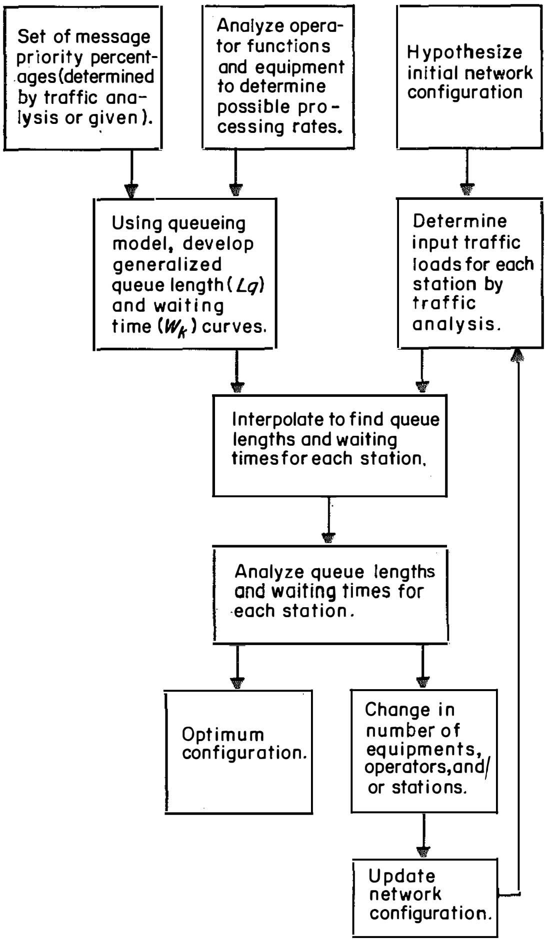


Figure 1—Procedure to determine optimum configuration within equipment and cost constraints.

representative message loads for each station in the proposed configuration as a function of time of day.

(**F**) Using the generalized set of curves, with the appropriate set of percentages for each station in the configuration, determine the expected queue lengths and waiting times.

(**G**) Analyze the expected queue lengths and waiting times for acceptability. If they are acceptable, determine whether the message loads are low using the present configuration. If so, modify the configuration to use less equipment; otherwise the chosen configuration is satisfactory. If the expected queue lengths and waiting times are not acceptable, modify the station configuration to use more equipment.

(**H**) If the station configuration is changed, return to (**D**) and proceed accordingly.

The described queueing-theory model may be used to test various network configurations in the iterative process of determining an optimum configuration, within time and equipment constraints, or it may be used as a model to determine expected queue lengths and expected waiting times of messages at any desired station(s).

Figures 2 and 3 are computer flow charts of the development of the generalized set of curves for queue lengths and waiting times.

## 4. Illustration of the Model

### 4.1 Application

The model was applied to the evaluation of a semiautomatic communication network designed for an Air Force base. Figure 4 shows a simplified configuration. The designed network consisted of a centrally located switching center at the base, several message centers, and the modified office typewriters (hereafter

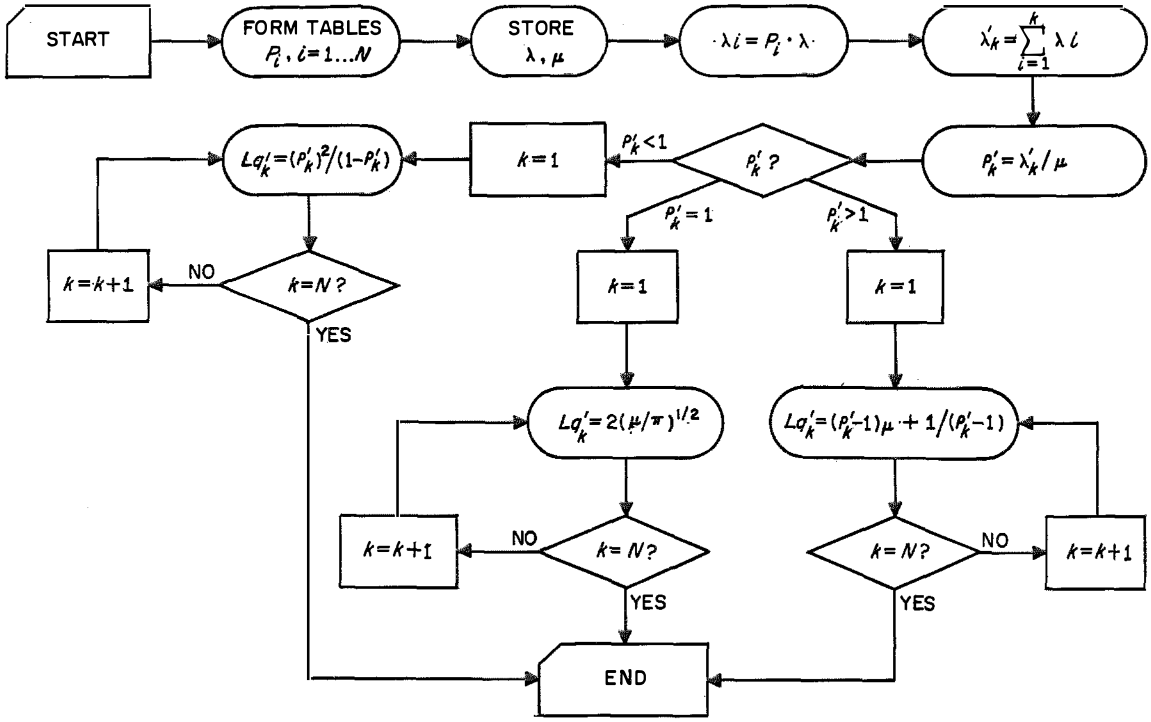


Figure 2—Program to develop generalized set of curves for queue lengths.

called typewriters) on which outgoing messages are prepared for automatic transmission to using offices.

All messages enter the base network at the switching center where they are automatically stored on tape. The information on the tape is then electrically transmitted to the proper message center. From here messages are hand-carried to the final destination. They are transported either by scheduled courier or are immediately hand-carried, depending on priority, location of the message center, and time of day.

Outgoing messages are originated at an office where, after proper coordination and authorization, each message is hand-carried to the typewriter servicing that office. This produces a paper-tape record that is to be transmitted and a hard copy of the message. The typewriter may be located in the office where the message originates, at the message center, or at some other location. The message, in paper-tape form, is then hand-carried to the message center associated with the originating office, where an operator electrically transmits the taped information to the switching center. The message is then automatically transmitted from the switching center to outside destinations over a communication network.

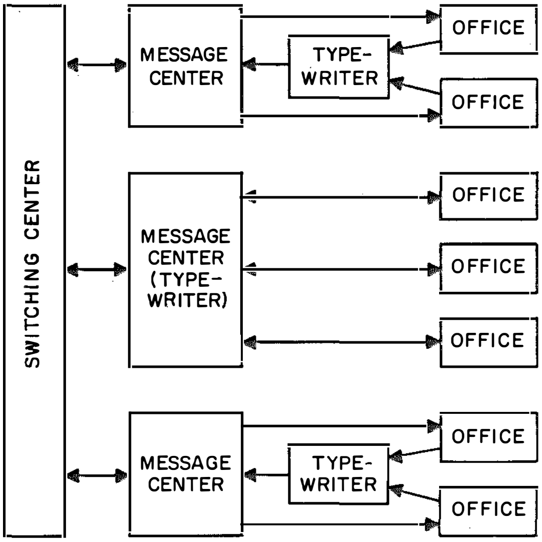


Figure 4—Simplified communication network.

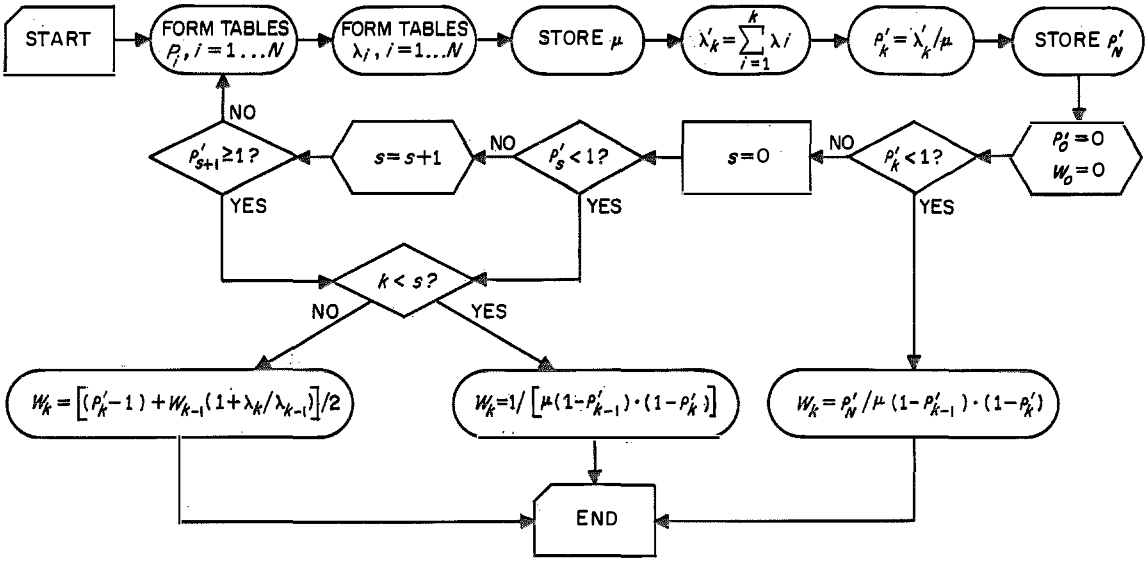


Figure 3—Program to develop generalized set of curves for waiting times.

### 4.2 Numerical Example

Studies of typing-speed requirements and other operator functions determined the processing rate for each typewriter to be 12 messages per hour. Where there are two typewriters, the processing rate is 24 messages per hour; that is, the interaction constant is equal to 1. At the message center, the processing rate was 30 incoming and 20 outgoing messages per hour. (The discrepancy in these processing rates is caused by different operator tasks in processing incoming and outgoing messages.) The switching center may process 80 incoming and 80 outgoing messages per hour.

Messages had one of four precedences: deferred, routine, priority, and operational immediate (in order of increasing precedence). A traffic survey was conducted at the base and the mean distribution of message traffic was found to be 40 percent deferred, 32 percent routine, 25 percent priority and 3 percent operational immediate. These percentages (variables in the described model and not necessarily constant over time) were obtained by averaging the traffic over all stations at all times of day. A generalized set of curves was developed for the expected queue lengths and waiting times at each type of station in the configuration (see Figures 2 and 3) based on the established percentages. Note that an exact set of percentages may be used for each station in the configuration, if desired, by developing different sets of generalized curves and using those appropriate to the particular station. Development of these curves for a set of priority percentages requires little computation and they may be obtained with a very limited amount of computer time. Generation of the curves shown in Figures 5 through 12 and similar curves for message centers and switching center requires less than 1 minute of *IBM* 7090 time.

Figures 5 through 8 and 9 through 12 are the generalized sets of curves showing expected queue lengths and waiting times, respectively, at a typewriter station for the percentages considered.

To illustrate the model, let us determine the expected queue lengths and waiting times for two typewriters serving a message center.

Figure 5—Typewriter queue length for operational-immediate traffic ($P_1 = 3$ percent of total).

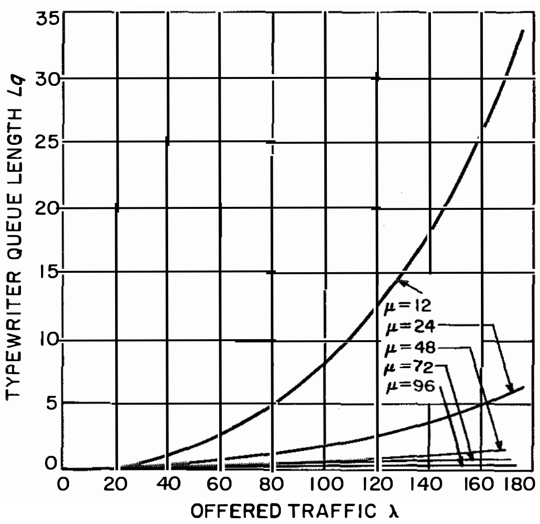


Figure 6—Typewriter queue length for priority traffic ($P_2 = 25$ percent of total).

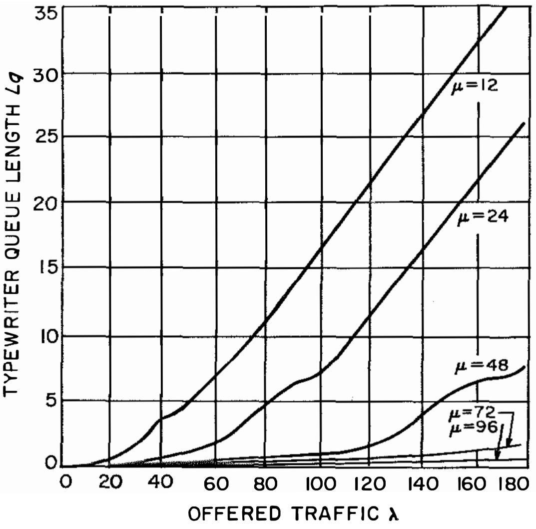


Table 1 lists their message loads as a function of time of day.

The processing rate used for a typewriter station is 12 messages per hour as determined by analysis of operation functions and other appropriate considerations. Using Figures 5 through 12, we interpolate to determine the readings in Tables 2 and 3.

The interpolation from Figures 5 through 12 is performed as follows:

TABLE 1

Message Loads for Typewriters 1 and 2

| Time | Typewriter 1 | Typewriter 2 |
|---|---|---|
| 0000–0100 | 0 | 0 |
| ⋮ | ⋮ | ⋮ |
| 0700–0800 | 0 | 0 |
| 0800–0900 | 4 | 1 |
| 0900–1000 | 4 | 1 |
| 1000–1100 | 8 | 2 |
| 1100–1200 | 8 | 3 |
| 1200–1300 | 2 | 1 |
| 1300–1400 | 15 | 10 |
| 1400–1500 | 19 | 7 |
| 1500–1600 | 11 | 1 |
| 1600–1700 | 7 | 1 |
| 1700–1800 | 0 | 0 |
| ⋮ | ⋮ | ⋮ |
| 2300–2400 | 0 | 0 |

$Lq_4$, for time 1400–1500 of Table 2, was determined by adding $Lq_4$ for time 1300–1400 to the message load input of time 1400–1500 and looking up this quantity in Figure 8. That is, finding $5 + 19 = 24$ along the horizontal axis of Figure 8 gives the queue length of 8 messages during this time interval. This queue length is added in turn to 11, the original message load during time 1500–1600, to determine the queue length of a fourth-precedence message during time 1500–1600, et cetera. Waiting times are determined the same way. For example, the mean waiting time for a third-precedence message during time 1500–1600 was found by adding the queue length for the third-precedence messages in time 1400–1500 to the original message load in time 1500–1600 and finding this quantity in Figure 11. That is, $4 + 11 = 15$ which from Figure 11 gives a mean waiting time for a third-precedence message of 0.46 hour.

If delays of 0.92 hour for a fourth-precedence message, 0.61 hour for a third-precedence message, 0.18 hour for a second-precedence message, and 0.09 hour for a first-precedence message are acceptable, then the typewriters are left unchanged; if not, then possibly some

Figure 7—Typewriter queue length for routine traffic ($P_3$ = 32 percent of total).

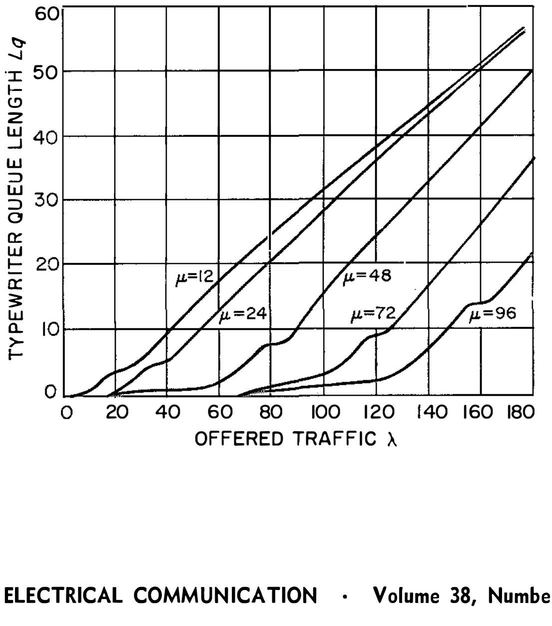
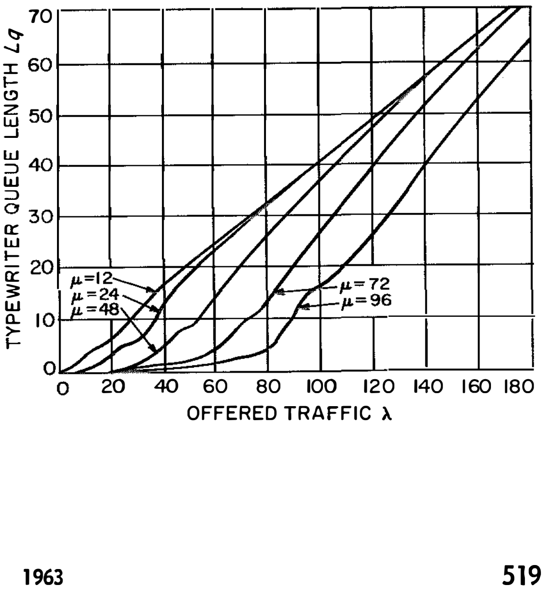


Figure 8—Typewriter queue length for deferred traffic ($P_4$ = 40 percent of total).



offices whose message traffic is included in the determination of the loads shown in Table 1 may be serviced by another typewriter.

This rearrangement would be made by detaching an office, on the basis of geographical location and its associated message load distribution, from typewriter 1 for service by another one associated with the same or with another message center. The message load distribution of the detached office is then subtracted from the load distribution of typewriter 1 in Table 1, taking into account the delivery time from the office to the typewriter. The message load distribution associated with

TABLE 2

EXPECTED QUEUE LENGTHS AND WAITING TIMES FOR TYPEWRITER 1

| Time | Message Load | $Lq_1$† | $Lq_2$ | $Lq_3$ | $Lq_4$ | $W_1$‡ | $W_2$ | $W_3$ | $W_4$ |
|---|---|---|---|---|---|---|---|---|---|
| 0000–0100 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ |
| 0700–0800 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| 0800–0900 | 4 | 0 | 0 | 0 | 0 | 0.02 | 0.03 | 0.14 | 0.16 |
| 0900–1000 | 4 | 0 | 0 | 0 | 0 | 0.02 | 0.03 | 0.14 | 0.16 |
| 1000–1100 | 8 | 0 | 0 | 0 | 2 | 0.05 | 0.07 | 0.25 | 0.28 |
| 1100–1200 | 8 | 0 | 0 | 0 | 3 | 0.05 | 0.07 | 0.25 | 0.38 |
| 1200–1300 | 2 | 0 | 0 | 0 | 1 | 0.01 | 0.01 | 0.08 | 0.18 |
| 1300–1400 | 15 | 0 | 0 | 2 | 5 | 0.09 | 0.13 | 0.46 | 0.72 |
| 1400–1500 | 19 | 0 | 1 | 4 | 8 | 0.09 | 0.18 | 0.61 | 0.92 |
| 1500–1600 | 11 | 0 | 0 | 2 | 6 | 0.07 | 0.12 | 0.46 | 0.78 |
| 1600–1700 | 7 | 0 | 0 | 1 | 4 | 0.04 | 0.05 | 0.30 | 0.58 |
| 1700–1800 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0.02 | 0.16 |
| 1800–1900 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ |
| 2300–2400 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |

† $Lq_i$ is defined as the expected queue length for a message of $i$th precedence entering the station during the specified time interval.

‡ $W_i$ is defined as the expected waiting time for a message of $i$th precedence entering the station during the specified time interval under consideration and is measured in hours.

Figure 9—Typewriter waiting time in hours for operational-immediate traffic ($P_1$ = 3 percent of total).

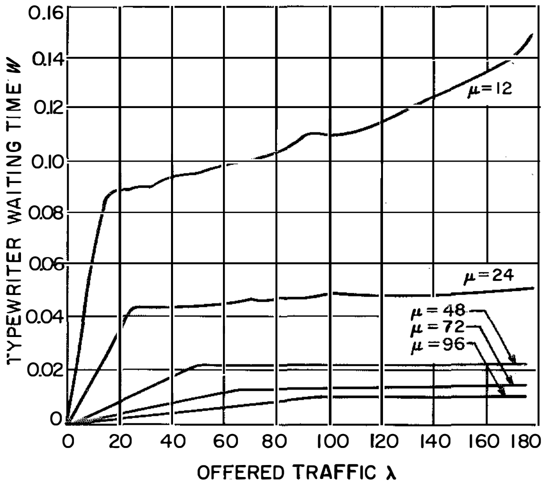
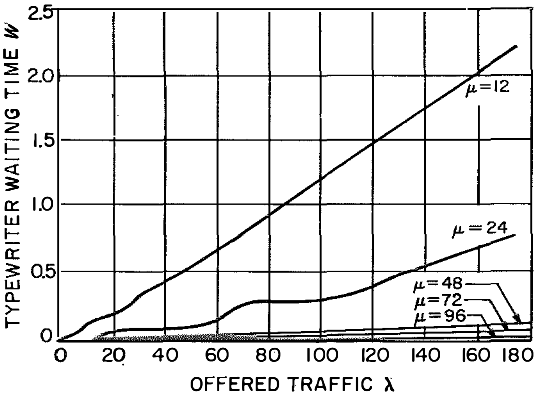


Figure 10—Typewriter waiting time in hours for priority traffic ($P_2$ = 25 percent of total).



the detached office is then added to the existing load distribution of the selected typewriter (again considering time displacement as shown in the following paragraph).

Assume that typewriters 1 and 2 are associated with the same message center and that a delay of more than 0.92 hour could be tolerated for a fourth-precedence message. Also assume that geographic considerations make it feasible to group the offices serviced by both typewriters and process the messages at typewriter 3. Assume further that typewriter 3 is to be located near the original ones to eliminate the requirement for adjusting the message traffic distributions. Therefore, the message load of typewriter 3 (refer to Table 4) for time 1300–1400, for example, is 15 + 10 = 25.

Since we are using a single typewriter, the processing rate is still considered to be 12 messages per hour. Taking care that a Poisson input distribution still holds, the expected queue lengths and waiting times for this

TABLE 3

Expected Queue Lengths and Waiting Times for Typewriter 2

| Time | Message Load | $Lq_1$ | $Lq_2$ | $Lq_3$ | $Lq_4$ | $W_1$ | $W_2$ | $W_3$ | $W_4$ |
|---|---|---|---|---|---|---|---|---|---|
| 0000–0100 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ |
| 0700–0800 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| 0800–0900 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0.02 | 0.06 |
| 0900–1000 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0.02 | 0.06 |
| 1000–1100 | 2 | 0 | 0 | 0 | 0 | 0.01 | 0.01 | 0.07 | 0.08 |
| 1100–1200 | 3 | 0 | 0 | 0 | 1 | 0.01 | 0.02 | 0.10 | 0.12 |
| 1200–1300 | 1 | 0 | 0 | 0 | 0 | 0 | 0 | 0.02 | 0.08 |
| 1300–1400 | 10 | 0 | 0 | 1 | 3 | 0.07 | 0.10 | 0.32 | 0.35 |
| 1400–1500 | 7 | 0 | 0 | 0 | 3 | 0.05 | 0.06 | 0.26 | 0.35 |
| 1500–1600 | 1 | 0 | 0 | 0 | 1 | 0 | 0 | 0.02 | 0.16 |
| 1600–1700 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ |
| 2300–2400 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |

Figure 11—Typewriter waiting time in hours for routine traffic ($P_3$ = 32 percent of total).

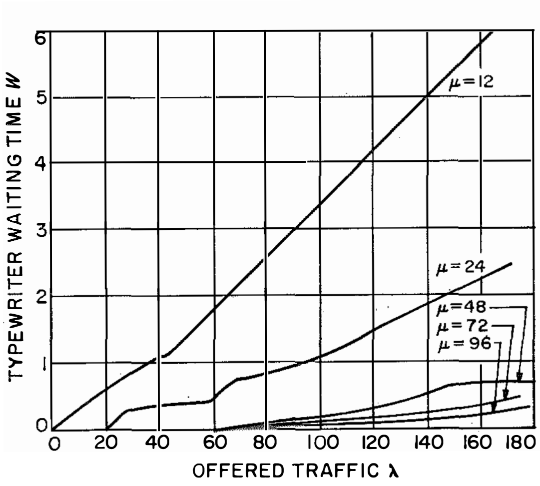
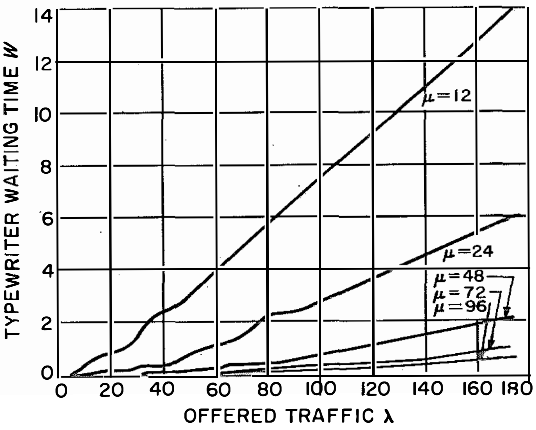


Figure 12—Typewriter waiting time in hours for deferred traffic ($P_4$ = 40 percent of total).



station are given in Table 5 (using Figures 5 through 12).

If the resulting mean waiting times of 2.00 hours for a fourth-precedence message, 0.85 hour for a third-precedence message, 0.26 hour for a second-precedence message, and 0.09 hour for a first-precedence message (during time 1400–1500) are acceptable, the configuration is left unchanged; if not, then the configuration should be modified to change the message load distribution of typewriter 3. This is done by having some of the offices, whose messages are processed by this typewriter, transfer their messages to another typewriter, if one exists, associated with the same message center or a typewriter associated with another message center.

TABLE 4
MESSAGE LOAD FOR TYPEWRITER 3

| Time | Typewriter 3 | = | Typewriter 1 | + | Typewriter 2 |
|---|---|---|---|---|---|
| 0000–0100 | 0 | = | 0 | + | 0 |
| ⋮ | ⋮ | | ⋮ | | ⋮ |
| 0700–0800 | 0 | = | 0 | + | 0 |
| 0800–0900 | 5 | = | 4 | + | 1 |
| 0900–1000 | 5 | = | 4 | + | 1 |
| 1000–1100 | 10 | = | 8 | + | 2 |
| 1100–1200 | 11 | = | 8 | + | 3 |
| 1200–1300 | 3 | = | 2 | + | 1 |
| 1300–1400 | 25 | = | 15 | + | 10 |
| 1400–1500 | 26 | = | 19 | + | 7 |
| 1500–1600 | 12 | = | 11 | + | 1 |
| 1600–1700 | 8 | = | 7 | + | 1 |
| 1700–1800 | 0 | = | 0 | + | 0 |
| ⋮ | ⋮ | | ⋮ | | ⋮ |
| 2300–2400 | 0 | = | 0 | + | 0 |

However, if we must choose between the queue lengths and waiting times shown in Table 5 and those shown in Tables 2 and 3, then we must weigh the value of a maximum mean waiting time of 2.00 hours and the cost of a single typewriter against a maximum mean waiting time of 0.92 hour and the cost of two typewriters.

## 5. Acknowledgements

The author acknowledges the many invaluable discussions held with members of the staff of the Advanced Research and Development Laboratory of ITT Kellogg, especially Dr. S. Veniar, Director of Advanced Projects Organization, and the constructive criticism of Dr. C. D. W. Thornton of International Telephone and Telegraph Corporation.

## 6. References

1. T. L. Saaty, "Résumé of Useful Formulas in Queueing Theory," *Operations Research*, volume 5, number 2, pages 161–200; 2 April 1957.

TABLE 5
EXPECTED QUEUE LENGTHS AND WAITING TIMES FOR TYPEWRITER 3

| Time | Message Load | $L_{q1}$ | $L_{q2}$ | $L_{q3}$ | $L_{q4}$ | $W_1$ | $W_2$ | $W_3$ | $W_4$ |
|---|---|---|---|---|---|---|---|---|---|
| 0000–0100 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ |
| 0700–0800 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| 0800–0900 | 5 | 0 | 0 | 0 | 1 | 0.03 | 0.04 | 0.17 | 0.20 |
| 0900–1000 | 5 | 0 | 0 | 0 | 1 | 0.03 | 0.04 | 0.17 | 0.22 |
| 1000–1100 | 10 | 0 | 0 | 1 | 4 | 0.07 | 0.10 | 0.30 | 0.37 |
| 1100–1200 | 11 | 0 | 0 | 1 | 4 | 0.07 | 0.12 | 0.37 | 0.68 |
| 1200–1300 | 3 | 0 | 0 | 0 | 1 | 0.02 | 0.03 | 0.14 | 0.16 |
| 1300–1400 | 25 | 0 | 1 | 4 | 9 | 0.09 | 0.24 | 0.72 | 1.04 |
| 1400–1500 | 26 | 0 | 1 | 5 | 14 | 0.09 | 0.26 | 0.85 | 2.00 |
| 1500–1600 | 12 | 0 | 0 | 3 | 9 | 0.07 | 0.14 | 0.52 | 1.04 |
| 1600–1700 | 8 | 0 | 0 | 1 | 5 | 0.05 | 0.08 | 0.36 | 0.75 |
| 1700–1800 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | 0.03 | 0.20 |
| 1800–1900 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ | ⋮ |
| 2300–2400 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 | 0 |

2. P. M. Morse, "Queues, Inventories, and Maintenance," John Wiley and Sons, New York; 1958.

3. A. B. Clarke, "The Time-Dependent Waiting Line Problem," University of Michigan Engineering Research Institute Report M720-1, R39; 1953.

4. A. B. Clarke, "A Waiting-Line Process of Markov Type," Annals of Mathematical Statistics, volume 27, number 2; 1956.

5. H. Kesten and J. T. Runnenburg, "Priority in Waiting-Line Problems," Math, Centrum; December 1956.

**Frank P. Randazzo** was born in Palermo, Sicily, on 25 November 1938. He received B.S. degrees in mathematics and in statistics from the University of Chicago in 1960.

In 1960, he entered the engineering department of ITT Kellogg Communications Systems and has recently joined its Advanced Projects Organization. He has been engaged in systems analysis and evaluation of command and control systems. He is presently working on communication requirements for earth-orbiting manned space stations.

Mr. Randazzo is a Member of the American Statistical Association and an Associate Member of the Operations Research Society of America.

# Traffic Simulation and Its Application in Telephony

G. DIETRICH
H. WAGNER
Standard Elektrik Lorenz AG; Stuttgart, Germany

## 1. Introduction

Theoretical computations of the traffic-handling capacity of telephone switching networks produce only approximate results, particularly for arrangements used in practical applications. The practical usefulness of the approximation equations must be verified by measurements performed with real or artificial telephone traffic.

Since measurements with real telephone traffic entail numerous difficulties, the idea of using so-called traffic machines [1–3] for the generation of artificial traffic was conceived several decades ago [4].

Lately, electronic computers frequently replace the traffic machine. Controlled by a program, the computer simulates the switching network to be tested as well as the artificial traffic flowing through it. This procedure does not require equipment investments, but merely writing a program. This program is known as a simulation program while the traffic trials are known as traffic tests or throwdowns.

Besides giving a general survey, this paper primarily describes the method of generating artificial telephone traffic and presents some examples. It also presents some known [5–7] and some new solutions to the problem based on accepted principles. Finally, some test results obtained with the aid of our *ER56* electronic computer are compared with theoretical results.

## 2. Simple Traffic Test

The principles underlying the generation of artificial telephone traffic, and consequently the investigation of switching networks, are explained with the aid of a simple example.

The number of calls rejected because no outlet is free is to be determined for a full-availability switching network with an unlimited number of inlets and three outlets (each inlet always having access to each free outlet), and with offered traffic $A = 2$ erlangs.

Although the beginnings and ends of the calls are random, probability theory establishes a certain order that permits mathematical definition of the telephone traffic. To simulate randomness, a random generator is required that will determine both the beginnings and ends of calls in the defined order.

For simple traffic tests, this can be achieved by repeatedly casting a six-sided die. We obtain a series of random numbers made up of 1, 2, 3, 4, 5, and 6, as for example 3 2 6 2 5 6 2 1 4 5 1 3 4 5 4 3 1 5 3 3 1 4 5 4 6 2 2 1 6 6. The random changing of the states of occupancy of the three outlets is now determined through the following use of these random numbers:

1, 2, and 3—releasing numbers—as soon as one of these numbers comes up, the corresponding outlet is released if occupied. If it is not occupied, the number has no effect.

4 and 5—call numbers—when one of these numbers appears, a call is originated and occupies a free outlet. If no outlet is free, a lost call is registered.

6—unused random number—this number has no function, and it causes no change in the state of occupancy.

This random distribution (refer to (6) with $R_0 = 2$) simulates offered traffic $A = 2$ erlangs. Figure 1 shows the changing states of occupancy caused by the above series of random numbers. The first few events are explained in Table 1. It is assumed that outlets

TABLE 1
CHANGING STATES OF OCCUPANCY

| Number Cast | Effect |
|---|---|
| 3 | Call on outlet 3 ends. |
| 2 | Random number without effect because outlet 2 is free. |
| 6 | Unused random number. |
| 2 | Random number without effect because outlet 2 is free. |
| 5 | Originated call finds outlet 2 accessible. |

1 and 3 are occupied at the beginning of the experiment.

Although the individual steps do not correspond to actual constant time intervals, it is evident from Figure 1 that the call duration varies considerably. One call may, for example, be longer than two other calls. In our experiment we offered $c_A = 10$ calls, and $c_v = 2$ of them, or 20 percent, were rejected.

To determine the loss

$$B = c_v/c_A \tag{1}$$

through such a traffic test with a sufficient degree of reliability, the test duration should be several thousand times longer than in this simple example to ensure a balance of statistical variations.

## 3. Generation of Random Numbers

A very-long series of random numbers is required to generate enough artificial telephone traffic for a meaningful result. This series, however, must be made up from many more random numbers than 1 to 6 because there are switching networks that have more than 100 outlets, and higher offered traffic $A$ that often have decimal positions. The numbers 1, 2, 3, . . . , do not suffice and must be supplemented by decimal numbers. The probability of any number in the series coming up next must be equal for all numbers and may not in any way depend on the preceding number (equal distribution and irregularity).

Some of the methods for generating a series of random numbers are particularly suitable for electronic computers. The mathematical instructions underlying these methods permit generating numbers $x_i$, starting with an arbitrary number $x_0$, according to the equation

$$x_i = f(x_{i-\nu}, \quad x_{i-\nu+1}, \cdots, x_{i-1})$$

or, for example, simply

$$x_i = g(x_{i-1}).$$

By omitting a certain number of decimal positions, the numbers $z_i$ are derived from $x_i$. They belong to the quantity of numbers $Z$ and form a series

$$Z_0, Z_1, Z_2, \cdots$$

If the mathematical equation for the generation of $x_i$ and the modification method to $z_i$ are suitably chosen, the series will have the desired properties although it is based on a formation law. Such a series of pseudo-random numbers has proved equal in many applications to a series of genuine random numbers. They have been used advantageously for traffic tests because they can easily and quickly be generated by electronic computers and are always reproduceable without having to be stored.

A description of the methods for generating pseudo-random numbers is beyond the scope of this paper. Some papers [8–10] containing detailed information are listed in Section 7, References.

## 4. Artificial Telephone Traffic

### 4.1 General

Switching networks in telephone systems are operated either on the busy-signal or the delay principle. Combining these two principles

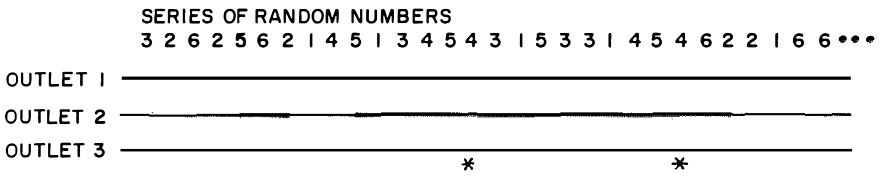


Figure 1—States of occupancy of 3 outlets.
* = lost calls.

is also possible and will be used whenever the number of waiting stores is insufficient or calls are terminated because of extended waiting times. In this paper, however, only the busy-signal principle is discussed, although simulation of a delay system or of a combination system is also possible.

Artificial telephone traffic must have the essential characteristics of real telephone traffic if it is to provide results reflecting practical operation. The tests simulate traffic that corresponds to the actual traffic during an arbitrarily determined number of main traffic hours (statistical equilibrium). This assumption underlies all conventional theoretical loss calculations.

When simulating systems that operate on the busy-signal principle, the duration of the different occupancy states of the switching network is insignificant. Merely the changes, that is, the probabilities of origination and termination of calls [7] are important. Actual timing is not simulated, not even in accelerated form.

The probability of a call originating is primarily dependent on the following:

(**A**) Number of sources or subscribers.

(**B**) Traffic offered and its distribution among the subscribers.

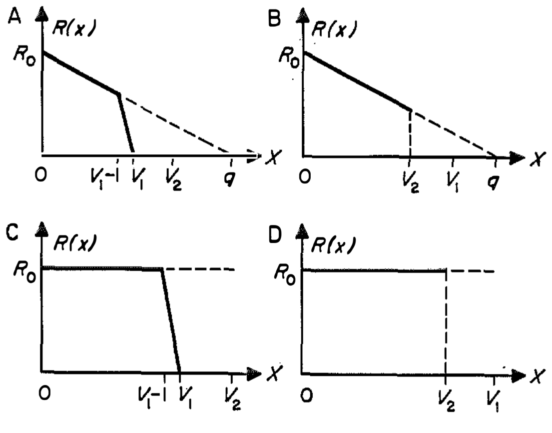


Figure 2—Possibilities of function $R(x)$.

(**C**) Type and capacity of the networks between subscribers and the switching network being investigated.

(**D**) Treatment of rejected calls, which either are to be registered as lost or are routed via another group of overflow trunks to their destinations.

Assuming an equal distribution of the offered traffic among all subscribers $q$, the probability of a call originating is directly dependent on the number of simultaneously busy subscribers $x$, since a busy subscriber definitely does not originate a call. It thus must be proportional to the function

$$R(x) = R_0\left(1 - \frac{x}{q}\right). \tag{2}$$

The factor $R_0$, which depends on the previously defined parameters, is explained as follows.

The number of sources $q$ and the maximum possible number of simultaneous calls $x$ influence the characteristics of the curve $R(x)$, which is defined for positive integral values of $x$ only. On the one hand, the number of sources $q$ may be finite or considered infinite; on the other hand, the variable $x$ may be limited either through the number of inlets of switching network $v_1$, or of outlets $v_2$, or of links $v_z$. These four possibilities are indicated in Figure 2, the discrete points being connected to form curves to simplify understanding.

The probability of a call terminating depends solely on the mean call duration.

The treatment of rejected calls (lost or overflow calls) is important only when considered with a finite number of sources. It then becomes significant whether a source is immediately released after an unsuccessful attempt to call or whether the call-originating probability is reduced for the intended duration of the call. In the following derivation of the equations, the mean duration $t_v$ of a rejected call is, for the time being, considered in a general way. It may later be selected as desired.

The equations required for the traffic tests are derived for condition $A$ in Figure 2. Equations for the other three conditions are computed from this one. Condition $A$ has little practical significance while the other conditions are of importance in practice. They are described in Sections 4.3 through 4.5.

### 4.2 DERIVATION OF CONDITION $A$

The probability of a call terminating within a sufficiently short time $\Delta t$ is

$$w_e = \Delta t / t_m \tag{3}$$

with $t_m$ being the mean call duration.

With $q$ as the equally loaded traffic sources, the probability of a call originating within $\Delta t$ is

$$w_a(x) = (q - x)\frac{\Delta t}{p_m} \tag{4}$$

if $x$ sources are busy. In this expression, $p_m$ denotes the mean time interval between successive calls (time of inactivity of sources).

As a source can only be either active or inactive, the following condition is met.

$$c_y t_m + c_v t_v + c_A p_m = q \tag{5}$$

where

$c_A$ = mean number of calls offered per busy hour.

$c_y$ = mean number of calls handled per busy hour.

$c_v$ = mean number of rejected calls per busy hour.

$t_v$ = mean duration of rejected calls.

$\approx 0$ if call is to be considered lost.

$= t_m$ if call finds an overflow path.

The equation for the mean time intervals between calls is obtained from (5).

$$p_m = \frac{1}{c_A}(q - c_y t_m - c_v t_v) \tag{6}$$

or, with the offered traffic $A = c_A t_m$ and the handling capacity $y = c_y t_m$, then

$$p_m = \frac{t_m}{A}(q - y - c_v t_v).$$

When examining condition $A$, it must be realized that the probability of a call originating is zero if all inlets $v_1$ of the switching network are busy. Assume the probability for this to be $w(v_1)$. The mean time interval between calls $p_m$ in (4) is the time during which free sources may be expected to originate calls. The mean time interval as expressed in (6) contains passive times for condition $A$ because an inactive source is not able to reach the switching network if all inlets $v_1$ are occupied. Only the mean active time interval $p_{m,a}$ however should be entered in (4).

We obtain

$$\begin{aligned} p_{m,a} &= p_m[1 - w(v_1)] \\ &= (q - y - c_v t_v)\frac{t_m}{A}[1 - w(v_1)] \end{aligned}$$

and therewith the probability of a call originating in condition $A$ is

$$w_a(x) = \frac{q - x}{q - y - c_v t_v} \cdot \frac{A}{1 - w(v_1)} \cdot \frac{\Delta t}{t_m}, \quad \text{for } x < v_1 \tag{7}$$

$$w_a(v_1) = 0.$$

In the simulation, the probabilities for the origination and termination of calls now have to comply with (3) and (7). For this purpose, a series of pseudo-random numbers is used that has the properties described in Section 3. The numbers are subdivided into releasing numbers $\mathcal{L}$, calling numbers $\mathcal{R}$, and not-used random numbers $\mathfrak{N}$, with the following agreement:

$$\mathcal{L} = [0, Z_1), \quad \text{with } 0 < Z_1$$

$$\mathcal{R} = [Z_1, Z_2), \quad \text{with } Z_1 < Z_2$$

$$\mathfrak{N} = [Z_2, N), \quad \text{with } Z_2 < N.$$

The mode of representation $[a, b)$ denotes that this quantity includes all numbers $z_i$ with $a \leq z_i < b$.

The limits $Z_1$ and $Z_2$ must be so chosen that the probability of a releasing or calling number appearing next in the series corresponds to (3) and (7). One of these two limits may be set arbitrarily.

To simplify writing the simulation program for the computer, it is advisable to set

$$Z_1 = v_2 \tag{8}$$

with $v_2$ denoting the number of switching network outlets. In this case, any individual outlet $j$ that is occupied is always released if a random number $z_i$, with

$$j \leq z_i < j + 1$$

appears in the series. If the numbering of the outlets starts with 0, for example, the random number $z_i$ will directly indicate by the number preceding its decimal point the outlet to be released. During the simulation process, the releasing number probability $w_1$ corresponds to the call-terminating probability $w_e$ and thus is proportional to 1.

$$w_1 \sim 1. \tag{9}$$

The probability $w_a(x)$ of a call originating is linearly dependent on the state of occupancy $x$, according to (7). The upper limit of the calling number quantity, therefore, must also vary with $x$. Let us assume it is $Z_2'(x) < Z_2$. The calling number probability $w_r(x)$ during the simulation corresponds to the call-originating probability $w_a(x)$ and is proportional to $R(x)$ in (2). Thus

$$\begin{aligned} w_r(x) \sim R(x) &= R_0\left(1 - \frac{x}{q}\right) \\ &= Z_2'(x) - Z_1. \end{aligned} \tag{10}$$

To align the proportionality factors of (9) and (10), the quotient of calling-number and releasing-number probability is set equal to the quotient of call-originating and call-terminating probability

$$\frac{w_r(x)}{w_1} = \frac{w_a(x)}{w_e}.$$

From (3), (7), (9), and (10) we thus obtain

$$\begin{aligned} R(x) &= R_0\left(1 - \frac{x}{q}\right) \\ &= \frac{q - x}{q - y - c_v t_v} \cdot \frac{A}{1 - w(v_1)}, \end{aligned} \tag{11}$$

for $x < v_1$

$$R(v_1) = 0$$

and for the unknown $R_0 = R(0) = Z_2 - Z_1$

$$R_0 = \frac{Aq}{[q - (y + c_v t_v)][1 - w(v_1)]}. \tag{12}$$

The limits for the quantities of random numbers are thus

$$\begin{aligned} Z_1 &= v_2 \\ Z_2 &= v_2 + R_0. \end{aligned} \tag{13}$$

The upper limit of the quantity of calling numbers, which depends on the state of occupancy, is, according to (10)

$$Z_2'(x) = v_2 + R_0\left(1 - \frac{x}{q}\right). \tag{14}$$

Practical applications of (12) are discussed in Sections 4.3 through 4.5.

With the offered traffic distributed equally to all sources, each successfully established call must reduce the upper limit by $R_0/q$ and each release must increase it by the same amount. If no source is active, $Z_2'(0) = Z_2$.

### 4.3 ERLANG TRAFFIC

If, for the number of outlets, a very-large number of traffic sources have unrestricted access to the switching network to be investigated, the transition $q \rightarrow \infty$ can be effected in (12), (Erlang traffic)

$$R_0 = \lim_{q\to\infty} \frac{Aq}{[q - (y + c_v t_v)][1 - w(v_1)]}.$$

We obtain

$$R_0 = \frac{A}{1 - w(v_1)}, \tag{15}$$

for $v_1 \leq \min(v_2, v_{Z1}, v_{Z2}, \cdots)$

and

$$R_0 = A, \tag{16}$$

for $v_1 > \min(v_2, v_{Z1}, v_{Z2}, \cdots)$

because $w(v_1) = 0$ in (16). If several groups of outlets are connected to the switching network, $v_2$ denotes the total sum of all outlets, and $v_{Z1}, v_{Z2}, \cdots$ denote the number of links between the stages if a link arrangement is tested.

The probability $w(v_1)$ can be determined mathematically or, if the simulation program is arranged correspondingly, through the test itself. It is merely necessary to count the calling numbers that appear in the condition $x = v_1$. The simulated offered traffic $A$ can be computed with (15) only on completion of the test.

When assuming $q \rightarrow \infty$, the inlets of the tested switching network are usually not connected to traffic sources. Instead, traffic has already been concentrated by means of other switching networks. The assumption of unlimited access of traffic sources to the inlets thus is seldom realized in practice because it would require a full-availability switching network as a connecting link. However, it is mandatory to assume such a full-availability switching link. Otherwise an entire switching office has to be simulated simultaneously. This assumption is also justified if the losses or the waiting times, respectively, are small in the preceding networks.

### 4.4 Engset Traffic with Overflow Provisions

When generating artificial telephone traffic, special attention must be paid to the treatment of rejected calls [11] if a finite number of traffic sources are to be simulated.

If a call is rejected by the tested group of trunks but is able to reach the destination via an overflow trunk, the mean duration of the rejected calls is $t_v = t_m$. The test result presents an overflow rate instead of a loss. To be able to set $t_v = t_m$, the overflow capacity must be sufficient to ensure that no loss occurs. In practice, this cannot be realized. Compared to the total traffic offered for the direction tested, however, losses are usually very small and the above assumption causes only negligible error.

If $y + c_v \cdot t_m = c_A \cdot t_m = A$, we derive from (12)

$$R_0 = \frac{Aq}{(q - A)[1 - w(v_1)]}, \qquad (17)$$

$$\text{for } v_1 \leq \min(v_2, v_{Z1}, v_{Z2}, \cdots)$$

and

$$R_0 = \frac{Aq}{q - A}, \qquad (18)$$

$$\text{for } v_1 > \min(v_2, v_{Z1}, v_{Z2}, \cdots).$$

### 4.5 Engset Traffic Without Overflow

If the calls rejected by the tested trunks are lost, the holding time $t_v$ of the switching network inlets is normally very short. If $t_v$ approaches zero [11] and if the unknown traffic-handling capacity is set at $y = A\ (1 - B)$, we derive from (12)

$$R_0 = \frac{Aq}{[q - A(1 - B)][1 - w(v_1)]}, \qquad (19)$$

$$\text{for } v_1 \leq \min(v_2, v_{Z1}, v_{Z2}, \cdots)$$

and

$$R_0 = \frac{Aq}{q - A(1 - B)}, \qquad (20)$$

$$\text{for } v_1 > \min(v_2, v_{Z1}, v_{Z2}, \cdots).$$

The loss $B$ is not known before the test has been carried out. Thus, an $R_0$ has to be computed first for the estimated loss and, on completion of the test, the actually simulated offer $A$, using (19) and (20).

Equation (20) is important for testing subscriber switching networks because the finite number of subscribers must be taken into consideration and because definitely $q = v_1 > v_2$.

### 4.6 Performance of Traffic Tests

Since a simulation program should be written for as many applications as possible, the freely selectable parameters must first be determined for a particular test. The following information must be supplied to the computer in addition to the program:

**(A)** Size of the switching network to be tested.

**(B)** Type of traffic to be simulated.

**(C)** Traffic offered.

**(D)** Traffic distribution to the inlets of the network.

The duration of the test depends, for example, on the number of calls offered. Since a statistical equilibrium has to be simulated, it is also necessary to achieve a mean state of occupancy before beginning to count calls handled and calls rejected.

With the aid of one of the equations (15) through (20), it is possible to compute, at least approximately, the factor $R_0$ for a predetermined offered traffic $A$. This simultaneously establishes the limits for the quantities of releasing and calling numbers (13). If the inlets of the switching network are divided into a number of subgroups, the quantity of calling numbers must be divided similarly, to be able to determine to which subgroup an originating call will be directed. If the traffic offered is distributed unequally to the subgroups, $R_{0i}$ and the corresponding values $q_i$, $A_i$, $w(v_{1,i})$, and $B_i$ must be computed using (15) through (20) for every subgroup. The calling number subquantities then are

$$[v_2, v_2+R_{01}), \quad [v_2+R_{01}, \quad v_2+R_{01}+R_{02}), \cdots,$$

$$[v_2+\sum_{i=1}^{m-1} R_{0i}, \quad v_2+\sum_{i=1}^{m} R_{0i}). \qquad (21)$$

The modulus of limit changes is also different for each subgroup, that is, $R_{0i}/q_i$. The loss $B_i$ must be determined separately for each subgroup so as to be able to compute, using $R_{0i}$, the actual simulated offered traffic $A_i$.

### 4.7 Reliability of Test Results

The simulation of busy-signal systems as described here presents, as the principal result of a test, the relative frequency of loss, expressed by the quotient of rejected calls $c_v$ and offered calls $c_A$

$$B = c_v/c_A. \qquad (1)$$

The limit of relative frequency for $c_A \to \infty$ would yield the true loss $\bar{B}$. The test, however, must be terminated after a finite time. The result achieved is only an approximation whose accuracy increases with the number of rejected calls $c_v$.

In sampling theory, it is customary to give so-called confidence intervals for measured results $(B - \Delta B,\ B + \Delta B)$ within which, with a certain statistical probability, the true value of the measured term $B$ lies. This procedure can be used to indicate the degree of reliability of the test results.

The number of calls rejected during a test is liable to deviate randomly from the true mean value. To determine the loss deviation $\Delta B$, the variance of $c_v$ must be known. It can be determined theoretically for full-availability switching networks [12]. For the normally encountered case of switching networks with limited availability, the empirical variance is computed by registering during a test $n$ intermediate results $c_{v,i}$ $(i = 1, 2, \cdots, n)$ after an equal number of calls offered $c_A/n$, and deriving $n$ losses $B_i$ therefrom. With these intermediate results, it is possible to calculate the empirical variance

$$\sigma^2 = \frac{1}{n-1} \sum_{i=1}^{n} (B_i - B)^2 \qquad (22)$$

with

$$B = \frac{1}{n} \sum_{i=1}^{n} B_i = \frac{c_v}{c_A}. \qquad (23)$$

With the aid of Student's $t$-distribution, it is possible to determine the deviation $\Delta B$ under certain conditions. The true loss $\bar{B}$ is then expressed with the statistical probability $S$ by

$$B - \frac{1}{n^{1/2}} \sigma t_{S,n} \leq \bar{B} \leq B + \frac{1}{n^{1/2}} \sigma t_{S,n}. \qquad (24)$$

The factor $t_{S,n}$ derived from the $t$-distribution depends on the required statistical probability

$S$ and the number of intermediate results $n$. The limits of the confidence interval approach almost linearly to the mean value $B$, with quadratic extension of the test time.

The confidence intervals may be determined for the total loss as well as for subgroup loss.

## 5. Computer as a Traffic Machine

The individual elements of a switching network must be simulated in the computer to the extent that they influence the traffic flowing through the network. The outlets, and perhaps also the links, of the network are represented by storage cells into which digital free or busy signals can be entered. The duties of registers and markers that hunt for free links or outlets and occupy them, if available, are performed by a subprogram. We have to distinguish between two cases.

(**A**) If the establishment of connections between switching stages (Figure 3) and the order in which the outlets of a multiple are occupied follow rules that apply to a large group of switching networks, these rules will be programmed.

(**B**) If connections are apparently arbitrary, however, or if the outlets of several switching multiples are combined by grading that may vary widely, the connecting or grading pattern—in addition to the normal program—will be entered in suitable form into the computer. This is accomplished by writing the addresses of the links or outlets accessible from the following switching stage into the storage cells representing the links. In both cases, the subprogram must determine the route through the network that would be used for the call, at the same state of occupancy, in technically realized arrangements. If a call is to be released on a specific outlet, the busy criteria must be cancelled in the corresponding storage cell and in the cells associated with the links used. Besides the busying criteria, the subprogram must therefore write into the storage cells the links used to reach the respective outlets.

During the test, the total number of calls offered $c_A$ and of calls rejected $c_v$ must also be counted for each subgroup, if required. At the same time, loss $B$ and loss deviation $\Delta B$ are determined. If Engset traffic has been simulated, the actual offered traffic $A$ is computed using one of the equations (17) through (20).

Consider the relatively simple example in Figure 3 of a 2-stage link arrangement without directional selection (relay-type register-finder). The simulation program should be able to generate optionally either Erlang or Engset traffic, and the size of the network should also be selectable within certain limits.

The program establishes the number of subgroups, of links per subgroup, and of outlets in each switching multiple of the second stage. The type of link arrangement shown in Figure 3 is thus fixed. The connecting pattern is so easy and regular that it is worthwhile to program it. The outlets are tested one after the other from a home position for their states of occupancy.

Figure 4 is the block diagram for this simulation process and gives an outline of the program structure. A detailed description would

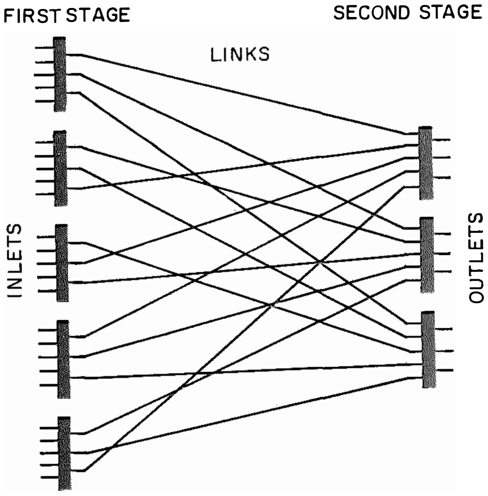


Figure 3—Two-stage link arrangement using 25 inlets, 15 links, and 9 outlets.

not be pertinent as it must be given for a specific computer model.

When designing a simulation program, there are additional considerations. It is desirable, for example, to work out programs for as large a group of switching networks as possible and to provide for all types of traffic; on the other hand, the computer time for a test should be minimized. A compromise must be found for these two contradictory demands.

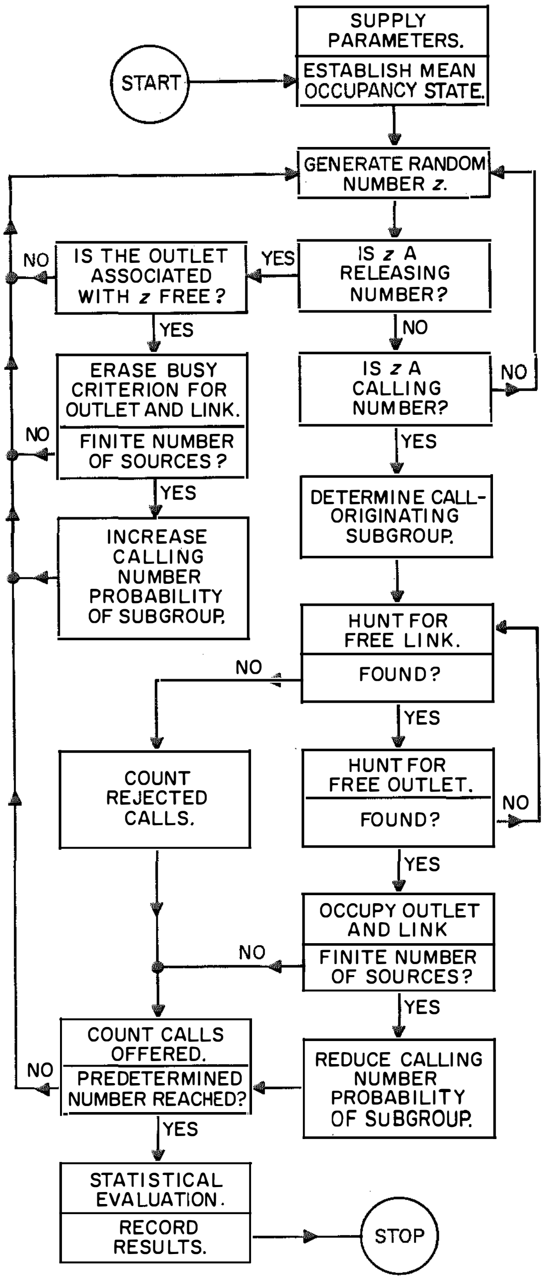


Figure 4—Block diagram of simulation program for 2-stage link arrangements.

## 6. Test Results

Test results were obtained with the program described in Section 5, using the *ER56* electronic computer. These results are only samples and were influenced by the method of generating the pseudo-random numbers and by the state of occupancy of the switching network at the beginning of the test. The program and the computer, however, did not influence the results. The switching network to be tested was chosen to permit a comparison of the test results with theoretical results verified in practical operation. The program for 2-stage link arrangements has been used here to test a 1-stage full-availability switching network having $v_2 = 27$ outlets and Erlang traffic (16). The first stage of the link arrangement in this case is 1 switching multiple with 27 outlets, each connected to only 1 outlet of a switching multiple of the second stage. This link arrangement theoretically corresponds to a 1-stage full-availability switching network.

The loss probability may be theoretically determined with the aid of the Erlang equation $B = E_{1,v2}(A)$ or may be obtained from [13].

In Table 2, test results and losses according to

TABLE 2

LOSS OF A FULL-AVAILABILITY SWITCHING NETWORK WITH $v_2 = 27$ OUTLETS (THEORETICAL AND TEST VALUES)

| Offered Traffic $A$ in Erlangs | Theoretical Loss $E_{1,27}(A)$ in Percent | Traffic Test | |
|---|---|---|---|
| | | Loss $B \pm \Delta B$ in Percent | Number of Offered Calls $c_A$ |
| 22.0 | 5.137 | 5.515 ± 0.742 | 20 000 |
| 19.5 | 2.205 | 2.140 ± 0.297 | 30 000 |
| 18.5 | 1.421 | 1.403 ± 0.290 | 30 000 |
| 17.5 | 0.852 | 0.862 ± 0.148 | 40 000 |
| 15.0 | 0.160 | 0.168 ± 0.067 | 50 000 |

[13] are compared for 5 different offers. Despite the relatively low number of calls offered, the results correspond well. The theoretical loss is never beyond the confidence interval and is usually very close to the mean value $B$.

## 7. References

1. J. Kruithof, "Rotary Traffic Machine," *Electrical Communication*, volume 23, pages 192–211; June 1946.

2. S. W. Broadhurst and A. T. Harmston, "Electronic Traffic Analyzer," *Post Office Electrical Engineers' Journal*, volume 42, pages 181–187; January 1950.

3. L. Kosten, "Beschrijving van een machine voor kunstmatig veerkeer op electronische basis," *De Ingenieur*, volume 63, pages 62–66; 1951.

4. R. W. Fraser and E. A. Elliman, "An Artificial Traffic Machine for Automatic Telephone Traffic," *Post Office Electrical Engineers' Journal*, volume 22, pages 31–39; 1929.

5. G. Neovius, "Artificial Traffic Trials Using Digital Computers," *Ericsson Technics*, volume 11, pages 279–291; 1955.

6. B. Wallström, "Artificial Traffic Trials on Two-Stage Link System Using a Digital Computer," *Ericsson Technics*, volume 14, number 2, pages 259–289; 1958.

7. G. Bretschneider and A. Wendt, "Datenverarbeitungsanlagen als Hilfsmittel für die Fernsprech- und Fernschreib-Verkehrsplanung," *Nachrichten-Technische Zeitschrift*, volume 14, pages 487–492; 1961.

8. J. Moshman, "Generation of Pseudo Random Numbers on a Digital Calculator," *Journal of Association for Computing Machinery*, volume 1, pages 88–91.

9. S. V. Hoerner, "Herstellung von Zufallszahlen auf Rechenautomaten," *Zeitschrift fuer Angewandte Mathematik und Physik*, volume 8, pages 26–52; 1957.

10. J. V. Neumann, C. E. Forsythe, and P. C. Hammer, "Monte Carlo Methods," United States Bureau of Standards, Applied Mathematics Series, number 12, pages 36–38; June 1951.

11. H. Wagner, "Über eine neue Verteilungsfunktion für den Fernsprechverkehr in vollkommenen Koppelanordnungen," *Archiv der Elektrischen Übertragung*, volume 14, pages 508–514; 1960.

12. L. Kosten, J. R. Manning, and F. Garwood, "On the Accuracy of Measurements of Probabilities of Loss in Telephone Systems," *Journal of the Royal Statistical Society*, Series B, volume 11, pages 54–67; 1949.

13. C. Palm, "Table of the Erlang Loss Formula," second edition, Stockholm; 1954.

**Günter Dietrich** was born at Liebichau, Kreis Pr. Stargard, Poland, on 25 July 1932. In 1955 he received a diploma in mathematics from the University of Rostock.

In 1955, he joined Wissenschaftlich-Technisches Büro für Gerätebau in Berlin, where he was engaged in work on the development of special electrical measuring equipment. In 1958, he joined Standard Elektrik Lorenz, and is now chief of the department for telephone traffic theory.

**Heinz Wagner** was born in Künzelsau, Germany, on 26 January 1927. After serving as a public-school teacher, he studied mathematics at the Technische Hochschule in Stuttgart from which he received a diploma in 1958.

In 1957, he joined Standard Elektrik Lorenz where he is now engaged in studies of telephone traffic theory.

# Automatic Switching for Telex–Operation

**H. L. HORWITZ**

American Cable and Radio Corporation; New York, New York

The 7E rotary switching equipment [1] was selected for inclusion in the New York and San Francisco exchange centers of American Cable and Radio Corporation to provide both expanded facilities and increased automation. Fully automatic operation had already been established for direct subscriber-to-subscriber service between the United States, United Kingdom, and France.

It was necessary in the over-all design and adaptation of the 7E equipment to maintain compatibility with the other United States telex * carriers and with the administrations of some 70 directly or transit-connected overseas countries. Numerous problems pertaining to mandatory or desirable operating features were resolved during the course of manufacturing the equipment and prior to the official cut-over on 1 September 1962. The more-important features of the installations are briefly described.

## 1. Prefix Numbering Plan

To provide for the recording of both fully and semiautomatic calls in the 7E registers, it was necessary to develop an international prefix numbering plan having an indicator for each country that provides telex service. The basic plan [2] was developed in cooperation with the other United States carriers.

A 3-digit prefix precedes the subscriber's assigned telex number. The 1st digit indicates the continent and the 2nd and 3rd digits are general location and country or city indicators, respectively. In countries served by more than one carrier, the 2nd digit also identifies the carrier. For economy and other reasons, the 1st digit, the continent indicator, is presently omitted for traffic exclusively within the continental United States.

## 2. Overseas Trunks and Signalling

Trunks to overseas networks are provided through frequency-modulated voice-frequency telephone cable or multiplex radio channels with automatic error correction utilizing a constant-ratio code between terminals of the various continents.

Incoming trunk calls require keyboard selection and *A*-type [3] signalling. The various signal modes of outbound trunk calls are designed to conform with the requirements of overseas administrations with keyboard or dial selection and *A* or *B* type of signalling, respectively.

Incoming trunk calls to subscribers are routed automatically through the use of the subscriber's listed number and the assigned 3-digit prefix.

A transit call to a subscriber of an overseas administration that is provided with automatic switching is routed through the 7*E* facilities without intervention of the telex operator. Alternatively, calls directed to networks without such facilities are automatically routed for completion to the operator of the gateway city of the terminating network.

## 3. Adaptation to United States Domestic Telegraph and Telephone Networks

Special adapter equipment is provided for connecting trunks of the dial-operated domestic telephone system, as these are normally channelled through the conventional telephone switching systems and terminate in special applique apparatus installed in the New York and San Francisco exchanges.

The operating characteristics of the telephone network as later described do not correspond with the specifications of the Comité Consultatif International Télégraphique et Téléphonique for telex service as employed by overseas administrations. Translators for interpreting variations in a number of keyboard characters, special convertors to accept speed character-

---

* Telex indicates private subscriber-to-subscriber teleprinter operation over the switching and transmitting systems of the public telephone and telegraph networks.

istics of alternative 45- and 50-baud systems, and many other devices of associated equipment had to be provided.

Of particular importance, special arrangements are provided for signalling capable of converting the conventional telephone tone signals to suitable control signals compatible with the international specifications. This permits overseas operators, without intervention of local operators, to complete calls directly into the telephone networks. An intercept board is provided for incompleted calls due to reception of telephone tone signals (busy, no answer, or no such number) that cannot be interpreted, the intercept operator informing the caller accordingly. Provision is also made so that this operator can re-establish the call if desired. If the overseas caller does not wish to select the telephone subscriber's number, the call is automatically routed to the intercept operator for completion.

Connections to the domestic telegraph system are provided for dial pulse and *B*-type signalling.

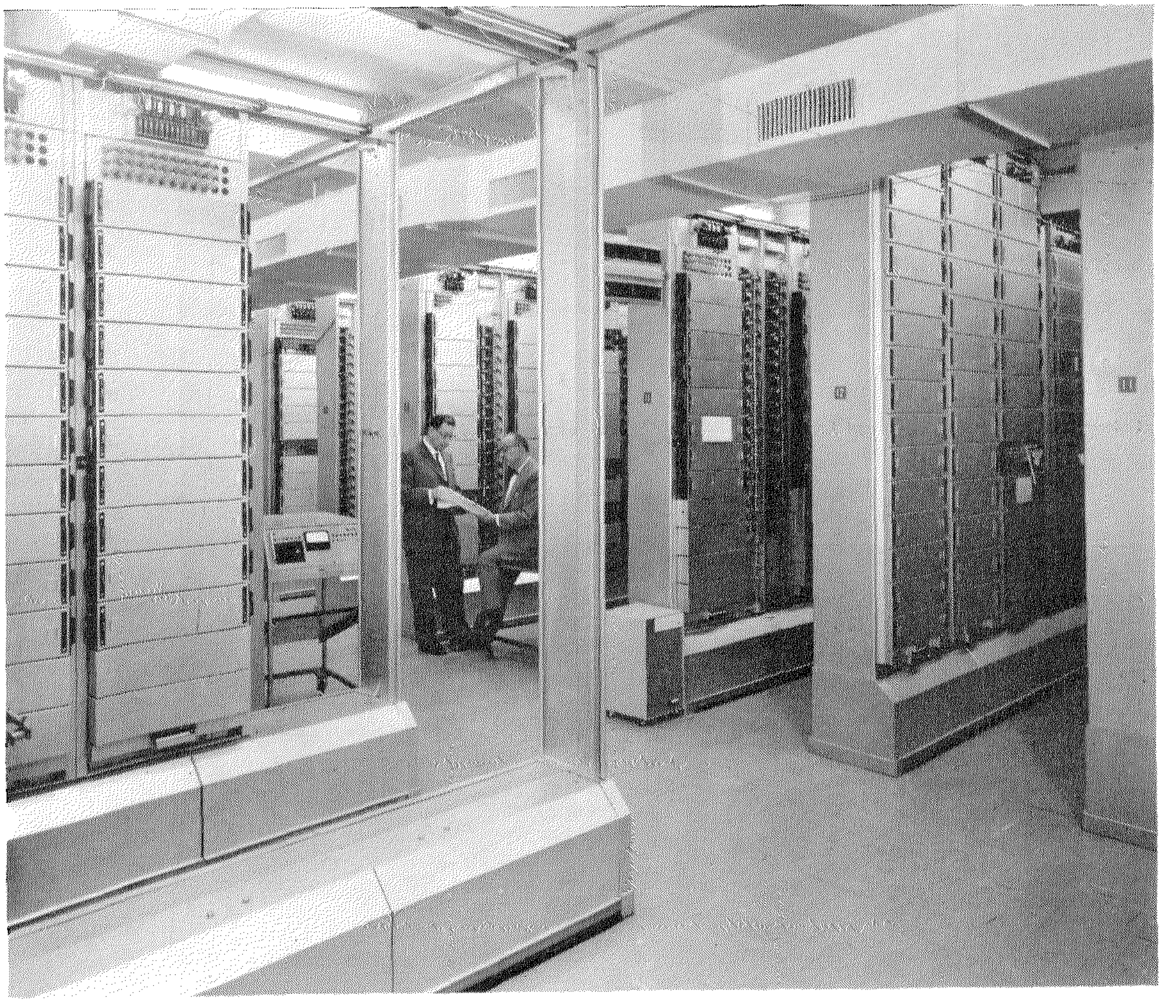


Figure 1—A view of the switching room at New York with the author (right) and Mr. T. Buyssens, project engineer for Bell Telephone Manufacturing Company (Antwerp).

On seizure of an incoming trunk from the telephone or telegraph system, no further selection is required by the calling party. Such calls are routed directly to telex operators, without register operation, for completion.

## 4. Trunk-Status Board

A trunk-status panel board, visible from all telex operator and overflow positions, is provided with nameplates for all connecting networks and their assigned trunk numbers. Two lamps, connected to each trunk as signals for out-of-order or busy conditions, serve as a ready indicator of trunk conditions.

## 5. Telex Operator Positions

Each telex operator position is equipped with 2 printers and 6 connecting circuits, each capable of handling 2 calls simultaneously and of establishing up to 6 calls. Incoming calls are automatically distributed and connected to idle positions. During intermittent peak traffic conditions when all operator positions are busy, further calls are automatically routed to overflow positions. The telex operators extend calls through the switching system by keyboard selection. These cordless positions are equipped with control keys and the indicating lamps necessary for operation and supervision.

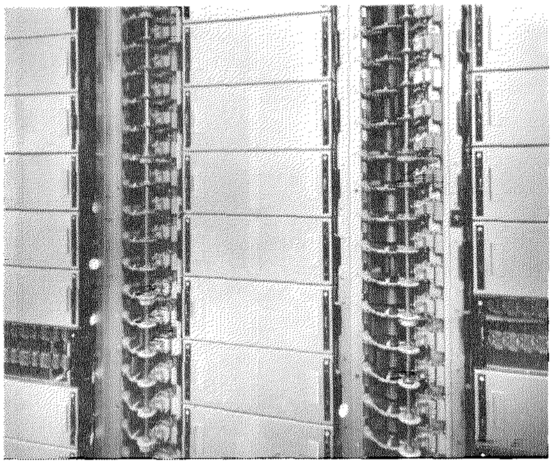


Figure 2—A typical rack of machine-driven selectors with their associated banks of relays.

Each position is fitted with electrically operated elapsed-time indicators. At termination of a call and release of the called party, provisions are made for the operator to transmit the charges to the calling party without setting up for a new call. In the arrangements for telephone and telegraph subscribers, however, where the equipment facilities prevent holding of the calling party, a device is provided that accepts a time-and-charge request from those subscribers in the form of an assigned telegraph code signal (5 times upper-case *M*) to alert the operator and for the release of the called party.

On a fully automatic call, the charges are automatically sent to the calling party at the termination of the connection without intervention of an operator.

## 6. Overflow Facilities

The overflow positions are similar to the regular operator positions. They are manned by supervisory personnel, are additionally provided with printers for each connecting circuit, and have arrangements to set up conference calls, to provide information to subscribers, and to provide other special services in addition to the handling of overflow calls, which are automatically diverted from fully occupied regular operating positions.

## 7. Message Traffic Facilities

All subscribers have facilities to use the telex printers for sending and receiving conventional overseas telegraph messages, which are billed at the established per-word rates. Message receiving and message sending positions are provided for this service.

### 7.1 RECEIVING POSITIONS

The message receiving positions, each equipped with one printer to an associated trunk and normally unattended, can be reached by any subscriber selecting the assigned code number. The calling party, receiving an automatically sent answer-back signal from an idle position, then proceeds to send his message. The calling party's number is also recorded for billing purposes. These positions are divided into 2 groups to facilitate routing of traffic in conformance with the requirements of alternative class of service, as specified.

### 7.2 SENDING POSITIONS

The message sending positions are provided with individual connecting circuits and, by means of keyboard selection, the operator may complete calls to any subscriber in the continental United States.

## 8. Subscribers' Equipment

Telex printer stations in subscribers' offices are equipped with control boxes containing a push button for initiating and releasing calls. Since keyboard selection is provided, service indicators, such as proceed to select, trunk line busy, called subscriber busy, trunk out of order, et cetera, are handled by typed messages. Some subscribers have provisions for off-line operation necessary for tape preparation. Such operation is interrupted in favor of incoming calls.

The proceed-to-select signal, containing an indicator for identification purposes, is terminated by upper-case *D,* which trips the subscriber's answer-back device. The *A/B* code contains the subscriber's number, which is extracted and stored in the *7E* register for billing purposes. The subscriber selects the called party's number, including the prefix. The register extends the call directly to the called subscriber of a fully automatic system or the call will be routed for completion to the telex operator of the gateway city.

## 9. Time and Date

The time-and-date information is transmitted to the calling and called subscribers on completion of a call. The data contains the progressive number of the day of the year; the carrier indication, which is essential for calls to telephone and telegraph subscribers to identify the carrier that routed the call; and Greenwich Mean Time.

## 10. Reperforator Equipment

Reperforators are provided to receive data in the 5-unit telegraph code from the memory circuits associated with the fully automatic connecting circuits. At the termination of each such call, which originates from any subscriber handled by the New York, San Francisco, or Washington offices, the memory circuit presents the following data to an idle reperforator: calling party's number, called party's number, time, date of call, and elapsed billable time.

## 11. Automatic Billing

Data from the punched tapes from the above reperforators is used for accounting and billing.

## 12. Monitoring Facilities

Banks of printers, each equipped with an automatic paper take-up spool arrangement, are connected to active trunk circuits to provide monitoring facilities for supervisory personnel. This provides for periodic visual inspection of operating conditions with hard copies of traffic for ready and future reference in case of errors.

## 13. Test Facilities

Automatic transmitters are provided for sending the "quick brown fox" sentence in biased or unbiased form. This test signal is automatically connected to any calling trunk or subscriber's line on reception of a special service code number assigned to this facility.

Mobile test sets equipped with printer, transmitter-distributor, and distortion-analyzer equipment are available in the automatic switchrooms for plug-in connection to any trunk, subscriber's line, or operator's position for analyzing faults that may occur.

An automatic routine test arrangement, incorporated in the *7E* switchroom apparatus, provides for fully automatic circuit checking with an associated paper-tape device for recording faults and their location. This equipment will be operated periodically under maintenance schedule.

## 14. Conclusions

This system is intended to provide ultimately connections between subscribers in the United States and all telex subscribers overseas without the intervention of operators. This service may be extended to the domestic telephone- and telegraph-network telex subscribers.

Arrangements have been completed to extend these automatic switching facilities through the New York exchange to subscribers in Washington, District of Columbia.

## 15. References

1. H. L. Van Lommel and P. J. Clemens, "Automatic 7E Rotary Teleprinter Switching System," *Electrical Communication,* volume 34, pages 38–55; March 1957.

2. "Question 8/1—World-Wide Telex Routing Plan," Comité Consultatif International Télégraphique et Téléphonique, Technical Committee of the United States Telegraphic Carriers, Committee I—14-E and Committee X—23-E, Release of 19 February 1962.

3. "Recommendations Concerning Telegraph Switching," Red Book, Comité Consultatif International Télégraphique et Téléphonique, volume 8, pages 119–120; March 1961.

**Henry L. Horwitz** was born on 30 March 1901 in Berlin, Germany. He received a telecommunication engineering degree from the Institute of Technology in Berlin in 1923.

He worked on automatic telephone switching systems in Europe including 15 years in Italy. In 1944, he joined Federal Telephone and Radio Corporation. In 1959, he was transferred to ITT Communication Systems and came to American Cable and Radio Corporation the following year as a specialist on automatic switching.

Mr. Horwitz holds 23 patents. He is a Member of the Institute of Electrical and Electronics Engineers.

# Automatic Switching for Telex–Equipment

**P. J. CLEMENS**

**Bell Telephone Manufacturing Company; Antwerp, Belgium**

As mentioned in the first part of this paper, the New York and San Francisco exchange centers of the American Cable and Radio Corporation use the *7E* rotary teleprinter switching system.

In calling, the need for a telephone-type dial is avoided by using the teleprinter itself. The numbers sent from its keyboard are stored in a register that supervises the subsequent switching. The search is under electronic control at the register, which performs several functions that in other designs are handled by line finders or selectors, with simplification of that corresponding apparatus.

The position of the single-motion selectors is under direct control of the register. A 450-hertz source produces 12 phases each 30 degrees apart at two voltages, 12 and 20 volts, providing 24 distinguishable conditions. Each such condition identifies a different group of outlets in the group-selector arcs and in the final selectors identifies the first digit of a decimal group and the individual outlets.

Single-motion selectors are mechanically simpler and are easier to maintain than two-motion switches. They provide greater flexibility in assignment of lines as their outlets are not divided rigidly into, say, 10 levels of 10 outlets each, but can be electrically divided into any number of groups with differing numbers of outlets in the groups.

The routing diagrams for the three types of calls, incoming, outgoing, and transit, are shown in Figures 1 through 3.

The power-driven single-motion rotary switch shown in Figure 4 is employed both as line finder and selector. It has 100 sets of outlets and 7 double brushes. Neither the selector nor the finder is provided with a home position, but remains on the terminal to which it was last positioned. This reduces hunting time as the switch positioned nearest to the new-call outlet will accept the assignment with minimum motion, thus reducing wear on all the hunting switches. A universal 2-relay circuit is associated with each switch regardless of whether it is used as a group or as a final selector.

The wire-spring relay shown in Figure 5 is similar to that developed by Bell Telephone Laboratories in the United States. Its efficient

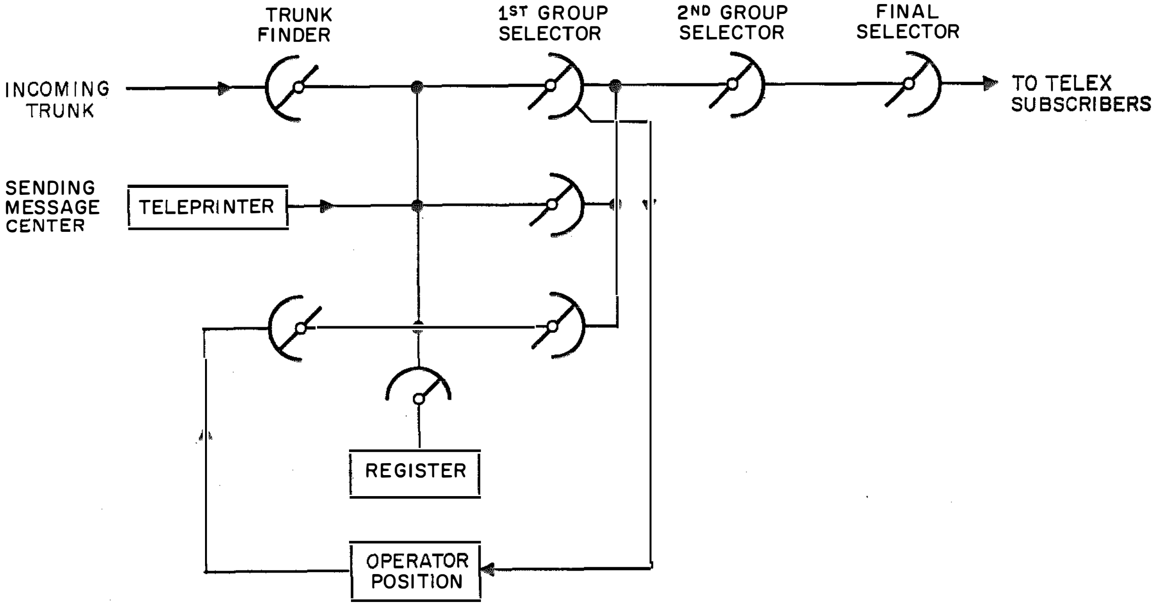
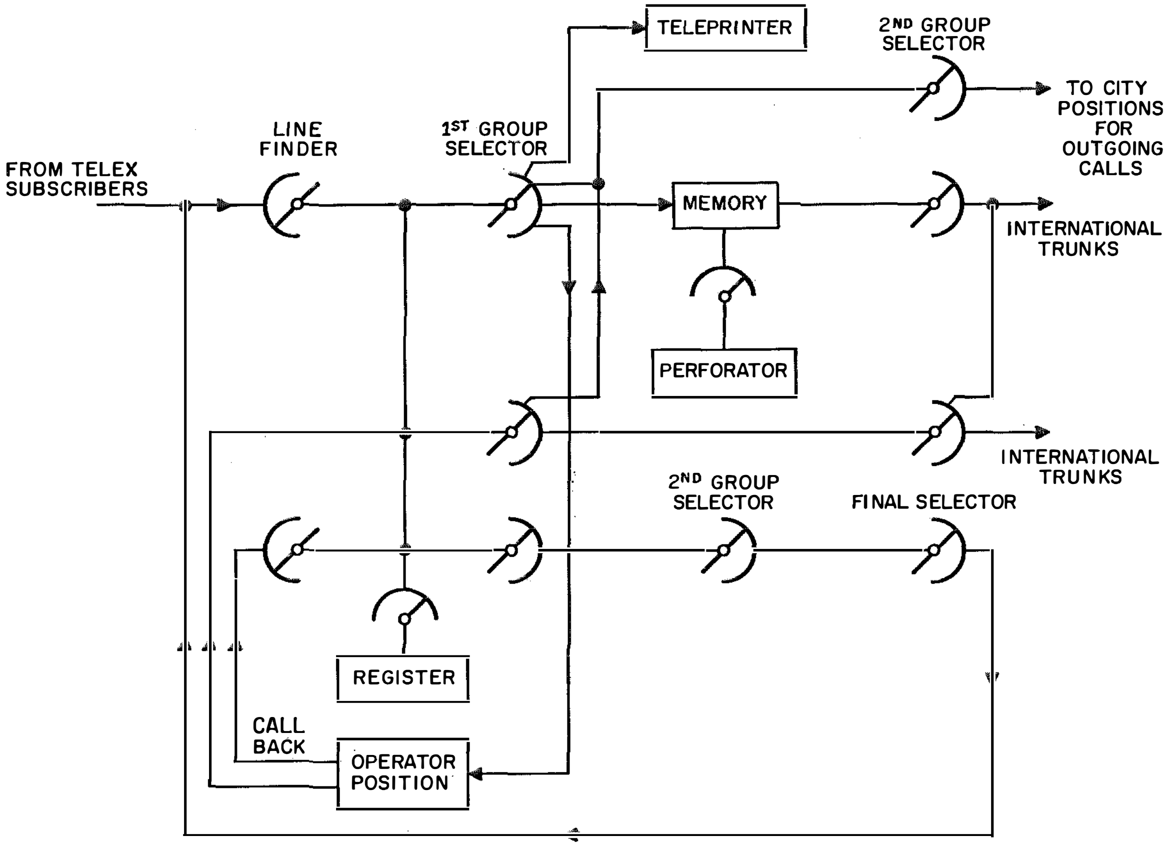


Figure 1—Routing diagram for incoming calls to telex subscribers.

Figure 2—Routing diagram for outgoing calls from telex subscribers.



magnetic circuit and pretensioned wire springs combined with card operation provide long life and fast operation with low actuating power.

Its excellent contact performance stems from the use of palladium, the equalizing of pressure on all contacts, complete independence of the twin contacts from each other, the special shape of the contacts, and the provision of contact covers. Palladium is used for the fixed contacts and palladium-gold for the moving contacts to improve operation under different load conditions.

Relays are mounted in completely dustproof enclosures. Those associated with the selector circuits are of the jack-in type to permit variation of the capacity of a group to meet traffic requirements.

A single switch rack accommodates equipment for 300 subscriber-station lines including the 300 line circuits, 3 groups of connecting circuits with repeaters, and 3 groups of final selectors. The interconnecting cables are in the base of the rack. An enclosed cable rack on the floor of the switchroom provides for interconnecting the racks.

Each subscriber's line circuit consists of 2 wire-spring relays and 2 adjustable resistors. The relays for 10 line circuits are grouped in one apparatus unit. The adjustable resistors are mounted at the top of the bay. These resistors control the line and bias currents of the repeater that is part of the connecting circuit and they must be adjusted before a line is put in service to compensate for the line characteristics.

Figure 3—Routine diagram for transit calls.

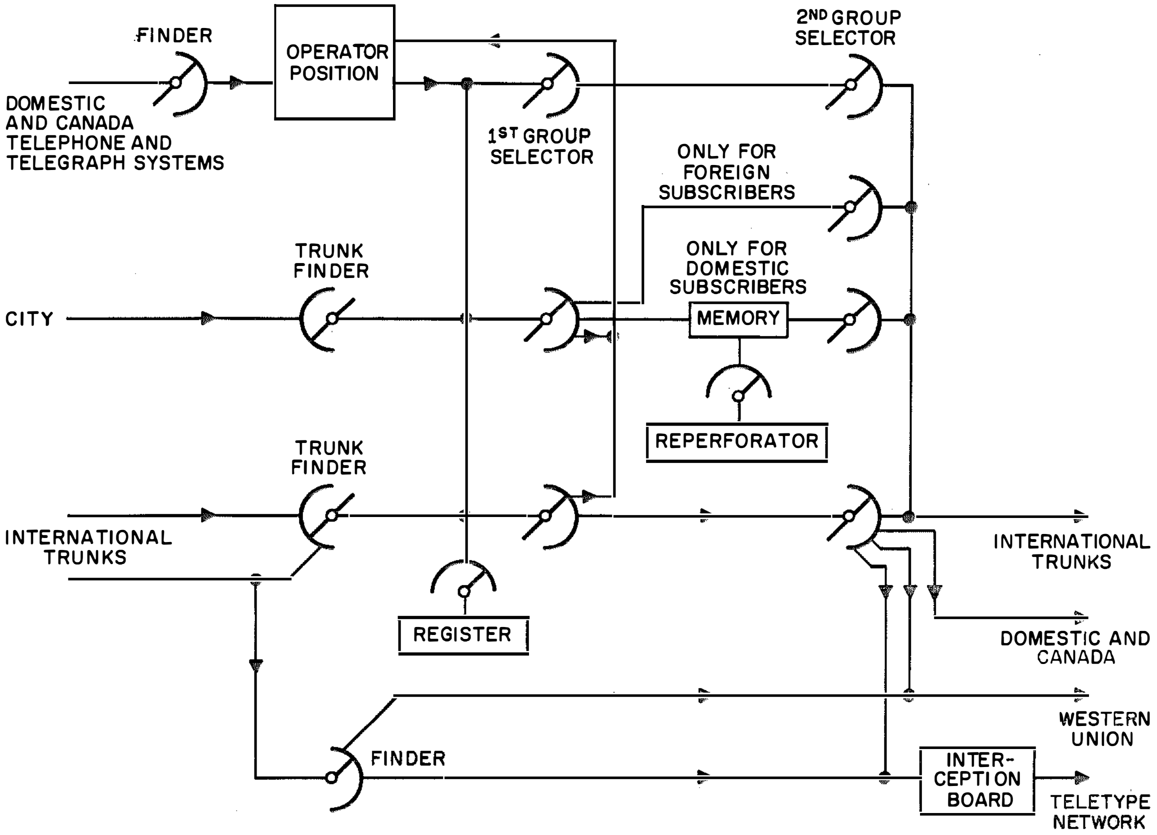


Figure 4—Rotary switch having 7 double-ended non-homing brushes and 100 sets of outlets.

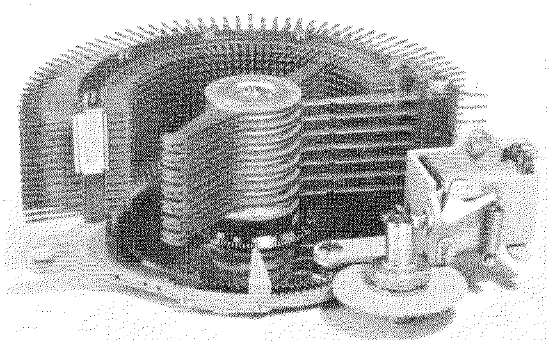
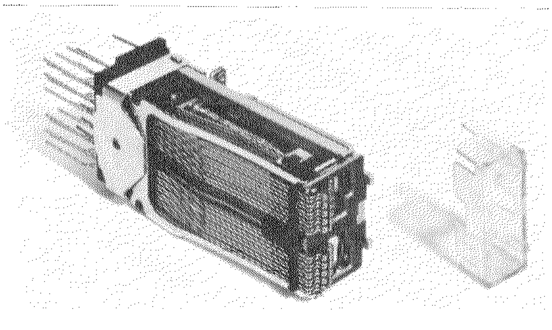


Figure 5—Wire-spring relay and contact cover.



**P. J. Clemens** was born in Utrecht, The Netherlands, on 3 May 1899.

In 1921, he joined Bell Telephone Manufacturing Company, Antwerp, as a student engineer. He was placed in charge of design and development of telegraph switching systems in 1936.

Mr. Clemens is now chief of the telegraph department.

# High-Power Precision-Stabilized Rectifiers

**K. KEMPER**

**Nederlandsche Standard Electric Maatschappij, N.V.; The Hague, Netherlands**

Certain physical experiments carried out in a strong magnetic field put stringent demands on the stability of the field and consequently on the current in the magnet coil. These demands are so rigid that formerly they could be met only by using batteries.

To avoid the disadvantages of batteries with regard to space, maintenance, efficiency, requirement of auxiliary equipment for precise regulation, et cetera, more-convenient solutions have been developed.

For example, there have been fairly successful attempts to provide motor-generator equipment with automatic control to obtain the constant current required. In this way it is possible to reach stabilities better than 1 in $10^3$, but many problems of space, efficiency, vibration, and ripple remain unsolved.

Using electronically controlled rectifier equipment did not seem very attractive at first, since the electron tube operates essentially at a high impedance level and is not suitable for currents of more than a few amperes directly.

The advent of power transistors offered important possibilities. These transistors are suitable for application as control devices because of their low-impedance characteristics, and they can be inserted directly in the main current path. A final development concept was reached based on the use of power transistors, silicon power diodes, and transistors in all associated electronic circuits.

As a result, a stability of 5 parts in $10^5$ of the selected current for an 8-hour period or 5 in $10^6$ for a period of several minutes with an effective ripple of 5 in $10^6$ can be guaranteed for equipments now being built.

In these equipments, the control circuit can cope with ±10-percent slow voltage variations, ±3-percent transient variations, and also resistance variations in the magnet coil.

Rectifiers are now being built with ratings from 1.4 kilowatts at a maximum of 8 amperes through 112.5 kilowatts at a maximum of 450 amperes. One of the 112.5-kilowatt units will be used to operate the famous "big magnet" in the Kamerling Omnes Laboratories at Leiden.

The essential problem in developing high-stability equipment of this type is not the circuit design, which is quite conventional, but the necessity to construct every detail with utmost care and to pay attention to minor considerations that are normally neglected.

Figure 1 shows the simplified circuit of the rectifiers. The 3-phase supply is connected to the motor-driven regulator transformers, which provide a voltage adjustable from zero to a level slightly above the supply voltage.

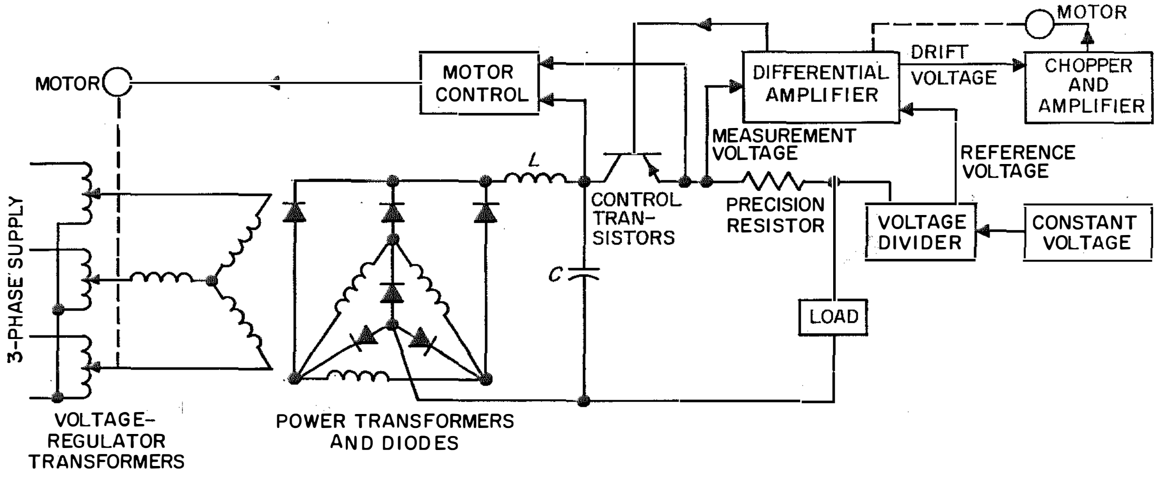


Figure 1—Simplified circuit of rectifier and control system.

This voltage is converted by the power transformers to the desired level and then rectified by water-cooled silicon diodes. The ripple is reduced to 1 percent effective by a smoothing filter consisting of inductance $L$ and capacitance $C$. The rectified current flows through control transistors, a precision resistor, and the load.

When the control circuit is inoperative, the output current is a function of the supply voltage and the load resistance. Stabilization of load current takes place as follows. The precision resistor provides a voltage proportional to the load current. This voltage is compared with the reference voltage obtained from a precision voltage divider operating from a constant-voltage source. The difference between the two voltages indicates the deviation of the measured value of the load current from the desired value. This difference is amplified and used to bias the control transistors to provide a limited control range capable of coping with fast variations. For economy, the number of transistors is related to the load current. Slow variations caused by selected changes in output current and by slow supply-voltage variations are corrected by simple motor-driven control of the voltage-regulator transformers. The average value of the transistor voltage is thus stabilized.

Proper construction of the precision resistor is very important to maintain current stability. Its resistance value should stay constant under all load conditions and it should be isolated from induced extraneous voltages. It is constructed as follows. Two copper bars separated by a thin layer of mica are firmly clamped together. The ends of a circularly bent sheet of manganin, which forms the actual shunt, are soldered to the narrow side of each bar. Thus the temperature difference across the manganin-copper junctions is minimized because of the very small heat resistance between the bars. As a result, the thermoelectric potential difference will not exceed 0.3 microvolt from zero to full load (1 volt at 450 amperes) across the resistor. The voltage terminals are mounted on the side of the shunt in direct contact with the manganin. One terminal wire runs along the inside of the manganin and then enters a common shield with the wire from the other terminal. In this way they are not exposed to induction fields. Furthermore, advantage is taken of the fact that the resistance-temperature characteristic of manganin follows a parabolic curve with its top at about 30 degrees centigrade (86 degrees fahrenheit). The whole unit is immersed in oil that is thermostatically controlled within ±0.5 degree centigrade (±0.9 degree fahrenheit) at this optimum temperature. Forced oil circulation keeps the temperature difference between oil and manganin within this tolerance. Surplus heat is extracted by water cooling. After careful adjustment and aging, the precision resistor can maintain a stability of 1 in $10^6$ for many hours.

Another direct influence on the stability of the output current is the reference voltage to which the shunt voltage is compared. This reference voltage, which can usually be adjusted between 0 and 1 volt, is obtained from a voltage-divider circuit connected to a very-stable voltage source consisting of 3 zener-diode stages in cascade that reduce supply-voltage variations to 1 in $10^6$. The ripple on the reference voltage is also 1 in $10^6$.

The only other serious problem is the fact that the parameters of the semiconductors, especially those in the final stabilization circuit, are temperature dependent. Three remedies effectively reduce this disadvantage.

**(A)** Careful compensation for positive and negative thermal effects.

**(B)** Enclosing parts within a copper bar to equalize their temperature variations.

**(C)** Placing the whole circuit in a thermostatically controlled oven to keep temperature variations within ±0.5 degree centigrade (±0.9 degree fahrenheit).

Thus the output voltage enjoys a stability in some units of 1 in $10^6$ for many hours. It is also possible to keep the ripple voltage from exceeding a few microvolts by rigid shielding and careful arrangement of parts.

Adjustment of the voltage-divider circuit controlling the reference voltage for various output currents may be based on the following considerations: stability, possibility for remote control, resolving power, accuracy of resetting, and absolute value of the setting. Ordinary multiturn wire-wound potentiometers are unsatisfactory because the sliding contact may touch more than one turn at a time and cause variations of resistance with vibration and shock.

The voltage-divider network used consists of a number (usually 4) of high-quality 10-position step switches with wire-wound precision resistors. Thus the output current can be digitally adjusted in 10 000 equal steps. For instance, selection of the number 3785, for a power supply with a maximum output of 100 amperes, equals an output current of 37.85 amperes. The output current deviation does not exceed 5 units in the last digit, or 50 milliamperes. It should be noted, however, that this concerns only the current-setting accuracy and has nothing to do with the actual stability of the output current, which could be of the order of 2 milliamperes.

The entire voltage-divider circuit is placed in a thermostatically controlled oven. The switches are motor-driven using simple servo circuits. The control switch for these servo circuits can be installed at the most convenient spot with regard to the experiment for which the stable current is needed.

Without precautions, the output current could make enormous jumps (10 amperes) because of the fast response of the control system if the digital switch is adjusted indiscriminately. This is avoided by preventing any change in output current until selection of the new current setting has been completed. The output current then smoothly adjusts itself to the newly selected value. This method of resetting also permits slow and continuous scanning of a given current range.

Finally, the differential amplifier is of critical importance to the stability of the output current. This direct-current amplifier suffers from drift and a gain characteristic that slopes downward toward the high frequencies, where a certain amount of phase shift occurs. Very-high gain is required. A typical input-signal variation of 10 microvolts across 1000 ohms produces 0.01 microampere and 0.1 picowatt to control output values across the control transistors of 4.5 volts and 450 amperes, or 2 kilowatts, indicating a power amplification of $10^{16}$. It is necessary, for example, to use as many as 4 voltage- and 5 current-amplification stages in cascade.

The required bandwidth is from direct current to 20 kilohertz, the limiting factor being the 10-millisecond time constant of the smoothing filter in the main rectifier circuit. As a result, higher frequencies in the output voltage influenced by a sudden supply variation are considerably attenuated. Current components containing frequencies higher than 5 kilohertz are sufficiently attenuated to reduce any variations in the output current considerably even without special means of control. Accordingly, the phase shift in the amplifier must be negligible from direct current to a few kilohertz, while at 20 kilohertz a 10-to-20-degree phase shift combined with a few decibels reduction in gain is acceptable. Therefore, the required characteristic can be met without too much difficulty because of the relatively small bandwidth, although the parts and wiring require careful arrangement.

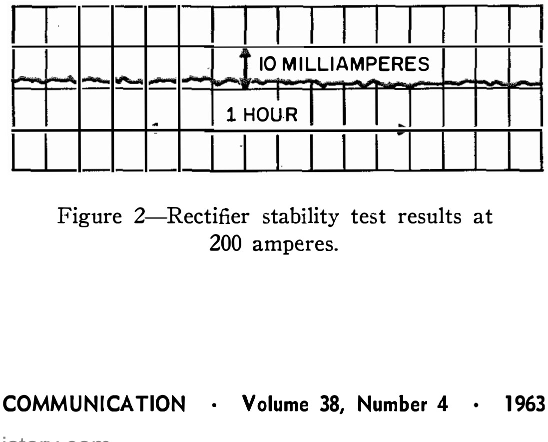


Figure 2—Rectifier stability test results at 200 amperes.

An additional difficulty in keeping the control system free of oscillation is that these rectifiers must be capable of energizing various types of magnet coils and other loads having a wide variety of impedances. The direct-current cable capacity may also be a problem. To guarantee proper operation under all conditions, the system has been provided with several feedback loops. One of these is obtained by re-injecting the ripple component of the output voltage via a simple network into the amplifier, resulting in effective ripple suppression.

The deviation of the output current from its average value under the influence of quick disturbances never exceeds ±5 parts in $10^5$.

With regard to drift, the amplifier output voltage should have a certain absolutely constant value under ideal conditions when the differential input terminals are short-circuited. In reality there are two deviations, drift and amplifier noise, the first causing a variation in the direct-current level and the second being an alternating voltage containing a certain frequency spectrum. The latter factor is negligible since the noise level at the amplifier input is less than 1 microvolt. To correct for drift, it is necessary to provide chopper stabilization for the amplifier. The differential-amplifier direct-voltage drift of several microvolts is chopped at 50 hertz by a precision switch and converted into a small alternating voltage.

The phase of the alternating voltage depends on the polarity of the drift voltage. After high amplification, it is used to control a small servomotor coupled to an adjustable resistor in the differential amplifier. If drift occurs, the servomotor drives the adjustable resistor in the direction to cancel the drift voltage. The drift of the differential amplifier is thus held to less than 1 microvolt. The cancellation rate is relatively slow and has a bandwidth of only a few hertz.

By summing the maximum errors of all individual circuits, the expected stability of the

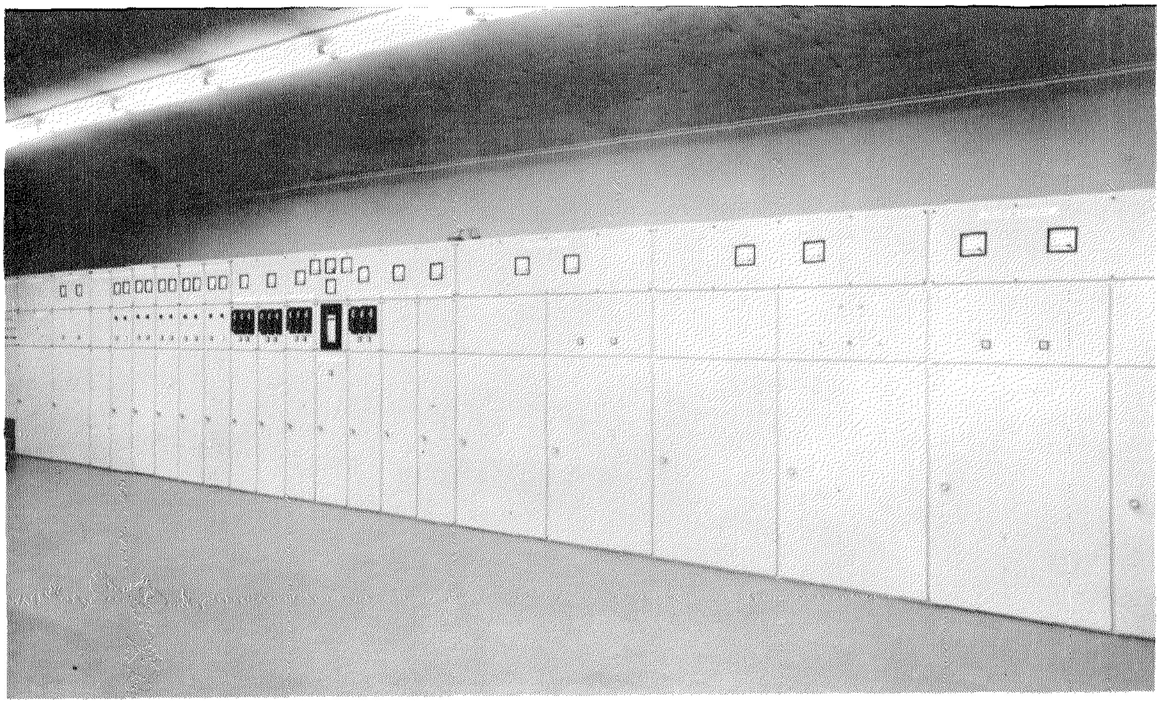


Figure 3—High-stability rectifiers located in the Physical Laboratory of the University of Groningen.

output current can be predicted. If we consider that in a certain rectifier installation the compared shunt and reference voltages vary between 0.2 and 1 volt, we find the following maximum errors: shunt, 1 microvolt; reference voltage, 3 microvolts; voltage divider, 3 microvolts; and differential amplifier, 1 microvolt. The total error is less than 10 microvolts or a stability of 1 in $10^5$ at the maximum output current, or 5 in $10^5$ at 0.2 of the maximum output current. An additional safety margin is that it is highly improbable that all errors are both maximum and additive.

Figure 2 shows part of a stability test record of one of the rectifiers. A distribution room in the Physical Laboratory of the University of Groningen, housing a series of high-stability rectifiers, is shown in Figure 3.

## Acknowledgement

The author is greatly indebted to his assistants, J. de Graaf and A. Cools, who have contributed substantially to the realization of the equipment outlined in this article. He also thanks Mr. W. A. Hansen of The Netherlands Government Building Authority and the directors of the various research institutes for their full cooperation and confidence.

**K. Kemper** was born on 8 January 1925, in Haarlem, Holland. He is a graduate engineer of the Technological University of Delft.

After several years of work on automatic traffic controls, he joined Nederlandsche Standard Electric in 1952. In 1954, he was placed in charge of circuit engineering for the rectifier department. Since 1957, Mr. Kemper has been responsible for engineering and development for the rectifier and automatic-controls department.

# United States Patents Issued to International Telephone and Telegraph System; August–October 1962

Between August 1962 and October 1962, the United States Patent Office issued 88 patents to the International System. The names of the inventors, company affiliations, subjects, and patent numbers are listed below.

P. R. Adams and E. C. L. DeFaymoreau, Federal Telecommunication Laboratories, System for Supporting and/or Controlling Bodies, 3 048 044.

R. T. Adams, Federal Telecommunication Laboratories, Crystal Controlled Oscillator Circuit with Frequency Control Means, 3 054 973.

F. J. Altman, ITT Laboratories, Signal Receiving System, 3 048 782.

B. E. Ash, Standard Telephones and Cables (London), Tension Control Mechanism for Winding Machines, 3 059 869.

D. R. Barber, Standard Telecommunication Laboratories (London), Magnetic Trigger Devices, 3 053 993.

J. A. Baudin, ITT Federal Laboratories, Pulse Modulator Circuit for Generating Paired Pulses, 3 053 999.

J. A. Baudin, ITT Federal Laboratories, Channel Selector Servo Circuit, 3 051 881.

C. B. Bendell, ITT Federal Laboratories, Transient Response Testing Apparatus, 3 061 777.

H. Benmussa, J. P. LeCorre, and C. Weill, Laboratoire Central de Télécommunications (Paris), Magnetic Circuits, 3 056 038.

H. Benmussa and J. P. LeCorre, Laboratoire Central de Télécommunications (Paris), Automatic Telephone Switching System, 3 053 935.

J. Bernutz and A. Gunst, Mix & Genest Werke (Stuttgart), Marking Spring Arrangement for Crossbar Switches, 3 059 059.

L. Bloomquist and E. Marsh, ITT Kellogg, Transistorized Generator of Telephone Ringing Current, 3 056 931.

D. J. Booty, ITT Kellogg, Desk Stand for a Combination Telephone Station and Switching Unit, Design 193 894.

C. Buff, Mackay Radio and Telegraph Company, Twinplex Telegraph Transmission, 3 048 658.

C. Buff, American Cable & Radio Corporation, Diplex Telegraph System Using Frequency Modulation, 3 048 657.

C. G. Byrd and W. K. Hagan, Farnsworth Electronics Company, Flip-Flop Varies Frequency of Blocking Oscillator, 3 061 795.

K. W. Cattermole, Standard Telecommunication Laboratories (London), Electric Signalling Circuits, 3 059 060.

B. Dal Bianco and M. Scata, Fabbrica Apparicchiature per Comunicazioni Elettriche Standard (Milan), Electrical Switching Device and Method for Making Same, 3 059 074.

B. Dal Bianco and M. Scata, Fabbrica Apparicchiature per Comunicazioni Elettriche Standard (Milan), Sealed Contact Device, 3 056 869.

## US Patents to ITT; August–October 1962

K. L. DeBrosse and J. M. Lewis, Farnsworth Electronics Company, Scanning Apparatus, 3 060 755.

E. C. L. DeFaymoreau, Federal Telecommunication Laboratories, Distance Measuring System Providing Interrogation Rate Control, 3 054 103.

E. C. L. DeFaymoreau and J. A. Baudin, ITT Federal Laboratories, Delay Compensation in Telemetering Systems, 3 051 948.

R. H. Duncan, T. P. Miller, A. V. Maciulis and S. L. Fudaley, ITT Kellogg, Conference Call Circuit, 3 060 265.

E. H. Eberhardt, ITT Federal Laboratories, Heat Imaging Device, 3 054 917.

J. S. Engel, Federal Telephone and Radio Company, Moving Target Selector, 3 051 947.

J. S. Engel, H. M. Bellis, and E. J. Huber, ITT Federal Laboratories, Wide Band Omnidirectional Beacon Antenna, 3 054 107.

R. C. Ferrar and H. A. French, Federal Telecommunication Laboratories, Communication System, 3 048 840.

F. Fichtmuller, Mix & Genest Werke (Stuttgart), Article Sorting Arrangement, 3 059 770.

W. R. Frame and D. K. Coles, ITT Laboratories, Joule-Thomson Effect Gas Liquefier, 3 048 021.

W. J. Frankton, W. N. Roseway, G. Atkinson, and W. J. Archibald, Standard Telephones and Cables (London), Carrier Current Communication Systems Incorporating Repeaters, 3 059 068.

K. N. Fromm, ITT Federal Laboratories, Radio Direction Finding System, 3 050 729.

K. N. Fromm, Farnsworth Electronics Company, Radio Direction Finding System, 3 061 831.

J. C. Gibson, ITT Kellogg, Translator-Supervisory Apparatus for Telephone System, 3 059 067.

A. Girinsky and P. C. A. Bacot, Compagnie Générale de Constructions Téléphoniques (Paris), Frequency Modulation System, 3 051 786.

F. P. Gohorel, Compagnie Générale de Constructions Téléphoniques (Paris), Telephone Systems, 3 049 594.

F. P. Gohorel, Compagnie Générale de Constructions Téléphoniques (Paris), Telephone Systems, 3 059 055.

T. J. Goldan, Federal Telecommunication Laboratories, Precision Approach Radar, 3 060 423.

V. J. Habisohn, ITT Kellogg, Signal Control System, 3 060 331.

L. B. Haigh, M. Bisk, and T. W. Tuttle, Federal Telecommunication Laboratories, PABX Switching System, 3 048 662.

L. B. Haigh, H. F. Herbig, and R. W. Blanchard, ITT Federal Laboratories, Answering Signal Detector for a Private Automatic Branch Exchange, 3 051 790.

J. W. Halina, ITT Federal Laboratories, Gain Regulation Circuit, 3 060 383.

R. Harding, Jr., International Telephone and Telegraph Corporation, High Speed Recording Apparatus, 3 051 785.

H. S. Haynes, Federal Telecommunication Laboratories, Pulse Waveform Synthesizer Using Plurality of Individually Charged Storage Means Sequentially Discharged Through Common Load, 3 051 906.

S. L. Hjertstrand, Standard Radio & Telefon (Bromma), Signalling Arrangement for Automatic Telephone Systems, 3 055 986.

K. G. Hoer, Creed and Company (Croyden), Facsimile Apparatus, 3 060 262.

O. H. A. Holstein and E. Gross, C. Lorenz (Stuttgart), Control Attachment for Electromechanically Operated Communication Apparatus Used in Pulse-Code Systems, 3 048 656.

R. W. Hutton, ITT Kellogg, Conversation-Timing Reverting-Call Telephone System, 3 056 859.

A. Kastler and M. Arditi, ITT Federal Laboratories, Frequency Selection System Utilizing a Plurality of Transitions, 3 054 069.

H. E. Krefft, Kuthe Laboratories, Anode Structure, 3 048 727.

L. Lamin and V. J. Habisohn, ITT Kellogg, Electronic Line Circuit for Code Signalling System, 3 056 856.

L. Lamin and M. Ribner, ITT Kellogg, Electrical Identification Circuit, 3 060 271.

L. Lamin and J. Karklys, ITT Kellogg, Automatic Teletypewriter Transmitter, 3 059 048.

E. C. Lee, Standard Telephones and Cables (London), Electrostatic Shield for High Voltage Cables, 3 051 771.

P. W. Lennox, Standard Telephones and Cables (London), Intelligence Storage Equipment, 3 051 931.

C. Lucanera, Federal Telecommunication Laboratories, Omnidirectional Beacon Antenna, 3 051 951.

M. MacPhee and D. E. Bolger, Standard Telecommunication Laboratories (London), 3 051 557.

T. J. Marchese, ITT Federal Laboratories, Pulsed Beam Tube, 3 051 865.

C. F. Mason, Farnsworth Electronics Company, Matrix Line Selector, 3 054 985.

F. A. Muller and D. K. Keel, Federal Telecommunication Laboratories, Method of Preparing Aminotriazines Useful for Tissue-Equivalent Plastic Composition, 3 055 898.

F. A. Muller and D. K. Keel, ITT Federal Laboratories, Tissue-Equivalent Plastic Composition, 3 055 843.

## US Patents to ITT; August–October 1962

R. K. Orthuber, ITT Federal Laboratories, Solid-State Radiation Amplifier, 3 054 900.

C. N. Pagano, E. H. P. Bigo, and R. C. P. Hinton, ITT Federal Laboratories, Electronic Relay Responsive to Instantaneous Polarity of Control Signal Whether Alternating Current or Direct Current, 3 054 003.

E. G. Parker, M. A. Karpeles, and R. W. Craine, ITT Laboratories, Instrument Landing System, 3 048 842.

E. G. Parker and G. C. Lehman, ITT Federal Laboratories, Directional Beacon Antenna, 3 054 106.

J. A. Peeters, Bell Telephone Manufacturing Company (Antwerp), Mechanism for the Automatic Processing of Documents, 3 053 530.

R. E. Poole, ITT Federal Laboratories, Variable Pulse Width Generating System, 3 048 714.

M. Press and R. C. Davis, ITT Federal Laboratories, Radar Tracking System, 3 054 102.

N. Raver, ITT Laboratories, Pulse Timing Control Circuit, 3 048 708.

E. Richert and H. Fliegner, Mix & Genest Werke (Stuttgart), Stacking Device, 3 051 332.

E. Richert and H. Fliegner, Informatik Division of Standard Elektrik Lorenz (Stuttgart), Storage Apparatus for Article Sorting System, 3 051 333.

A. Rothbart, ITT Federal Laboratories, Electromechanical Delay Device, 3 051 916.

H. W. G. Salinger, ITT Federal Laboratories, Magnetic Focusing Device, 3 050 653.

R. O. Schildknecht, ITT Laboratories, Multiplex Communication Receiver, 3 048 661.

H. A. Snow and H. W. Kettmann, ITT Laboratories, Broadband Diode FM Modulator, 3 048 796.

B. R. Stack, ITT Federal Laboratories, Carrier Signalling System, 3 060 269.

S. B. Silverschotz and S. Schreiber, ITT Federal Laboratories, Electric Motor, 3 054 011.

K. Steinbuch, R. Braun, and G. Merz, Mix & Genest Werke (Stuttgart), Telephone Line Supervisory System, 3 053 936.

F. Steiner and E. Kramar, C. Lorenz (Stuttgart), Radio Direction Finding System, 3 054 105.

F. Steiner, C. Lorenz (Stuttgart), Circular Antenna Array System Scanning Switch, 3 048 843.

J. F. Sullivan, ITT Federal Laboratories, Pulse Pair Decoder, 3 051 928.

E. Touraton and J. P. LeCorre, Laboratoire Central de Télécommunications (Paris), Switching Systems Between Multiplex Communication Channels, 3 049 593.

E. G. Trost, Schaub Apparatebau (Pforzheim), Video-Frequency Contrast-Control Arrangement for Television Receivers, 3 051 780.

F. Ulrich, Mix & Genest Werke (Stuttgart), Electronic Switching Unit for the Construction of Information Storage Devices, Counters and the Like, 3 056 045.

H. K. M. Verhoeven, Bell Telephone Manufacturing Company (Antwerp), Opening Device for Document Carriers, 3 050 917.

M. Wagner and M. Kolb, Suddeutsche Apparatefabrik (Nürnberg), Electrical Condenser, 3 054 029.

T. B. Warren and J. B. Harvey, ITT Laboratories, Signal Receiving System, 3 048 783.

J. M. Wilson, Standard Telephones and Cables (London), Methods of Producing Silicon of High Purity, 3 053 631.

S. J. Worley, ITT Federal Laboratories, Radio Direction Finding System, 3 050 728.

E. P. G. Wright, D. S. Ridler, and R. Grimmond, Standard Telecommunication Laboratories (London), Intelligence Storage Equipment with Independent Recording and Reading Facilities, 3 048 827.

### Intelligence Storage Equipment With Independent Recording and Reading Facilities

3 048 827

E. P. G. Wright, D. S. Ridler, and R. Grimmond

This is a storage arrangement that receives and stores a block of data consisting of a number of words, each of which is received in binary code in parallel fashion over a number of channels equal to the number of elements in the word. This block of data is stored and then read out either immediately or subsequently, at a predetermined rate, which might be different from the rate at which it is stored. The recording and reading means are entirely separate so that both processes can be in progress concurrently.

### Circular Antenna Array System Scanning Switch

3 048 843

F. Steiner

A rotating-pattern Doppler-type circular antenna array is provided with a scanning switch to selectively actuate the antennas of the array to produce a rapid oscillation of the pattern that progresses at a slow rotation rate around the circular array. This is accomplished by a variation in the spacing of stator and rotor coupling elements of the scanner. A vernierlike incidence of coupling is provided.

### Telephone Systems

3 049 594

F. P. Gohorel

This switching circuit arrangement, designed for the Pentaconta system, allows a considerable simplification in the design of the registers, multiplying circuits, and supervisory pulse senders. The system features common and individual devices for controlling selection and certain auxiliary functions controlled by trunk bundles designed to associate the devices temporarily. The number of these trunk bundles is relatively small because of their short holding time.

### Electrical Condenser

3 054 029

M. Wagner and M. Kolb

This is a capacitor having an anode of a base material such as aluminum or tantalum with a coating of aluminum or tantalum oxide and a plurality of layers between the coating and the cathode. The layers are of semiconductor material of the same conductivity type to avoid formation of barrier layers. At least two of the layers have temperature coefficients of conduction of opposite sign.

### Intelligence Storage Equipment

3 051 931

P. W. Lennox

A magnetic storage unit for building up a matrix of rows and columns, with a control access magnetic core, is the subject of this patent. The unit provides a series of magnetic cores, which form a matrix row, around the periphery of a somewhat larger core that serves as a switch to control access to the individual components of the storage array.

### Sealed Contact Device

3 056 869

B. Dal Bianco and M. Scata

This is a sealed relay switch, operable by magnetic effects. It consists of two conductors of magnetic material and a magnetic conductive element to bridge between the two conductors. The bridging element is pivoted on one of the conductors, and in the absence of an operating magnetic force opens the switch to an extent determined by the enclosing envelope.