MR. T. H. WALKER

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Welding the connecting wires to *1A* thermistors.

Thermistor Production*

By W. T. GIBSON

Standard Telephones and Cables, Limited; London, England

TO SATISFY certain circuit requirements it may be convenient to make use of a device whose resistance rapidly falls to a low value as its temperature increases. Such a device, known as a thermistor, can be made from semi-conducting materials, and this article gives a brief account of its production on a commercial scale. Two important circuit applications are referred to, one using a bead-type thermistor and the other a rod-type.

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Material used in electrical circuits can be divided into three classes: conductors, semi-conductors, and insulators. While the first and third groups have been used widely in the whole of electrical history, only a limited use of the second group has been made until recently, though the rate of application is increasing rapidly.

A thermistor is a device made from one form of semi-conductor whose specially interesting characteristic is that its rate of change of resistance with temperature is much greater than that of conductors, and so great that many useful applications can be made of it. While the resistance of a conductor increases with rising temperature, generally proportionally to the absolute temperature, the resistance of a semi-conductor decreases with rising temperature according to a relationship:—

$$R_T = R_0\, e^{b/T},$$

where

R_T = the resistance at temperature T
b = a constant
e = the base of Napierian logarithms.

While many semi-conductors may show very interesting and useful properties by having non-ohmic resistances, the thermistors that have been developed are essentially ohmic, that is E/I (where E is the applied voltage and I the current) is constant for all values of E at constant temperature.

* Reprinted from *The Post Office Electrical Engineers' Journal,* Part 1, volume 46, pages 34–36; April, 1953. Figures 5 and 7 did not appear with the original paper.

1. Thermistor Material

The development of thermistors within the laboratories of Standard Telephones and Cables, Limited, has been carried out continuously over a period of nearly 20 years, and many materials have been investigated and tried out in fairly large-scale tests during that period. Today all the various types of material except one have been discarded for one or another reason, and large-scale production of a wide variety of thermistors is now carried on with this successful material. The basic material is a mixture of metallic oxides that is fired in air to a very high temperature to give it the desired characteristics. The metals used are ones having several valencies, such as nickel, and by the use of appropriate binary or ternary mixtures, and by variations of firing technique, a range of final materials can be produced, each of whose resistivity varies over a ratio of the order of 10 000 within practicable operating limits of temperature. More important is the fact that any desired result can be regularly and reliably achieved with a reasonable and practical amount of technical control.

These materials have very useful characteristics. For those with the higher values of resistivity the temperature coefficient of resistance at room temperature is of the order of 4 per cent per degree centigrade, this coefficient being smaller for the lower-resistivity materials. They are free from all rectification effects in alternating-current circuits and are unchanged by heating in air or oxidising atmospheres, but may be adversely affected by heating in reducing atmospheres. The materials are also non-magnetic and their alternating-current and direct-current resistances are alike up to very high frequencies. Most types of thermistors show a good degree of stability of resistance with time.

2. Types of Thermistor

There are three main types of thermistor in regular production: bead types, pellet types, and

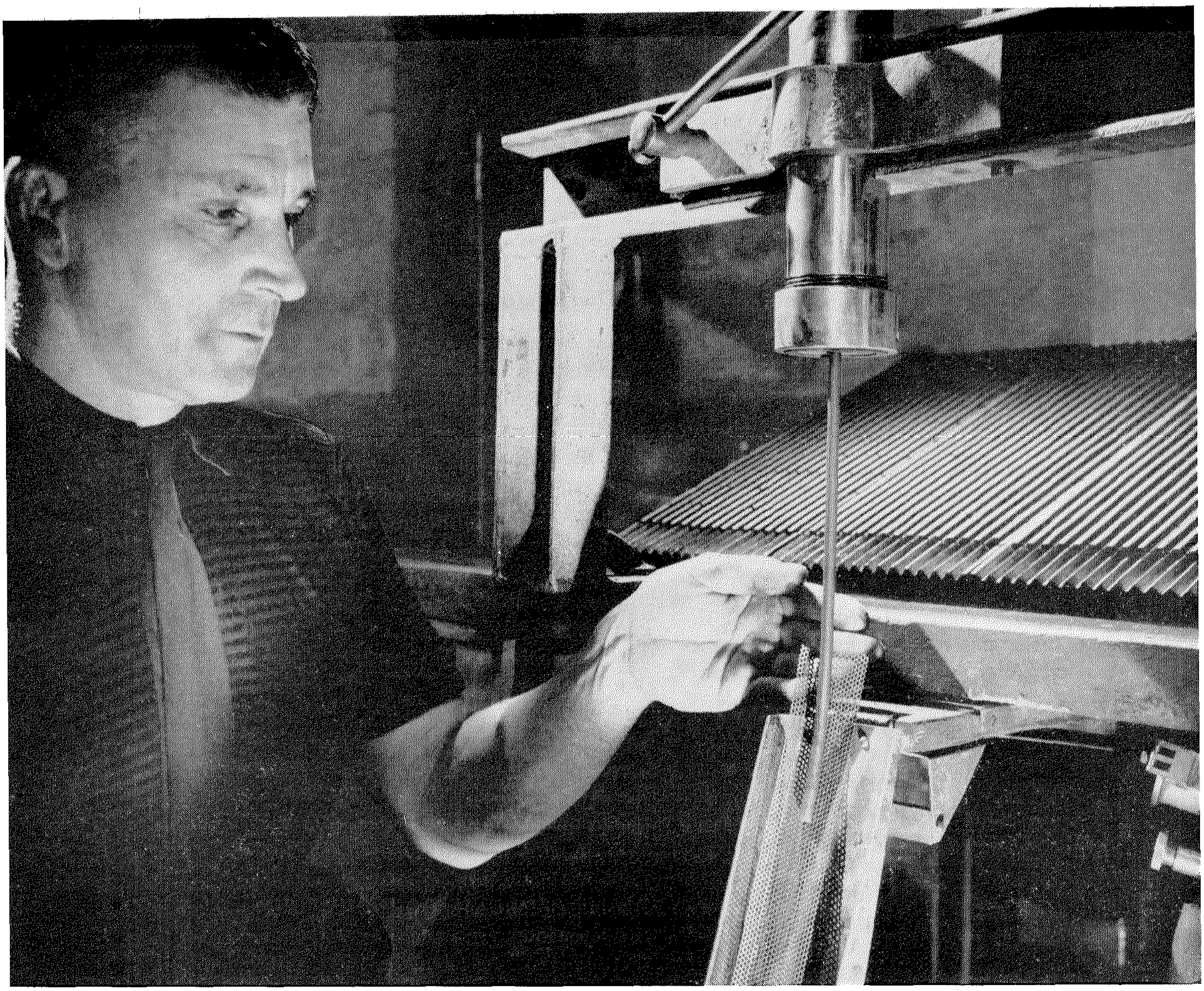

Figure 1—(Above) material for Brimistors being extruded vertically.

Figure 2—On the opposite page, trays of Brimistor material after curing.

rod types. Only the methods of producing the first and last will be shortly described here.

The bead types are many in number. All essentially consist of a minute bead of thermistor material formed on to two closely adjacent platinum wires. They may finally be of directly heated or indirectly heated design, but the main production processes being similar, the directly heated type will be described here.

The pellet types are, as the name describes, pellets of material usually up to about 0·5 inch (1·3 centimetres) in diameter and up to 0·13 inch (0·32 centimetre) thick, silvered on the flat faces for connection easily into circuits, and in some cases soldered on to brass base-plates.

The rod types are circular rods up to 0·25 inch (0·64 centimetre) in diameter and any desired length, usually up to about 1·5 inch (3·8 centimetres). These may be silvered at the ends for insertion into clips, or they may be supplied with wire ends.

3. *Uses of Thermistors*

It is not proposed to describe the very wide field of application of thermistors, but some uses have developed to the extent where very large quantities are used and large-scale manufacture has been introduced. For illustration, therefore, two cases are taken to show the diversity of application.

Firstly, we can consider the type of rod thermistor known as the Brimistor. This has been specially developed to be used in broadcast-receiver sets and television sets, in which valves

are heated in series. As is well known, the ratio of the hot to cold resistance of a valve heater is quite large, and at the instant of application of voltage to a cold heater a heavy current surge passes. This in itself may be of little importance, but, when valves are heated in series, conditions can often arise where some heaters may be temporarily very severely overheated and short life can result. The use of a Brimistor in series, which has initially a resistance high enough to suppress the surge, will enable all the heaters to come to working temperature without any danger of quick-heating valves being overheated. After a few seconds the resistance of the Brimistor falls to a value that is negligible in this circuit. In this case the device has to be suitable to carry the heater current continuously and massive enough to take some seconds for the resistance to fall to the working value.

Secondly, we can look at a very widely different application in the British telephone network, where the British Post Office thermistor type *1A* is employed in shared-line service to suppress bell tinkling. In this application, the thermistor has to remain at a sufficiently high resistance to suppress bell tinkling while dialling impulses are incoming, but fall to such a low resistance as not to interfere with bell ringing within a ringing cycle. This requires that the thermistor shall be very quick heating, so the bead must be very small and the rate of loss of heat by conduction down the leads must also be reduced.

4. Manufacture of Brimistors

The raw materials for the rods are nickel, manganese, and copper oxides, which are prepared in the correct chemical form and in a fine state of division, and these properties are first checked.

The materials are then weighed out in the necessary quantities and mixed with distilled water to a slurry that ensures thorough blending of the components. When mixing is complete, the water is filtered off and the filter cake dried and pulverised, and stored in hoppers.

When required, the dried powder is mixed with suitable binding agents and thoroughly worked

Figure 3—Firing ovens.

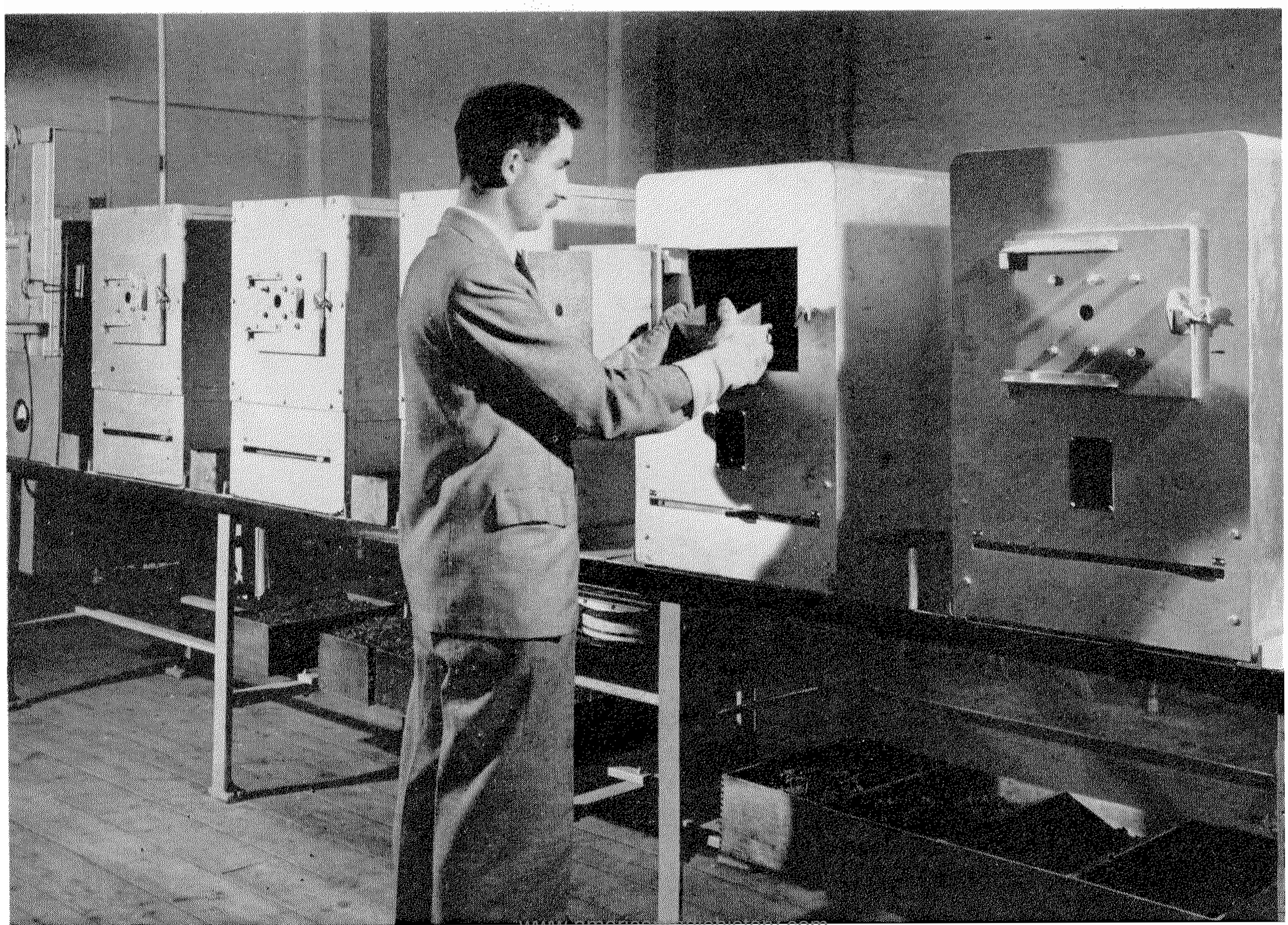

Figure 4—Soldering wires to Brimistors.

to a stiff dough. This is transferred to an extruding machine where it is vacuum de-aired prior to extrusion. Figure 1 shows a length of material being extruded vertically. At this stage the material is soft and putty-like, and is easily cut into short lengths suitable for the finished Brimistor. A few days' curing changes it to a dry material hard enough to withstand subsequent handling. Figure 2 shows trays of cured material.

From this stage the material proceeds to the firing that is carried out in two stages, according to a carefully controlled time and temperature schedule, in the ovens shown in Figure 3. During this process the material shrinks considerably, becomes so hard that a rod can no longer be broken by hand, and acquires the desired electrical properties.

To be able to connect the thermistors into circuits, they are then silvered at each end, the silver being fired on in a conveyor furnace, and wires are soldered to the silvered ends, as shown in Figure 4. Manufacture is completed by mechanical tests for dimensions, strength of end connections, etc., electrical tests of performance and resistance, followed by packing (Figure 5).

5. *Manufacture of 1A Thermistors*

In this thermistor the circuit requirements can only be achieved with a bead, roughly spherical in shape and only a few thousandths of an inch (hundredths of a millimetre) in diameter. Because of the firing conditions, which are in the order of 1200 degrees centigrade in air, the only practical supporting wires are platinum or platinum alloys. The time constants desired in the thermistor, and other characteristics necessary to ensure that the thermistor works properly in the subscriber's set over a very wide range of ambient temperatures, impose further limitations on the particular platinum alloy, its diameter, and length. It is possible to use platinum wires of about 0·002-inch (0·051-millimetre) diameter, or an alloy such as platinum–iridium with a diameter of only 0·001 inch (0·025 millimetre).

Two of these wires about 6 inches (15 centimetres) long are mounted on a frame that spaces them the correct small distance apart, and holds them parallel under light tension.

The process of making the beads is carried out by hand. It is a highly skilled one and is illustrated in Figure 6. The oxides are mixed as fine powders in the correct proportions, together with suitable binders. These are kept mixed in a small agate mortar and the consistency is continually adjusted as necessary to maintain the proper viscosity and surface tension. Minute drops of the mixture are picked up on a glass rod and applied at about 0·75-inch (2-centimetre) spacing to the platinum wires. If the physical properties have been properly adjusted, then the drop can be made to take up an ellipsoidal shape symmetrically placed on the wires and of the correct size. The close requirements on the final article allow very little latitude in the consistency of the material.

After drying, the beads are stiff enough to permit the wires being cut from the sprung frames, and the beads are then fired by pulling them through a vertical tubular furnace with platinum heating elements. When they have been fired, the beads are extremely hard and have shrunk appreciably. The wires are now cut close to one side of each bead so that there are two short wire leads with a bead joining their ends. The ends of the wires are welded to an alloy hairpin under a spot welder, thyratron controlled to give exact electrical conditions, and pressure controlled to give exactly reproducible mechanical conditions. This operation is shown in the frontispiece. The loops are then sealed into small glass tubes and tipped off, Figure 7, and the bend of the hairpin is cut off.

To stabilise the thermistors, they are aged under conditions more stringent than those met with in service, and are then ready for final inspection. This is shown in Figure 8. The inspection process includes mechanical inspection of the bead and welds under a binocular microscope and electrical testing appropriate to its special application.

Throughout all the operations very close attention to cleanliness and technical control of conditions is essential to ensure meeting the requirements. The rate of change of characteristics with chemical condition of the bead, firing schedule, and other variables is so rapid that, without such control, regular production of these thermistors would not be a practical proposition.

6. *Acknowledgments*

The author would like to thank colleagues associated in this work, and especially Mr. H. Wolfson, for their great assistance.

Figure 5—Packaging the finished Brimistors.

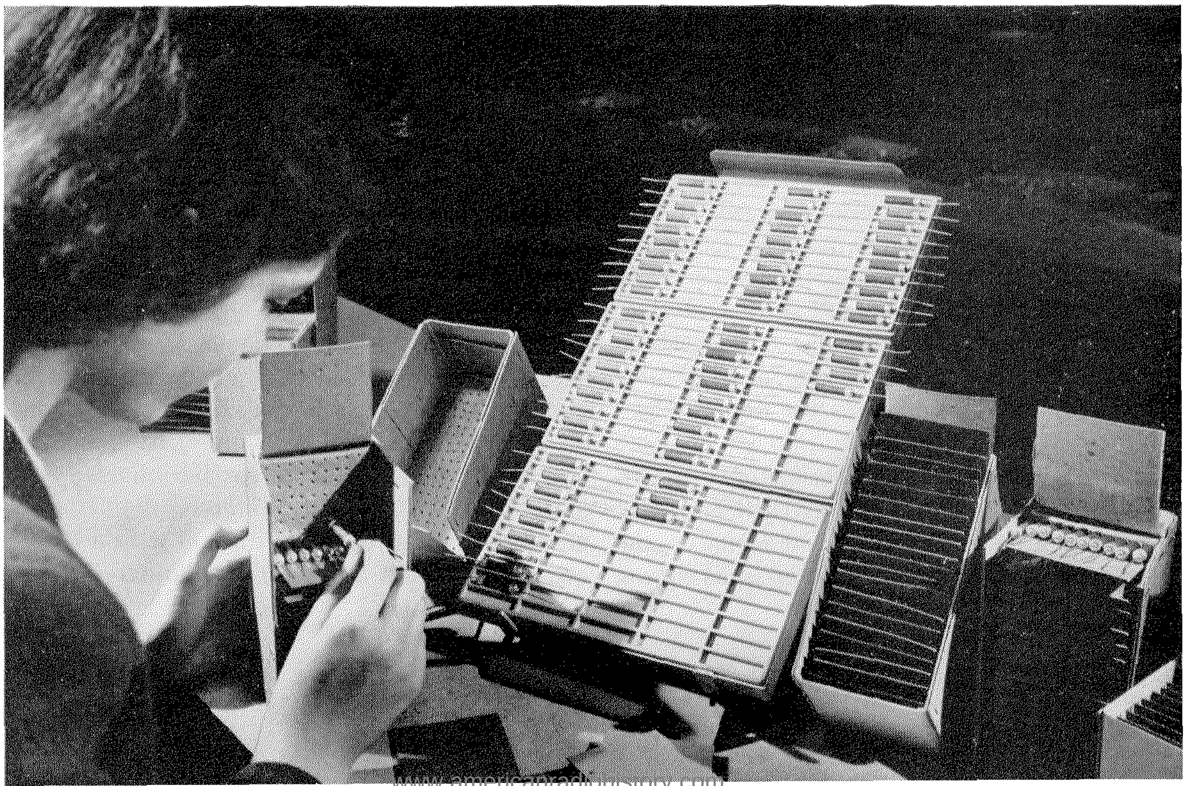

Figure 6—Placing the mixture to form beads for *1A* thermistors.

Figure 7—Above, sealing *1A* thermistors into protective glass tubes.

Figure 8—Below is shown the final inspection of *1A* thermistors.

Automatic-Tuning Communication Transmitter*

By M. C. DETTMAN

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AFTER the experience of the second world war with communication transmitters in the low- and medium-frequency ranges, and similar experience with much-later-design very-high-frequency equipment, it became evident that a modernization program was highly desirable for communication transmitters operating in the medium- and high-frequency bands. Accordingly, a development program was instituted to build a modern communications transmitter that could effect communication under the rigorous conditions encountered in the services, with precision and reliability, without the necessity of preliminary calling, and with absence of interference with other units of the communications system. It was desirable that this transmitter be rather versatile so that it might be used on a variety of surface and undersurface vessels.

A development program was instituted at Federal Telephone and Radio Company in 1947, which was subsequently transferred to Federal Telecommunication Laboratories. This project was to develop a transmitter meeting the needs indicated above. The first development model of this transmitter was presented by the contractor in the latter part of 1948. A development model substantially the same as the final production model was submitted in August, 1950. Extensive laboratory tests were made on this unit. After the laboratory tests were completed, the equipment was installed on vessels and seven months of operational evaluation followed.

1. Specifications

This new transmitter meets the following broad specifications. The transmitter may be automatically tuned and matched to specified antennas at any frequency between 300 kilocycles per second and 26 megacycles per second. The operator may either select any one of 10 preset channels or may manually set up any frequency in the operating range by manipulating a group of knobs that directly indicate the frequency. The remainder of the tuning is accomplished automatically, including the antenna matching. Tune-up time for most frequencies with proper antennas in use is approximately 30 seconds. Maximum time is 90 seconds. In the event that the operator wishes to tune the equipment under manual control, he may do so. The nominal power output of the transmitter is either 100 or 500 watts depending on the type of equipment and the radio frequency in use. Power output is limited to 100 watts nominal over the range of 300 kilocycles to 2 megacycles. From 2 megacycles to 26 megacycles, power output may be either 100 watts or 500 watts as desired by the operator. The transmitter may be keyed at speeds up to 600 words per minute. Either polar or neutral keying signals are acceptable. Radiophone operation using a standard carbon or dynamic microphone is available. Provision is made for facsimile operation with a 600-ohm input circuit and utilizing frequency-shift modulation. Frequency-shift keying with or without phase modulation is also available. All the aforementioned features are integral with the equipment.

In addition to these features, compactness of size and flexibility of installation were required. To this end, the entire equipment is broken down into groups as follows.

A. The transmitter group, which itself is broken down into a basic 100-watt equipment and additional frames and chassis capable of raising the nominal power level to 500 watts. (Figure 1.)

B. The antenna-control group, which may be mounted at any convenient spot but preferably adjacent to the transmitter bay. (Figure 2.)

C. The antenna-tuning unit and capacitor assembly, which is always mounted at the base of the antenna to be used with the transmitter. (Figure 3.)

* Reprinted in part from *Convention Record of the I.R.E. 1953 National Convention*, Part 2—Antennas and Communications, pages 137–144. Information on the operating procedure and on the frequency-determining system, which were formerly classified, have been added and several illustrations appearing in the original paper have been omitted. Presented at the National Convention of the Institute of Radio Engineers in New York, New York, on March 25, 1953.

D. The remote channel selector consisting of a waterproof case containing a telephone dial for selecting preset channels and an indicator showing the channel in use. (Figure 4.)

2. *Basic Units*

These basic units may be assembled into three standard equipments known as *AN/URT-2*, *AN/URT-3*, and *AN/URT-4*.

The *AN/URT-2* consists of the basic 100-watt transmitter bay shown in Figure 5, the antenna-control group, antenna tuner, capacitor assembly, and remote channel selector. The *AN/URT-3* consists of an *AN/URT-2* 100-watt system plus a booster section composed of a high-level modulator and a high-voltage rectifier with associated frames, capable of raising the operating power output to 500 watts, nominal. Figure 1 shows the transmitter bay. The *AN/URT-4* consists of two *AN/URT-2* systems and a booster section. The frames of the transmitter group are bolted together forming an integral assembly as shown in Figure 6. The *AN/URT-4* makes available two complete systems capable of simultaneous operation on two channels at 100-watt level. At the option of the operator, power may be boosted to 500 watts on either one of these two channels provided the operating frequency is above 2 megacycles. Automatic interlocking is incorporated to ensure that power is limited to the 100-watt level below 2 megacycles even though the booster is selected for 500-watt operation.

Now that we have viewed and discussed the over-all equipment, let us see what the individual chassis look like. Starting at the top of the main transmitter frame, is the radio-frequency-amplifier unit (Figure 7). This unit is basically designed for 500-watt operation. The vacuum-tube lineup includes a *6AG7* feeding a *5933* (*807W*). These two stages are gang-tuned, driven by a tuning motor that in turn is controlled by a servo system when automatic tuning is used. The power-amplifier tube is a *4-400A* whose output circuit is designed to feed a 50-ohm line. This power amplifier has its own tuning motor and servo for automatic tuning. Included on the chassis are the servo-system tubes and keying tubes. Most of the relays and control circuitry used for automatic tuning sequencing are also located on this chassis.

The next drawer down is the low-level modulator and is shown in Figure 8. This unit includes the audio-amplifier tubes, squelch circuits, clipper tubes, and automatic-volume-control circuits that are used for radiophone operation. The audio circuits have sufficient power output to modulate the transmitter in 100-watt operation. A group

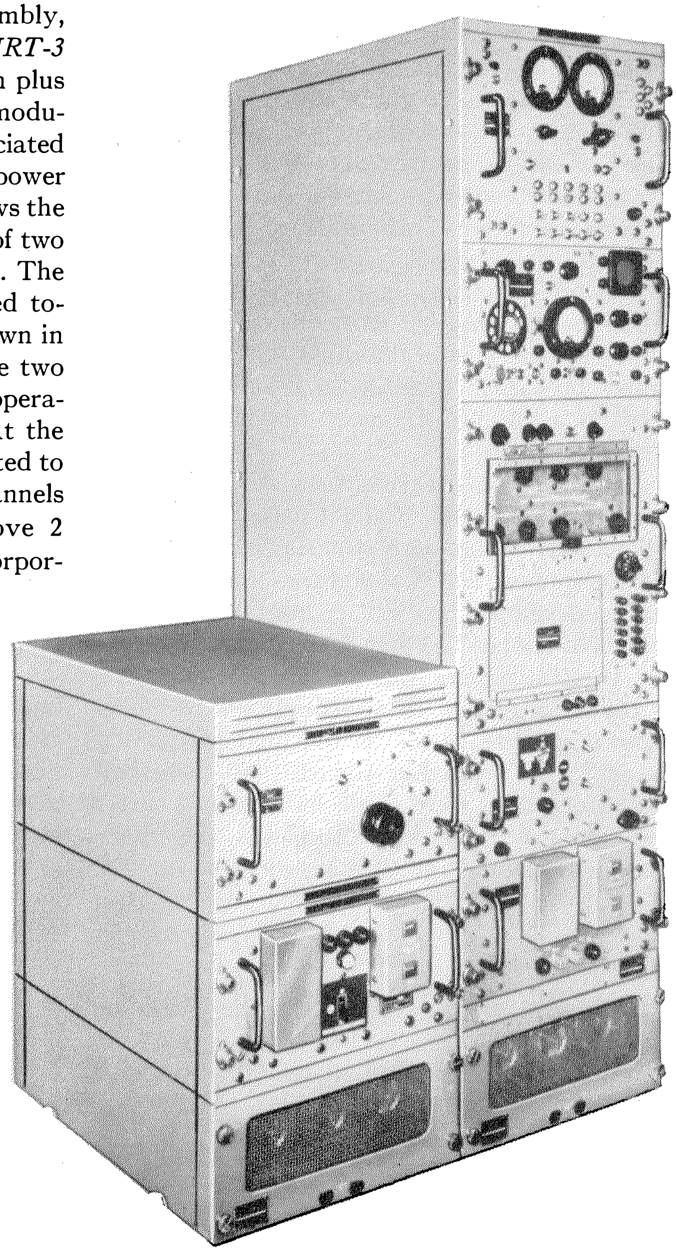


Figure 1—*AN/URT-3* transmitter bay.

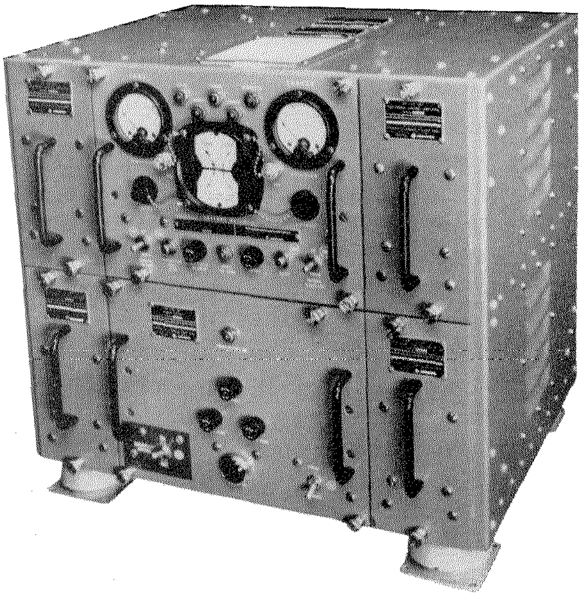


Figure 2—Antenna control group.

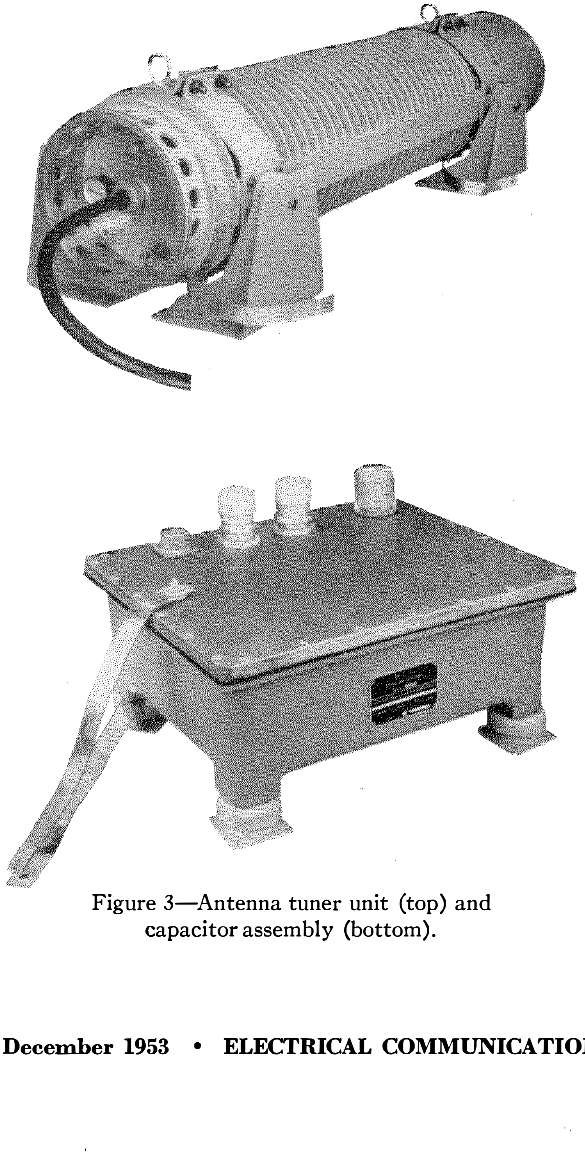


Figure 3—Antenna tuner unit (top) and capacitor assembly (bottom).

of tubes are utilized to take either hand keying or teleprinter keying voltages and transform them into suitable signals for either *A1* emission or frequency-shift operation.

Another section of this chassis contains the equivalent of a standard two-inch test oscilloscope with circuit-sampling switches for checking the operation of various sections of the equipment. Additionally, two power supplies are included on the chassis. One is a regulated 250-volt power supply and the other, a 12-volt direct-current supply for standard radiophone control-unit remote-box operation. A large number of the operating controls are centralized on the front panel of this unit.

Figure 9 shows the radio-frequency oscillator. Its function is to provide the fundamental radio frequency required at a 2-volt signal level sufficient to drive the radio-frequency amplifier unit. It must do this with a high degree of frequency precision, in the order of plus or minus 1.5 parts per million +20 cycles under all operating conditions. Included on the front panel are the preset switch banks, which allow presetting to any desired 10 channels. Selection of any of these channels is by means of a telephone dial and an associated channel-indicator meter over a two-wire line. A direct-reading dial indicating amount of frequency shift is on the front panel. This is used on frequency shift and facsimile to set the amount of deviation.

The chassis below this is the low-voltage power supply and includes 24-

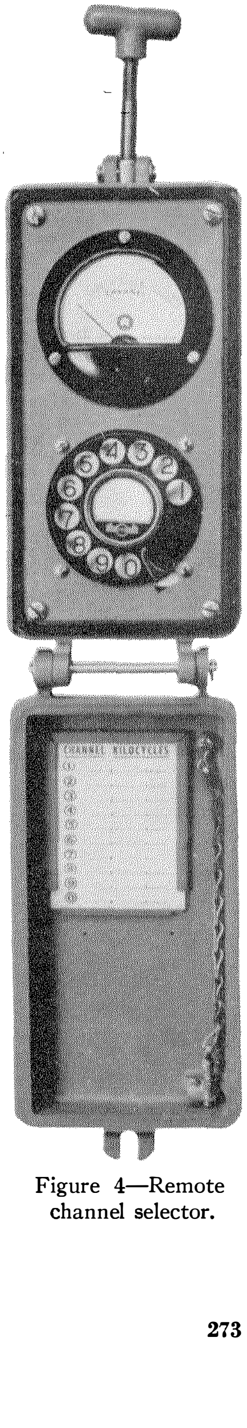


Figure 4—Remote channel selector.

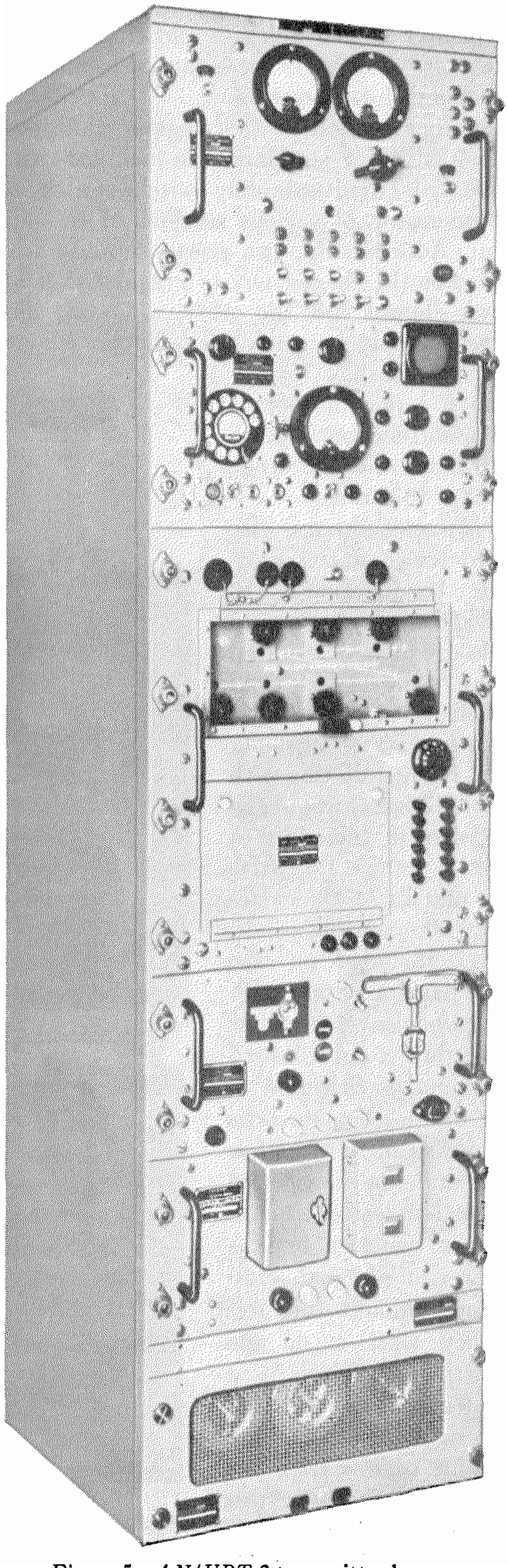


Figure 5—*AN/URT-2* transmitter bay.

volt direct-current supplies for control-circuit relays and tuning-motor operation, bias, 250-volt, and 300-volt supplies.

The bottom chassis in the basic 100-watt frame is the medium-voltage power supply. It supplies 500 volts for the modulator tubes and radio-frequency driver tubes and either 1100 volts or 1300 volts as required for the radio-frequency-amplifier tube for phone or continous-wave operation.

The five individual chassis just discussed are mounted in the 100-watt frame. This frame contains all interconnecting wires and necessary connectors for plug-in operation of the chassis. Terminal boards are provided at the bottom for tieing in the external cables to make an installation. For the *AN/URT-2* 100-watt installation, this frame is mounted on a standard base mount. This mount contains the main blowers that supply the necessary filtered air for cooling the units of the 100-watt frame. Four shock mounts at the bottom of the base mount support the frame. Two additional mounts at the top of the 100-watt frame prevent excessive sway.

To provide operation at a 500-watt power level at frequencies above 2 megacycles, the plate voltage of the final stage of the radio-frequency amplifier is supplied from a three-phase rectifier. The output of this unit is 2400 volts direct current for phone operation and 3000 volts direct current for other types of service. In order to plate modulate the final radio-frequency amplifier at this level, a high-level modulator is used. A pair of *4-125*'s are used as modulators. Part of the high-voltage-rectifier filter is in this chassis. The power selector switch for 100- or 500-watt operation is on the front panel. The high-voltage rectifier and the high-level modulator are located in two frames, making up the booster section. These may be mounted one above the other on top of the standard base mount, and when used in conjunction with a single 100-watt frame the *AN/URT-3* is formed.

The radio-frequency power developed in the transmitter passes via a 50-ohm coaxial cable to the antenna-tuning unit, mounted at the base of the antenna. An exploded view of this unit is shown in Figure 10. The function of this unit is to match the antenna automatically to the 50-ohm transmission line with a normal standing-wave-ratio performance of better than 2:1 and a

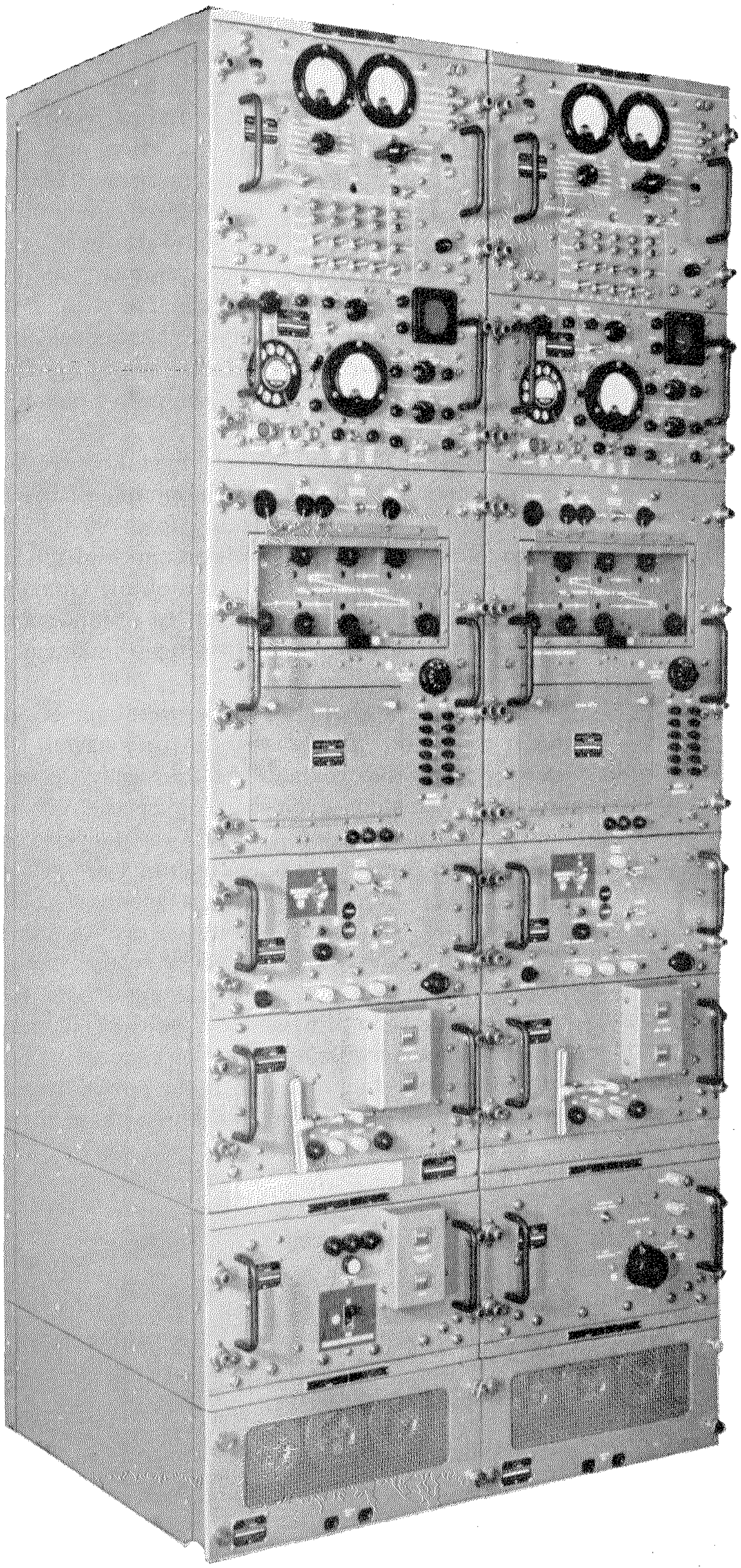


Figure 6—*AN/URT-4* transmitter bay.

specification limit of 4:1. The antenna tuner is effectively a coaxial stub with a lumped center conductor. Its electrical length is about 18 degrees at 300 kilocycles. Two special broadband iron-core transformers are used in this unit; one steps up the impedance of the 50-ohm circuit to 160 ohms and the other is used in the sensing bridges. These bridges supply information to the antenna-control group for automatically tuning the antenna to the frequency in use.

To tune some specified antennas over the wide frequency range, an additional unit, the capacitor assembly, is utilized. This assembly varies the electrical length of the antenna as presented to the antenna tuner when the tuner fails to resonate the load with the capacitor assembly out of the circuit.

The operation of the antenna-tuning unit and the capacitor assembly is determined by the antenna-control group. Automatic-tuning-sequence information from the transmitter bay, capacitor-assembly step-position information, sensing-bridge, and tuning-element-position information from the antenna-tuning unit are collected in the antenna-control group. The input information thus collected is used to operate two 2-phase motors in the antenna-tuning unit in such a manner as to accomplish the necessary antenna match. In the event that a full scan of the tuner fails to resonate the antenna, the antenna-control group changes the impedance of the antenna by controlling the capacitor-assembly step position. The antenna-tuning unit then cycles again. When the antenna circuit is resonated and matched, information is sent to the transmitter bay that proper tuning has been accomplished and communications may proceed. In the event that the antenna can not be resonated, the antenna-control group sends this information to the transmitter bay and the operator may then take appropriate action.

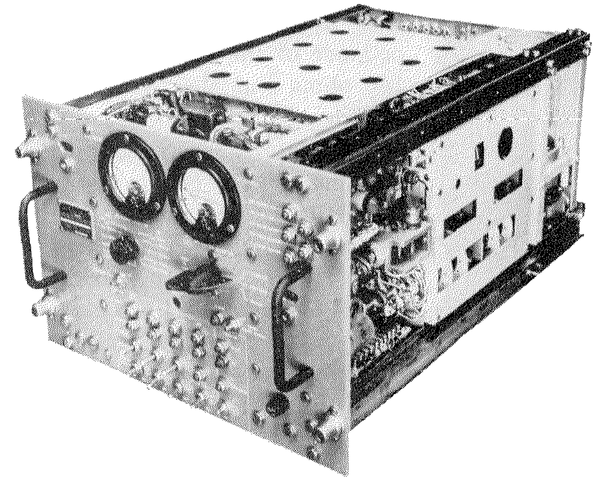


Figure 7—Radio-frequency amplifier.

The antenna-control group is composed of a control and indicator unit. Two meters indicate the positions of the tuning elements. A special dual-movement meter indicates the output of the sensing bridges during the tuning process. When tuning is completed the meter indicates standing-wave ratio on the 50-ohm line. A power supply for operating control relays is included on this chassis. There are two identical preamplifier units. After the sensing-circuit bridge information has passed through the control and indicator unit, it goes to the proper preamplifier unit, where it passes through a chopper converting it to a 60-cycle-per-second signal and is amplified. These signals are then passed to the proper power amplifier, where the signal is further amplified to a level sufficient to drive the appropriate servo motor in the antenna-tuning unit. One amplifier chain controls the tuning servo motor and the other controls the coupling servo motor of the antenna-tuning unit. A dual regulated 300-volt power-supply chassis feeds the vacuum tubes of the antenna-control group.

3. Operation

Actual operation of the *AN/URT-3* is accomplished in the following manner. Normally, the equipment is set for fully automatic operation. The type of emission is determined by the setting of a rotary switch on the front panel of the low-level modulator. For the purposes of illustration, radiotelephone operation will be assumed. After checking all emergency power switches and selecting the desired maximum operating power, say the 500-watt level, the operator will then turn the equipment ON by pressing the main power switch. He then dials one digit on the telephone-type channel selector to choose one of 10 available operating frequencies previously set up in the preset panel of the radio-frequency oscillator.

Starting with a 100-kilocycle crystal-controlled source and a high-stability inductance–capacitance-tuned oscillator covering from 90 to 100 kilocycles, by frequency division and multiplication there may be obtained suitable frequency components that can be mixed to produce any desired transmission frequency to within 5 cycles of the desired value.

Setting up a channel on the preset panel entails no more effort than adjusting seven 10-position switches to read the frequency desired to the nearest 10-cycle-unit figure, i.e., if an output frequency of 1.230718 megacycles is desired, the switches for one channel are set to read 0, 1, 2, 3, 0, 7, and 2, respectively. On dialing the channel, a series of events takes place that culminate in a "carrier ready" signal some 30 seconds later indicating that the operator, or remote operator if so desired, may start voice transmission.

The action of closing the main power switch operates the single-phase and three-phase main

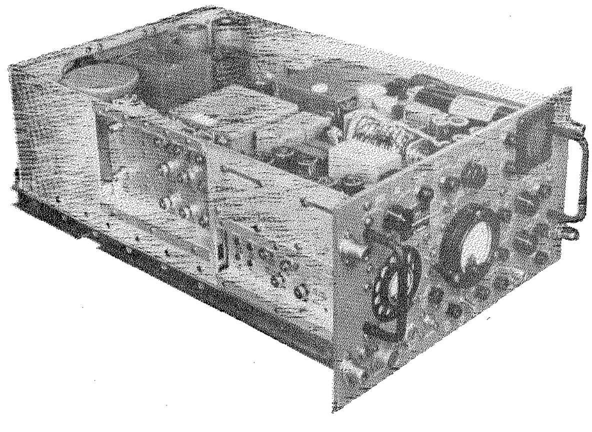


Figure 8—Low-level modulator.

power contactors, which in turn energize all power supplies and power transformers except the medium- and the high-voltage plate transformers. These are under the control of suitable time-delay relays, which also delay one of the 24-volt control circuits. This insures that the

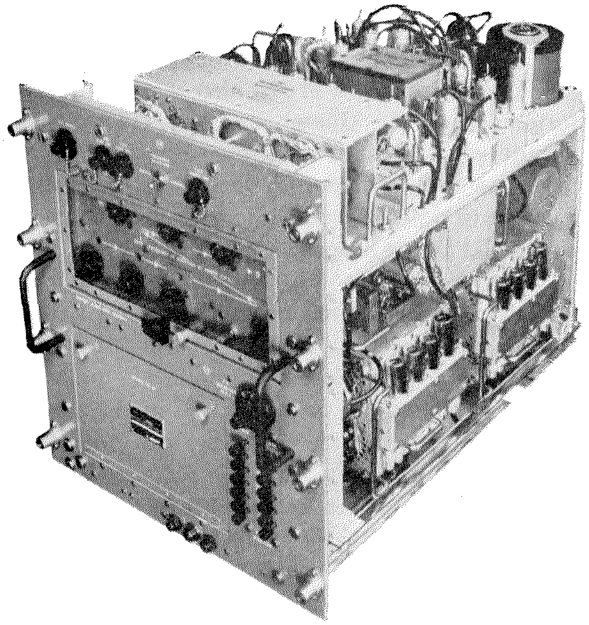
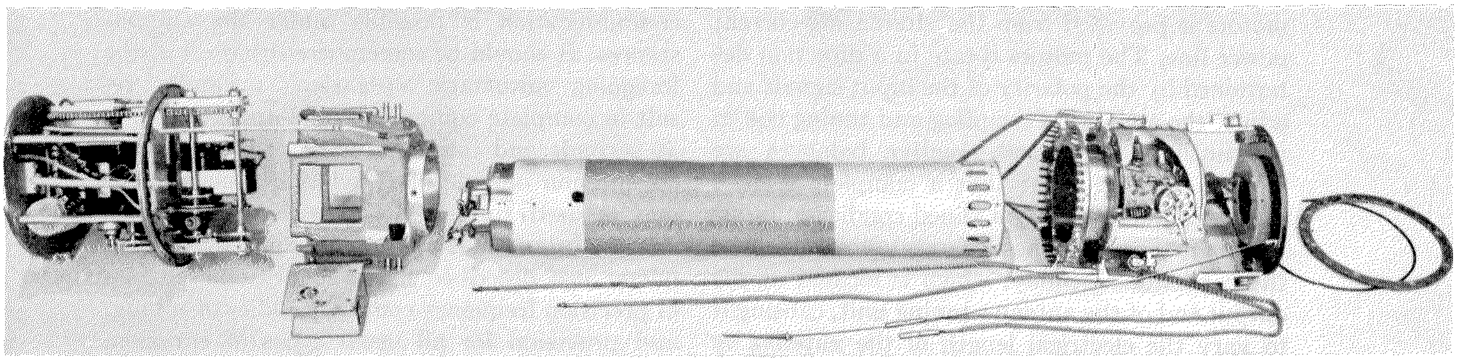


Figure 9—Radio-frequency oscillator.

intermediate- and power-amplifier circuits of the transmitter will not try to tune before plate and screen voltages are available and, incidentally, allows other supply voltages to stabilize. On dialing the desired channel, impulses are sent via repeater relays to the radio-frequency amplifier and the antenna-control group, causing the tuned circuits to go to a "home" position in preparation for the tuning cycle. Additionally, the repeater impulses actuate a 10-position stepping switch located in the radio-frequency oscillator. This switch in turn picks out the desired bank of seven preset 10-position switches that determine the settings of various continuously rotatable switches in the radio-frequency oscillator. The rotary switches, powered by "Ledex" type stepping motors, are operated to the correct positions required to connect in the necessary frequency multipliers, dividers, mixers, and amplifiers from which the output frequency of the radio-frequency oscillator is synthesized. On the completion of frequency selection, a control signal directs the radio-frequency amplifier to proceed to tune up. Correct bandswitch-position information is also supplied to the amplifier at this time.

The radio-frequency amplifier now rotates the bandswitch to select the correct one of 6 available bands and gives information to the power supply and modulator control circuits indicating the maximum permissible operating power level, i.e., 100 watts or 500 watts. The control circuits now wait for information that the time delays have run out and plate and screen voltages are available for tuning up the radio-frequency circuits. On receipt of this information, the power supplies are turned on and the intermediate-power-amplifier tuning motor starts to scan from the high-frequency to the low-frequency end of the band. A diode across the intermediate-power-amplifier output tank circuit rectifies the radio-frequency voltage appearing there and charges a "memory" capacitor to the peak value of the voltage as the circuit passes through resonance. At the completion of this memory scan, the amplifier is again tuned through the band, starting from the low-frequency end. A second diode now rectifies the

Figure 10—Antenna tuner unit (exploded view).



radio-frequency voltage appearing across the output tank circuit and this voltage is compared with that stored on the memory capacitor. When these voltages are nearly equal, a thyratron tube is triggered, operating a relay that stops the motor and sends a signal to the following power amplifier instructing it to tune. The power amplifier proceeds to tune at a 100-watt level into a 50-ohm resistive dummy load in a manner analagous to that of the intermediate power amplifier. After the power amplifier is tuned, a signal is sent to the antenna-control group.

If the antenna-control group is ready to tune, it utilizes the control-circuit information from the power amplifier, requests radio-frequency power in the 50-ohm line at the 100-watt level, and proceeds to tune. The radio-frequency power flows from the 50-ohm line to the antenna-tuning unit. Another wide-band autotransformer located in the antenna-tuning unit converts from 50 to 160 ohms to supply power to an adjustable coupling coil, which transfers power to the antenna circuit. A low-impedance wide-band transformer samples the current in the 160-ohm circuit to operate a sensing bridge that controls the tuning motor.

The circuity of the sensing bridges is arranged to provide a direct-current output. The resistance-bridge section of the sensing circuit is adjusted to give a resistive balance of zero output for 160 ohms. The reactance-bridge section is adjusted to give zero direct-current output for zero reactance. The two outputs thus indicate that the antenna is tuned and matched within the over-all limits of accuracy of the tuning system. The outputs of these bridges are supplied to the antenna-control group, where they are amplified to a level sufficient to drive one phase of their respective 2-phase servo motors located in the antenna-tuning unit. The other phase of these motors is provided from the alternating-current power line. The motors rotate in a direction determined by the polarity of the input signals and adjust the amount of coupling and tuning coil in use until resistive and reactive balances are reached. In the event that a complete scan is made without reaching a tuned condition, information is sent from the antenna-control group to the capacitor assembly, which is located at the output end of the antenna-tuning unit, causing it to vary the electrical length of the antenna by switching in a series or shunt capacitor in a predetermined order. The antenna-tuning scan is then repeated. On reaching a tuned condition, the antenna-control group switches circuits to utilize the output of a standing-wave-ratio monitor located in the antenna-tuning unit so as to monitor the standing-wave ratio continuously. If the ratio exceeds 4:1, an alarm signal is sent to the transmitter bay warning the operator. The antenna-control group also sends tuned information to the transmitter bay. The control circuits then actuate the proper relays that connect the required modulator, apply the necessary plate voltages, and inform the operator by signal light that he may start to transmit.

For the example originally chosen, 500-watt radiotelephone operation, this last tuned information would set up the low-level modulator as a driver for the high-level modulator, make the necessary change in gain required in the speech-amplifier circuits, connect in the high-level modulator, turn off the 1100/1300-volt plate supply, turn on the 2400/3000-volt supply to deliver 2400 volts for phone operation (3000 volts are used for all others), and also notify the operator to proceed to communicate. In the event that the output frequency was below 2 megacycles, the circuits would be automatically set up to connect the low-level modulator (as a modulator) directly to the final radio-frequency-amplifier plate circuit, the high-level modulator would be disconnected, the 1100/1300-volt supply connected to deliver 1100 volts for phone operation, the 2400/3000-volt supply disconnected, and the operator informed that he may proceed to communicate. In the event that the antenna cannot tune, a signal is sent to the operator at the transmitter bay and he may then investigate the cause or elect manually to force the transmitter on the air to effect whatever communication is possible under the circumstances. It should be remembered that all of the foregoing automatic operations normally result in complete equipment tune-up in less than 30 seconds and for the worst condition, where an antenna outside of the design specifications may be used, tune-up occurs within 90 seconds.

The automatic-tuning communications transmitter just described has met the requirements of precision frequency control, rapidity of tuning, and provision for all usual types of emission.

Automatic Control System with Provision for Scanning and Memory*

By N. H. YOUNG

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THIS PAPER describes a system for adjusting a control element to produce a maximum or minimum value of a resulting parameter. The control element is scanned through its entire range of adjustment and returned to the position resulting in the desired maximum or minimum effect. In a system where several maxima or minima are present, the tuning equipment reliably selects the extreme value. The apparatus required is simple and easily understood and may be maintained by relatively unskilled personnel.

• • •

In automatic control systems, it is frequently desired to adjust a control element to produce a maximum value of a resulting effect. (A minimum value may also be used; throughout this paper maximum only will be used for simplicity of expression.) Systems sensitive to the slope of the effect are common, where the sensitive element of the system examines the slope of the curve at the point where it happens to be resting when the tuning cycle begins, and adjusts the control element to increase the result. Such systems reliably determine the peak of any curve having only one maximum, and having no points of such low slope that the system cannot determine the proper direction of motion to correct.

In the system described in this paper, these limitations are not present. The system will accommodate response curves having more than one peak; in such cases the system will adjust the control element to the highest peak within the range. In addition, the system will operate correctly with curves having long, relatively flat sections within which the slope, if any, is so small as to be indeterminate. Curves of this type are often found in tuning radio transmitters and receivers. Secondary responses are occasionally produced by harmonics or image responses, and in receivers the system may be used to select the strongest of several competing signals.

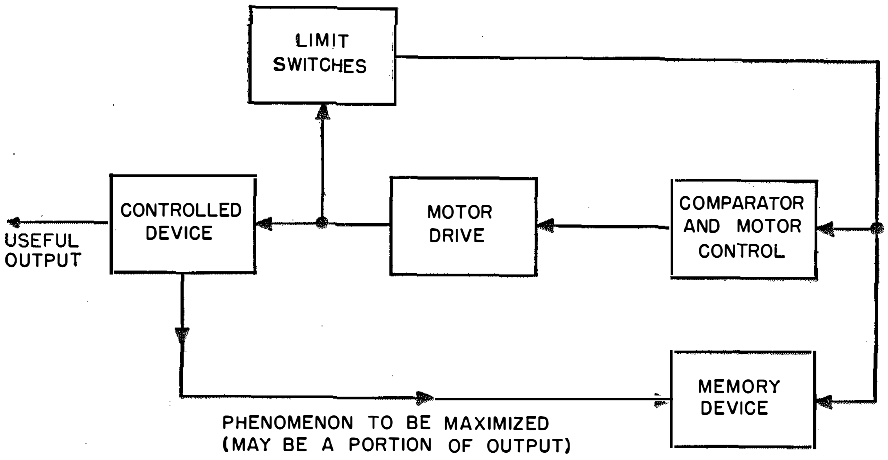

Figure 1—Block diagram of the system.

The sequence of operation is exactly like the operations performed by a careful human operator in tuning an equipment having the type of response described. Such an operator, when asked to readjust the system to tune to a new response point, would adjust the control element throughout its entire range, noting the amplitude of any

* This paper is based on development work sponsored by the Bureau of Ships of the United States Department of the Navy. It was presented at the American Institute of Electrical Engineers, Summer General Meeting in Atlantic City, New Jersey, on June 17, 1953 and published in *Electrical Engineering,* volume 72, pages 782–784; September, 1953: and in the *Transactions of the AIEE,* volume 72, pages 392–395; September, 1953.

peaks observed, then returning the control to the position producing the highest response. The automatic system proposed, which we have called the "memory-scanning" system, follows exactly this procedure.

The elements of this system are shown in Figure 1. The postulated response curve of the controlled device is shown as the solid line in Figure 2. The input to the sensitive element of

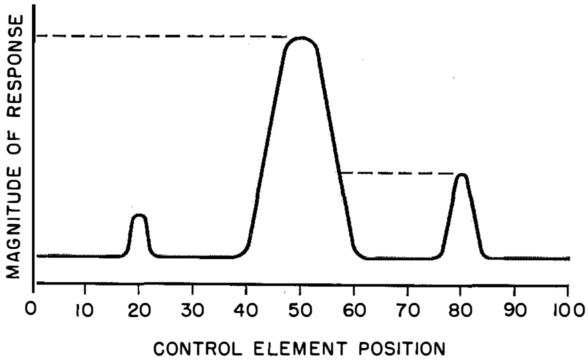

Figure 2—Assumed response curve of a typical controlled element.

the control system is assumed to be a unidirectional voltage of varying magnitude. When a tuning cycle is initiated, the controlled element is moved to one end of its range (say position 100 in Figure 2). During this time the switch *S* in the system shown in Figure *3* is in position *1*, shorting the memory capacitor *C*. When the controlled element has reached the end of its range, a limit switch is operated, causing the switch *S* to go to position *2* and causing the motor to move the controlled element through its entire range (say to position zero in Figure 2). During this scan, the signal representing the response of the controlled element is being impressed through a diode *D* on the memory capacitor *C*. The potential on this capacitor will rise when the input signal rises, but will not be able to drop when the input signal drops because of the unidirectional conducting properties of the diode *D*. Thus, the potential on the memory capacitor will follow the *dotted* lines in Figure 2, and at the end of this memory scan will be equal to the highest potential the input signal had reached during the scan.

At the end of this scan, another limit switch operates causing the switch *S* to be advanced to position *3*. In this position, the memory capacitor is connected to one of a balanced pair of tubes, while the instantaneous input signal from the controlled element is applied to the other tube of the pair. The motor is also caused to begin to return the controlled element to the other end of its range. At the instant that the potential delivered by the controlled element equals the remembered potential stored on the memory capacitor *C*, the signals applied to the two tubes will be equal and their plate potentials will be equal. (Necessary adjustments to affect such balance initially are assumed.) A sensitive relay connected from plate to plate will thus release just at the time this balance is achieved. This relay is arranged to stop the motor moving the controlled element. Thus the controlled element is stopped

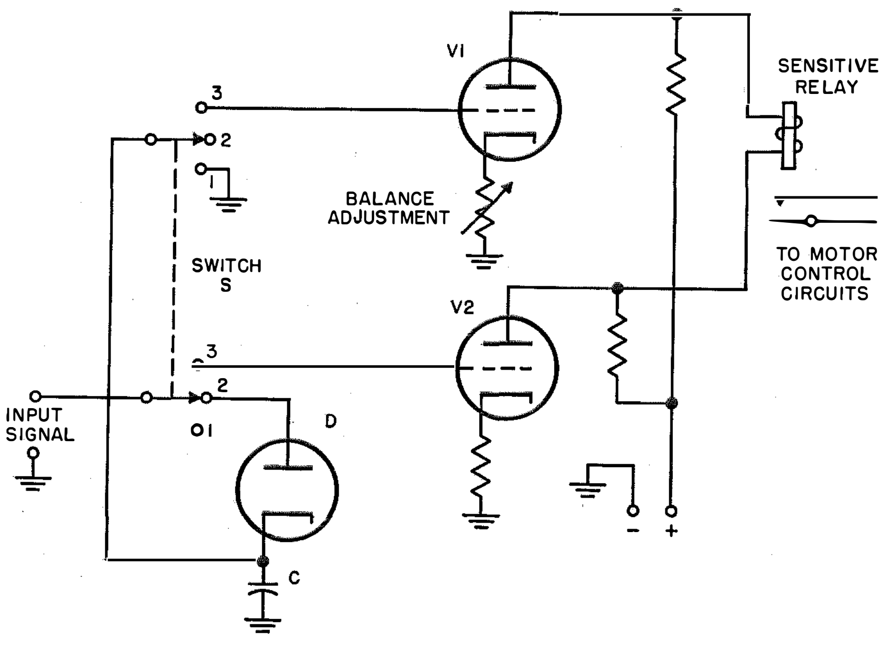

Figure 3—Simplified circuit diagram of the system.

just at the point giving the maximum value of the desired effect.

Practical design points affecting the accuracy of the foregoing statement are evident by examination of the system. In the first place, if exact matching of potentials is required to effect stopping of the motor, even a minor change in sensitivity of the system between the memory scan and the final tuning scan might prevent the motor from stopping. Secondly, if the motor stopping impulse were initiated at the moment of perfect tuning, there would inevitably be some overshooting of the moving parts of the system, introducing errors of setting. Fortunately these two effects may be balanced against each other to a degree that depends principally on the speed of tuning operation required. For example, if it is possible to take a long time for the final scanning operation, the time of action of the relays involved and the time to operate an electromagnetic brake to clamp the motor shaft will be negligibly short. In this case, almost exact matching of potentials will be desirable. If instabilities of the phenomenon being adjusted make accurate match difficult, it may be desirable to cause the system to stop when the instantaneous voltage is equal to only 0.8 of the remembered voltage, and possibly introduce an intentional time lag in the application of the braking force so that the normal stopping position of the system will be very near the peak. In other cases, where the time available for the tuning cycle is very limited, it may be necessary to anticipate the point of balance as above indicated especially in order to allow time for the motor-stopping action. If necessary, the motor speed may be reduced for the final scan to increase the accuracy of adjustment. By suitable choice of design factors, a wide range of applications may be accommodated.

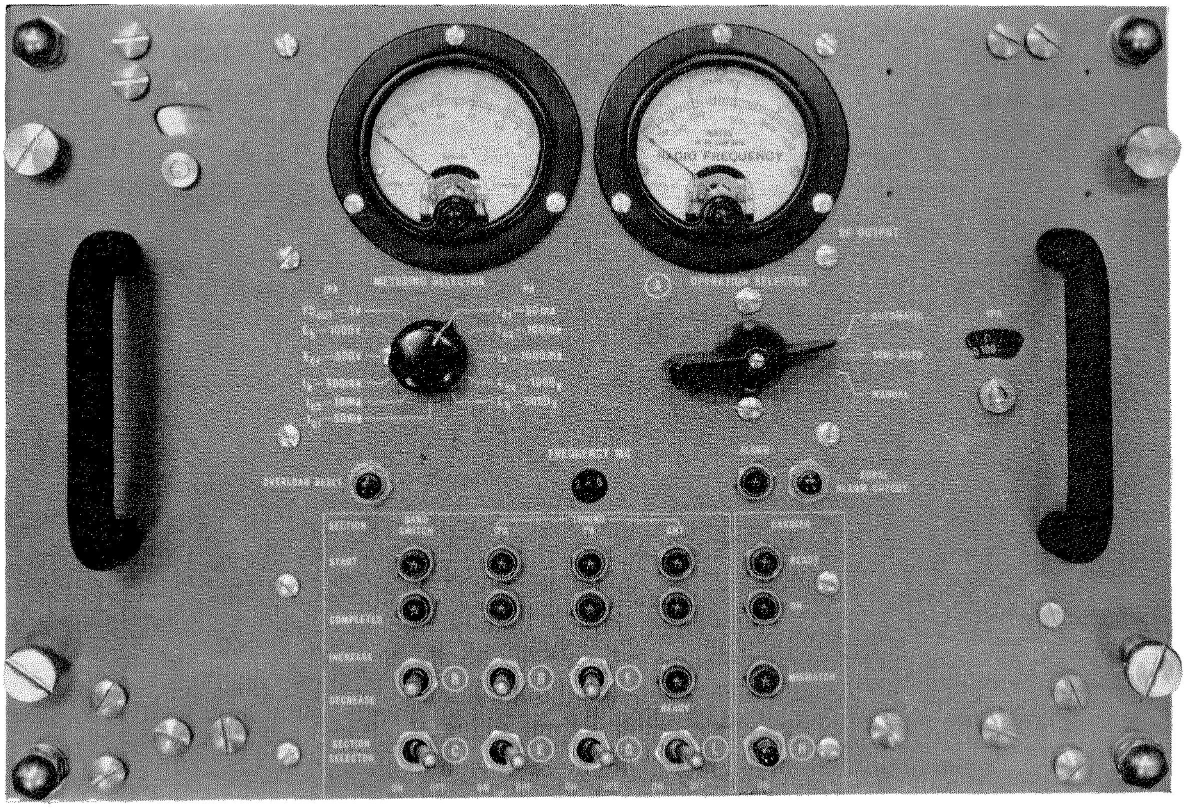

Figure 4—Front panel of radio-frequency amplifier of transmitter embodying the system.

One of the principal advantages of this system is the simplicity of the theory of operation, and its similarity to the actions of a human operator. This makes it possible for the usual technicians to understand fully what is going on and to feel confident in maintaining and trouble shooting the equipment. No knowledge of the effects of phase on servo-motors is required. No vacuum-tube amplifiers are needed. The drive motors may be of conventional types, either alternating or direct current, with no unusual problems of checking torque or maintaining slip rings. There is no direct limit to the size of the drive motors or the power that may be applied to the controlled element. (Some inertia problems will become apparent in larger sizes of apparatus.)

One embodiment of this principle that has reached production is the automatic tuning of the amplifier stages of a radio transmitter after the exciting frequency has been determined. One important advantage of this system over older systems having mechanical memory of the positions of shafts is in the indefinitely large number of predetermined channels that may be tuned. Any input frequency may be used, without regard to whether it has ever been set up before. Once the input is provided, the automatic system will resonate each of the amplifier stages to the proper frequency quickly and accurately.

The front panel of the radio-frequency amplifier section of this transmitter is shown in Figure 4. It will be noticed that there are no tuning controls in the usual sense. The knobs shown are for switching the panel meter to the proper circuits and for limiting the fully automatic action of the tuning system to slow down its tuning cycle for simpler analysis of troubles and for corrective maintenance. Jog switches are also provided to tune the circuits if the automatic system should fail, but they are not normally used. Two separate cycles of the above automatic tuning procedure are employed in this application. One motor drives the two low-level amplifier stages, which are ganged to a common shaft, while another motor drives the output amplifier, which supplies approximately 500 watts to a special automatic antenna-resonating system. (The final amplifier cannot be ganged to the low-level stages because of the reaction of the antenna system on its resonance point.) The tuning cycle for the smaller unit driving the low-level stages requires approximately 6 seconds, while that of the higher-powered stage requires from 8 to 14 seconds, depending on the frequency band in which the equipment is being operated.

The system described leads to simple, easily maintained equipment capable of setting any adjustable control to the point in its range producing a maximum (or minimum) of the effect desired. It is capable of an indefinitely large number of settings, depending on the input conditions. If the phenomenon being adjusted is susceptible to drift with time, repeating the tuning cycle will readjust to the new correct position. Its low cost and applicability to a wide range of effects makes it a valuable tool in the automatic adjustment of complex equipment.

Hyperbolic Protractor

A recent paper[1] describes a special protractor that simplifies and expedites calculations on microwave impedance measurements. These protractors are now available and may be purchased with an instruction booklet for $2.50, or for $2.00 each in lots of 12 or more to a single address. Orders and remittances in New York funds should be sent to Treasurer's Department, International Telephone and Telegraph Corporation, 67 Broad Street, New York 4, New York.

[1] G. A. Deschamps, "New Chart for the Solution of Transmission-Line and Polarization Problems," *Transactions of the IRE Professional Group on Microwave Theory and Techniques*, volume 1, pages 5–13; March, 1953: and *Electrical Communication*, volume 30, pages 247–254; September, 1953.

Experimental Determination of the Properties of Microstrip Components*

By MAURICE ARDITI

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MICROSTRIP lines have been developed as a substitute for waveguides and coaxial lines.[1-3] The microstrip lines take the form of thin strips of copper, aluminum, or silver bonded to a slab of dielectric.

Figure 1 shows different geometrical configurations. Various dielectrics have been tested: polyethylene, Fiberglas, and teflon-impregnated Fiberglas.

Several microwave components have been developed using microstrip.[1-3] However, at present, an exact theory of microstrip or other open lines has not been completely developed and no theoretical formulas are available for the characteristics of the line or for the characteristics of obstacles in the line. An experimental investigation of these characteristics was undertaken with the aim of accumulating sufficient data to permit the compilation of a microstrip design handbook.

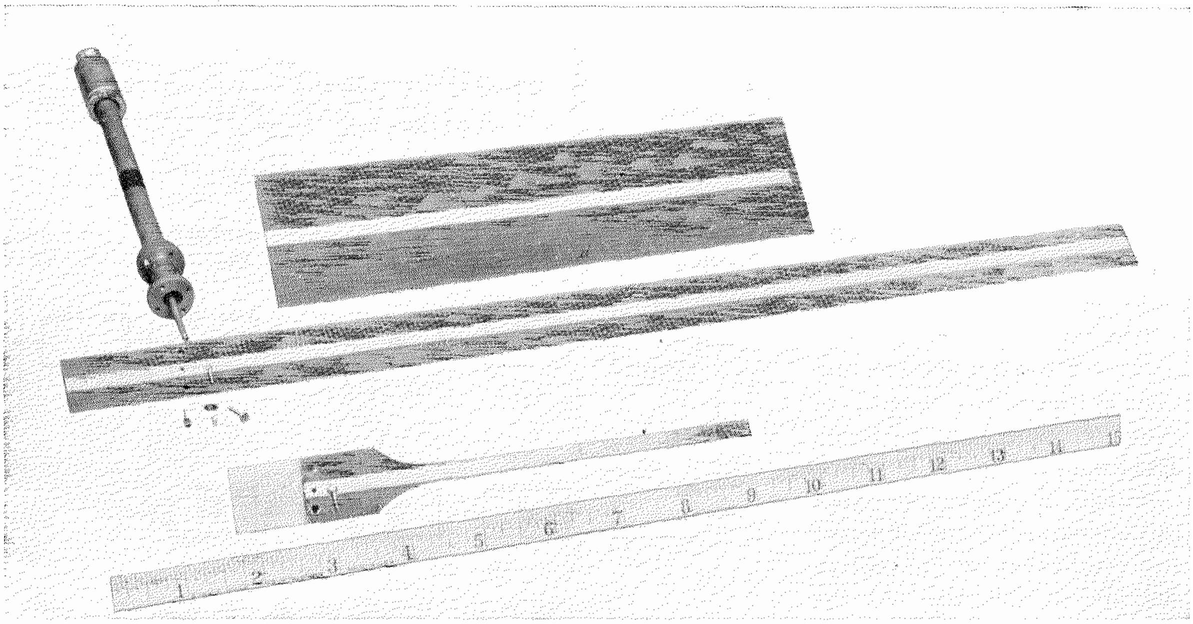


Figure 1—Microstrip lines.

The characteristics of transducers, bends, transverse posts, offset junctions, step discontinuities, gaps or slots in the strip conductor, parallel-coupled junctions, and directional couplers are given for some particular geometrical configurations.

Examples of resonant sections made with microstrip will be briefly discussed. As an example

* Reprinted from *Convention Record of the 1953 I.R.E. National Convention*, Part 10—Microwaves, pages 27–37. Presented at the National Convention of the Institute of Radio Engineers in New York, New York, on March 26, 1953. Figures 12 and 13 of the original paper have been omitted.

[1] D. D. Grieg and H. F. Englemann, "Microstrip—A New Transmission Technique for the Kilomegacycle Range," *Electrical Communication*, volume 30, pages 26–35; March, 1953: also, *Proceedings of the IRE*, volume 40, pages 1644–1650; December, 1952.

[2] F. Assadourian and E. Rimai, "Simplified Theory of Microstrip Transmission Systems," *Electrical Communication*, volume 30, pages 36–45; March, 1953: also, *Proceedings of the IRE*, volume 40, pages 1651–1657; December, 1952.

[3] J. A. Kostriza, "Microstrip Components," *Electrical Communication*, volume 30, pages 46–54; March, 1953: also, *Proceedings of the IRE*, volume 40, pages 1658–1663; December, 1952.

of practical realization, various microstrip components have been assembled into a wideband microwave receiver, complete with automatic frequency control. A comparison with a similar receiver built with conventional rectangular waveguide shows the savings in material and the simplification of manufacture that microstrip can provide.

1. Method for Experimental Determination of Characteristics of Microstrip Components

Recently, a method has been described by G. A. Deschamps[4–7] for the determination of the reflection coefficients and insertion losses of a junction in closed waveguides. This method has been adapted to microstrip and the details have already been disclosed in a previous paper.[8] The experimental setup is shown schematically in Figure 2. Briefly the principle of the method is as follows.

The microstrip specimen is connected through the junction to an ordinary coaxial or waveguide standing-wave machine. The length of the line is varied over half a wavelength either by moving a short-circuit across the conductors or by cutting the line at its end, and the corresponding reflection coefficient is measured at the input in the standard manner. The successive positions of the short-circuit are arranged to differ by $\lambda/8$ (as measured in the microstrip line). When plotted on the complex plane these reflection coefficients fall on a circle, assuming the losses in the line are small within half a wavelength. The chords joining opposite points intersect in a point that is related in a simple manner to the reflection and transmission coefficients of the junction.[4–7]

By repeating the measurements for two different lengths of the line with the same junction, the losses in the line per unit length can be obtained. The scattering-matrix coefficients and

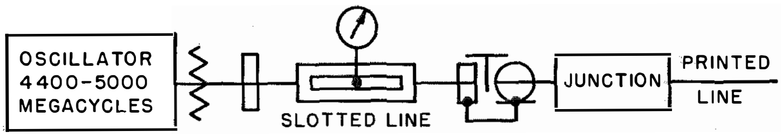


Figure 2—Schematic of experimental setup.

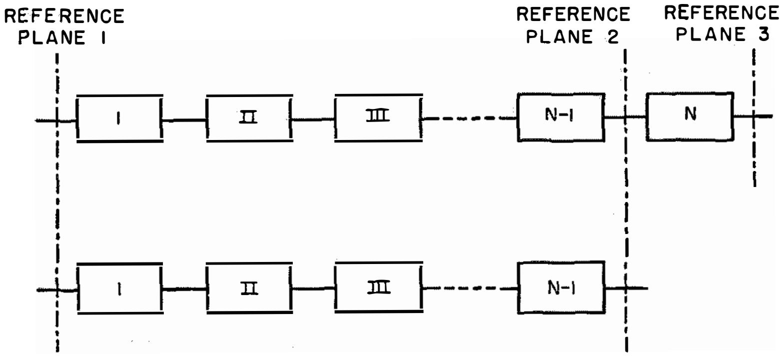


Figure 3—Junctions or discontinuities in series.

[4] G. A. Deschamps, "Application of NonEuclidian Geometry to the Analysis of Waveguide Junctions," presented at the Joint Spring Meeting of the American Section of the International Scientific Radio Union and the Institute of Radio Engineers in Washington, District of Columbia; April 23, 1952. Published in part under the title, "Determination of the Reflection Coefficient and Insertion Losses of a Waveguide Junction" in *Journal of Applied Physics,* volume 24, pages 1046–1050; August, 1953.

[5] G. A. Deschamps, "Geometric Viewpoints in the Representation of Waveguides and Waveguide Junctions," *Proceedings of the Symposium on Modern Network Synthesis,* pages 277–295; September 30, 1952. Presented at the symposium sponsored by the Polytechnic Institute of Brooklyn and the Office of Naval Research in New York, New York, on April 18, 1952.

[6] G. A. Deschamps, "New Chart for the Solution of Transmission-Line and Polarization Problems," *Electrical Communication,* volume 30, pages 247–254; September, 1953: also, *Transactions of the IRE Professional Group on Microwave Theory and Techniques,* volume 1, pages 5–13; March, 1953.

[7] G. A. Deschamps, "Accurate Comparison of High Standing-Wave Ratio and Its Application to the Determination of Attenuation Constant of a Waveguide," presented at the Ultra-High-Frequency Measurements Conference in Washington, District of Columbia, on January 16, 1953. Unpublished.

[8] M. Arditi and J. Elefant, "Characteristics of Microwave Printed Lines with Application to Bandpass Microwave Filter Design," presented at the Joint Spring Meeting of the International Scientific Radio Union and the Institute of Radio Engineers in Washington, District of Columbia, on April 23, 1952. Unpublished.

[9] S. E. Miller and W. W. Mumford, "Multi-Element Directional Couplers," *Proceedings of the IRE,* volume 40, pages 1071–1078; September, 1952.

[10] H. F. Englemann, "Microstrip—A New Microwave Transmission Technique," presented at the Symposium of the Institute of Radio Engineers Professional Group on Microwave Theory and Techniques in New York, New York, on November 7, 1952. Unpublished.

the losses of the junction itself can then be deduced. Also, for a given length of line, by plotting the two circle diagrams corresponding to an open end or a short-circuited end of line, the reflection coefficient at the open end of the line can be obtained in magnitude and phase. From the circle diagrams, the wavelength or phase velocity of the dominant mode can be easily deduced.

When measuring a discontinuity or junction N (see Figure 3) through a series of junctions I, II . . . N-1, N connected through lengths of line, the method is to consider first all the junctions I, II . . . N-1, N in series as a two-port linear network, then disconnect junction N and consider junctions I . . . N-1 in series as a two-port linear network. From these two measurements, the characteristics of junction N can be deduced. Here again, the computations are considerably simplified by a series of simple graphical constructions suggested by Deschamps.[4]

Besides, the method has the considerable advantage of not disturbing the line under test: there is no danger of either spurious coupling or field perturbation as may occur if a traveling probe is used.

Although, theoretically, some caution should be exercised in the extension of the method to open structures, experience shows that the results obtained with microstrip can be used with a very good degree of accuracy in the design of microwave components.

2. *Characteristics of Microstrip Components*

2.1 General Considerations

Commercially available waveguides are usually dimensioned for a frequency band such that only the dominant mode of transmission is above cutoff. For this mode, the attenuation of the waveguide is well defined. In the case of microstrip, the theory of all possible modes of transmission along the line is not yet completely developed, and it is difficult at this time to specify the attenuation in the line for any given mode. The presence or absence of the various modes will depend on the method of excitation and on the geometry of the line.

In the following experiments, the microstrip line was excited through a coaxial feeder to be described later. With this particular transducer, the line appears to carry only one dominant mode, and application of Deschamps' method was possible. Because of the possibility of excitation and propagation of several modes in the line, caution should be exercised in applying the results to situations where the line is excited in another way. Some obstacles in the line can also excite higher modes that can propagate along the line. Examples of this effect will be shown later. Care should be taken to avoid such obstacles because they complicate considerably the design of microstrip components.

Unless specifically mentioned, most of the results given below are related to a microstrip line having the following characteristics:

Dielectric—Fiberglas (supplier: Formica Company, code number *G-6*). $\epsilon' = 4.2$ at 4700 megacycles per second (loss factor 0.0063, power factor 0.0015 at 1 megacycle).

Conductors—Copper strips 0.00135-inch (0.038-millimeter) thick (i.e., about 40 times the skin depth at 4700 megacycles).

Most of the experiments have been carried over the frequency band from 4400 to 5000 megacycles.

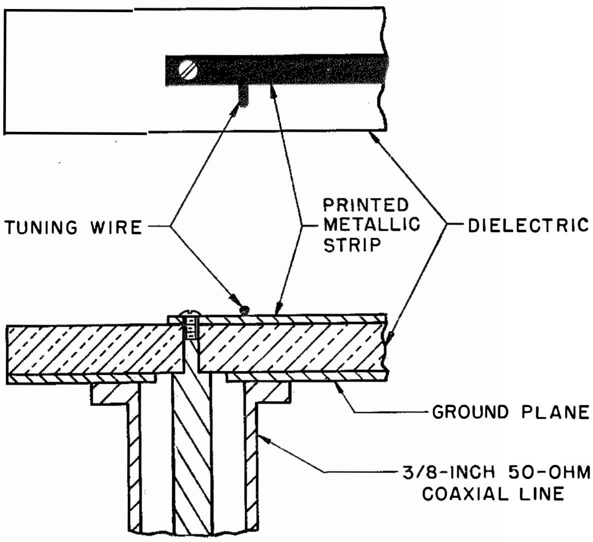


Figure 4—Wideband transducer from coaxial to microstrip.

In certain cases, the losses in the junction are so small that they can be neglected and an equivalent circuit has been drawn. No particular attempt was made to derive an equivalent circuit for lossy junctions because it is equally easy to work with the scattering-matrix coefficients directly given by Deschamps' method.

2.2 Wideband Transducer

A coaxial-to-microstrip transducer has been developed (Figure 4). This transducer is easy to match over a wide band of frequencies. The matching is achieved by properly locating a small piece of wire on the strip line near the junction (see also Figure 1). The input voltage standing-wave ratio of this transducer in the frequency range from 3800 to 5400 megacycles is less than 1.4, the average value being 1.2. The insertion loss of this transducer is of the order of 0.3 decibel. With this type of transducer, the following characteristics have been obtained for the microstrip line.

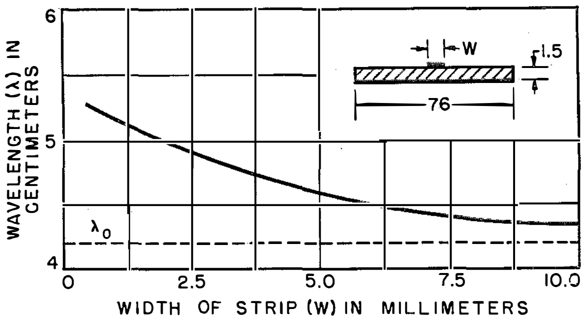


Figure 5—Wavelength of dominant mode in polyethylene-dielectric microstrip. The dimensions are given in millimeters. The frequency is 4700 megacycles.

The wavelength λ of the dominant mode is greater than the value λ_0 corresponding to TEM-mode propagation between infinite parallel plates. Figure 5 shows the variation in λ with change in width of strip conductor. As the width increases, the wavelength approaches the value of λ_0. In the frequency range from 4400 to 5000 megacycles, the line is not dispersive and this is probably true for a larger frequency range.

Smaller values of losses have been obtained with microstrip made of teflon-impregnated Fiberglas: at 4700 megacycles the losses are of the order of 1.4 ± 0.2 decibels per meter.

2.3 Right-Angle Bends

Figure 6 shows the variations of the input voltage standing-wave ratio produced by different cuts in a right-angle bend in a line terminated by a matched load. As can be seen, a 45-degree cut or a smooth round corner will produce no appreciable reflection, and this has been checked over the band from 4400 to 5000 megacycles. The radiation losses in the case of a round corner are of the order of 0.15 decibel.

2.4 Transverse Posts

The variation in shunt susceptance with the diameter of a round metallic post transverse to

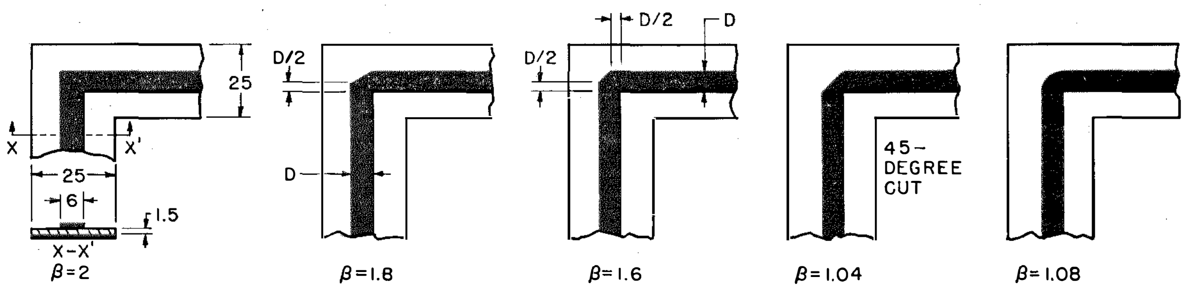


Figure 6—Right-angle bends in microstrip. The dimensions are given in millimeters. Frequency is 4700 megacycles. β is the input voltage standing-wave ratio.

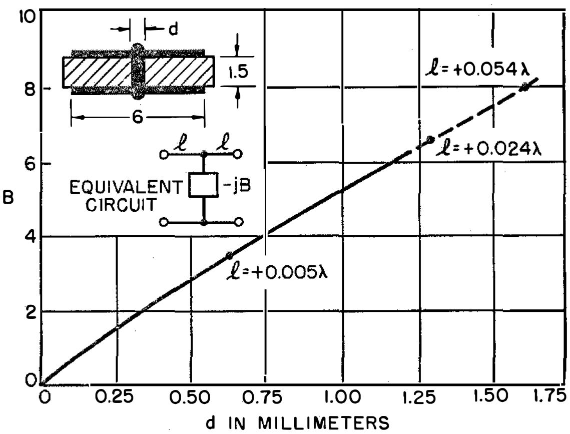


Figure 7—Variation of shunt susceptance with diameter of a post. The dimensions are given in millimeters. The frequency is 4700 megacycles. The reference plane is at the axis of the post.

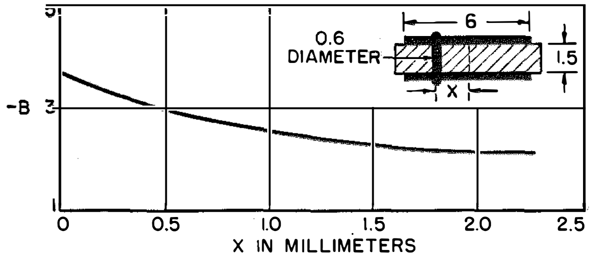


Figure 8—Variation of shunt susceptance with location of a post. The dimensions are in millimeters.

the line on the axis and soldered to the conductors, is shown in Figure 7. Figure 8 shows the variation in shunt susceptance with the location of the post across the line. For values of susceptance B larger than -6, the radiation losses are of such an order that it is preferable to specify in terms of the scattering coefficients, rather than the lossless equivalent circuit. These posts can excite different modes in the line, and care should be taken when using these elements in the design of a resonant section.

2.5 Offset Junctions

Figure 9 shows several configurations of an offset junction with an equivalent circuit representation. The values of the shunt susceptance produced by this type of junction are relatively small, even for large offsets. The radiation losses in all cases are smaller than 0.1 decibel.

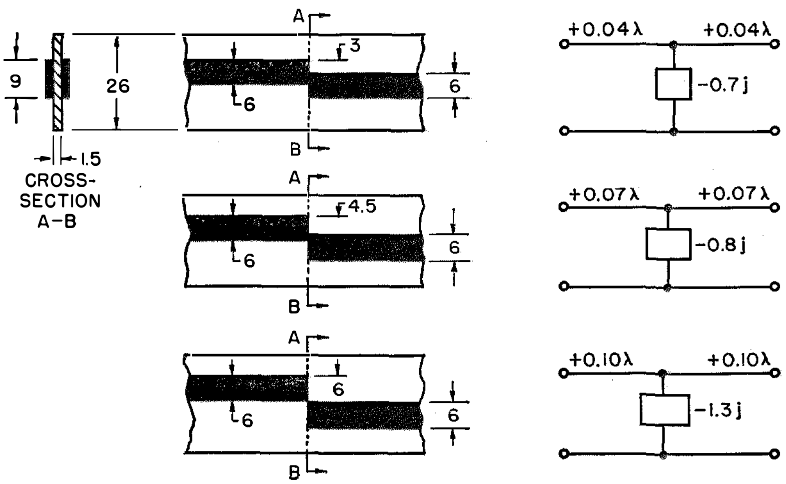


Figure 9—Microstrip offset junction using strip conductors on both sides of the dielectric. The dimensions are given in millimeters. The frequency is 4700 megacycles. In the equivalent circuit, the reference plane is at AB.

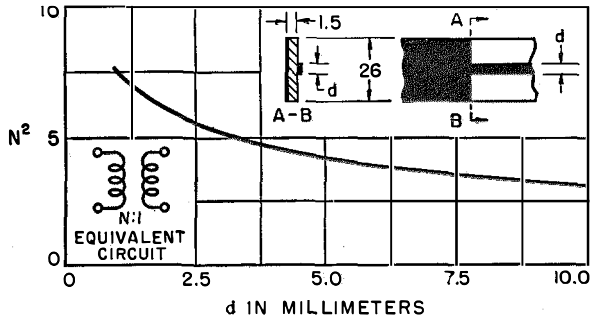


Figure 10—Microstrip step discontinuity and equivalent transformer circuit. In the equivalent circuit, the reference plane is at AB. The dimensions are given in millimeters. The frequency is 4700 megacycles.

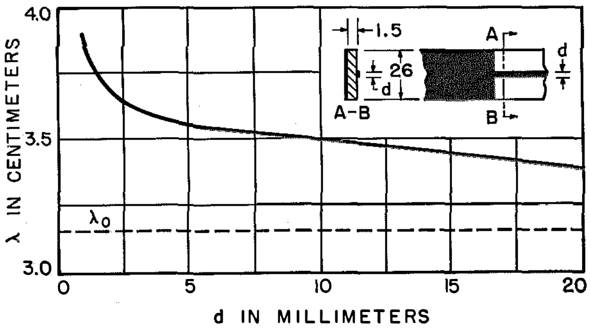


Figure 11—Line wavelength versus line width for a microstrip step discontinuity. Dimensions are in millimeters. The frequency is 4770 megacycles.

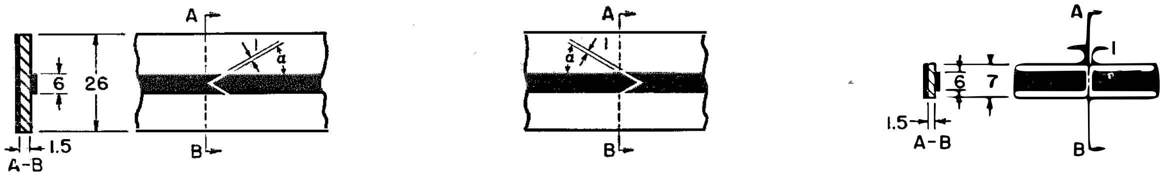
2.6 Step Discontinuity

The experiments show that the equivalent circuit of a microstrip step discontinuity is an ideal transformer, provided the losses in the line are small. Figure 10 shows the values of N^2, the square of the turns ratio of the equivalent transformer for different widths of one of the lines. Figure 11 shows the corresponding variation in the line wavelength. In some cases, multistep discontinuities of short length can produce multimoding.

2.7 Gaps or Slots in Microstrip

Like the step discontinuity or the parallel coupling, this type of obstacle is perfectly adapted to any form of printed-circuit technique, since the additional element is contained in two dimensions on the surface of the dielectric.

It was found that for a given gap spacing the values of the reflection coefficients are a function of the angle of the slot in the strip line (Figure 12). The values of the transmission coefficient $|S_{12}|^2$ in decibels are given in Figure 13 for different gap spacings.



Scattering-Matrix Coefficients (Reference plane at AB)

$$\alpha = 17 \text{ degrees} \begin{cases} S_{11} = -0.35 - j0.68 \\ \pm S_{12} = -0.53 + j0.29 \\ S_{22} = -0.28 - j0.68 \end{cases} \qquad \alpha = 40 \text{ degrees} \begin{cases} S_{11} = 0.72 - j0.65 \\ \pm S_{12} = 0.08 + j0.24 \\ S_{22} = 0.96 + j0.12 \end{cases} \qquad \alpha = 90 \text{ degrees} \begin{cases} S_{11} = 0.98 - j0.12 \\ \pm S_{12} = 0.03 + j0.20 \\ S_{22} = 0.93 - j0.16 \end{cases}$$

$$\alpha = 45 \text{ degrees} \begin{cases} S_{11} = 0.45 - j0.86 \\ \pm S_{12} = 0.06 + j0.29 \\ S_{22} = 0.59 + j0.70 \end{cases}$$

Figure 12—Gaps in microstrip. The dimensions are in millimeters. The measurement frequency is 4700 megacycles.

2.8 Parallel–Coupled Junctions

Figure 14 shows several configurations together with the corresponding values of the scattering-matrix coefficients. Large values of susceptance can be produced by this type of junction, but the radiation losses at the ends of the lines limit their use in the design of high-Q resonant sections.

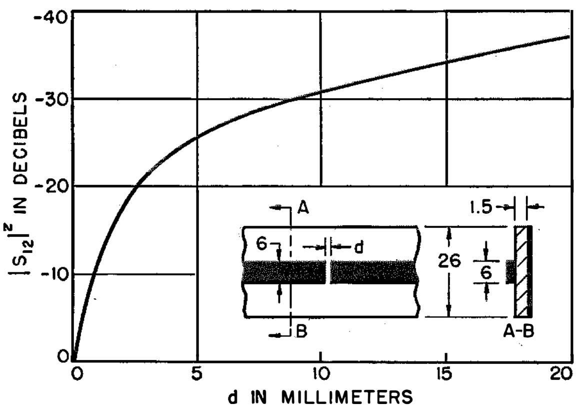


Figure 13—Gaps in microstrip—transmission coefficient. The dimensions are in millimeters. The measurement frequency is 4700 megacycles.

2.9 Directional Coupler and Magic T

Microstrip hybrid ring structures have been described in previous papers.[1, 3, 10] Directional couplers based on the properties of parallel adjacent lines have also been constructed; the principle of operation is identical to that of parallel-coupled waveguides.[9] When the coupling interval is several wavelengths long, broadband directivity can be achieved.

Figure 15 shows the results obtained with several types of microstrip directional couplers. Each termination was matched to a coaxial feeder with a voltage standing-wave ratio of 1.08 or better. The directivity and coupling are a function of the angle of approach α of one line with respect to the other.

When $\alpha = 90$ degrees, the coupling is about -10 decibels with a directivity of 3 decibels.

When $\alpha = 30$ degrees, the coupling is about -1 decibel with a directivity of 26 decibels.

Such a directional coupler can be used as a broadband magic T when α is somewhat smaller than 30 degrees. Effectively, in that case, when all arms of the coupler are connected to matched loads, terminals *2* and *4* will receive an almost equal amount of power from the input signal source at *1*, and terminal *3* will receive an amount of power 30 decibels below those of *2* or *4*. Moreover, it has been checked that these properties apply almost equally well when any of the terminals *1*, *2*, *3*, or *4* is taken as the input.

2.10 Resonant Sections

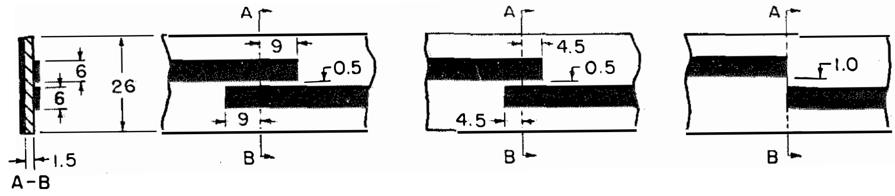
In the previous paragraphs, the characteristics of some obstacles, discontinuities, or junctions in microstrip have been described. By placing two such obstacles at the proper distance, a resonant section can be obtained. In the design of such resonant sections in microstrip, the following properties have to be taken into account:

A. The line is not dispersive.

B. The line can support several modes.

C. The losses in the line and the losses in the obstacles are not negligible. In a given line, the circle diagram of the obstacles permits the determination of the input voltage standing-wave ratio and the insertion losses of the resonant section.[8]

For the step discontinuity, the transverse post, or the straight gap in the upper conductor, it was found that the scattering coefficients did not vary much in the frequency range from 4400 to 5000 megacycles. For the parallel-coupled junction, the length of the coupling sections makes this



Scattering-Matrix Coefficients (Reference plane at AB)

$S_{11} = -0.95 - j0.09$	$S_{11} = -0.93j$	$S_{11} = 0.95$
$\pm S_{12} = 0.07 - j0.19$	$\pm S_{12} = 0.36$	$\pm S_{12} = 0.30j$
$S_{22} = -0.94 - j0.09$	$S_{22} = -0.89j$	$S_{22} = 0.90$

Figure 14—Microstrip parallel-coupled junctions. All dimensions are given in millimeters. The frequency is 4700 megacycles.

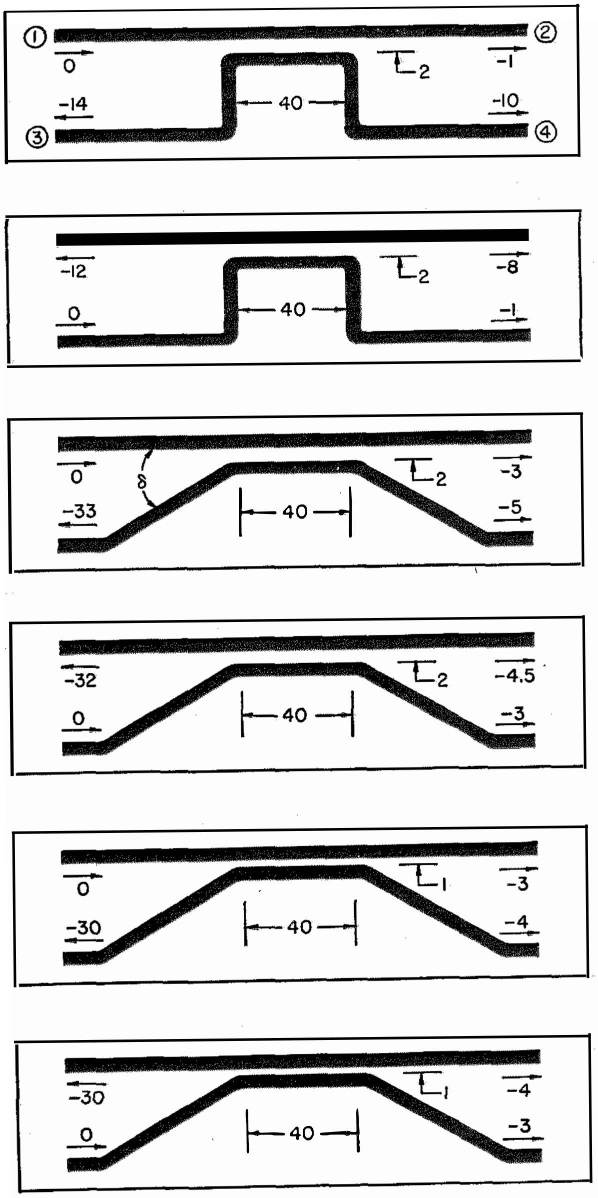


Figure 15—Left, microstrip directional couplers. Dimensions are given in millimeters. The numbers associated with the directional arrows indicate the power level, zero level being the input. The terminal numbering is shown in circles for the uppermost drawing. Angle δ is 30 degrees.

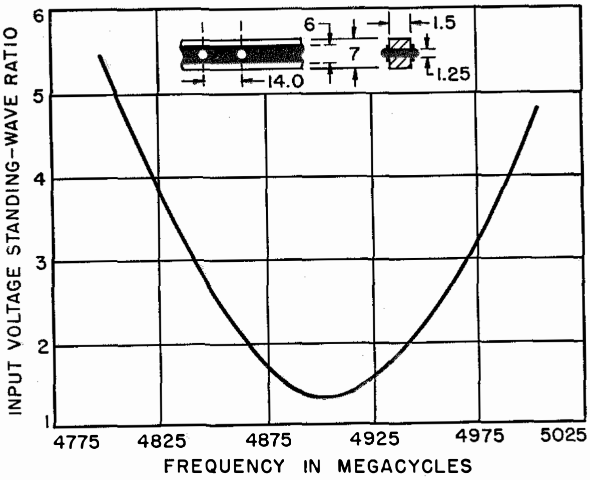


Figure 16—Microstrip resonant section with transverse posts. All dimensions are in millimeters. The insertion losses at center frequency were 1.4 decibels.

type of junction frequency sensitive, and this effect has to be included in the design of a resonant section.

Figure 16 shows the response curve of a resonant section made with posts transverse to the line. Figure 17 shows a similar curve for a resonant section made with two gaps in the line, and Figure 18 a resonant section made with parallel-coupled junctions.

The radiation losses from the obstacles and the losses in the line are a limitation in obtaining higher Q values.

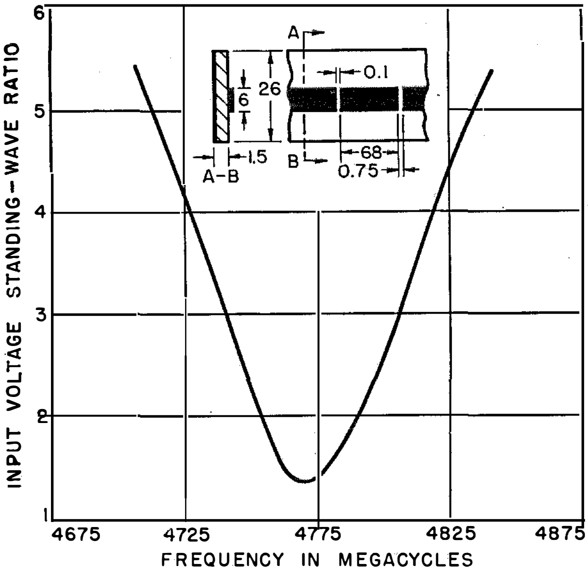


Figure 17—Microstrip resonant section with gaps in the line. All dimensions are in millimeters. The insertion losses at center frequency were 6.5 decibels.

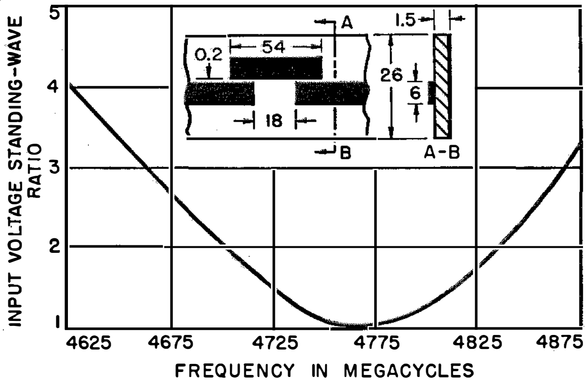


Figure 18—Microstrip resonant section with parallel-coupled junctions. All dimensions are given in millimeters. The insertion losses at center frequency were 0.8 decibel.

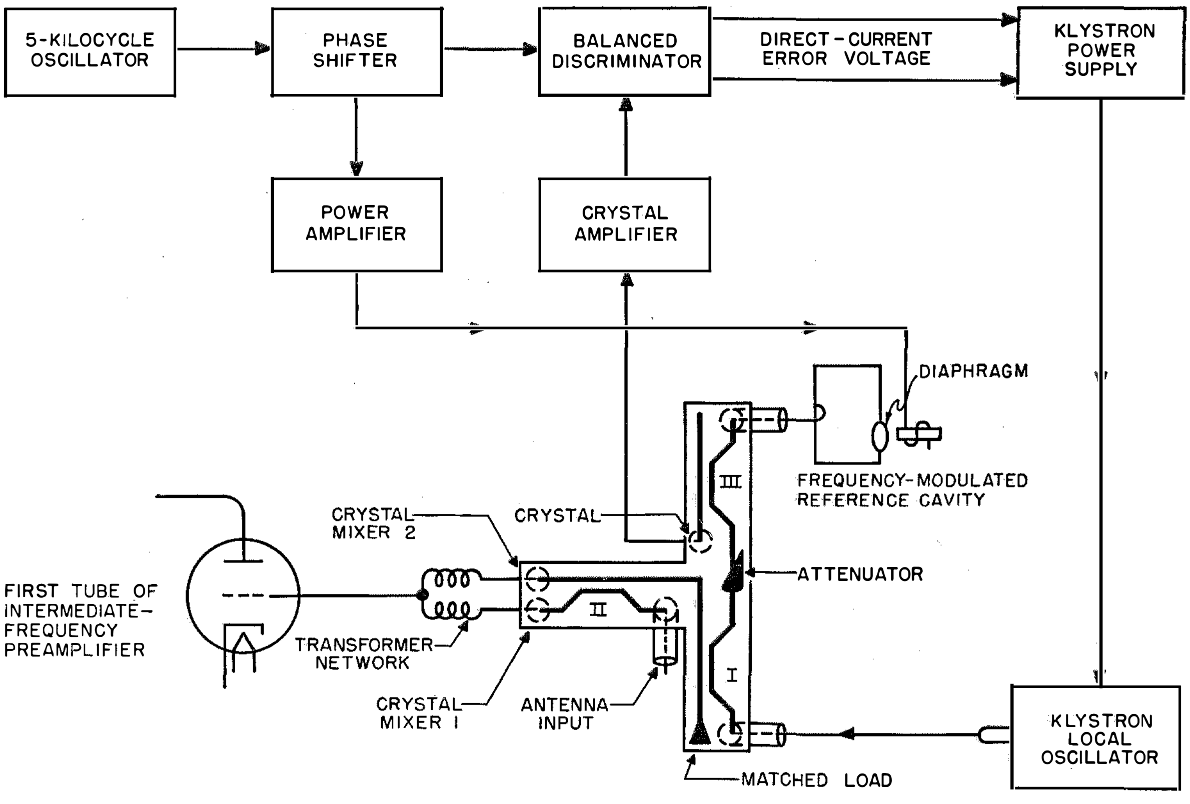


Figure 19—Schematic of microstrip wideband microwave receiver with automatic frequency control.

3. *Microwave Receiver Using Microstrip Components*

As an example of a practical realization using various microstrip components previously described, Figure 19 shows a schematic of the radio-frequency head of a broadband superheterodyne microwave receiver with an automatic-frequency-control system using a modulated reference cavity. Except for the high-Q cavity resonator, the whole radio-frequency circuit is printed in microstrip.

The system uses three microstrip magic T's of the type previously described.

The local oscillator feeds magic T *1*, which splits the microwave power between magic T *2* and magic T *3*. The antenna is also connected to

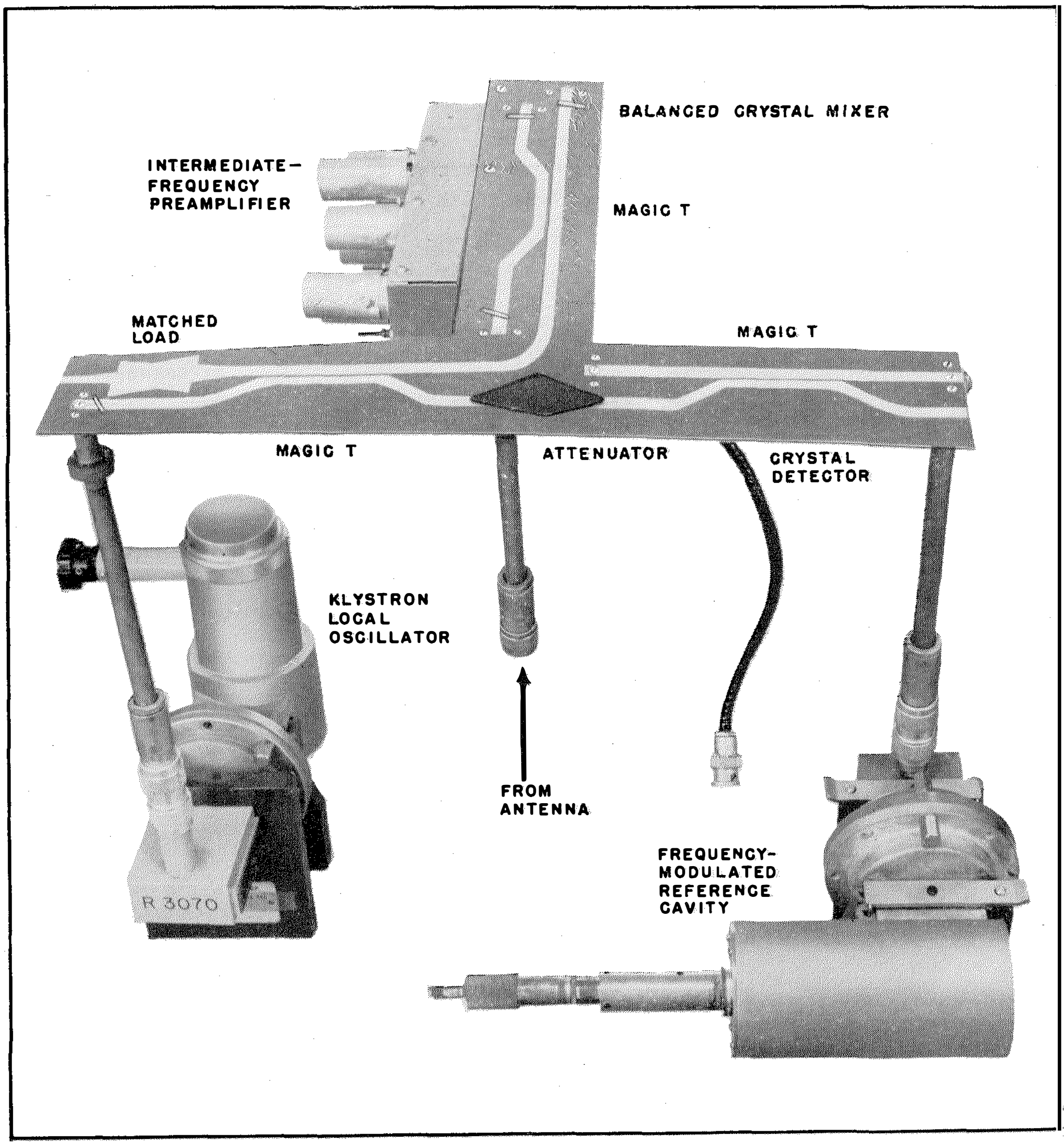


Figure 20—Microstrip receiver—front view.

magic T *2*, and the signals from the antenna and the local oscillator are mixed in balanced crystal mixers placed at the ends of the lines of magic T *2*. The resultant intermediate-frequency signal is amplified through an intermediate-frequency preamplifier of the cascode type.

The signal coming from the local oscillator splits between the arms of magic T *3*. Reflections from the open line and the modulated reference cavity are mixed in a crystal detector. The resultant current is amplified and used to unbalance a balanced detector. The amplitude and phase of the crystal output determine the nature of the unbalance. The relation of the center frequency of the cavity to the frequency of oscillation of the klystron local oscillator determines the sign of the direct-current output of the discriminator. The automatic frequency control is realized by connecting the discriminator output in series with the klystron's repeller supply.

Figures 20 and 21 show the actual realization. The saving in components and space can be appreciated by comparing with Figure 22, which shows a similar receiver made with conventional 2-by-1-inch (5-by-2.5-centimeter) rectangular waveguide.

In the frequency range from 4400 to 5000 megacycles, the noise figure of this receiver with

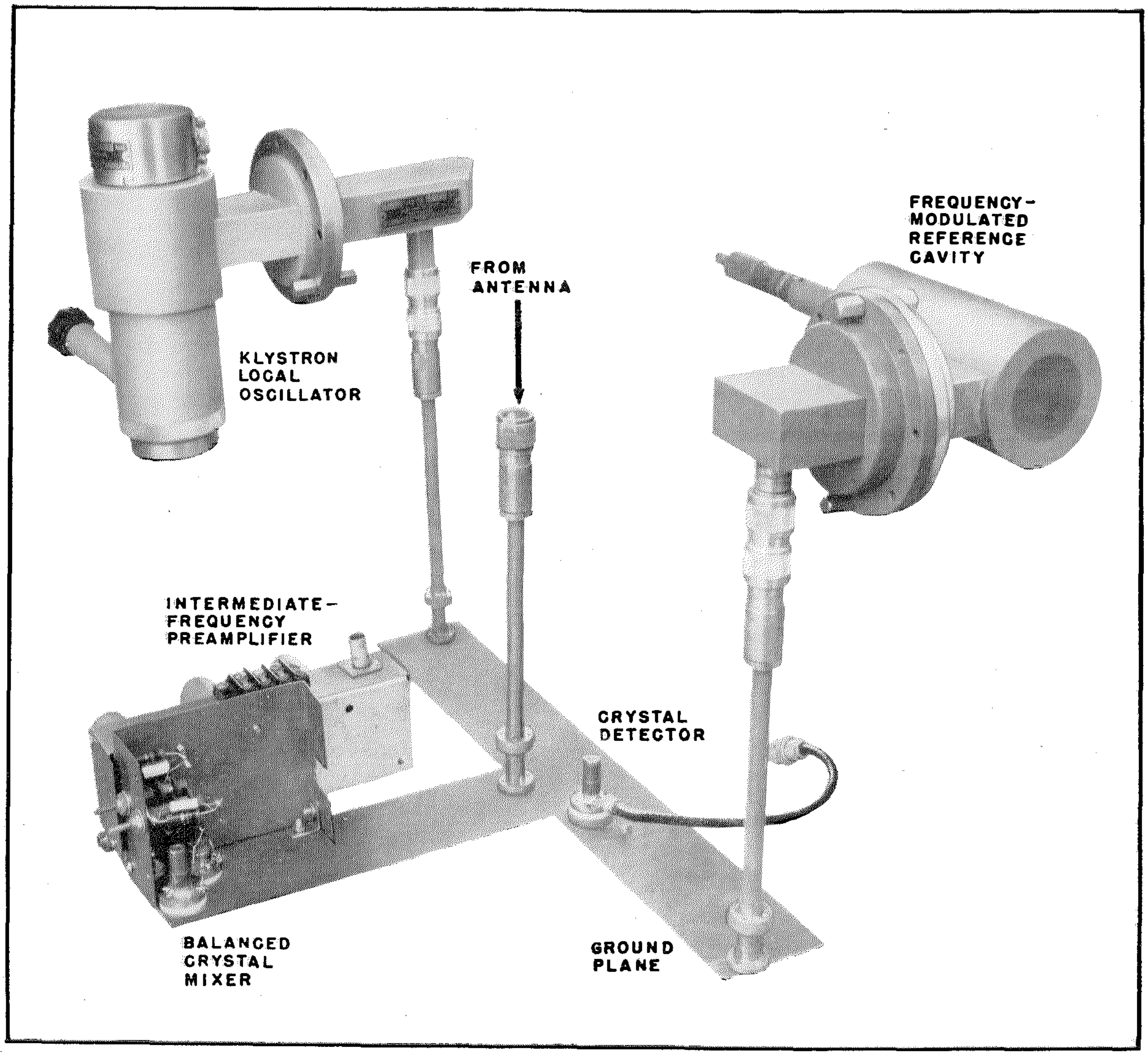


Figure 21—Microstrip receiver—back view.

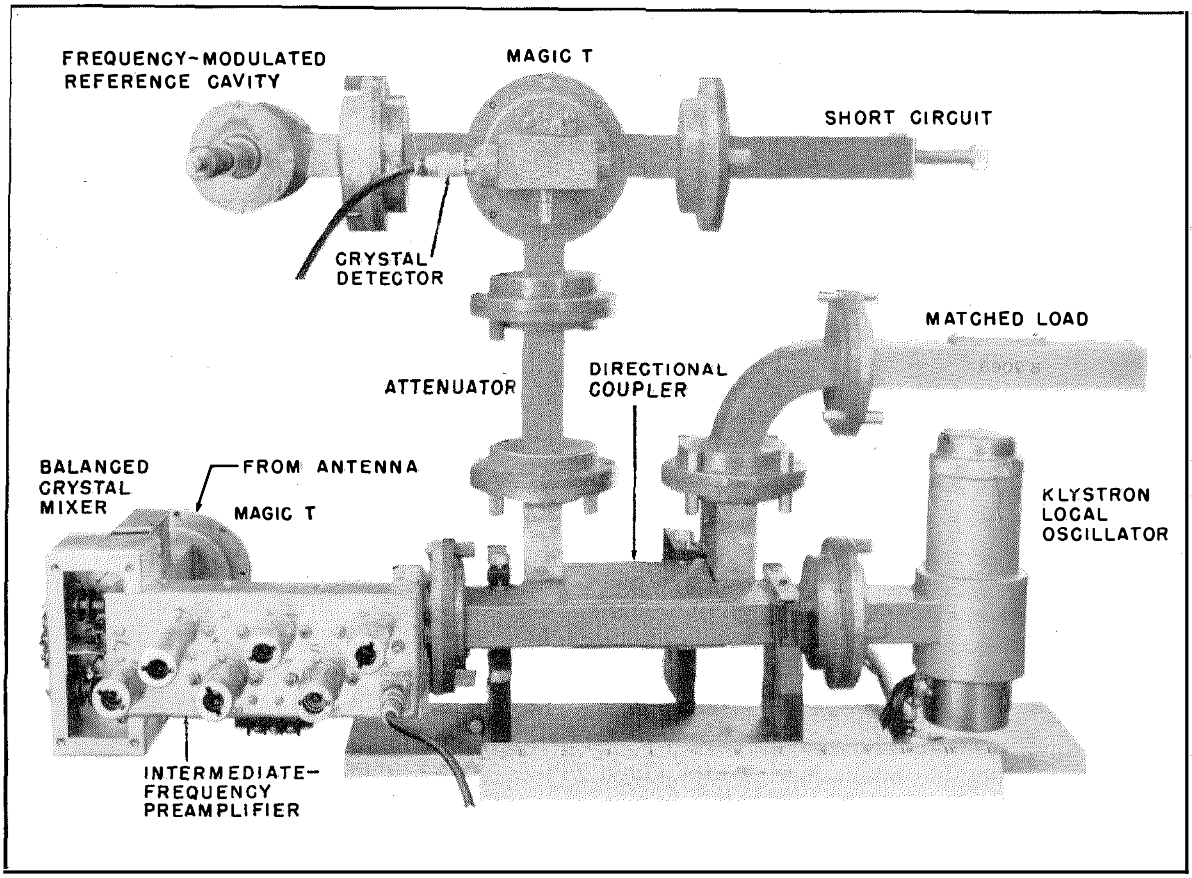


Figure 22—Conventional microwave receiver using rectangular waveguide.

an intermediate frequency of 100 megacycles and a 3-decibel bandwidth of 30 megacycles was about 16 decibels and quite comparable with a similar receiver using rectangular waveguides. The performance of the automatic-frequency-control system was also similar to that obtained with rectangular waveguides.

4. Conclusions

The types of microstrip structures described, namely transducers, magic T's, and discontinuities of various kinds, represent only a small fraction of the designs possible. The ease with which microstrip can be manufactured and tested facilitates the accumulation of a large amount of experimental data useful in the development of microwave components. If an equipment is designed so as to make full use of both microstrip and conventional printed-circuit techniques, a considerable saving in space, weight, and cost can be achieved.

Wide-Frequency-Range Tuned Helical Antennas and Circuits*

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TRANSMISSION-LINE circuit elements or, more generally, circuit elements with distributed constants, are used advantageously at very- and ultra-high frequencies. At lower frequencies, except for antennas and occasionally in associated networks, their use is not common for the very good reason that such circuit elements become prohibitively bulky. In antenna applications the large size is tolerated in order to achieve a useable radiation efficiency.

This condition prevails as long as the type of distributed-constant circuit elements under consideration is limited to the conventional coaxial- or parallel-transmission-line type. There is, however, a class of distributed-constant circuit elements, made up of helical conductors with or without cylindrical outer conductors, that can be used with considerable advantage at all radio frequencies up to and including the ultra-high-frequency band. Of particular interest for the present discussion is the case of the helix with a diameter small compared to a wavelength.

Though this class of helical circuit element has seen limited use in applications similar to the ones under discussion at present, it is felt that their potentialities have not been fully appreciated and utilized nor has their theory and fundamental properties been adequately discussed in technical literature.

The helical antenna, with diameter considerably less than a wavelength, radiating predominantly vertical polarization in the normal mode (peak of radiation pattern normal to the helix axis) has several advantages over a conventional dipole when the height must be considerably less than a quarter wavelength. The radiation resistance is about 50-percent greater than a dipole of the same height and the input impedance can be made real so that the problem of matching to a transmission line is simplified. In addition, the helix can be matched by grounding the base of the helix and connecting the transmission line to the proper tap point so that no external matching networks are required.[1]

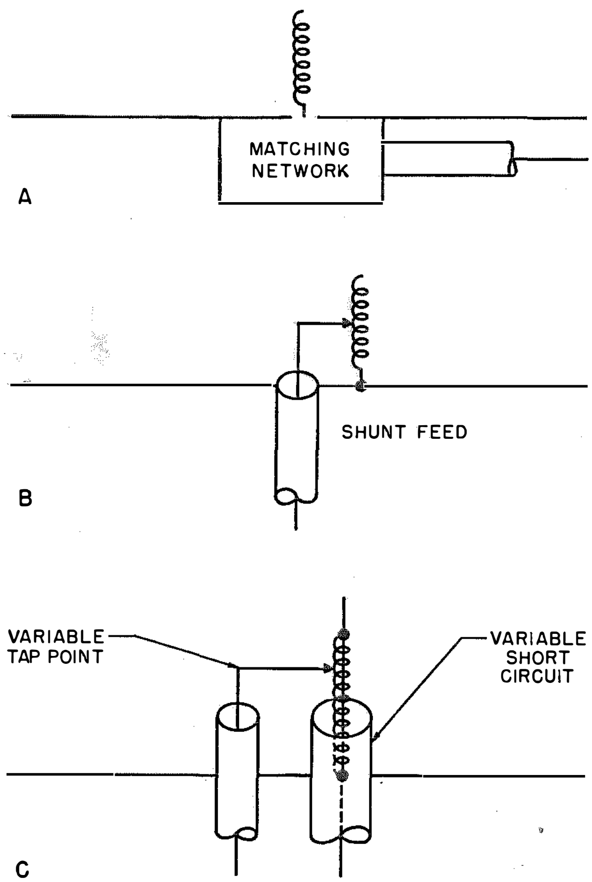

Figure 1—Helical antenna types.

The basic idea of the short resonant helical antenna is not new; for example, Wilson[2] used a helix 2 feet (61 centimeters) high with a 6-foot (183-centimeter) whip to make an efficient 4-megacycle-per-second mobile antenna. It is understood that some radio amateurs use a type similar

* Reprinted from *Convention Record of the IRE 1953 National Convention,* Part 2—Antennas and Communication, pages 42–47. Presented at the National Convention of the Institute of Radio Engineers in New York, New York, on March 23, 1953.

[1] J. F. Morrison and P. H. Smith, "Shunt-Excited Antenna," *Proceedings of the IRE,* volume 25, pages 673–696; June, 1937.

[2] W. R. Wilson, "Solenoid-Whip Aerial," *Electronics,* volume 14, pages 56–64; January, 1941.

to Wilson's or a complete helix on the 4-megacycle and 14-megacycle bands. However, the small-diameter helix does not seem to have been discussed extensively in the literature, in contrast with the work of Kraus[3] and others on the large-diameter helix.

The two general types are shown in Figure 1. Type *A* generally requires an external matching network and for some applications is less desirable than type *B*. Type *B* uses shunt feed so that the matching network is part of the radiating structure. In addition, type *B* can be made tunable over wide frequency ranges as shown in Figure 1C by using a variable tap point and a variable short circuit. The lower frequency limit is obtained when the length of wire in the helix is approximately a quarter wavelength, while the upper limit is obtained when the height is a quarter wavelength. With proper design, tuning ranges up to 100 to 1 can be obtained.

1. Velocity of Propagation

The phase velocity along the axis of the helical antenna can be obtained from the previous work on coaxial lines with helical inner conductors and also from the work on the helix for use in traveling-wave tubes.[4–6] The velocity is given by

$$(c/V)^2 = 1 + (M\lambda/\pi D)^2. \qquad (1)$$

M is a function of the helix diameter, frequency, and number of turns per unit length, and is given in Figure 2. Also shown is the ratio (approximate) of the apparent velocity along the wire to the velocity of light, V_w/c. This ratio is close to 1 for the helices used in traveling-wave tubes ($M \gg 1$), less than about 1.3 for the usual coaxial line with helical inner conductor, but may be greater than 2 for some helical antennas. The exact expression for V_w/c is

$$V_w/c = \left[\frac{1+(N\pi D)^2}{1+(M\lambda/\pi D)^2}\right]^{1/2}. \qquad (2)$$

[3] J. D. Kraus, "Antennas," McGraw-Hill Book Company, Incorporated, New York, New York, First Edition; 1950.

[4] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Company, Incorporated, New York, New York, First Edition; 1950: page 229.

[5] J. R. Pierce, "Theory of the Beam-Type Traveling-Wave Tube," *Proceedings of the IRE*, volume 35, pages 111–123; February, 1947.

[6] L. L. Chu and J. D. Jackson, "Field Theory of Traveling-Wave Tubes," *Proceedings of the IRE*, volume 36 pages 853–863; July, 1948.

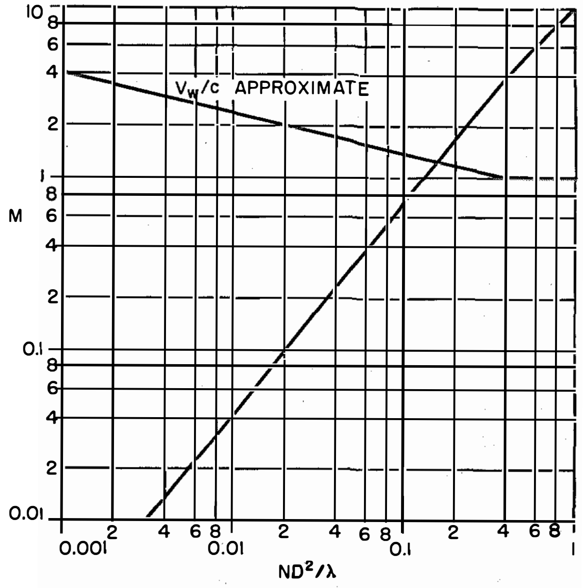

Figure 2—Relative velocity along the wire (V_w/c) and quantity *M* used in calculating the velocity along the axis.

The approximation in Figure 2 is obtained by neglecting (1) in comparison with the other factors.

When the helix is used with a ground plane, the height must be a quarter wavelength at the velocity given by (1) to obtain a real input

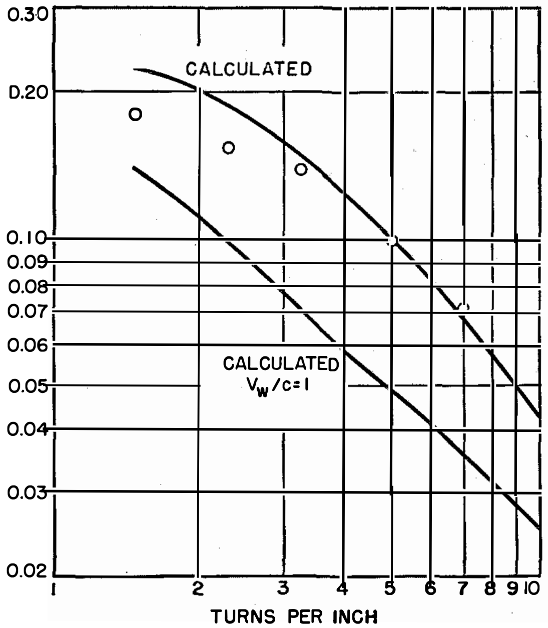

Figure 3—Helical antenna above a ground plane—height in wavelengths at resonance versus turns per inch.

impedance

$$\frac{h}{\lambda}=\frac{1}{4c/V}=\frac{1}{4[1+20(ND)^{2.5}(D/\lambda)^{1/2}]^{1/2}}. \quad (3)$$

The last expression holds when $ND^2/\lambda \leqslant 1/5$. Figure 3 shows measured and calculated values of height versus number of turns per inch. The curve near the measured points was calculated using (3), while the lower curve was calculated assuming that $V_w/c = 1$. As can be seen from the figure, the measured height is sometimes twice the height predicted using $V_w/c = 1$.

Figure 4 shows how the calculated resonant frequency varies with the number of turns per inch while holding the wire length constant. If V_w/c were a constant, each curve would be a straight line with zero slope.

2. *Radiation Resistance and Effective Height*

The effective height[7] of a short dipole above a perfect ground is $l_{\text{eff}} = h/2$. The effective height of a resonant helical antenna is $l_{\text{eff}} = 2h/\pi$ because the current distribution is sinusoidal instead of linear. The radiation resistances are given by

$$\left.\begin{aligned} R_r &= (20h/\lambda)^2 \text{ for short dipole,} \\ R_r &= (25.3h/\lambda)^2 \text{ for resonant helix.} \end{aligned}\right\} (4)$$

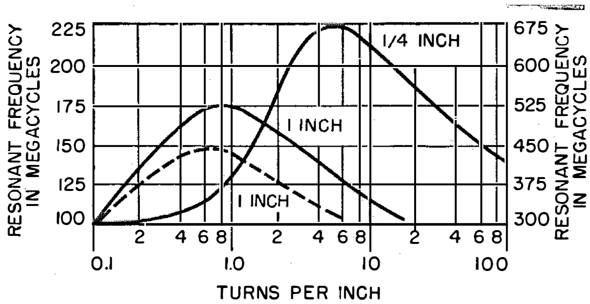

Figure 4—Resonant frequency versus turns per inch with constant wire length. Helix diameter is shown on the curves. The resonant-frequency scale for the dashed curve is at the right; for the other curves use the left scale.

For heights near a quarter wavelength, the numerical factor in the resonant helix equation should be changed to 24 in order to get 36 ohms when the height is a quarter wavelength.

[7] See, for example, S. A. Schelkunoff and H. T. Friis, "Antennas, Theory and Practice," John Wiley and Sons, Incorporated, New York, New York, First Edition; 1952: page 309.

3. *Polarization*

For a helix in free space, the ratio of the vertically polarized field to the horizontally polarized field is given by

$$\frac{E_v}{E_h}=\frac{J_0(\pi D/\lambda)}{(N\pi D)J_1(\pi D/\lambda)}\approx\frac{\lambda}{5ND^2}. \quad (5)$$

The approximate result, good for small diameters, has been given by Wheeler.[8] If the helix is resonant, circular polarization is obtained when the over-all height is about 0.9 times the diameter.

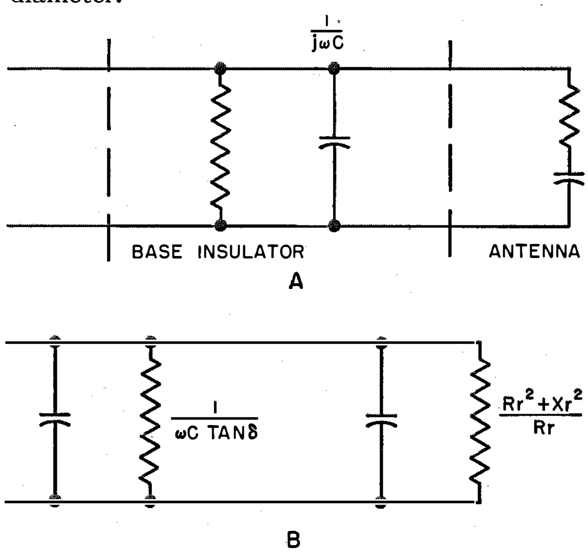

Figure 5—Circuits of a short dipole and base insulator. *A*—actual circuit. *B*—equivalent circuit.

When the helix is used with a ground plane, the horizontal polarization is weakened considerably if the height of the helix is small. For this case the pattern of the horizontally polarized field is

$$E_h' \approx E_h\ (2\pi h/\lambda) \sin\theta \cos\theta. \quad (6)$$

The maximum value of this pattern occurs at $\theta = 45$ degrees. The pattern of the vertically polarized field is

$$E_v' \approx E_v\ \pi \cos\theta. \quad (7)$$

The ratio of the horizontally polarized field at $\theta = 45$ degrees to the vertically polarized field at $\theta = 0$ degrees is

$$\left(\frac{E_h'}{E_v'}\right)_{\max} \approx \frac{5ND^2h}{\lambda^2}. \quad (8)$$

[8] H. A. Wheeler, "Helical Antenna for Circular Polarization," *Proceedings of the IRE*, volume 35, pages 1484–1488; December, 1947.

Thus, if circular polarization were obtained with a helix 0.1 wavelength high in free space, the horizontally polarized field at $\theta = 45$ degrees would be 20 decibels below the vertically polarized field at $\theta = 0$ degrees, when the helix is used with a ground plane.

4. Losses

4.1 Base-Insulator Losses

Low efficiency may be obtained with short dipoles because of power loss in the insulator used to support the dipole. The choice of dielectrics is limited because of mechanical requirements and a power factor less than 10^{-3} is probably unobtainable at present. The equivalent circuit is shown in Figure 5. For this circuit,

$$P_l/P_r = (\omega_c \tan \delta)\ (R_r^2 + X_r^2)/R_r. \quad (9)$$

For a 35-foot (10.6-meter) whip 2 inches (5 centimeters) in diameter, below about 3 megacycles, with a 40-micromicrofarad base insulator of power factor 10^{-3}, and using the following,

$$\left.\begin{aligned} R_r &= (20\ h/\lambda)^2, \\ X_r &= -\ Z_0 \cot\ (2\pi h/\lambda), \\ Z_0 &= 60\ [\ln\ (4h/d) - 1], \end{aligned}\right\} (10)$$

this becomes

$$P_l/P_r \approx 1.25/F_{\text{mc}}^3. \quad (11)$$

At 1 megacycle the efficiency is about 50 percent, while at 300 kilocycles the efficiency is about 2 percent (17 decibels loss).

In addition to the base-insulator loss, there will be further losses in the network used to match the antenna to a transmission line. In contrast, a helix 35 feet (10.6 meters) high made of 0.5-inch (12.7-millimeter) wire wound on a form one foot (30 centimeters) in diameter with 1.6 turns per inch (0.63 turns per centimeter) has a calculated total efficiency of about 5 percent (13 decibels loss) at 300 kilocycles.

4.2 Helix Losses

The ohmic losses in the metal of a short dipole are ordinarily quite small, but the losses in the helical antenna may be appreciable because the wire diameter is usually much smaller than the diameter of a dipole of the same height. The loss can be calculated[9] assuming that the current distribution is sinusoidal and neglecting proximity effects.

$$\frac{P_l}{P_r} = \frac{210^{-4}(V_w/c)}{dF_{\text{mc}}^{1/2}(h/\lambda)^2}. \quad (12)$$

This equation gives the ratio of power dissipated in the copper to the power radiated. The efficiency of the helix is therefore $1/(1 + P_l/P_r)$. Figure 6 is a plot of height versus resonant frequency for 50-percent efficiency assuming that $V_w/c = 1$.

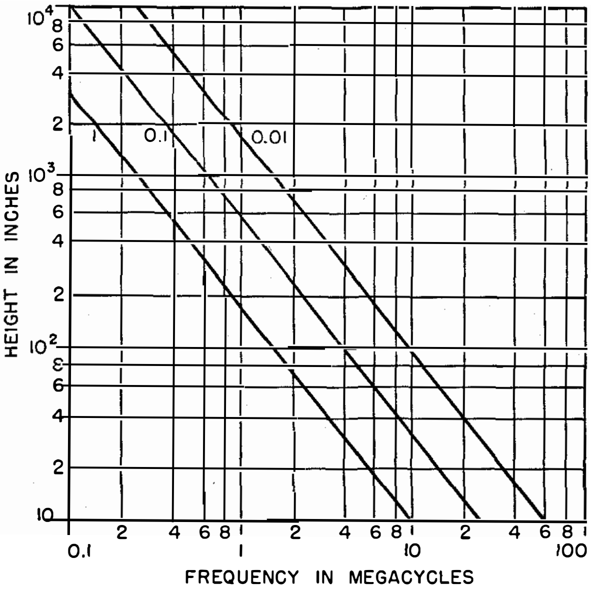

Figure 6—Height versus resonant frequency for 50-percent efficiency, assuming $V_w/c = 1$. Wire diameter is shown on the curves.

Measurements were made at 100 megacycles of the power radiated by helices of various heights and wire diameters compared to the power radiated by a matched quarter-wavelength dipole. The base of each helix was soldered to the ground plane and the inner conductor of the 50-ohm transmission line connected to the proper tap point on the helix to get a standing-wave ratio less than 1.5. The results are shown in Figure 7. The solid curves are calculated using (9) above. In most cases, the agreement is fair except for the thinnest wire, where the measured points are consistently low.

5. Q and Tap Point

The Q of an antenna is of interest because it limits the bandwidth that may be used. The Q

[9] See, for example, footnote reference 7, page 338.

of the helical antenna[10] can be calculated if some assumptions and approximations are made.

The antenna is assumed to be equivalent to a resonant line a quarter-wavelength long. The input resistance ($\lambda/4$ away from the open circuit) is[11]

$$R_{base}/Z_0 = \pi/4Q. \qquad (13)$$

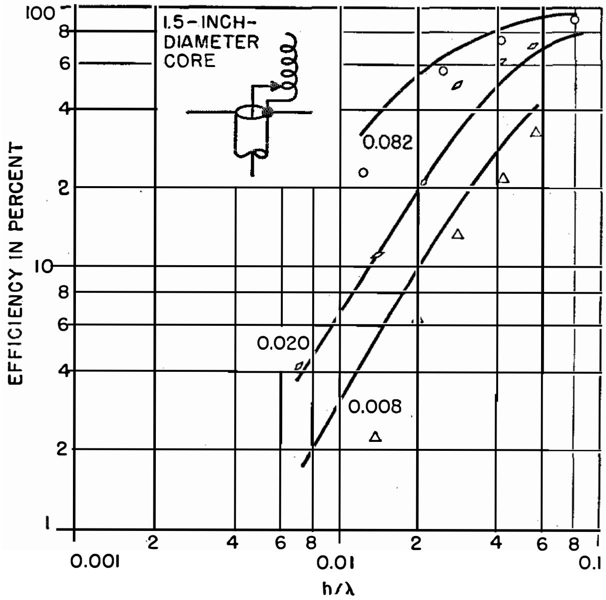

Figure 7—Measured efficiency versus height. Wire diameter is shown on the curves.

R_{base} is the sum of radiation resistance and wire resistance, referred to the base of the antenna. For a resonant antenna, $R_w = R_0\, l/2 = R_0\, (\lambda/8) \times (V_w/c)$.

$$R_{tap} = \frac{Z_0^2 \sin^2 \theta_1}{R_{base}} = \frac{4}{\pi} Q Z_0 \sin^2 \theta_1. \qquad (14)$$

The characteristic impedance that best fits the experimental data is the helical-transmission-line characteristic impedance derived previously.[12] This characteristic impedance is also given by

$$Z_0 = 60\, [\ln(4h/D) - 1]\, (c/V). \qquad (15)$$

[10] In this paper, Q means unloaded Q. When driven by a generator with zero internal resistance, the radiated power is 3 decibels below the maximum radiated power at two frequencies F_0/Q apart. When driven by a generator with internal resistance equal to the antenna resonant resistance, the radiated power is 3 decibels below the maximum radiated power at two frequencies $2F_0/Q$ apart.

[11] "Reference Data for Radio Engineers," Federal Telephone and Radio Company, New York, New York, Third Edition; 1949: page 312.

[12] W. Sichak, "Coaxial Line with Helical Inner Conductor," to be published in *Proceedings of the IRE.*

The first factor in this equation is the familiar expression that predicts the reactance of ordinary dipoles. The (c/V) factor takes account of the fact that the axial velocity of the helical antenna is less than that of the ordinary dipole.

Figure 8 shows how the Q varies with h/λ for one of the antennas used in obtaining the data for Figure 7. The agreement is fair using (13). When (14) is used to predict the tap point θ_1, fair agreement is also obtained except when the diameter is large and/or c/V is large.

6. *Circuit Applications*

Important applications of helical transmission lines are delay lines, wide-tuning-range resonant circuits, and the extension of transmission-line techniques to frequencies as low as 300 kilocycles.

A delay line $1\frac{5}{8}$ inches (4.13 centimeters) in diameter for use at 10 megacycles was built with the following characteristics: characteristic impedance, 5000 ohms; a delay of 0.2 microsecond per foot (0.66 microsecond per meter), and an attenuation of 0.9 decibel per microsecond.

Figure 9 shows the resonant frequency versus linear movement of a short-circuiting plunger for five quarter-wavelength resonators. The helical inner conductors were made with different configurations—some were wound with constant pitch, some with a constantly varying pitch, and others with sections of different pitch. One of the

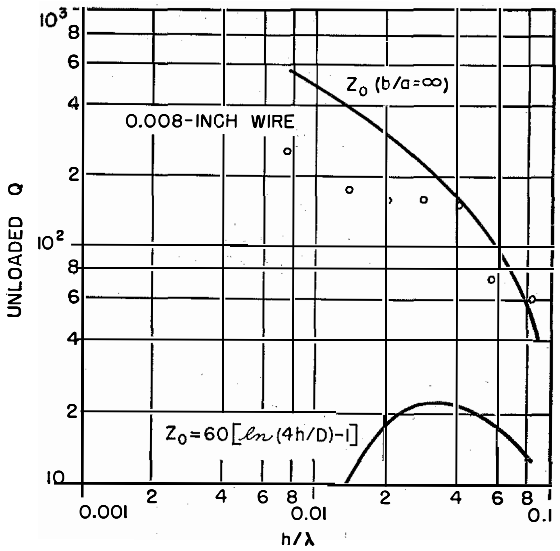

Figure 8—Measured Q versus height.

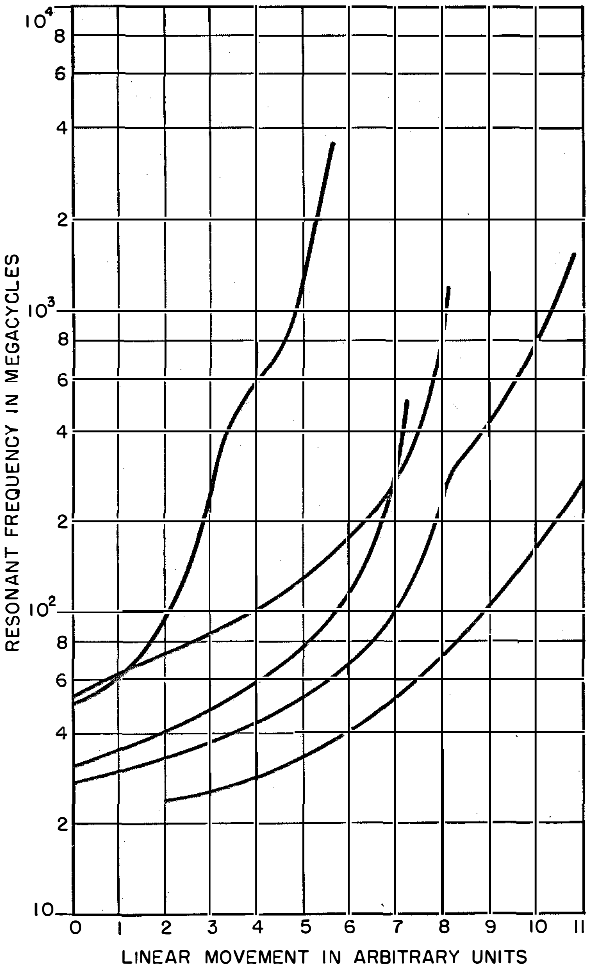

Figure 9—Resonant frequency versus plunger movement for five quarter-wavelength resonators.

resonators shown tuned from 50 megacycles to 3500 megacycles—a range of 70/1.

The tuning range is, approximately:

$$\frac{f_{\max}}{f_{\min}} = \frac{\lambda_{\max}}{\lambda_{\min}} \approx \frac{l}{h} \approx c/V. \tag{16}$$

Thus with proper design, tuning ranges of 100/1 are possible.

7. Acknowledgment

Acknowledgment is due to R. A. Felsenheld, L. Juhas, and W. Huey for making the measurements reported in this paper.

8. Glossary

c	= velocity of light
d	= wire diameter in inches
D	= mean helix diameter in inches
F_{mc}	= frequency in megacycles
h	= height
l_{eff}	= effective height
n	= number of turns per inch
P_l/P_r	= ratio of power dissipated to power radiated
R_r	= radiation resistance
R_0	= resistance per unit length
V	= phase velocity along the axis
V_w	= apparent phase velocity along the wire
ω	= $2\pi \times$ frequency
X	= reactance
Z_0	= characteristic impedance
$\tan\delta$	= power factor
λ	= wavelength
θ_1	= angular distance from short-circuited end to tap point.

Coaxial-to-Helix Transducers for Traveling-Wave Tubes*

By R. E. WHITE

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A MAJOR problem in the design and construction of traveling-wave tubes is that of the transition from waveguide or coaxial cable to the helical slow-wave structure usually employed in this type of tube. Many factors indicate the desirability of providing coaxial input and output circuits. The characteristics of an ideal transition section may be closely approximated by a new type of coaxial-to-helix matching section. Complete design data and procedures for such sections have been worked out, and transducers giving good matches over two-to-one frequency ranges have been constructed and incorporated in traveling-wave tubes. In tubes employing large-diameter helices, transition sections providing four-to-one frequency ranges have been obtained.

1. Introduction

A slow-wave structure for the propagation of radio-frequency energy is an inherent component of a traveling-wave tube. Conventionally, radio-frequency energy is carried by either coaxial cables or waveguides. Accordingly, in order to operate such a tube, a transition between the two propagating structures is always required at both the input and output of a traveling-wave tube. Since coaxial input and output connections permit utilization of the maximum available bandwidths, which are very large in traveling-wave tubes, and offer other advantages of size, weight, and simplicity of magnetic focusing, this paper will discuss only coaxial-to-helix transducers. An ideal transition from coaxial line to helix would have the following characteristics.

A. The operating bandwidths should be equal to or greater than the operating bandwidth of the traveling-wave tube.

B. The transition should add no axial length to the over-all tube structure.

C. The transition should not radiate and should be shielded from outside radiation.

D. The helix should be insulated from the tube structure and have external connections suitable for monitoring current or for handling heating currents during processing.

In practice, requirement A is often defined as a standing-wave ratio of 1.5 to 1 over a bandwidth of 2 to 1. Requirements B, C, and D, however, are taken very literally by the design engineer.

2. Design Considerations

2.1 Measurement of Impedances

The first step in the design of networks to transform from one impedance to another is to determine each of the impedances between which the transformation is desired. The circuit impedance of the coaxial line is easily defined. On the other hand, the circuit impedance of a helix is a somewhat indefinable term. The impedance of a helix as related to the interaction between the propagated wave and the electron beam has been treated by Pierce,[1] Tien,[2] and many others. However, the specific impedance of interest in the design of matching sections is that impedance existing at the end of a helix in close proximity to all portions of the mechanical structure required to support the helix and connect the coaxial line to the helix. This impedance is only vaguely related to the ratio of fields existing on a long helix and varies radically with specific physical orientations of tube components.

A series of measurements taken with the circuit shown in Figure 1 has served to determine the magnitude of the end impedance of several small helices and to indicate methods of modifying this impedance to a value more suitable for matching to a coaxial line. These measurements were made by connecting a helix directly to the center con-

* Reprinted from *Convention Record of the 1953 I.R.E. National Convention*, Part 10—Microwaves, pages 42–45. Presented at the National Convention of the Institute of Radio Engineers in New York, New York, on March 26, 1953. This work was sponsored in part by the United States Signal Corps Engineering Laboratories; Fort Monmouth, New Jersey.

[1] J. R. Pierce, "Traveling-Wave Tubes," McGraw-Hill Book Company, Incorporated, New York, New York, First Edition; 1950.

[2] P. K. Tien, an unpublished report.

ductor of a coaxial line and measuring the magnitude and the phase of the discontinuity set up at this junction. Such measurements allow one

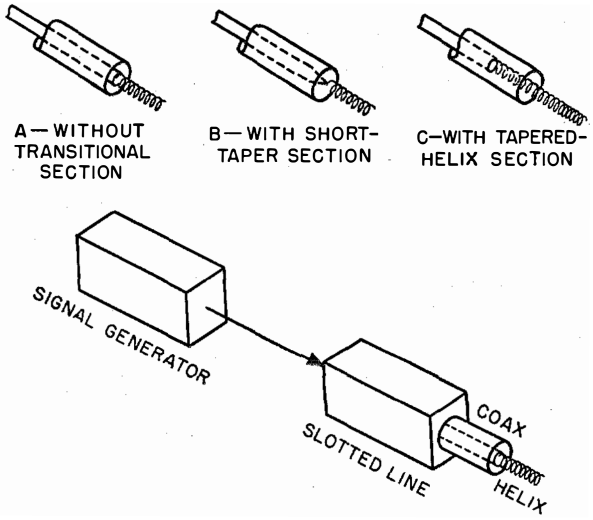

Figure 1—Test circuit for measuring helix impedance.

to plot the impedance of the helix in the particular physical configuration employed in the measurement.

When the helix was connected directly to the center conductor, as diagramed in Figure 1A, with no transition section, the transfer of radio-frequency energy to the helix was negligible. With the inner conductor tapered rapidly to the size of the helix wire as in Figure 1B, the mode transition was considerably improved. For this arrangement, the impedance plotted as curve A in Figure 2 was obtained. When the transition section shown in Figure 1C, that is, a short section of "tapered-pitch" helix, was employed, an impedance plotted as curve B in Figure 2 was measured. An examination of data obtained in this fashion indicates that over quite-wide frequency ranges the helix may be represented by a series RLC circuit with the magnitude of the resistive component being markedly increased by a tapered-helix transition section.

2.2 Calculation of Matching Circuits

Over a wide frequency range, an impedance of this kind may be transformed to a normalized value near $1 + j0$ by a combination of line transformer with a parallel transmission-line section. This transition may also be accomplished by a combination of line transformer and series transmission-line sections.

Figure 3 indicates the equivalent circuits of the forementioned types of matching sections. Here Z_h represents the end impedance of the helix, l_1 a length of transmission line of characteristic impedance Z_0 connected in series between the input terminals and the helix, and l_2 a short-circuited length of transmission line connected in shunt or in series at the junction of the input terminals and the line transformer section. If the attenuation is assumed negligible, the impedance at the junction of the parallel-line network may be

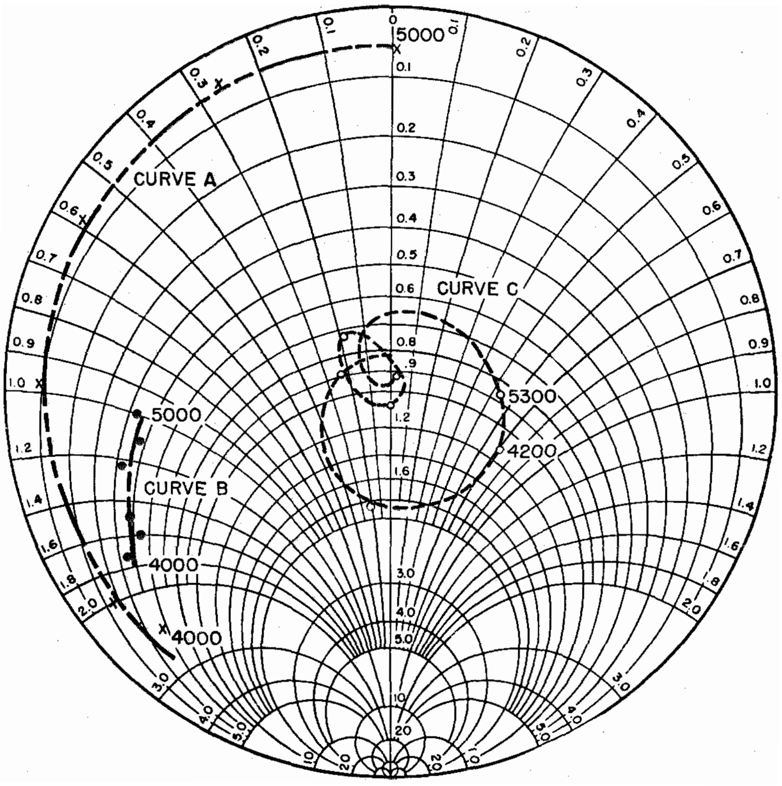

Figure 2—Impedance curves of transitional sections.

written as

$$Z_{\text{in}}=\frac{Z_1Z_2\tan\beta l_1\tan\beta l_2-jZ_hZ_2\tan\beta l_2}{\dfrac{Z_2Z_h}{Z_1}\tan\beta l_1\tan\beta l_2-Z_h-jZ_2\tan\beta l_2}.$$

Such an expression is solvable for specific choices of β, l_1, l_2, Z_1, Z_2, and Z_h only. Thus, it is not

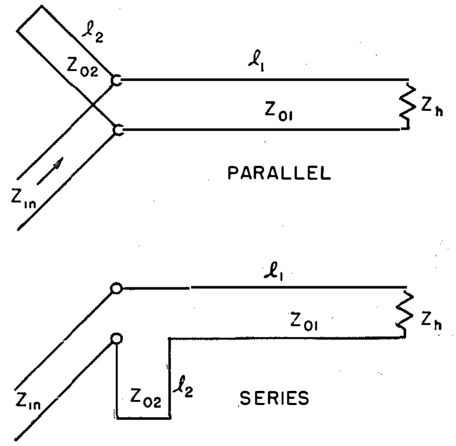

Figure 3—Equivalent circuits of matching sections.

amenable to easy use by the designer. Nevertheless, for any specific set of values, Z_{in} may be obtained in a straightforward manner by the use of a transmission-line calculator. The sequence of operation is as follows.

Using a Smith-chart calculator, one enters the chart at the helix impedance Z_h. This impedance is then rotated a distance l_1 toward the generator, and the impedance at the end of the transformer section is then read on the chart. To find the impedance of the short-circuited stub section, one enters the chart at the zero-impedance point and rotates a distance l_2 toward the generator. The impedance at the input to the stub section may then be read. A simple combination of these two impedances in parallel will give the input impedance.

The expression for a series line section with negligible attenuation may be written as

$$Z_{\text{in}}=\frac{Z_h+Z_h\tan^2\theta}{1+\left(\dfrac{Z_h}{Z_{01}}\tan\theta\right)^2}+j\frac{Z_{01}\tan\theta-\dfrac{Z_h^2}{Z_{01}}\tan\theta+Z_{02}\tan\phi+\dfrac{Z_{02}Z_h^2}{Z_{01}^2}\tan^2\theta\tan\phi}{1+\left(\dfrac{Z_h}{Z_{01}}\tan\theta\right)^2}.$$

Again, such an expression is of little use to a designer, but as before, the impedance for any given set of values of θ, ϕ, Z_{01}, Z_{02}, and Z_h is easily computed with the help of a transmission-line calculator. In this case, the impedance at the input of the line transformer section is obtained as in the previous case and the impedance of the short-circuited section is calculated as before. These impedances must then be added to form the input impedance. Through the use of the simplified calculation procedures described above, a process of successive approximations will lead to a design of a matching section for the helix impedances commonly encountered. The details of such a matching section utilized in one of the experimental tubes are shown in Figure 4.

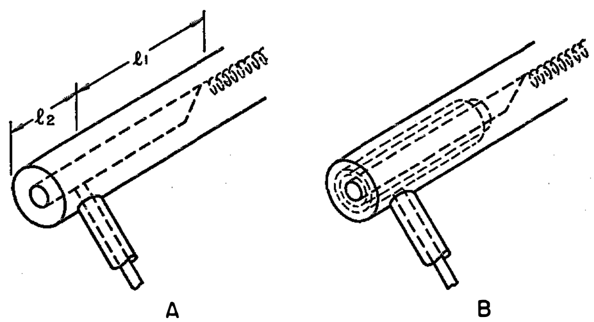

Figure 4—Grounded *A* and insulated *B* coaxial matching sections.

2.3 Design of Matching Sections

Figure 4A illustrates the equivalent construction of such a section from an electrical viewpoint. Here the helix is connected to a tubular section, hereafter referred to as the ferrule, that comprises the inner conductor of a coaxial-line transformer, l_1 in length, between the helix and the input coaxial line. A short-circuited section of transmission line, l_2 in length, is connected at the junction of the coaxial line and the line transformer section. Since isolation of the helix is desirable in order to monitor the helix current, the actual physical structure was built as shown in Figure 4B. This type of construction provides the same electrical behavior as the section diagramed in Figure 4A, but the ferrule has been insulated from the input coaxial line by a thin sheath of dielectric material that offers a very-low radio-frequency but infinite direct-current impedance. Because the base of the cylinder is at zero radio-frequency potential, a wire may be connected at this point for the purpose of monitoring the helix current. A curve of the impedance versus frequency for a section of this type is plotted as curve *C* in Figure 2.

Since a long drift tube was used as part of the noise-reducing scheme in this particular tube, the axial length of the input matching structure could be utilized without adding appreciably to the tube length; but in other types the axial length of such a matching structure would add directly to the length of the tube. To overcome this disadvantage, a structure of the type shown in Figure 5 has been developed. This structure has an equivalent circuit identical with that of the coaxial structure. However, the use of a section

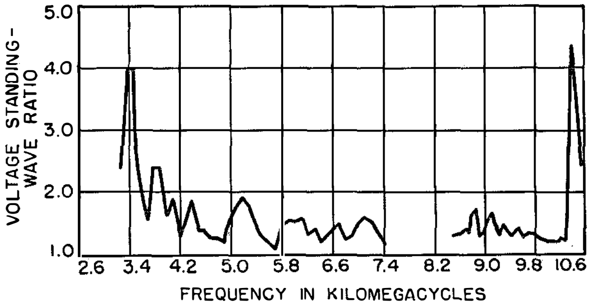

Figure 6—Matching characteristics of strip-line section.

of strip-over-ground-plane line in place of the coaxial structure has made it possible to introduce the matching structure at right angles to the helix, thus greatly reducing the axial length required.

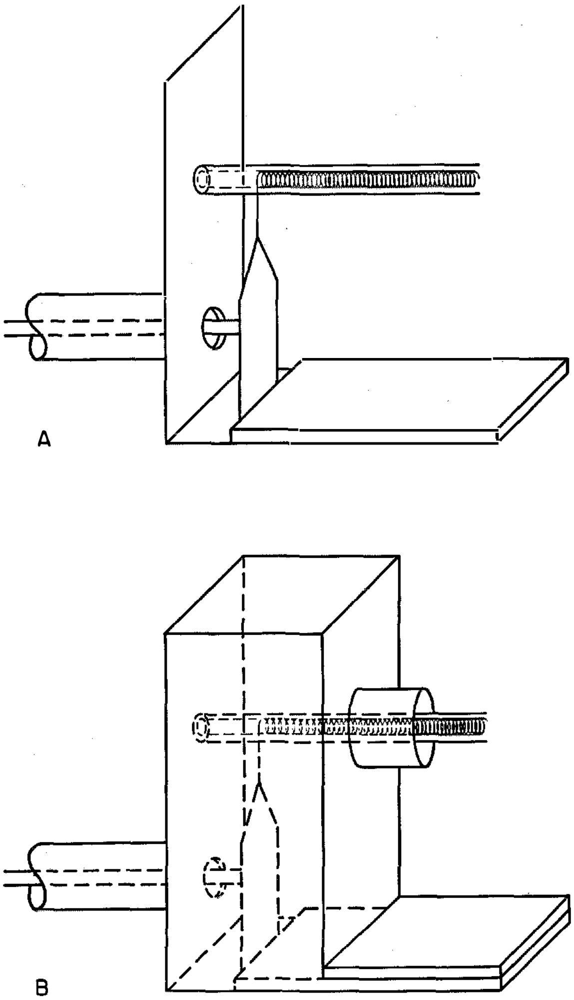

Figure 5—Unshielded *A* and shielded *B* strip-line matching sections.

The simple mechanical construction of this transition section leads to several other advantages. An impedance taper from the tap point to the helix is obtained by a smooth taper of the width of the strip line, and some variation of the impedance of the entire line may be obtained by changes in the spacing of the line over the plane. The impedance match obtained by using the transition section shown in Figure 5 is plotted as voltage standing-wave ratio versus frequency in Figure 6. Since all of the fields in a strip-over-ground-plane transmission line are not confined between the line and the plane, shielding must be provided for this type of section. A completely shielded design is diagramed in Figure 5B. This structure offers great promise and will be included in several models of experimental tubes.

Both the particular structures illustrated here have been of the parallel-line type, which is used in our present type of traveling-wave tubes because of constructional advantages. However, certain helix impedances can be best matched through the use of a series reactive element. Undoubtedly some of our future designs will employ the series type of section.

Although the matching sections outlined above are quite useful with small helices at frequencies above 1000 megacycles per second, a quite-different problem exists in the case of large-diameter helices operating at low frequencies and over bandwidths in excess of 2 to 1. Under these latter conditions, another technique, which will produce excellent matches over bandwidths of 3 or 4 to 1, may be obtained with only a slight transgression against the commandment that the transition should add no axial length to the helix structure. The matching device employed under these conditions is electrically a long taper that has, of course, excellent broadband impedance-matching characteristics. However, to avoid the large physical dimensions associated with a long taper, it is accomplished in the slow-wave structure.

An example of such a matching structure is sketched in Figure 7. The helix is surrounded by a metallic shell brought very close to the helix until each wire of the helix now acts like a line-over-ground-plane transmission line with a characteristic impedance equal to that of the incoming coaxial line. This surrounding cylinder is then flared away from the helix over a period of

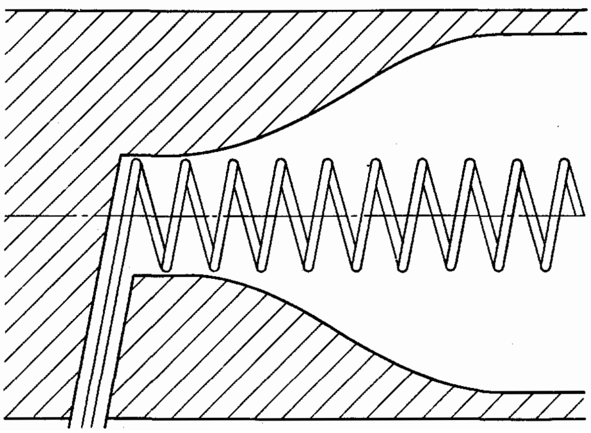

Figure 7—Tapered-shell matching section.

approximately one-half of a helix wavelength at the lowest frequency and a long electrical taper results. As indicated in the graph of Figure 8, this type of taper provides excellent matches over extremely wide frequency ranges. The limitation of this type of matching section lies in the physical dimensions and tolerances involved, but its use is recommended wherever physical dimensions allow.

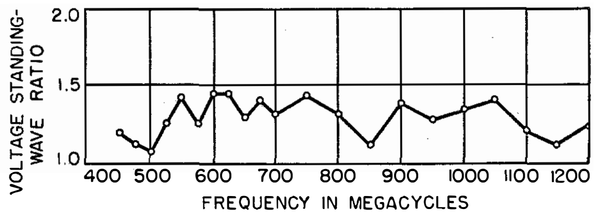

Figure 8—Matching characteristics of tapered-shell section.

3. *Design Procedure*

In summary, the design procedures that have been developed to match small helices are: *A*. Visualize a mechanical construction with the dimensions and demands of the particular tube under consideration. *B*. Measure the helix impedance in this physical configuration. *C*. Compute the combination of line transformer section with either series- or parallel-line section, to provide the best match over the frequency range under consideration. *D*. Construct the section and test to determine if the actual physical embodiment has affected the initial helix impedance. If so, a second approximation will usually suffice to produce the results predicted by the calculations. For large helices, the tapered section described above may be used with the length of the taper determined by the frequency range involved.

Telemetering for System Operation*

By R. H. DUNN and C. H. CHAMBERS

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THE PAPER outlines the requirements of power-system control engineers in respect of telemetered information to facilitate load dispatching.

The problems associated with providing these continuous readings are dealt with, and a new electronic telemeter system is described that has advantages with regard to response and accuracy, and permits several simultaneous readings over a restricted bandwidth.

Equipments of the type described have received field trials and are in commercial use.

• • •

1. *Introduction*

A modern large supply system consists of a number of generating stations that are coupled together by a transmission network to convey the generated power to suitable points for bulk supply to local distribution networks to meet the consumer demand.

The balance between generated power and consumer demand is effected by the system control engineer, who carries responsibility for continuity of supply to the consumer, transmission of power, and the efficient loading of generators. To enable the control engineer to perform these duties, he requires instrumental aids to give detailed information of the essential electrical conditions of the network.

A large network is, for convenience, divided into a number of areas or groups, each with an operation staff responsible for detailed control, coordinated by a central or master operation staff. The instrument requirements of group and master control engineers are therefore technically similar but operationally different.

The control rooms are generally established at points convenient to the communication network, but nevertheless involving distant transmission of load, voltage, and frequency readings from generating and switching stations. This distant transmission presents problems of communication as well as of measurement.

The main telemetering problem for system control is to provide sufficient information of acceptable quality with due regard to capital and maintenance costs. A recent paper[1] on load dispatching gave improved instrumentation as an essential requirement for improved system operation.

In the subsequent consideration of these instrument requirements, the term "instrumentation" is restricted to the problems associated with transmission and indication of load and frequency measurement as distinct from supervisory indication and telephone equipment, which may well justify a separate paper. The authors have no operating experience, but are concerned with the design of communication and telemetering equipment to meet the specifications and requirements of operating authorities.

2. *Instrument Requirements of the Control Engineer*

2.1 General

The purpose of instrument indications in the control room is to present the system control engineer with an up-to-the-minute picture of the magnitude and direction of power flow, together with the magnitude and trends of system load and frequency. To give some idea of the required overall picture of the system, Figure 1 shows the communication network whereby metering information is transmitted from the generation and switching stations into the group control room, and then selected information is transmitted from a number of these group centres to the master control room.

At each group control room the information on generated power may be concentrated on a loading desk or display panel, whilst information associated with magnitude and direction of power

* Reprinted from *Proceedings of the Institution of Electrical Engineers*, Part 1, volume 100, pages 39–51; March, 1953.

[1] A. R. Cooper, "Load Dispatching and the Reason for it, with Special Reference to the British Grid System," *Journal of the Institution of Electrical Engineers*, Part 2, volume 95, pp. 713–732; December, 1948.

flow in feeders may be presented on a quasi-geographic display. Associated with each of these displays are instruments that give the resultant generated load and transferred load on a group or

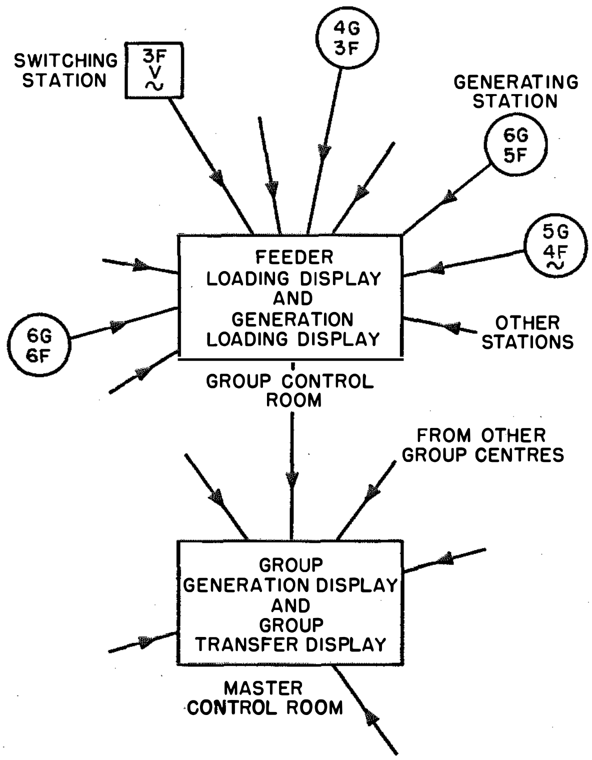


Figure 1—Typical metering network. *5G* = Number of generators. *4F* = Number of feeders. *V* = Voltage indication. ~ = Frequency indication. → = Communication channel.

area basis. In addition, there are indicators for the system voltage and frequency at selected points. These resultant loads, together with certain selected indications of load, voltage, and frequency, have to be transmitted to, and indicated in, the master control room.

The general requirements for both the group and master control rooms are summarized in Table 1. From this table it will be seen that instrument requirements fall into five main groups, some of which are required at the group or master control rooms only, whilst others are required at both points. To obtain a general idea of these instrument requirements, each of these five main groups must be considered independently, together with the manner in which the indications are to be derived and finally displayed in the control rooms.

2.2 Frequency and Time

A mains-frequency indicator is required at both group and master control rooms, and associated with this frequency indicator a rate-of-change indicator, scaled in cycles per second per minute, is required. Such an instrument is desirable for periods of system stress, and to indicate the effect of remedial action such as load shedding.

Standard time and system time are derived respectively from pendulum-controlled and synchronous-motor units, a combination of these units being required for an indication of time

TABLE 1

Typical Quantities to be Indicated in Group and Master Control Rooms

Instrument Group	Required at Group Control Rooms	Required at Master Control Room
1. Frequency and Time		
System Frequency	×	×
Rate of Change of Frequency	×	×
Time Difference	×	×
System Time	×	×
Standard Time (Greenwich)	×	×
2. Generation		
Generated Active Power (per Station)	×	
Generated Reactive Power (per Station)	×	
Generated Active Power (per Group)	×	×
Generated Reactive Power (per Group)	×	×
Generated Active Power (Total)		×
Generated Reactive Power (Total)		×
Voltage at Selected Points	×	×
3. Interconnection		
Group-Interconnection Active Power (Import/Export)	×	×
Group-Interconnection Reactive Power (Import/Export)	×	×
Net Group Transfer Active Power	×	
Net Group Transfer Reactive Power	×	
4. Feeder-Flow		
Feeder Active Power with Direction	×	
Feeder Reactive Power with Direction	×	
Feeder Current or Overload Alarm	×	
5. Emergency		
Voltage from Selected Points		×
Frequency from Selected Points		×
Feeder Active Power from Selected Points		×
Feeder Reactive Power from Selected Points		×

difference. These indications are derived locally and do not involve distant transmission.

2.3 Generation Metering

Figure 2 shows the arrangement for generation summation in each power station, the summation of power stations in each group, and the summation of groups for the master total-generation indication. The indications at the master control rooms are a copy of those in the group control rooms. The response of the telemeters should be adequate for the maximum rate of machine loading. The accuracy should be of the highest practical order, maximum accuracy being required at maximum reading.

2.4 Interconnector Metering

The magnitude and direction of power transfer in each group interconnection are required in the associated group control rooms. Figure 3 shows the arrangement of three interconnectors as indicated in one group control room, summed for net transfer and repeated at the master control room. In addition, indications are required of the magnitude and direction of reactive power in each interconnector.

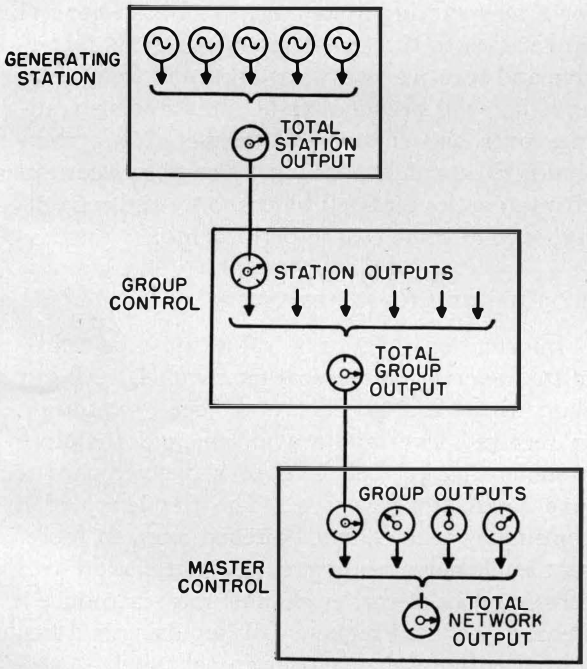


Figure 2—Summation and display for total generated output of active or reactive power.

The telemeters should be capable of following variations in average load and should have maximum accuracy at low and full-scale readings.

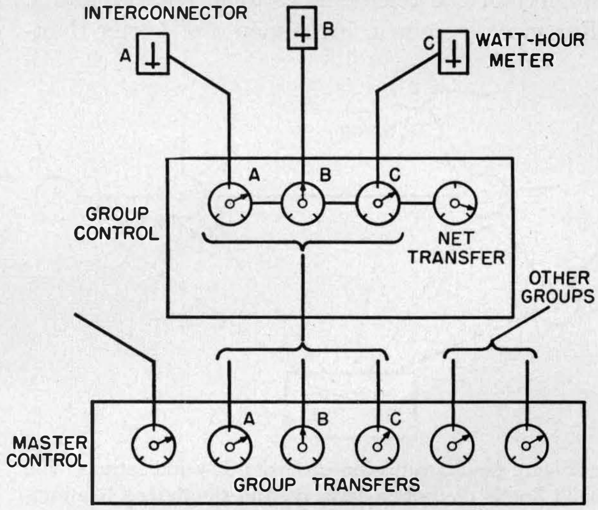


Figure 3—Interconnector and transfer metering.

2.5 Feeder–Flow Metering

In each group control room, indications are required for system supervision of magnitude and direction of active and reactive power, or current, with direction of power flow in each feeder. In addition, an audible alarm of current overload may be required to draw attention to points of excessive loading. Alternative routing of the meter indications and alarm signals is desirable, and Figure 4 shows an arrangement where the active- and reactive-power indications for any feeder are transmitted to the group control room by one route and an overload alarm by another. Thus loss of one communication channel does not leave a feeder entirely without means of indicating abnormal conditions.

2.6 Emergency Group

In addition to the total readings from the group control rooms, the master control engineer must be able to observe, on selection, the voltage and frequency at key points in the network, together with the power flow at selected points. Thus, in an emergency, close observation can be maintained on points of system stress.

2.7 Method of Presentation

A telemeter system can be much more accurate than a normal switchboard-type indicating instrument, although errors in the indicating instrument together with parallax can degrade the quality of the telemeter to unacceptable limits. For most purposes, instruments of better-than-normal accuracy with non-parallax scales are desirable, preferably of the long-scale type permitting a reading accurate to 1 per cent of full scale. Certain quantities have to be indicated and recorded. For this purpose separate indicating and recording instruments are sometimes required; each instrument is of service to the operating engineer, the indicator giving the instantaneous value and the recorder giving the trend. An instrument that would serve equally well for both requirements and could be incorporated in a diagram might meet with approval.

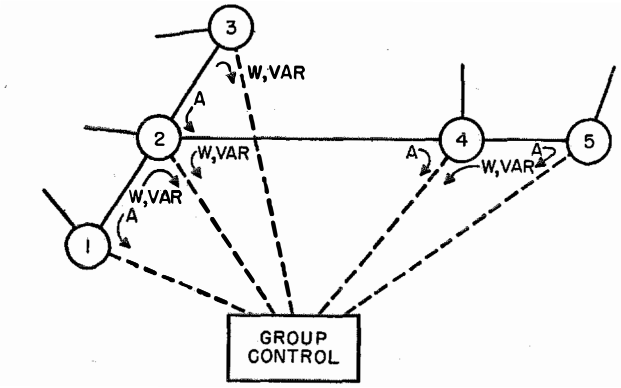


Figure 4—Transmission of feeder-flow indications. The solid line is the power network, and the dashed line indicates the communication network. A = Current indication or overload alarm. W = Watt indication. VAR = Reactive-power indication.

Indicating instruments are generally grouped according to function, or arranged in a simple geographical diagram. Figure 5 shows a suggested feeder-flow diagram incorporating the directional meters and associated alarms with switch indications.

For the convenience of the operating engineer responsible for supervising the load programme, the station generation instruments can be grouped in a desk form of panel, each instrument being in juxtaposition to the load-instruction transmitter, if provided, so that instructed and actual loads can be readily compared. Group-interconnection instruments can be associated with the feeder-flow diagram, and can possibly be repeated in proximity to the generation group. In addition, a prominent indication is required of system frequency, rate of change of frequency, standard time, and time difference.

At the master control room the inter-group transfer indications may be displayed in a single-line diagram showing the interconnections of groups, and thus facilitating the observation of load-transfer programmes. Group-generation meters may be arranged on one panel, and the total summation may be displayed in a prominent position, with system frequency, rate of change of frequency, standard time, and time difference. The emergency group of meters with associated selection controls may be incorporated in a desk, to permit close observation.

3. *Essential Characteristics of a Telemetering System*

3.1 General

The foregoing requirements result in the need for a telemetering system that can have general application to the transmission of signals for active and reactive power, voltage, and frequency, and that will permit several simultaneous readings over one channel. A telemetering system that is to be widely used must incorporate certain characteristics dictated by economy and reliability, even at some cost in performance.

3.2 Channel Requirements

Information defining a value must be transmitted over a communication channel. All but short-distance channels are subject to variation in received level and waveform, and therefore systems that rely on level or waveform cannot give satisfactory service. The frequency of a continuous tone cannot be relied upon to represent a telemeter reading, since transmission over certain types of carrier channel may introduce a frequency shift. Frequency of impulsing is transmitted without change over any channel.

Channels designed for impulse transmission are available for renting in many countries, and such channels can be obtained by well-established means over direct wires, speech, carrier, and radio circuits. Impulse frequency is thus available as a means of transmitting information without channel error, and with unrestricted choice of channel.

3.3 Bandwidth Requirements

A speech circuit transmitting the normal band of speech frequencies will yield a number of impulsing channels by frequency division.[2] This technique is in general use to provide 12 or 18 telegraph channels on one pair of wires, each employing a particular frequency within the audible band. Telemetering systems occupying one such channel for each indication have been used in the past.

The rental or provision of long-distance circuits represents a major item of expense for a control system, and if any appreciable number of meter readings is to be provided, it is essential to transmit the information for them in the minimum number of channels. In general, one or at most two such channels may be used, since the remainder of the frequency spectrum is required for speech communication and other signals. The adoption of impulsing at a slow rate enables a number of telemeters to share one voice-frequency channel by means of time-division technique.

3.4 Continuous Operation

A telemeter should be continuously dependent on the incoming signal rather than intermittently adjusted, whether automatically by changes in the value, or periodically in time. A reading maintained by continuous signalling gives an impression of a live instrument whilst indicating, and on the interruption of signalling it registers zero. An intermittently adjusted reading must be locked to each value until the next is received, thus giving a dead quality.

3.5 Flexibility

The basic telemeter should be adaptable to a variety of different readings such as power, current, voltage, frequency, etc., to enable similar equipment and technique to be used throughout a system, thus simplifying manufacture and maintenance.

3.6 Stability

A telemeter, once installed, should not require frequent adjustment or tedious recalibration. An electronic system should not be dependent on valve characteristics that change with time, or on power-supply voltages and frequency that are subject to variation. The system should not rely on highly-stabilized main power supplies, since such supplies are expensive in initial cost and add to the work of maintenance. Electronic circuits can be designed to restrict changes due to these factors to a negligible value, and such designs should be used.

3.7 Past Practice

Telemetering systems available in the past have been mainly of types where a primary element controls an electrical output suitable for remote transmission, or have been devices for transmitting a pointer position. Most of these are unsuitable for operation over a communication circuit with repeaters, although a few employ voice-frequency transmission and have been used for limited applications, such as the variable voice-frequency telemeter[3] and the photo-telemeter.[4] However, these are unsuitable for summation due to the lack of linearity, and do not

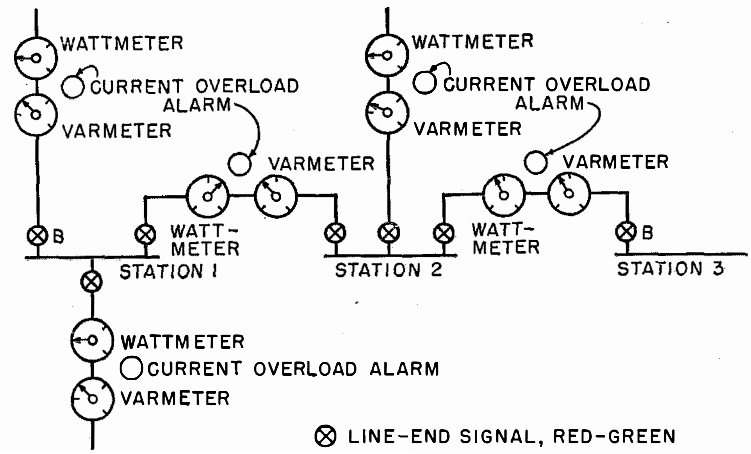


Figure 5—Part of suggested feeder-flow diagram.

[2] B. S. Cohen, "A Handbook of Telecommunications," First Edition, Sir Isaac Pitman and Sons, Limited, London, England; 1946: pages 353–355.

[3] C. G. White, "Control Station Apparatus," *Electrical Communication,* volume 25, pages 43–49; March, 1948: also, *British Engineering Export Journal,* volume 28, pages 666–671 and 784–788; March and April, 1946.

[4] G. A. Burns and T. R. Rayner, "Remote Control of Power Networks," *Journal of the Institution of Electrical Engineers,* volume 79, pages 95–125; July, 1936: see page 112.

readily lend themselves to multi-channel working on a restricted bandwidth.

As distinct from conversion systems changing to an electrical quantity, pulse-count systems[5, 6] are in use, where pulses generated by watt–hour meters are integrated over periods of minutes and the resultant is displayed in terms of watts on a conventional indicating instrument. Such systems are used for generation and feeder-flow metering.

4. Impulse-Frequency Telemetering System

4.1 General

The frequency of impulses derived from the rotation of an integrating meter is a measure of the instantaneous value being integrated, so that a watt–hour meter will yield a measure of power. The pulse-count system was an attempt to use this principle, the impulsing frequency being measured by a count over a definite time interval.

Experience suggested that this method did not use the information to the best advantage, and that a system measuring average impulse frequency directly would be more efficient. Elec-

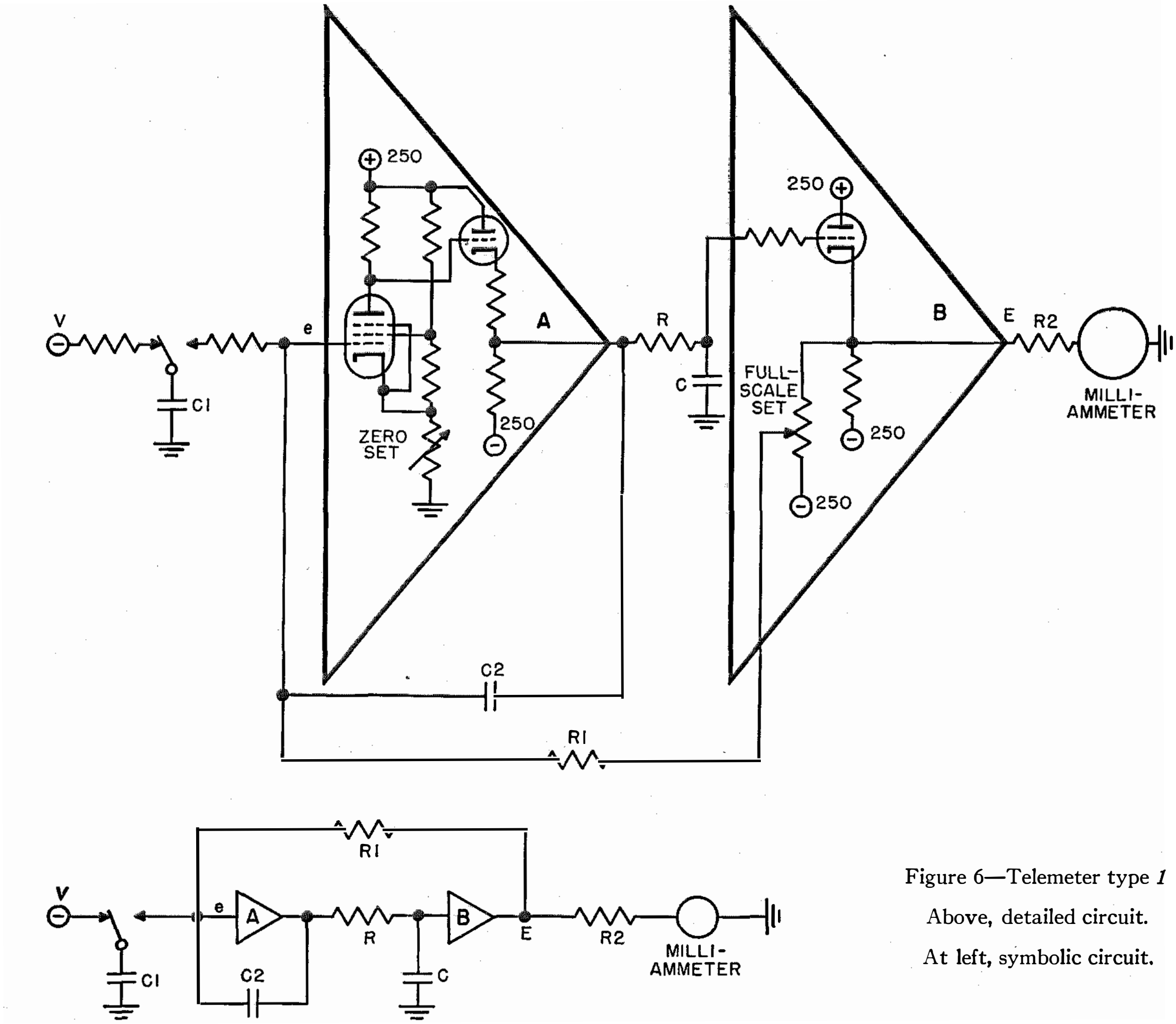


Figure 6—Telemeter type *1*
Above, detailed circuit.
At left, symbolic circuit.

[5] E. M. S. McWhirter and C. H. Chambers, British Patent 553 433; May 21, 1943.

[6] T. R. Rayner and G. A. Burns, British Patent 553 503; May 25, 1943.

tronic circuits to do this were evolved and they showed that for comparable performance the improved method required an impulse frequency of about one-third of the value previously required. Alternatively, for a given impulse frequency, the performance of the system is appreciably improved.

The integrating meter provides an accurate means of interpreting active and reactive power as an impulse frequency. The meter can be adapted to handle voltage and current, and a special instrument has been designed to deal with system frequency. An electronic device that will translate a direct voltage into a corresponding impulse frequency is also available for use as a transmitter for water-level and similar position devices.

4.2 Circuit Principles

4.2.1 Telemeter for General Application

The circuit is shown schematically in Figure 6, and uses electronic techniques.[7] A capacitor *C1* is charged to a potential $-V$ and discharged into capacitor *C2* at each operation of the impulsing relay. An amplifier A, of high voltage gain, connected through a time-constant network *CR* to an amplifier B (a cathode-follower) produces a voltage E at the output of amplifier B as a result of the average change in potential of *C2*. This voltage causes current to flow through *R1* into *C2*.

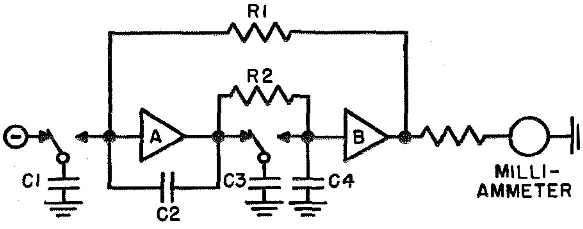


Figure 7—Telemeter type *2*.

Successive operations of the relay may be considered to produce a current from *C1* into *C2* that is balanced by the current through *R1* for a value of E that represents the frequency of impulsing f. In the following analysis of operation, e is the voltage change at the input of the amplifier having a voltage gain G, when an output voltage E is developed.

[7] E. M. S. McWhirter and R. H. Dunn, British Patent 629 244; September 15, 1949.

Input current from $C1 = C_1 f(-V-e)$.

Feedback current through $R1 = (E - e)/R_1$, and $e = -E/G$.

The sum of the currents at the amplifier input must be zero. Then

$$C_1 f\left(-V + \frac{E}{G}\right) + \frac{1}{R_1}\left(E + \frac{E}{G}\right) = 0.$$

Thus

$$f = \frac{E}{C_1 R_1}\left(\frac{1 + 1/G}{V - E/G}\right)$$

$$= \frac{E}{VC_1R_1}\left(1 + \frac{1}{G}\right)\left\{1 + \left(\frac{E}{VG}\right) + \left(\frac{E}{VG}\right)^2 + \ldots\right\}.$$

This expression gives a relation between f and E that tends to become linear as G is increased.

This may be considered as a linear system in which $f = E/VC_1R_1$, and the amplification is made high enough to bring the errors within the required limit. The first-order error term is $(1/G)(1 + E/V)$, and with normal values of G greater than 300 and $E/V = 0{\cdot}5$, the error is less than 0·5 per cent. The amplification can fall appreciably before the error increases to significant values, e.g. an error of 2 per cent is obtained when the amplification falls to 75.

The amplifier A is operated at very low ratings, which ensures an extremely long life of acceptable operation.

4.2.2 Telemeter for Interconnector Loads

The operational requirements for interconnector-load readings necessitate that readings near zero shall be steady so that the floating condition can be seen, and so that sudden changes in load, such as load trip or sudden overload, will be apparent. A modification to the circuit enables these conditions to be met and provides a telemeter of improved performance[8] where transmission of impulses does not introduce appreciable distortion.

The circuit, shown in Figure 7, differs from the original circuit of Figure 6 in that an additional contact of the impulsing relay switches a capacitor *C3* from one terminal to the other of the resistor *R2*. To obtain steady readings at low impulse frequency, *R2* is made large to increase the time-constant *CR*, and the capacitor *C3* has

[8] R. H. Dunn and P. K. Dyer, British Patent 667 989; March 12, 1952.

no effect in a steady state with no potential difference across *R2*. During a change in reading, a potential appears across *R2* as *C4* is charging to a new value, and *C3* by carrying charges to *C4*

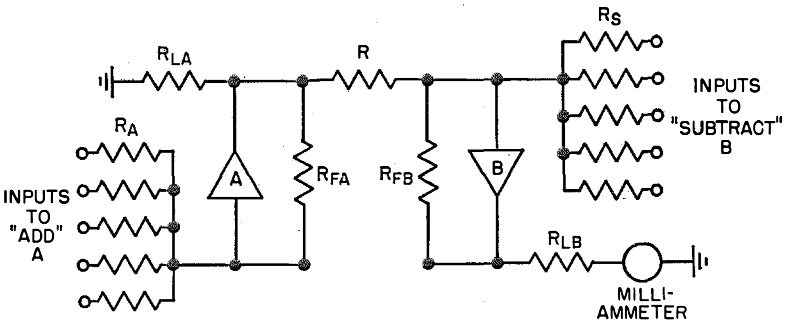


Figure 8—Summation amplifier.

effectively shunts *R2* and reduces the time-constant. This permits the circuit to operate with high time-constant during steady conditions and low time-constant during changing conditions.

4.2.3 Summator

The telemeters generate a voltage proportional to the value metered, and the values may be added by using a high-gain amplifier to balance a total input current obtained from the various telemeter outputs against a current generated in the summator.

The arrangement is shown in diagrammatic form in Figure 8; the amplifier A, with positive potential on any "input A" terminal, generates a negative potential across R_{LA} that feeds current through R_{FA} to balance the input current. The potential developed by the amplifier A is also applied to the resistor R as an input to a similar amplifier B, so that a positive potential is generated across R_{LB} to operate the indicating instrument. Amplifier B restores the polarity reversal of amplifier A, so that the summator gives a positive output for positive input.

Resistors R_S are provided at input B, so that positive potential on input B will subtract from positive potential on input A.

The summator operates on the same principle of current balance as the telemeter and its error term is the same, i.e. $(1/G)(1 + E/V)$. The amplification is made higher, and for typical values of $G = 1000$, and $E/V = 1$, the overall error is 0·2 per cent.

The arrangement permits all telemeters to generate the same output potential and all instruments to be of the same sensitivity but scaled to suit the primary value. Resistances R_A and R_S are made proportional to the inverse of the scale value to obtain the correct sum.

4.2.4 Impulse Generator

The telemeter uses a circuit in which the impulse frequency is arranged to control an input current balanced by a current generated in the circuit. This action may be reversed so that a steady current applied from an external source is balanced by a current due to an impulse frequency generated within the circuit.[9]

The arrangement is shown schematically in Figure 9. An amplifier A of high voltage gain follows the potential of the capacitor *C2* that is being charged by current through *R1*. An amplifier B passes from a conducting to a non-conducting state at a certain voltage and releases the oscillator that operates the relay. The operation of the relay transfers a charge from *C1* to *C2* to cause amplifier B to stop the oscillator. In this way, impulses are generated at intervals to balance the applied current, and the impulse frequency generated is proportional to the applied current.

The oscillator is arranged to have a free-running frequency rather higher than the maximum normally required. This circuit has a fundamental difference from the telemeter in that the telemeter output can become positive or negative and zero can be defined, whereas the impulses generated cannot change sense and zero is indeterminate. The

[9] R. H. Dunn, British Patent 629 254; September 15, 1949.

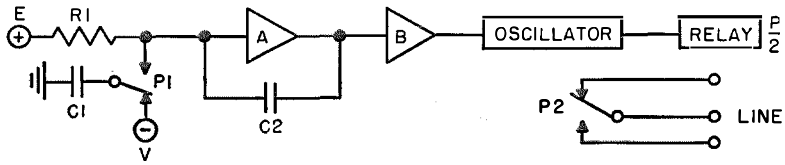


Figure 9—Impulse generator.

circuit is therefore arranged with a small fixed input that gives a minimum frequency at zero input current, and compensation is provided in the associated telemeter.

In operation this circuit makes a voltage excursion between fixed limits independent of the applied current. The telemeter, on the other hand, makes a voltage excursion about a mean potential that changes with the reading. The impulse generator then has no change of input potential due to the reading but a change of input potential that varies during each impulse and acts as an average constant potential modifying the applied potentials. The normal adjustment of full scale and zero eliminates its effect.

The circuit may be analysed as follows:

Input current from source $= E/R_1$,
Input current from zero setting $= V_0/R_0$,
Feedback current $= -fC_1V$.

For zero setting, $E = 0$ and $(V_0/R_0) - f_0C_1V = 0$. Then

$$f_0 = V_0/R_0C_1V.$$

For any other value of E,

$$\frac{E}{R_1} + \frac{V_0}{R_0} - fC_1V = 0,$$

$$\begin{aligned} E &= R_1(fC_1V - V_0/R_0) \\ &= R_1C_1V(f - V_0/R_0C_1V) \\ &= R_1C_1V(f - f_0). \end{aligned}$$

Thus the characteristic is linear between f_0 and f and there is no error term attributable to the normal action of the circuit.

4.2.5 *Differentiator*

The rate of change of a telemetered value can be obtained by means of a feedback amplifier and is a useful facility. The arrangement is shown in Figure 10. A changing direct potential E must modify the charge in C through the circuit C, $R1$, $R2$, and this circuit has a time-constant $C(R_1 + R_2)$. The potential across the resistance is then equal to the change in E over a time equal to $C(R_1 + R_2)$. Part of this potential operates the amplifier to generate an output voltage that operates the indicating instrument and provides feedback over R_F and r to maintain the amplifier input potential at the small value required. Choice of resistor values modifies the sensitivity of the circuit.

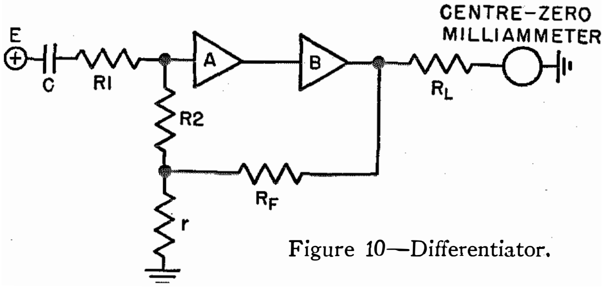


Figure 10—Differentiator.

4.2.6 *Accelerator*

If a power station is equipped to provide impulses for telemeters for group control, a high-accuracy indication of the load is then available to the station supervisor. Moreover, it is desirable that the station supervisor's instrument should give the same indication as the distant loading-engineer's instrument. A telemeter may be fitted for this purpose, and this circuit will also operate slave instruments in other parts of the building. For some purposes, such as boiler-house indication, a faster response is required. This can be done where it is possible to obtain a summed current that represents the station load, and although it is less accurate than a reading derived from the impulse source, it may be used to provide a potential during load changes. This is effected by an arrangement shown in Figure 11 in which a current transformer T and a rectifier M provide a direct voltage proportional to the load current. This voltage is differentiated by C,R, and integrated by $C1,R1$, and the amplifier A passes a potential to the amplifier B that produces a current related to the rate of change and has a time characteristic that will tend to

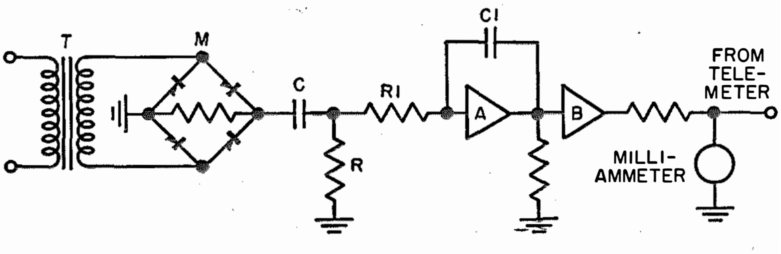


Figure 11—Accelerator.

compensate for the characteristic of the telemeter. The output of this circuit is fed into the indicating instrument in parallel with the telemeter current, and adds during increasing load and subtracts during decreasing load. Curves of the response are shown in Figure 12, from which it will be seen that good compensation has been effected. The compensation can be made as complete as desired by the application of these principles, but the circuits become complicated and for practical purposes the compensation obtained with the simple circuit used is adequate.

4.3 Multiplex Working

4.3.1 General

To achieve economy in the use of channel space, a tone in the audio range may be used to transmit impulses that, in turn, may be used to serve a number of telemeters. This method is practicable when the voice-frequency channel will transmit impulses at a frequency much higher than those of the telemeter impulses, and the voice-frequency bandwidth is governed by the impulse frequency it has to accommodate. In practice a telemeter impulsing at up to $66\frac{2}{3}$ impulses per minute can be accommodated with tolerable distortion on a 50-baud* telegraph channel that normally occupies a bandwidth of 120 cycles per second. (The actual figure of $66\frac{2}{3}$ impulses per minute is due to established metering practice, e.g. a load of 100 megawatts on a meter operating at 25 units per impulse gives an impulse frequency of $66\frac{2}{3}$ impulses per minute.)

Traffic on a channel when used in this way is continuous, and electronic equipment is necessary to provide satisfactory operation at the speed required and to eliminate the wear and tear that electro-mechanical apparatus would suffer. Recent developments in cold-cathode tubes have made available a choice of tubes of high reliability and long life, which enables the electronic-circuit problems to be solved and provides equipment with adequate operating margins for commercial use.

* The signals of the standard telegraph alphabet are composed of significant elements with theoretical duration equal to, or multiples of, the duration of a "unit time interval" determining the rate of formation of the signal.

The rate of formation of the signal shall be called "modulation rate" or "telegraph speed" and shall be measured in "bauds" by the reciprocal of the duration in seconds of the unit interval. (Recommendation 301, sixth meeting of the Comité Consultatif International Télégraphique; May, 1948.)

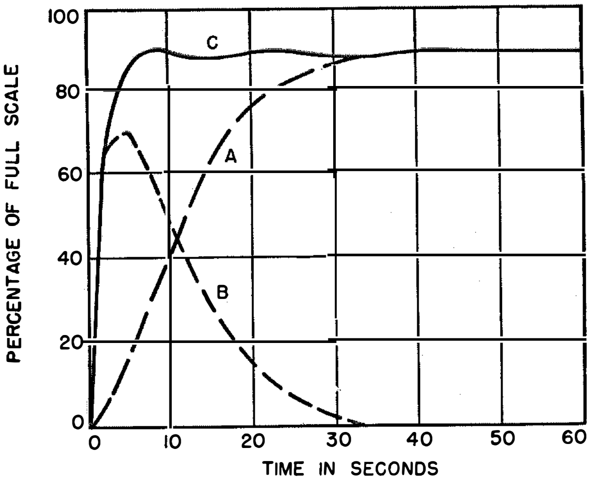


Figure 12—Accelerator response curve. A = Telemeter normal response, $66\frac{2}{3}$ impulses per minute. B = Accelerator output. C = Telemeter response with accelerator.

4.3.2 Teleprinter Coding

The teleprinter uses a 5-unit binary combination to yield 32 codes[10] and operates at a signalling speed of 50 bauds. Each signal includes a start and a stop unit making 7 units in all occupying 140 milliseconds, and therefore 7 signals per second can be transmitted. Each meter of a group

[10] E. Missen, "Elementary Telegraphy," First Edition, George Newnes, Limited, London, England; 1946; pages 14–19.

TABLE 2

Variation of Number of Meters per Channel with Pulse Rate

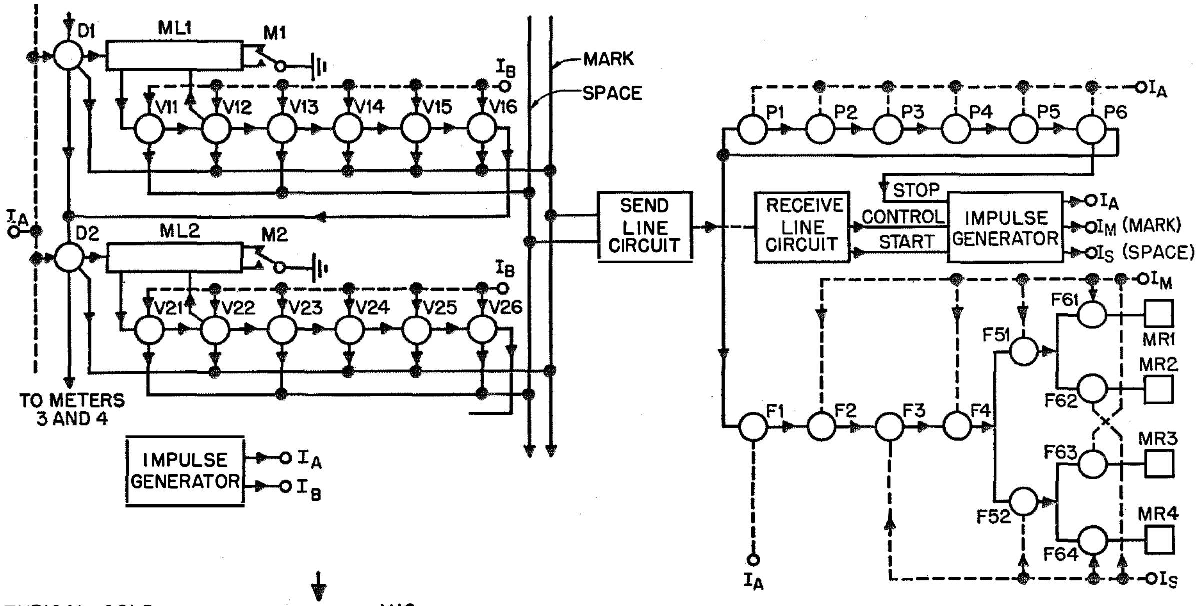
Pulse Rate in Impulses per Minute	Number of Meters per Channel
$66\frac{2}{3}$	6
$33\frac{1}{3}$	12
25	17
20	21

TABLE 3

Pulse Rates and Number of Related Meters per Channel (e.g. Active and Reactive Power)

Pulse Rate in Impulses per Minute	Pairs of Meters per Channel	Meters per Channel
$66\frac{2}{3}$	4	8
$33\frac{1}{3}$	9	18
25	12	24
20	15	30

can be allotted a different code, and a common sender can be arranged to originate the appropriate code each time a meter generates an impulse. The receiving equipment can be arranged maximum rates cannot be obtained simultaneously and the maximum combined impulse rate occurs at a phase angle of 45 degrees when each meter runs at 0·7 of the full rate. Under these



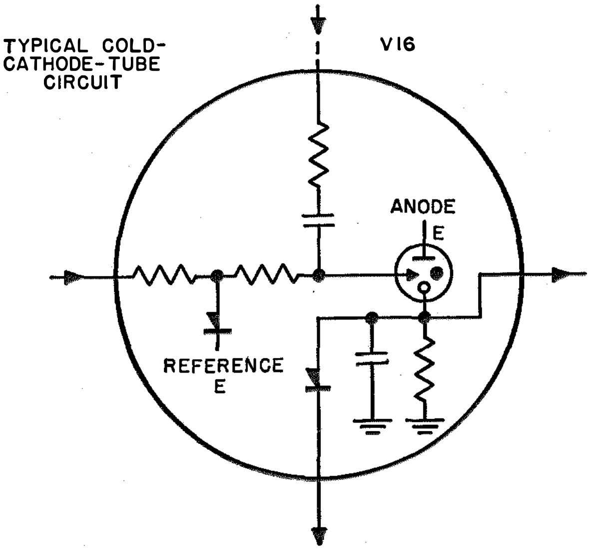


Figure 13—Multiplex teleprinter coder.

to respond to the codes and to operate the corresponding telemeters accordingly.

The number of telemeters that can be accommodated by such a system depends on the total number of impulses that must be transmitted. The channel can transmit a total of 428 impulses per minute, and therefore can carry independent meters as shown in Table 2.

Where meters are related, e.g. active and reactive power on the same circuit, the individual

conditions a channel will carry meters as shown in Table 3.

Where readings of meters of mixed rates are to be transmitted, the sum of the individual simultaneous maximum rates must not exceed 428 per minute, e.g.

2 pairs of meters for active and reactive power at $66\frac{2}{3}$ impulses per minute each	$= 186\frac{2}{3}$
2 meters independent at $66\frac{2}{3}$ impulses per minute each	$= 133\frac{1}{3}$
1 meter independent at 25 impulses per minute	= 25
3 meters independent at 20 impulses per minute	= 60
	405

Where a telemeter is required to read an import/export load, two codes can be alloted and the telemeter receives continuous indication of direction. Since only one code of the pair is in use at a time, the number of meters accommodated on a channel is not reduced on this account.

The arrangement of the circuit for teleprinter-code multiplex working is shown diagrammatically in Figure 13. Each meter is associated with a meter-link circuit *ML1*, *ML2*, etc., that is operated by a meter impulse and stores the signal awaiting transmission. A distributor, comprising a closed ring of cold-cathode tubes *D1*, *D2*, etc., energized in turn by continuous impulses I_A, is arranged to initiate operation of the associated coder group on coincident operation of a tube *D* with the corresponding circuit *ML*. Thus with *ML1* storing an impulse and *D1* operating in due course, the coder tube *V11* is prepared. The coder group comprises an arrangement of cold-cathode tubes *V11* to *V16*, *V21* to *V26*, etc., connected to be energized in succession by impulses I_B and started when *V11*, *V21*, etc., is prepared. The effect of operation of a coder group is to divert the distributor sequence via the code group; e.g., with *ML1* operated, the normal sequence *D1–D2–D3*, etc. becomes the sequence *D1–V11–V12*. . . *V16–D2–D3*, etc. During operation of a coder group the corresponding *ML* circuit is reset, and a 7-unit signal coded in accordance with connections from the tubes in the group is transmitted to line.

At the receiving point the line circuit responds to the incoming signal and starts an impulse generator that produces impulses to step a group of cold-cathode tubes *P1* to *P6* which stops the impulse generator after six impulses have been delivered. During the sequence of impulses the line circuit acts to repeat impulses on wire I_M or I_S, according to the coding of the signal. These keyed impulses operate a "fan" of tubes, *F1*, *F2*, etc., with the outlet at *F6* in such a way that a marker circuit *MR1*, *MR2*, etc., is operated only if the coding corresponds to that circuit. The diagram shows a fan with four outlets operated by codes

S M S M M M
S M S M M S
S M S M S M
S M S M S S

respectively. Any other code will fail to operate the fan to an outlet. The marker circuit *MR1*, *MR2*, etc., includes relays to provide impulses to operate the telemeter relays.

The operation of these circuits performs electrically the coding and decoding functions of a teleprinter. The coded signals may be initiated from a teleprinter keyboard and will operate the receiver, whilst the sender will operate teleprinter machine that will print the corresponding characters. This is a useful feature for checking the terminal equipment and line transmission during maintenance.

4.3.3 Distributor

A transmission channel may be shared between a number of impulsing meters by connecting the channel to each meter in turn in a continuously operating sequence. The cyclic time of the system must be such that a number of cycles elapse during a meter impulse; the greater the number the less the distortion of the meter impulses. During each cycle a signal must be transmitted to the line for each meter connected, and the number of meters that can share a channel depends on the ratio of channel/meter impulsing frequencies. This system requires switching devices at both terminals maintained in synchronism, and the principle of the start–stop teleprinter may be used.

A telegraph channel operating at 50 bauds will transmit ten code elements together with start-stop elements in a cycle time of approximately 0·25 second that will define meter impulsing at $66\frac{2}{3}$ impulses per minute with adequate fidelity. The equipment must operate continuously, and circuits using cold-cathode tubes will operate at the speeds required and eliminate wear and tear.

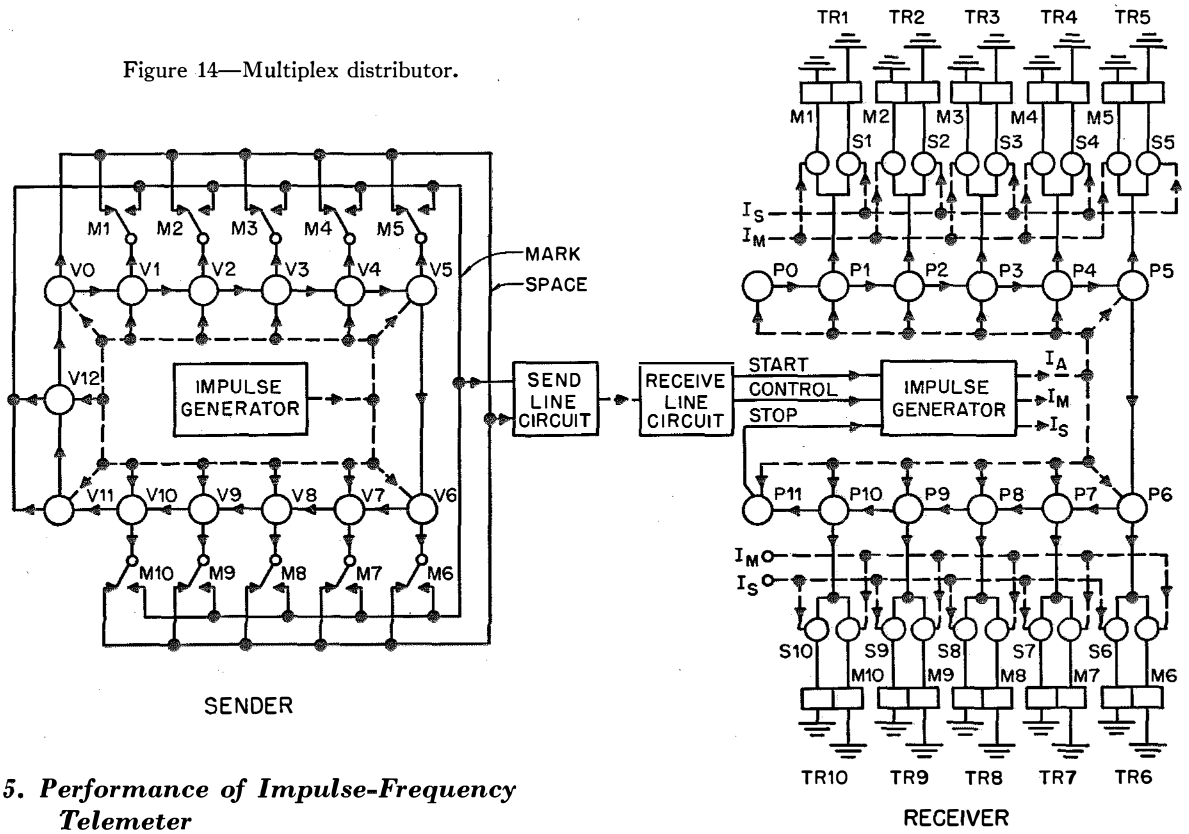
Meters giving an import/export indication require two positions of the signal elements, thus reducing the capacity of the system. Alternatively, the direction indication may be transmitted over another system, for example one used for circuit-breaker indication.

The circuit arrangement is shown in schematic form in Figure 14. The sender comprises a closed counting ring of 13 cold-cathode tubes, *V0–V12*, operated in sequence by continuous impulses. Ten of the tubes, *V1–V10*, are connected over contacts of meter relays *M1–M10* to control the line circuit to cause "mark" or "space" to be transmitted to line according to the position of the relay contact. The tube *V0* is connected to send "space" and tubes *V11–V12* to send "mark."

The receiver line circuit operates a start–stop impulse generator to start on a space signal and deliver impulses at I_A that step a circuit of 12 cold–cathode tubes *P0–P11*. Tube *P11* stops the impulse generator and the next space signal restarts it, the interval being nominally one signa unit—the 13th unit of the sender. The line circuit also controls the impulse generator to deliver impulses I_M or I_S corresponding to "mark" or "space" received from line. Tubes *P1–P10* prepare pairs of marker tubes *M1, S1–M10, S10*, one of each pair being operated by I_M or I_S, depending on the corresponding signal element, and the marker tubes control relays *TR1–TR10*.

In this way the sender examines the relay contacts, transmitting signals corresponding to the contact positions, whilst the receiver operates relays in accordance with the signals. The receiver relays then repeat the condition of the sender contacts. Auxiliary circuits can be included to give an alarm if the scanning sequence is interrupted or if synchronism is not maintained.

Figure 14—Multiplex distributor.



5. *Performance of Impulse-Frequency Telemeter*

5.1 General

The performance of a telemeter operated by impulse frequency depends upon the range of frequencies used. When multiplex signalling over lines is required the impulse frequency must be low, but where single readings are acceptable the impulse frequency may be much higher. Response time and response error are increased by reducing the frequency range, but system error is unaffected.

5.2 Accuracy

The error due to the action of the circuit has been shown (Section 4.2.1) to be 0·5 per cent. This error is eliminated in setting full scale and zero, but since it is due to valve characteristics it is important to obtain a low value. The circuit also contributes error due to variation of the contact potential of the input valve that modifies the applied potentials. This error can be kept below 1 per cent by maintaining the operating power supplies within $\pm$ 10 per cent.

5.3 Response Time

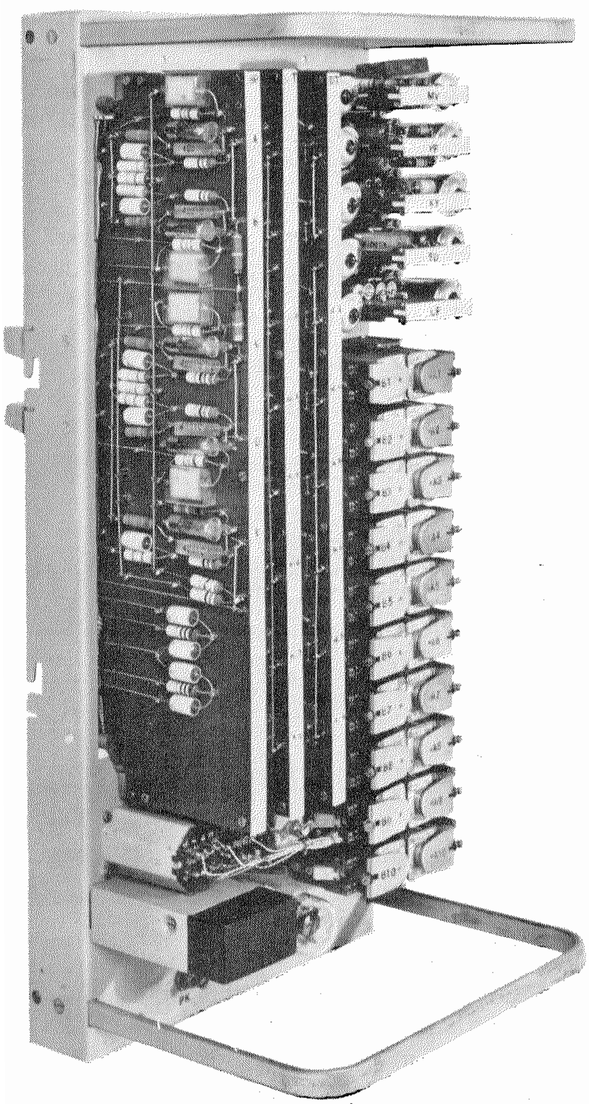
The response of a telemeter incorporating *CR* time-constants is logarithmic in form, and the response time may be defined as the time taken to reach 80 per cent of the final value after an instantaneous change. The response time is inversely proportional to the impulse frequency;

times for a selection of telemeters are shown in Table 4. The response curve is shown in Figure

TABLE 4

Telemeter Type 1 Response Times for Different Pulse Rates

Pulse Rate in Impulses per Minute	Response Time in Seconds
10	90
15	60
20	45
$33\frac{1}{3}$	30
$66\frac{2}{3}$	15
300	3
1200	1



Typical code-sender panel incorporating hot- and cold-cathode valve circuits and associated relays.

15, from which it will be seen that the change becomes apparent in a much shorter time. This is psychologically important where a change has been ordered and a response of the telemeter is expected.

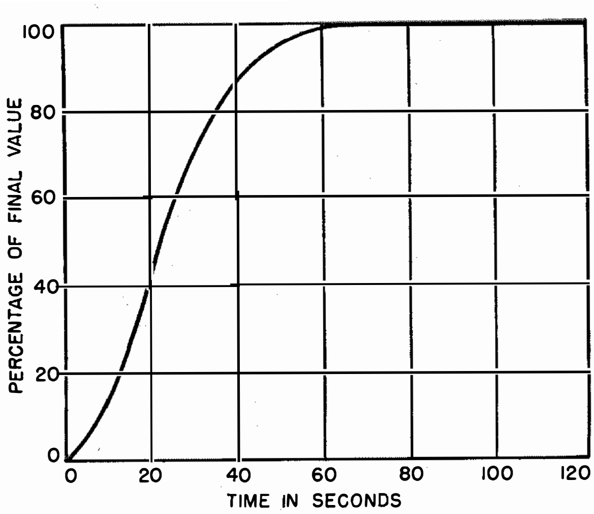


Figure 15—Response curve for impulse-modulated telemeter at $33\frac{1}{3}$ impulses per minute.

5.4 Response Error

In practice a telemeter is used to observe a condition of an electrical system that is continuously varying, and the response to these variations is more important than the response to the infrequent instantaneous changes. A telemeter that responds with time delay will show a varying value with a time lag, and the indicated reading will differ from the true reading by an amount dependent on the time lag and the rate of change of reading. This difference may be termed the response error.

Where the rate of change of a telemetered value is comparatively low the telemeter lags behind the true reading by a time that is approximately one-half of the response time. The response error is then the amount by which the reading is changing during this time. For example, a telemeter may follow slow changes with a time lag of 10 seconds, in which case a reading changing at the rate of 10 per cent per minute is indicated with an error of 1·66 per cent.

5.5 Irregular Impulsing

Impulses obtained from a single source are regularly spaced, and minor irregularities introduced by multiplex transmission can be accepted. A telemeter can be arranged to operate from a number of sources, giving a reading of the total by providing a number of input capacitors, each with a capacitance C/N. The random operations of the sources are then accepted if the time-constant of the telemeter is based on the impulse frequency of one source, since if all bunch continuously, the telemeter will then receive a value of $N \times C/N = C$ at each impulse, and this corresponds to single-source operation. The principle is unchanged by using impulses transmitted from a tariff summator. The telemeter time-constants are fixed by the impulse frequency of the individual sources and not by the impulse output of the summator.

In practice it is necessary to fix a minimum time interval between consecutive impulses to permit transmission over a shared channel. A reduced telemeter time-constant is then permissible, since the closer spacings are opened out.

6. *Type of Equipment*

6.1 General

Where telemetering equipment is associated with communication systems it is reasonable to use conventional Post Office type apparatus mountings. These mountings take the form of open-type racks or enclosed cubicles with horizontal shelves into which the individual telemeter units plug, thus facilitating easy maintenance and extension of the system. When only single telemeters are required that are not associated with communication equipment, self-contained units in wall- or panel-mounting cases may be used.

6.2 Telemeter Units

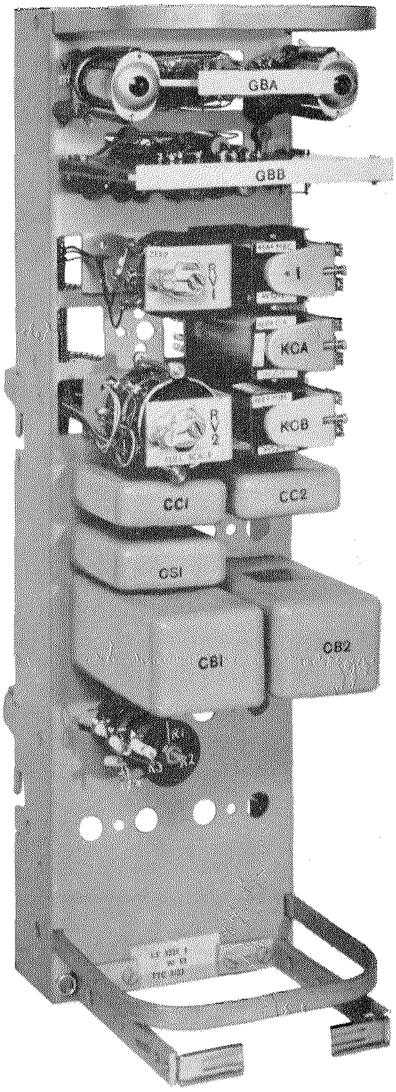
The individual telemeter panels or units are based on standard Post Office panel equipment. These units are arranged for multiple-plug connections and can accommodate relays, capacitors, valves, etc., directly mounted, and smaller components on sub-assemblies.

When these units are used for telemeters, voltages higher than those for which the equipment was originally designed are used, and therefore additional plug connections with improved insulation are required.

The sub-assemblies for components associated with the thermionic and cold-cathode tubes are boards of insulating material punched to facilitate the mounting and wiring of the resistors, capacitors, etc. Each board is double-sided, and when mounted occupies the space normally taken by two relays on the standard mountings.

6.3 Components

Special consideration has been given to the use of most suitable components in the telemeter



Electronic pulse-frequency-telemeter panel.

unit. With reference to Figure 6, the capacitor *C1* is of high capacitance stability, whilst *C2* is designed for high insulation resistance. The feedback and output resistors *R1* and *R2* are both of high stability, but the other components are of normal commercial limits, e.g. ± 10 per cent tolerance.

Only two types of commercial thermionic valves are used, a pentode and double-triode. The cold-cathode tubes used in the multiplex and distributor are restricted to one type which is a wire-ended miniature tube.

7. *Application of Impulse-Frequency Telemeter*

7.1 General

The impulse-frequency telemeter is applicable to all the electrical quantities required in the control room of a system operation centre, and since several readings can be transmitted over a restricted bandwidth by means of the electronic distributor or teleprinter sender it satisfies the economic requirements of line rental.

Considerations concerning each quantity to be telemetered are given in the following sections.

7.2 Generated Output

Tariff-metering installations in power stations provide high-grade prime movers for generation metering. At small stations, individual watt–hour meters fitted with pulse contacts can be used, whilst in larger stations output pulses from a mechanical summation unit actuated by the individual generator meters can be used.

At stations where mechanical summation equipment exists, pulses can be derived as shown in Figure 16. The magnetically polarized relay

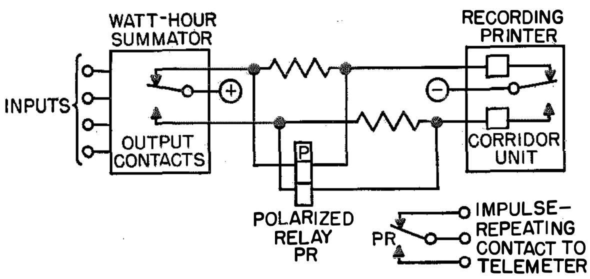


Figure 16—Metering from mechanical summator.

PR is actuated in parallel with resistors inserted in the operating-coil circuits, thus obviating additional pulse-generation equipment.

When measurement of a single circuit only is required, the arrangement shown in Figure 17 is suitable. The corridor relay *CR* responds to the meter contacts and signals to a telemeter as in Figure 6. Relay *CR* is differentially operated and eliminates the effect of possible "chatter" of meter contacts.

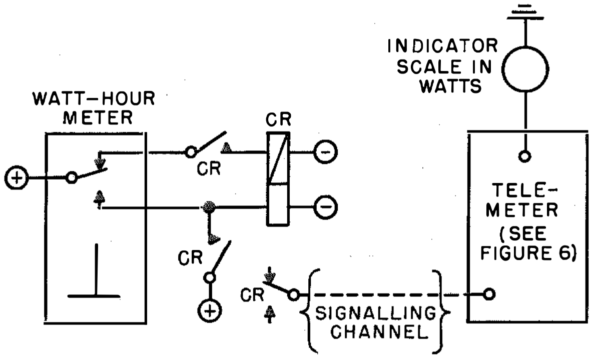


Figure 17—Load metering from watt–hour meter.

7.3 Summation

When two or more single-direction telemeter outputs of equal magnitude are to be summed, simple parallel-current summation can be employed as shown in Figure 18. In this case the total indicator has a full-scale deflection corresponding to the sum of the individual circuit currents. Such an arrangement is not convenient when circuits of different magnitudes are to be summed, since special instrument calibration is involved. When loads of different total magnitudes require summation, e.g. generators of differing sizes or generating stations of differing capacity, a summation amplifier is used as shown

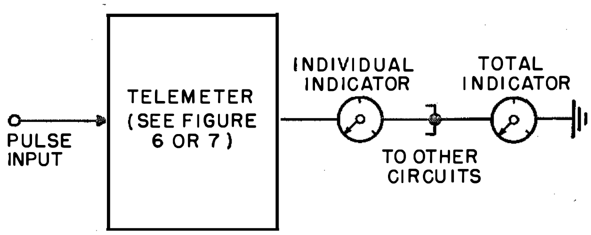


Figure 18—Summation of single-direction circuits of the same scaling.

in Figure 19. This is an arrangement of parallel-current summation, and in this case the input resistors *R1*, *R2*, etc., from each circuit are of values inversely proportional to the full-scale value of each individual circuit, thus permitting the use of a standard 5-milliampere instrument for the indication of the total, and the addition or subtraction of circuits without modification to the total meter.

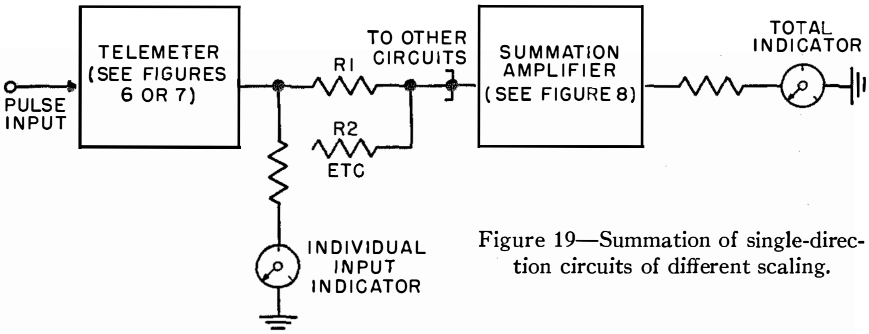


Figure 19—Summation of single-direction circuits of different scaling.

For the addition of two or more 2-direction telemeter outputs of equal magnitude, an arrangement such as that of Figure 20 is suitable. In this case the direction relay *DR* reverses the individual instrument deflection and gives the correct direction sense to each input to the total instrument.

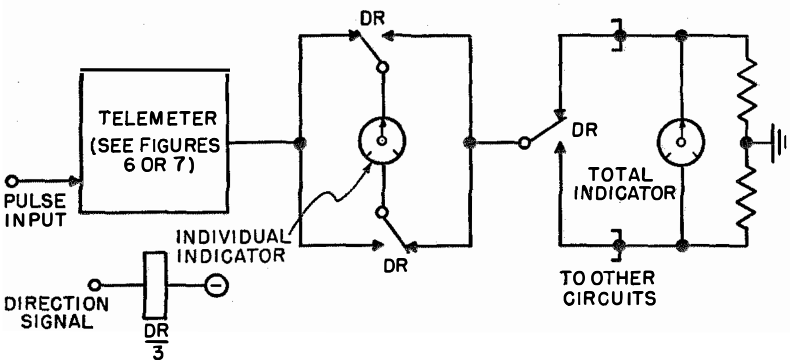


Figure 20—Summation of two-direction circuits of the same scaling.

Figure 21 shows the arrangement for the summation of 2-direction quantities of different magnitude. The values of *R1*, *R2*, etc., are made inversely proportional to the full-scale value of the individual circuits; the direction relay *DR* reverses the individual instrument and directs the input to the "add" or "subtract" side of the summator.

In cases where a chart recording of summed quantities is required it is desirable to use the summation amplifier to provide adequate output for the recorder and indicating instruments. Alternatively a sensitive recorder may be used.

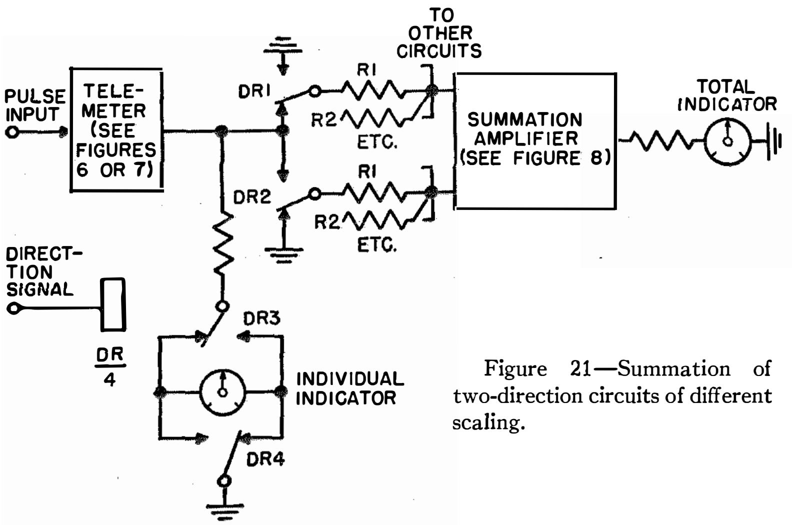


Figure 21—Summation of two-direction circuits of different scaling.

7.4 Feeder Flow

Since feeder metering involves both magnitude and direction, a watt-hour meter has been developed that is provided with three sets of impulse contacts as shown in Figure 22. This arrangement makes it possible to detect by simple relay technique the speed and direction of disk rotation.

It is necessary to transmit this direction indication together with the pulse rate to the central station. When the electronic teleprinter sender is used, the character can be changed to give direction, whilst in the case of the distributor an alternative position in the cycle is utilized.

A pulse rate of $66\frac{2}{3}$ impulses per minute is practical for the meter manufacturer, and employing the modified telemeter described in Section 4.2.2, response times of 80-per-cent change in 12 seconds and 100-per-cent change in 24 seconds are achieved without impairing the stability of readings near zero.

In the case of group interconnectors, a summator panel together with sense-of-direction indications can be used to give the net group transfer.

7.5 Voltage and Current

A modified form of integrating meter arranged with two sets of either voltage or current coils can generate pulses proportional to the square of the quantity. Such an arrangement has advantages in the fact that the prime mover is similar to an ordinary watt-hour meter and results in a telemeter scale shape that opens out in the upper part of the scale. The standard telemeter as described in Section 4.2.1 is suitable for use with these quantities.

For the generation of pulse rates directly proportional to voltage or current an impulse generator as described in Section 4.2.4 may be used, operating on rectified voltages derived from secondary instrument transformers.

7.6 Frequency

When it is necessary to obtain power-frequency measurement from remote points of the network, a new type of rotating frequency meter that is provided with impulse contacts can be employed. This meter embodies tuned circuits that make the speed of rotation a function of frequency. Two ranges are available:

45–52 cycles per second, corresponding to $66\frac{2}{3}$–$8\frac{1}{3}$ impulses per minute.

48–52 cycles per second, corresponding to $66\frac{2}{3}$–$8\frac{1}{3}$ impulses per minute.

It will be seen that zero pulse rate has been set at a corresponding frequency higher than 52 cycles per second in order to be outside the range of frequency normally experienced in power systems. Since the pulse rate is inversely proportional to the frequency, a reversed-zero-indicating instrument is used with the telemeter, which is of the basic type described in Section 4.2.1. A frequency meter of this type has an accuracy of 0·05 cycle per second over a restricted scale range.

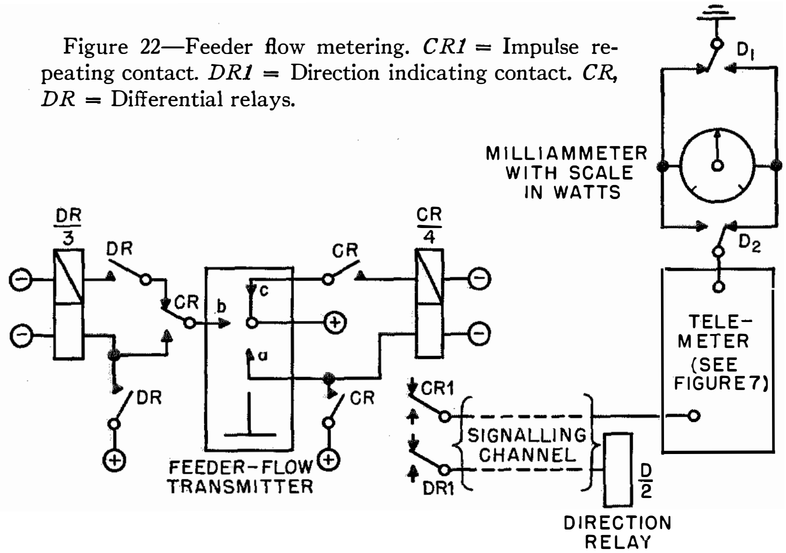


Figure 22—Feeder flow metering. *CR1* = Impulse repeating contact. *DR1* = Direction indicating contact. *CR*, *DR* = Differential relays.

7.7 Rate of Change of Frequency

The differentiator described in Section 4.2.5 provides a convenient device to indicate continuously the rate of change of frequency from a direct voltage on the output of a telemeter measuring frequency.

Figure 23 shows in block-schematic form the arrangement for a frequency meter and a rate-of-change meter. A centre-zero indicating instrument is used and a scaling of 0·2 cycle per second per minute is available. The output circuit of the differentiator can be used to operate a number of indicating instruments in series, thus permitting simultaneous observation on loading and switching-control displays.

7.8 Retransmission

When meter indications displayed in one control room have to be indicated in another control room remote from the first, two courses are available. If the reading is available in pulse form, as received for an individual circuit, it is possible to retransmit the pulses as generated by means of either the teleprinter sender described in Section 4.3.2 or the distributor of Section 4.3.3.

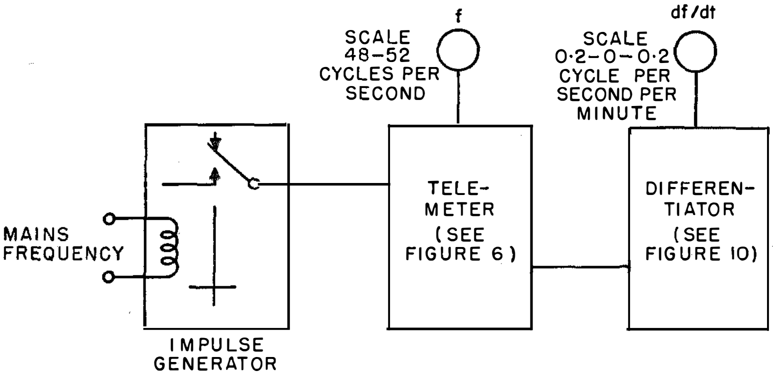


Figure 23—Arrangement for indicating frequency and rate of change of frequency.

Should the quantity required for retransmission be derived by summation at the display point, an impulse generator (Section 4.2.4) may be used to provide a pulse rate proportional to the direct-current input. This pulse rate can then be transmitted by the technique previously mentioned, measured by a standard telemeter unit at the receiving end and, if required, summed with other quantities. Thus it is possible to start with a pulse frequency, change to proportional direct current for display, and revert to pulse frequency for convenience of retransmission.

8. Conclusions

Telemetering systems that operate over any appreciable distance should be based on a variable frequency, and an impulse-frequency arrangement gives the greatest flexibility. Such a system, whilst not capable of handling readings of all possible quantities, can indicate load, voltage, current, and frequency, and can repeat readings from other known systems that change some quantity into a voltage.

Impulse transmission lends itself to time-sharing-multiplex arrangements that obtain the maximum economy in bandwidth and line rental. Electronic systems have now reached a stage of reliability that permits their use for power-system control. Cold-cathode tubes of long life are available and in use, and development is proceeding rapidly. Electronic technique has passed the stage of special engineering, and equipments are now presented in a form suitable for quantity production with interchangeability.

9. Acknowledgments

The authors are indebted to Mr. P. J. Squire, chief system operation engineer of the British Electricity Authority, for his help and guidance on matters of system operation.

10. Additional References

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Two New Equations for the Design of Filters*

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1. Introduction

THERE are some situations where a selective circuit that is equivalent to an inverse arm, i.e. a constant-K-configuration filter, is to be designed, *and the unloaded Q of the elements to be used is sufficiently high for them to be considered "nondissipative."* This paper presents two equations that for the nondissipative case, specify the exact element values required for the filter to produce that attenuation shape having the highest possible rate of cutoff, i.e., the Chebishev attenuation shape.

2. Examples of Nondissipative Equivalent Inverse-Arm Filters and the Three Circuit Constants That Must Be Correctly Adjusted

Because so many of the selective circuits now being used, or designed, seem physically so different from the basic inverse-arm configurations, many engineers new to the field do not realize that the design equations for the constant-K configuration can be applied.

It thus seems worthwhile calling attention to a few of these equivalent inverse-arm filters to stress the wide applicability of the two design equations to be presented.

It will be noticed that with one exception, the band-pass examples are from the ultra-high-frequency and microwave regions, because it is mainly in these regions that the ratio of unloaded Q to fractional midfrequency $Q_0/[f_0/(\mathrm{BW})]$ is high enough for the elements to be considered nondissipative.

Figure 1A shows a common direct-coupled waveguide band-pass filter using four resonators; in another language it would be called a quadruple-tuned band-pass filter. The equivalence of this to the fundamental constant-K configuration (either band-pass or low-pass) has been excellently described in W. W. Mumford's paper.[1]

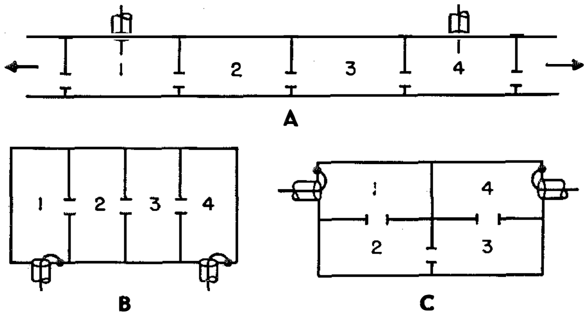


Figure 1—Three ways of arranging 4 waveguide resonators to produce a quadruple-tuned band-pass filter. When small-percentage bandwidths are used, these filters can be designed from the equations for the simple constant-k-configuration low-pass filter of Figure 5.

In the "language" used in this present paper, the design information that the engineer must possess (and which is required for all equivalent constant-K-configuration filters) is:

A. The required coefficient of coupling $K_{r(r+1)}$ between adjacent resonators. This fixes the size of the opening that must be made in the wall between adjacent resonators, and as is well known, this opening can take the form of a slot parallel to the electric-field vector, which will give the equivalent of mutual-inductance coupling between resonators; a slot perpendicular to the electric-field vector, which will give the equivalent of "low-side" capacitive coupling between resonators; a post parallel to the electric-field vector, which will give the equivalent of self-inductance coupling between resonators; or, in general, any kind of opening that will allow some of the electric and/or magnetic field of one resonator to enter the adjacent resonator.

B. The required resonant frequency (f_0) of each resonator. This fixes the distance between the walls of each resonator. As is well known, the coefficient-of-coupling mechanism must be correctly considered a part of each resonator to which it is connected;

* A condensed version of this paper appeared in *Convention Record of the 1953 I.R.E. National Convention,* Part 5—Circuit Theory, pages 44–47. This is the full version of the paper as presented at the National Convention of the Institute of Radio Engineers in New York, New York, on March 23, 1953.

[1] W. W. Mumford, "Maximally-flat Filters in Waveguide," *Bell System Technical Journal,* volume 27, pages 684–713; October, 1948.

otherwise the pass-band midfrequency will not coincide with the resonant frequency.

C. The required singly loaded Q (Q_1) of the first resonator (produced by correctly coupling the generator to this first resonator); and the required singly loaded Q (Q_n) of the last resonator (produced by correctly coupling the load to this last resonator).

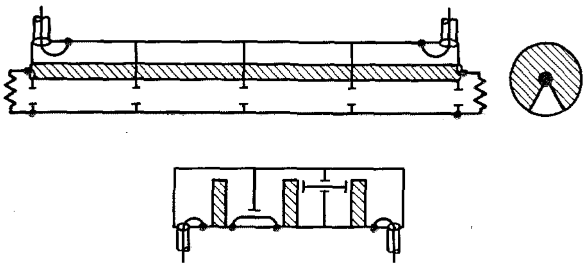


Figure 2—Two of the many ways of using coaxial resonators to produce a small-percentage-band-pass filter. Here, also, the low-pass-design information is applicable.

If a terminated waveguide is used on each side of the filter, this then fixes the size of the opening in the first and last wall of the structure of Figure 1A; or, if desired, these first and last walls can be completely closed off and, as Figure 1A attempts to show, the generator and load can be capacitively coupled to the first and last resonators by probes (or magnetically coupled by loops). Whatever the method used, this generator and load coupling must be adjusted until the first and last resonators, respectively, have the required singly loaded Q_1 and Q_n.

The above three well-known circuit constants have been discussed in a previous paper,[2] and methods of measuring and adjusting them have also been presented.[3]

Continuing with some other examples of equivalent constant-K structures, Figures 1B and 1C show that by discarding the waveguide concept in favor of the coupled-resonator concept, additional useful, and different-looking, filters can be built with the same four resonators. Figure 1B shows the four resonators of Figure 1A rotated by 90 degrees and placed together in such a way that the openings between adjacent resonators produce the equivalent of "high-side" capacitive coupling. Figure 1C shows the same four resonators arranged in yet another physical configuration that will still produce the same small-percentage bandwidth filtering action: there is equivalent "high-side" capacitive coupling between resonators *1* and *2*, mutual-inductance coupling (due to a vertical slot) between resonators *2* and *3*, high-side capacitive between resonators *3* and *4*; the generator sets Q_1 by inductive coupling to the first resonator and the load sets Q_4 by inductive coupling to the last resonator.

Figure 2 is included to stress the fact that the "different-looking" filters produced by using coaxial resonators are also equivalent to constant-K-configuration filters insofar as band-pass response and required circuit constants are concerned.

Figure 3 shows a triple-tuned band-pass filter that, while in no way physically resembling the classical inverse-arm structure, is still described

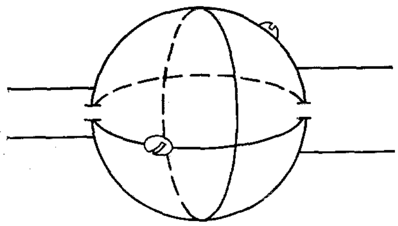


Figure 3—A 3-mode single-cavity microwave filter. While bearing no physical resemblance whatsoever to the classical inverse-arm filter, the low-pass-design information is also applicable to this structure, when small-percentage bandwidths are required.

by exactly the same design constants as the inverse-arm structure. It is the spherical resonator that is so designed that three of its resonant modes occur at the same frequency, i.e., are degenerate. The two screws shown project into the cavity and correctly adjust K_{12} (the coefficient of coupling between the first resonance and the second resonance), and K_{23} (the coefficient of coupling between the second resonance and the third resonance). The opening on the left is of the proper size to allow the terminated waveguide shown to load properly the first resonance, i.e., to set Q_1; and the opening on the right allows the terminated waveguide shown there to load

[2] M. Dishal, "Design of Dissipative Band-Pass Filters Producing Desired Exact Amplitude—Frequency Characteristics," *Electrical Communication*, volume 27, pages 56–81; March, 1950: also, *Proceedings of the I.R.E.*, volume 37, pages 1050–1069; September, 1949.

[3] M. Dishal, "Alignment and Adjustment of Synchronously Tuned Multiple-Resonant-Circuit Filters," *Electrical Communication*, volume 29, pages 154–164; June 1952: Addendum, volume 29, page 292; December, 1952: also, *Proceedings of the I.R.E.*, volume 39, pages 1448–1455; November, 1951.

properly the last resonance, i.e., to set Q_3. Finally, Figure 4 shows a three-resonator filter using mechanical resonators for the filter elements. Here, the coefficient of couplings K_{12} and K_{23} are set by the material, diameter, and "tap" point used for the quarter-wavelength-long (approximately) thin rods that connect two adjacent resonators. Q_1 of the first resonator is correctly set by the thin low-Q resonant rod connected to the first resonator, and the last resonator is similarly correctly loaded by the low-Q rod connected to it. The coils, by magnetostrictive action, convert the electric energy to mechanical energy and then vice versa; because of the unfortunately poor coupling produced by this phenomenon, there is usually negligible electrical loading coupled into the first and last resonators.

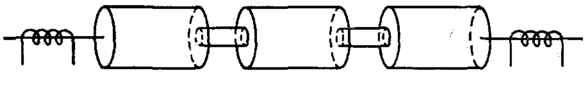


Figure 4—A triple-tuned mechanical filter made with half-wave slugs and quarter-wave coupling necks. The low-pass-design equations apply here also.

There are many other examples of filters that at first glance do not resemble the basic inverse-arm configuration, but that actually are equivalent to it; and in all of these many filters, the design engineer must know the required numerical value for all the coefficients of couplings in the structure; the required numerical value of the singly loaded Q of the first resonator and that of the last resonator; and the proper element values or physical lengths to produce the proper midfrequency (or design information exactly equivalent to these three quantities).

3. "*Incorrect*" *Coefficients of Coupling and End Q's Called for by Classical Filter Theory*

It will now be assumed that the reader realizes that, within the small-percentage-band-pass approximation, the transfer equation for the circuits of Figures 1 to 4 in terms of the frequency variable

$$\left(\frac{f}{f_0} - \frac{f_0}{f}\right) = \frac{f_2 - f_1}{(f_2 f_1)^{1/2}} = \left(\frac{\mathrm{BW}}{f_0}\right)$$

is identical in form to that of the low-pass ladder of Figure 5, when the frequency variable for the latter is radian frequency ω.

To clarify the meaning of coefficients of coupling and end Q's as applied to the low-pass inverse-arm ladder, classical filter theory will be applied to the ladder of Figure 5 to obtain the design values called for by this theory. It will be recalled that the two basic facts of this theory are as given in (1) and (2).

The full series-arm reactance and the full shunt-arm reactance must be so related that (1) is true at the desired cutoff frequency ω_c.

$$\frac{Z_{\text{series}}}{4Z_{\text{shunt}}} = -1. \tag{1}$$

The impedance Z_0, which must be used to terminate the ladder and which unfortunately cannot be physically realized, is given by (2).

$$Z_0 = (Z_{\text{series}} Z_{\text{shunt}})^{1/2}(1 + Z_{\text{series}}/4Z_{\text{shunt}})^{1/2}. \tag{2}$$

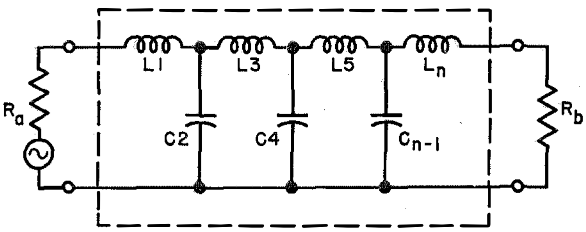


Figure 5—The inverse-arm low-pass ladder whose design equations apply to the small-percentage band-pass filters of Figures 1 to 4. The unloaded Q's of the elements are infinite.

From (1), we obtain (1A),

$$\left.\begin{aligned} 1/(LC_{\text{internal}})^{1/2} &= 0.50\omega_c, \\ \frac{1}{(L_1C_1)^{1/2}} = \frac{1}{(C_{(n-1)}L_n)^{1/2}} &= 0.707\omega_c, \end{aligned}\right\} \text{(1A)}$$

and from (2), we obtain

$$\frac{Z_0/L_{1,n}}{\omega_c} = \left[1 - \left(\frac{\omega}{\omega_c}\right)^2\right]^{1/2}. \tag{2A}$$

The required resonant radian frequencies given by (1A) for the adjacent arms of the low-pass ladder of Figure 5 are exactly equivalent to the well-known coefficients of coupling of the small-percentage-band-pass networks of Figures 1 to 4, when the frequency variable ω and BW/f_0 are used, respectively. Thus by (1A), classical filter theory requires that all the internal coefficients of coupling be made equal to 0.50 times the "cutoff" frequency variable, and that the coefficient of coupling between the first and second elements and between the next to the last and last elements be made 0.707 times the cutoff-frequency variable. Simultaneously, (2A) must be satisfied, and this is unfortunately impossible for it demands that at *zero frequency* $R_0/L_{1,n}$ —which is exactly equivalent to the well-known decrement $(1/Q)$ of the small-percentage-bandwidth networks—must equal 1.0 times the cutoff-frequency variable but then over the pass and attenuation band, the termination must vary in the way specified by the right-hand side of (2A); this is impossible to achieve.

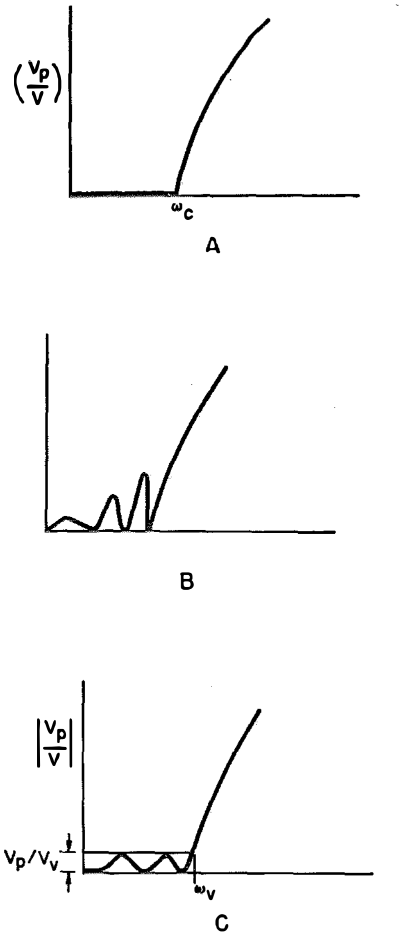


Figure 6—*A* shows the physically unrealizable attenuation shape of a Z_0-terminated constant-*k* filter. $(V_p/V) = \exp\,(n-1)\cosh^{-1}(\omega/\omega_c)$. When physical resistors are used instead of the unrealizable Z_0, the attenuation shape of *B* is obtained. *C* shows the optimum Chebishev shape, which can be obtained when the element values are correctly modified from the constant-*k* values.

$$|V_p/V|^2 = 1 + [(V_p/V_v)^2 - 1]\cosh^2[n\cosh^{-1}(\omega/\omega_c)].$$

If it were possible to obtain this termination and the above values of coefficient of coupling were used, then the response obtained would be that shown in Figure 6A.

The procedure usually resorted to is to approximate Z_0 by a resistance equal to the zero-frequency value of Z_0, and thus the classical filter is terminated in a fixed resistance of value $R/L_{1,n} = 1.0\omega_c$, i.e. an input and output decrement of 1.0 times the frequency variable is used. The coefficients of coupling between all internal elements are still maintained equal to 0.50 times the frequency variable, and the coefficients of coupling between the first and second elements and next to last and last elements are kept at 0.707 times the frequency variable. When this design is used, it is now well known that the type of response obtained is that[4] shown in Figure 6B. The increasing peak-to-valley ratio in the pass-band makes this type of response objectionable if it is necessary to use a large number of elements to obtain a high rate of cutoff.

4. *Optimum Constant-K-Configuration Attenuation Shape of Modern Filter Theory*

When there are stringent requirements on the pass-band tolerance and rate of cutoff of a filter, the approximate design procedure of image-parameter theory is usually discarded and the much-more-exact insertion-loss design procedure is used. It is interesting to note that although this procedure was originated more than 15 years ago,[5–7] most practicing engineers are still not familiar with it.

By a series of steps, the history of which is not very clear, it was realized that the nonoptimum

[4] G. L. Ragan, "Microwave Transmission Circuits," McGraw-Hill Book Company, New York, New York; 1948: chapter 9 by R. M. Fano and A. W. Lawson; pages 576 and 577.

[5] E. L. Norton, "Constant Resistance Networks with Applications to Filter Groups," *Bell System Technical Journal*, volume 16, pages 178–193; April, 1937.

[6] S. Darlington, "Synthesis of Reactance 4-Poles," *Journal of Mathematics and Physics*, volume 18, pages 257–353; September, 1939.

[7] E. L. Norton, United States Patent 1 788 538; January, 1931.

response of Figure 6B could be modified so that the ripples in the pass band were all of equal level. The obvious application of the known method of approximating a constant by means of Chebishev polynomials then led to the following basically important equation for the optimum response shown in Figure 6C. For the band-pass case, BW/f_0 would simply be used instead of ω.

$$\left|\frac{V_p}{V}\right|^2 = 1 + \left[\left|\frac{V_p}{V_v}\right|^2 - 1\right] \cosh^2\left(n \cosh^{-1} \frac{\omega}{\omega_v}\right). \tag{3}$$

Where amplitude filtering only is concerned, i.e. no phase- or time-response considerations are involved, the attenuation shape of Figure 6C is optimum in the sense that for a given allowable ripple in the pass band, (3) produces the maximum possible rate of cutoff for a given number of elements.

In (3), n is the number of arms in the ladder of Figure 5 and ω_v is the radian frequency at that point on the skirt where the attenuation is the same as the peak-to-valley ratio.

When the ripple in the pass band is made to become zero decibels, then by correctly approaching this limit, (3) becomes (4).

$$\left|\frac{V_p}{V}\right|^2 = 1 + \left(\frac{\omega}{\omega_{3db}}\right)^{2n}, \tag{4}$$

where ω_{3db} is the radian frequency at that point on the skirt that is 3 decibels down from the peak response.

The problem now is to apply the procedures of modern network theory to the attenuation shapes (3) and (4) and to synthesize the network and element values that will produce this optimum filtering shape.

5. *Two Basic Synthesis Procedures of Modern Network Theory*

Given the equation for a desired attenuation shape (it must be a rational function, of course), there are at least two basic procedures for synthesizing a corresponding network. For want of a better name, the first procedure will be called the direct method and as far as this writer is aware, Norton was the first to describe this method.[7] The second method is properly called the Darlington method and is basically described in reference 6.

5.1 Direct Method

Step 1. Pick a network configuration that is known (somehow) to be capable of producing the desired attenuation shape.

Step 2. By means of Kirchhoff's laws, write the complex equation for the network response to be synthesized. For the low-pass ladder, this equation will be in the form of the ratio of two polynomials in $j\omega$ of highest power n, and with constant coefficients made up of complicated combinations of the L, C, and R elements of the arms of the network.

Step 3. Solve the desired attenuation-shape equation (e.g. (3) or (4)) for its $2n$ complex zeros; multiply these ω zeros by j to make them functions of $j\omega$, and then combine the n left-half-plane zeros to obtain the complex-numerator polynomial in $j\omega$ of exactly the same form as that obtained in Step *2* but with numerical coefficients. Next, solve for the $j\omega$ attenuation infinities and use them to form the denominator polynomial.

Step 4. Compare the equations of Step *2* and Step *3* and equate the coefficients of identical powers of $j\omega$. This will result in n simultaneous equations that must be simultaneously solved for the n unknown element values.

Note: This method was described in detail in reference 1 at a time when the author did not know of the existence of references 5 and 7.

5.2 Darlington Method for a Uniformly Dissipative Filter

Step 1. Solve the desired attenuation-shape equation ((3) or (4)) for its $2n$ complex zeros. Multiply these ω/ω_v zeros by j so that they are functions of $j(\omega/\omega_v)$.

Step 2. Pick the n left-half-plane zeros and reduce the magnitude of the real component of these zeros by an amount equal to the normalized decrement of each arm, i.e. by an amount $(R/L)/\omega_v = (G/C)/\omega_v$ for the low-pass ladder.

Step 3. Use these modified n left-half-plane zeros to form a complex-numerator polynomial in $j(\omega/\omega_v)$ having numerical coefficients.

Solve the desired attenuation-shape equation for its $j(\omega/\omega_v)$ infinities and use these infinities to form a denominator polynomial in $j(\omega/\omega_v)$.

Our modified complex-shape equation will now be in the form of (5).

$$\frac{V_p}{V} = \frac{1}{|\Delta|_{\min}} \left\{ \frac{\left(j\frac{\omega}{\omega_v}\right)^n + U_{n-1}\left(j\frac{\omega}{\omega_v}\right)^{n-1} + U_{n-2}\left(j\frac{\omega}{\omega_v}\right)^{n-2} + \ldots + U_0}{\left[\left(j\frac{\omega}{\omega_v}\right)^2 + \left(\frac{\omega_{\infty 1}}{\omega_v}\right)^2\right]\left[\left(j\frac{\omega}{\omega_v}\right)^2 + \left(\frac{\omega_{\infty 2}}{\omega_v}\right)^2\right]\left[\cdots\right]} \right\} \tag{5}$$

with $|\Delta|_{\min}$, which is the square root of the minimum magnitude of the bracketed polynomial, as yet undetermined.

Step 4. Take the sum of the squares of the real and imaginary parts of the bracketed polynomials to form the magnitude and by differentiation, or plotting versus the frequency variable, find the minimum numerical value of this magnitude. This minimum value is $|\Delta|^2_{\min}$.

Step 5. The modified magnitude equation $|V_p/V|^2$ is now equal to the magnitude polynomial formed in Step *4* divided by the numerical value $|\Delta|^2_{\min}$.

Step 6. Subtract from the modified magnitude equation of Step *5* the numerical value P_p/P_m. This quantity is the ratio of the power delivered to the load at the peak response frequency to the maximum power available from the resistive generator. For a resistive generator and a resistive load, an impedance-matched output is usually desired, so for this case P_p/P_m is usually set equal to 1.0; however P_p/P_m may be set equal to any numerical value from 1.0 down to zero. Absorb this numerical value in the modified magnitude equation of Step *5* to form a new numerator polynomial. (The denominator polynomial will remain unchanged.)

Step 7. Solve the numerator polynomial obtained in Step *6* for its $2n$ complex zeros. Multiply these ω/ω_v zeros by j to make them functions of $j(\omega/\omega_v)$.

Step 8. Use the n left-half-plane zeros to form a complex numerator polynomial in $j(\omega/\omega_v)$ having numerical coefficients. This complex polynomial is the numerator of (6).

$$\left\{\left|\frac{V_p}{V}\right|^2 - \left|\frac{P_p}{P_m}\right|\right\}\left(j\frac{\omega}{\omega_v}\right) = \frac{1}{|\Delta|_{\min}} \left\{ \frac{\left(j\frac{\omega}{\omega_v}\right)^n + V_{n-1}\left(j\frac{\omega}{\omega_v}\right)^{n-1} + V_{n-2}\left(j\frac{\omega}{\omega_v}\right)^{n-2} + \ldots + V_0}{\left[\left(j\frac{\omega}{\omega_v}\right)^2 + \left(\frac{\omega_{\infty 1}}{\omega_v}\right)^2\right]\left[\left(j\frac{\omega}{\omega_v}\right)^2 + \left(\frac{\omega_{\infty 2}}{\omega_v}\right)^2\right]\left[\cdots\right]} \right\}. \tag{6}$$

Step 9. From the numerator polynomial formed in Step *3* and that formed in Step *8*, form the function

$$\left\{ \frac{2\left(j\frac{\omega}{\omega_v}\right)^n + (U_{n-2} + V_{n-2})\left(j\frac{\omega}{\omega_v}\right)^{n-2} + \ldots}{(U_{n-1} + V_{n-1})\left(j\frac{\omega}{\omega_v}\right)^{n-1} + (U_{n-3} + V_{n-3})\left(j\frac{\omega}{\omega_v}\right)^{n-3} + \ldots} \right\}. \tag{7}$$

Step 10. The function formed in Step *9* is any one of the following four input immittances of the required lossless network: when n is odd, it is $Z_{\text{in sc}}/R_a$ and $Y_{\text{in oc}}/G_a$; when n is even, it is $Y_{\text{in sc}}/G_a$ and $Z_{\text{in oc}}/R_a$; where R_a is the resistive termination on the left side of the network. Knowing the locations of the real frequency infinities as given in the denominator of (5), expand the function in continued-fraction form to obtain the network and element values in terms of R_a. When performing the expansion, it is necessary to know all the frequencies of infinite attenuation.

Step 11. When the right-half-plane zeros of Step *7* are used to form the polynomial of Step *8*, then the 2nd, 4th, 6th, et cetera, terms will be negative. If we use this polynomial to form the input immittance function, we obtain

$$\left\{\frac{2\left(j\frac{\omega}{\omega_v}\right)^n + (U_{n-2}+V_{n-2})\left(j\frac{\omega}{\omega_v}\right)^{n-2} + \ldots}{(U_{n-1}-V_{n-1})\left(j\frac{\omega}{\omega_v}\right)^{n-1} + (U_{n-3}-V_{n-3})\left(j\frac{\omega}{\omega_v}\right)^{n-3} + \ldots}\right\}. \tag{8}$$

Step 12. This function gives the expansion of the network in Step *10* from the other end, i.e. for n odd, this function is $Z_{\text{in sc}}/R_b$ and $Y_{\text{in oc}}/G_b$, respectively, and for n even, this function is $Z_{\text{in oc}}/R_b$ and $Y_{\text{in sc}}/G_b$, respectively. Assigning the infinite attenuation frequencies to the same network elements as in Step *10*, expand this function in continued-fraction form to obtain the same network as that in Step *10* but with element values given in terms of R_b, the termination on the right side of the network.

Step 13. Equating the two expressions for the same element, i.e. one in terms of R_a and the other in terms of R_b, will give the required ratio of R_b/R_a; all the reactive-element values and the terminations have now been synthesized. When the uniform loss used in Step *2* is now added to this network, it will produce the desired attenuation shape that was used in Step *1*.

Note: If the normalized decrement of Step *2* is very much smaller than the smallest real coordinate obtained in Step *1*, i.e. if the network elements to be used are essentially lossless as is the case in this paper, then Steps *2*, *4*, and *5* can be omitted, which materially reduces the amount of numerical work that must be done.

It should also be noted that when n is larger than 5 or 6, a large number of significant figures must be maintained in the simple partial-fraction expansion of Step *10* and Step *12*. For the cases of large n, the procedure outlined in Table 1 on page 298 of Dr. Darlington's paper[6] should be used.

6. *Need for Closed-Form Design Equations*

The procedures outlined in Sections 5.1 and 5.2 enable the engineer to synthesize a filter of any number of arms n if the numerical values contain sufficient significant figures; practically, this requires the use of a calculating machine and the time necessary to do the routine work.

The need for a large number of significant figures for all numerical values is particularly annoying and so it would be of real service to the practicing engineer if it were possible to obtain from the procedures of Sections 5.1 and 5.2 closed-form general solutions for the required element values.

These closed-form solutions would be obtained by using the general expressions for the roots in Step *3* of both procedures so that general expressions are obtained for the coefficients in the polynomials of (5) and (6), then, when the concluding steps are performed, it may be possible to recognize a law of formation for the required element-value equations.

7. *Element Values Required to Produce the Butterworth Attenuation Shape of* (4)

In January of 1931, E. L. Norton of the Bell Telephone Laboratories was granted a patent[7] showing that he had accomplished this general solution discussed in Section 6 for the attenuation shape of (4), when the low-pass network of Figure 5 has a resistance on one side only, the other side of the network being open-circuited if it ends in a capacitance or short-circuited if it ends in an inductance. He used the procedure of Section 5.1, and for the band-pass case his solution is as given in (9),

$$\frac{d_1}{\mathrm{BW}_{3\mathrm{db}}/f_0} = \frac{1}{\sin(90°/n)}, \quad d_{2\to n} = 0, \tag{9A}$$

$$\left(\frac{K_{r(r+1)}}{\mathrm{BW}_{3\mathrm{db}}/f_0}\right)^2 = \frac{\cos^2[r(90°/n)]}{[\sin(2r-1)(90°/n)][\sin(2r+1)(90°/n)]}. \tag{9B}$$

Then in March of 1932, W. R. Bennett also of Bell Telephone Laboratories was granted a patent[8] showing that he had accomplished the general solution for the attenuation shape of (4), when the network has equal resistances on both sides. He also used the procedure of Section 5.1 and for the band-pass case his solution is as given in (10).

Using the procedure of Section 5.1 and keeping the expressions for the polynomial coefficients as general as possible, the writer was not able to recognize the law of formation for the coefficient-of-coupling values. However, using the procedure of Section 5.1 as described in detail in reference 1 and using numerical values for all the angle functions, it was very simple to obtain the relations, given in Tables 1 and 2, between the coefficients of coupling required for the Butterworth shape and those required for the Chebishev shape.

$$\left(\frac{d_{1,n}}{\mathrm{BW}_{3\mathrm{db}}/f_0}\right) = \frac{1}{2\sin(90°/n)}, \quad d_{2\to(n-1)} = 0, \tag{10A}$$

$$\left(\frac{K_{r(r+1)}}{\mathrm{BW}_{3\mathrm{db}}/f_0}\right)^2 = \frac{1}{4[\sin(2r-1)(90°/n)][\sin(2r+1)(90°/n)]}. \tag{10B}$$

8. Element Values Required to Produce Chebishev Attenuation Shape of (3)

In the two decades that have passed since Norton and Bennett achieved their closed-form solutions for the Butterworth response shape, no one accomplished—or at an rate no one published—the closed-form solution for the more-general Chebishev response shape of which the Butterworth shape is the limiting case.

In a series of letter discussions with Mr. V. D. Landon of Radio Corporation of America during the summer of 1952, this problem was considered and the general solution for the Chebishev attenuation shape was obtained by the following means.

The relations in Table 1 are obtained by first solving the "Design Equations—Group 3" of reference 1 for the coupling values required for the Butterworth attenuation shape with the 3-decibel-down bandwidth as the reference bandwidth and with d_2 to d_n, inclusive, set equal to zero. Then the "Design Equations—Group 5" in reference 1 were solved for the coupling values required for the Chebishev attenuation shape—also with d_2 to d_n, inclusive, set equal to zero. Here, it should be noted the reference bandwidth is the "valley-decibel-down" bandwidth, not the 3-decibel-down bandwidth.

[8] W. R. Bennett, United States Patent 1 849 656; March, 1932.

TABLE 1

COEFFICIENTS OF COUPLING FOR THE REACTIVE-LOAD CASE

$n = 2$	$\dfrac{(K_{12}/F_v)_C^2}{(K_{12}/F_{3\mathrm{db}})_B^2} = S_2^2 + \cos^2(90°/2)$
$n = 3$	$\dfrac{(K_{12}/F_v)_C^2}{(K_{12}/F_{3\mathrm{db}})_B^2} = S_3^2 + 0.2500$
	$\dfrac{(K_{23}/F_v)_C^2}{(K_{23}/F_{3\mathrm{db}})_B^2} = S_3^2 + 1.000$
$n = 4$	$\dfrac{(K_{12}/F_v)_C^2}{(K_{12}/F_{3\mathrm{db}})_B^2} = S_4^2 + 0.1464$
	$\dfrac{(K_{23}/F_v)_C^2}{(K_{23}/F_{3\mathrm{db}})_B^2} = S_4^2 + 0.500$
	$\dfrac{(K_{34}/F_v)_C^2}{(K_{34}/F_{3\mathrm{db}})_B^2} = S_4^2 + 0.8535$

TABLE 2

COEFFICIENTS OF COUPLING FOR THE RESISTIVE-LOAD RESISTIVE-GENERATOR CASE

$n = 2$	$\dfrac{(K_{12}/F_v)_C^2}{(K_{12}/F_{3\mathrm{db}})_B^2} = S_2^2 + 1.000$
$n = 3$	$\dfrac{(K_{12}/F_v)_C^2}{(K_{12}/F_{3\mathrm{db}})_B^2} = S_3^2 + 0.7500$
	$\dfrac{(K_{23}/F_v)_C^2}{(K_{23}/F_{3\mathrm{db}})_B^2} = S_3^2 + 0.7500$
$n = 4$	$\dfrac{(K_{12}/F_v)_C^2}{(K_{12}/F_{3\mathrm{db}})_B^2} = S_4^2 + 0.500$
	$\dfrac{(K_{23}/F_v)_C^2}{(K_{23}/F_{3\mathrm{db}})_B^2} = S_4^2 + 1.000$
	$\dfrac{(K_{34}/F_v)_C^2}{(K_{34}/F_{3\mathrm{db}})_B^2} = S_4^2 + 0.500$

The relations in Table 2 are obtained by first solving the "Design Equations—Group 3" of reference 1 for the coupling values required for the Butterworth attenuation shape with the 3-decibel-down bandwidth as the reference bandwidth and with d_2 to d_{n-1}, inclusive, set equal to zero. Next the "Design Equations—Group 5" for the Chebishev shape were solved with the "valley-decibel-down" bandwidth as the reference bandwidth, and d_2 to d_{n-1} set equal to zero.

With the above relations established and realizing that the numerical values in the relations are formed from combinations of the sin and cos of multiples of $(90°/n)$, it was a relatively simple matter to discover that function of $(90°/n)$ that produced the numerical values in the two tables.

For the reactive-load case, the relation for the double-tuned circuit, i.e. $n=2$, could actually be made in closed form as given in Table 1, and using this as the starting point it required a relatively small amount of work to see that all the ratios in Table 1 were given by the simple function of (11).

$$\frac{(K_{r(r+1)}/F_v)^2_C}{(K_{r(r+1)}/F_{3\mathrm{db}})^2_B} = S_n^2 + \sin^2\left(r\frac{90°}{n}\right). \quad (11)$$

Actually, due to the form of the double-tuned-circuit solution, the angle function first obtained was $\cos^2(n-2r)(90°/n)$; it was then realized that this was identical to the function given in (12).

Since Norton and Bennett had obtained the solutions for $(K_{r(r+1)}/F_{3\mathrm{db}})_B$ as given in Section 7, the relations of (11) and (12) now give us immediately the desired solutions for the coefficients of coupling required for the general Chebishev attenuation shape.

Actually, due to the hint given by the double-tuned-circuit relation, the angle function first obtained was $\cos^2(n-r)(90°/n)$; it was then realized that this was identical to the function given in (11).

For the resistive-load resistive-generator case, it was immediately noted that the numbers in Table 2 were the squares of 1.0, 0.866, and 0.707, which in conjunction with the hint given by the form of the solution for the double-tuned circuit led to the realization that all the ratios in Table 2 were given by the simple function (12).

$$\frac{(K_{r(r+1)}/F_v)^2_C}{(K_{r(r+1)}/F_{3\mathrm{db}})^2_B} = S_n^2 + \sin^2\left(2r\frac{90°}{n}\right). \quad (12)$$

TO OBTAIN THE SHAPE $\left(\frac{V_p}{V}\right)^2 = 1+\left[\left(\frac{V_p}{V_v}\right)^2-1\right]\cosh^2\left\{n\cosh^{-1}\frac{\mathrm{BW}}{\mathrm{BW}_v}\right\}$

RESISTIVE GENERATOR AND RESISTIVE LOAD

$$\frac{Q_{1,n}}{f_0/\mathrm{BW}_v} = \frac{2\sin\theta}{S_n} \qquad Q_{2\to(n-1)} = \infty$$

$$\left[\frac{K_{r(r+1)}}{\mathrm{BW}_v/f_0}\right]^2 = \frac{[S_n^2+\sin^2 2r\theta]}{4\{\sin(2r-1)\theta\}\{\sin(2r+1)\theta\}}$$

RESISTIVE GENERATOR AND REACTIVE LOAD (OR VICE VERSA)

$$\frac{Q_1}{f_0/\mathrm{BW}_v} = \frac{\sin\theta}{S_n} \qquad Q_{2\to n} = \infty$$

$$\left[\frac{K_{r(r+1)}}{\mathrm{BW}_v/f_0}\right]^2 = \frac{[S_n^2+\sin^2 r\theta]}{\sec^2(r\theta)\{\sin(2r-1)\theta\}\{\sin(2r+1)\theta\}}$$

$$\theta = \frac{90°}{n} \qquad S_n = \sinh\left\{\frac{1}{n}\sinh^{-1}[(V_p/V_v)^2-1]^{-1/2}\right\}$$

Figure 7—The new design equations for nondissipative inverse-arm filters.

For the required input and output decrement values, the same procedure used to obtain the values given in Tables 1 and 2 showed immediately that for both the reactive-load case and the resistive-load resistive-generator case the relation between the decrement required for the Chebishev shape and that required for the Butterworth shape is as given in (13).

$$\frac{(d/F_v)_C}{(d/F_{3\text{db}})_B} = S_n. \tag{13}$$

Equations (11), (12), and (13) combined with the solutions of Section 7 are the desired design equations, and they have been combined to form Figure 7; they are the two new sets of equations for the design of the constant-k-configuration filters that are referred to in the title of this paper.

As indicated at the top of the figure, the attenuation shape that will be obtained will be the optimum Chebishev attenuation shown in Figure 6C. It is important to realize that the design equations of Figure 7 are given in terms of the "valley bandwidth" (BW_v) which is the bandwidth between the points on the skirt that are down by the same number of decibels as the peak-to-valley ratio V_p/V_v. The quantity S_n is a function of the number of resonators n used and the peak-to-valley ratio desired, and the reader should note that as V_p/V_v approaches unity, S_n becomes very large and therefore the required K's will be a large number of times the fractional *valley* bandwidth. However, a required bandwidth at some other decibels down rather than at the valley-decibels down is very often specified, and it is therefore necessary to get the numerical relation between the valley-decibels-down bandwidth and the specified-decibels-down bandwidth by using the shape equation at the top of Figure 7.

9. *Application of Design Equations to the Low-Pass Ladder*

As previously indicated, if one writes the transfer-impedance equations for the low-pass ladder, compares them to those obtained for the band-pass case, and uses a suitable normalizing procedure, it is found that in the low-pass case the frequency variable ω (i.e., $2\pi f$) is exactly equivalent to the band-pass-case frequency variable

$$\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \equiv \frac{\text{BW}}{f_0},$$

and in the low-pass case the quantity

$$1/(L_{\text{series}}C_{\text{shunt}})^{1/2}$$

is exactly equivalent to the well-known coefficient of coupling of band-pass-coupled-circuit theory; and the quantity L/R in a series arm and RC for a shunt arm are exactly equivalent to the well-known resonant-frequency Q of band-pass circuit theory. Thus to apply the equations of Figure 7 to a low-pass ladder, we do the following:

In place of BW_v/f_0 use ω_v.

In place of Q_1 use L_1/R_1 or R_1C_1.

In place of $K_{r(r+1)}$ use

$$1/(L_rC_{r+1})^{1/2}$$

or

$$1/(C_rL_{r+1})^{1/2}.$$

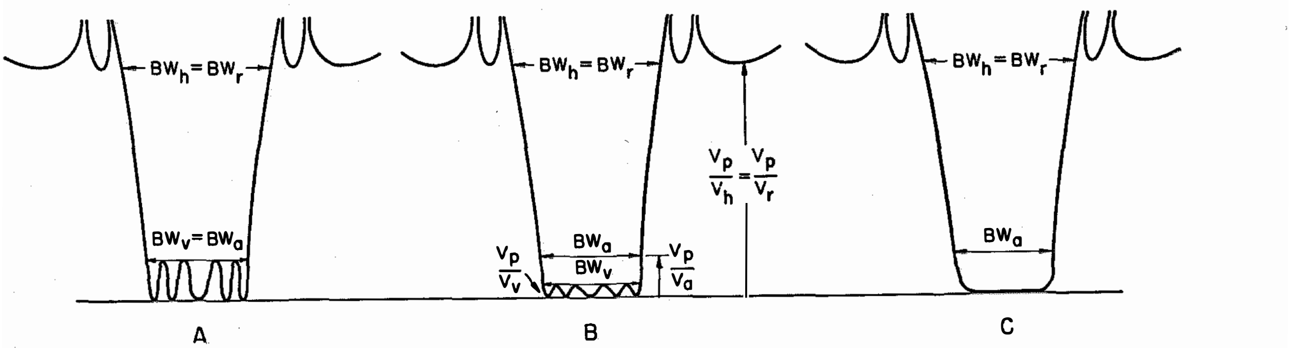


Figure 8—The optimum elliptic-function response shapes for m-derived-configuration filters. See Section 10.

10. *Corresponding Problem for m-Derived Configurations*

Exactly equivalent to the optimum attenuation shape of (3) for constant-k-configuration filters, there is an optimum attenuation shape for m-derived-configuration filters. This shape is shown in Figure 8 and is optimum in the sense that for a given allowable ripple in the accept band, a given minimum attenuation in the reject band (for a width equal to 10 accept bandwidths), and a given number of arms, it produces the sharpest possible rate of cutoff between the accept and reject bands.

Still to be discovered are the general design equations equivalent to those of Figure 7, which give the element values required to produce the optimum shapes to be described below. Just as the Chebishev shape of (3) changes in form for the limiting case of 0-decibel ripples and becomes (4), so does the optimum m-derived-shape equation change in form. Section 10.1 will consider the general equation and Section 10.2 will consider the limiting case of 0-decibel ripples in the pass band.

10.1 Optimum Attenuation Shape for m-Derived-Configuration Filters

All the following discussion will relate to the band-pass case and it will be assumed that the reader realizes it is immediately applicable to the low-pass case when the frequency variable ω is used in place of BW/f_0. As evolved by Norton and Darlington, plus a simple modification, the optimum shape is given by (14).

$$\left|\frac{V_p}{V}\right|^2 = 1 + \left(\left|\frac{V_p}{V_v}\right|^2 - 1\right)\left(\frac{cn}{dn}\right)_v \left[n\frac{K_v}{K_f}\left(\frac{cn}{dn}\right)_f^{-1}\frac{\mathrm{BW}}{\mathrm{BW}_v}\right], \quad (14A)$$

where

$$v = \left[\frac{(V_p/V_v)^2 - 1}{(V_p/V_h)^2 - 1}\right]^{1/2} \quad (15)$$

$$f = \frac{\mathrm{BW}_v}{\mathrm{BW}_h}. \quad (16)$$

For a given number of arms n, in (14), v and f cannot be picked independently but must always satisfy the relation between the three quantities given by (17).

$$\log q_v = n \log q_f, \quad (17)$$

where q is the so-called modular constant of the modulus given by the subscript. The function $\log q_k$ is tabulated on pages 49–51 of the 1945 edition of Jahnke and Emde; and the Smithsonian Elliptic Function Tables compiled by S. W. and R. M. Spenceley contains a very useful short appendix dealing with the numerical computation of the various elliptic functions. (It will be noted that for modulus values k less than 0.1 (say), the corresponding modular constant is simply $q_k = (k/4)^2$.) In (14), the symbol cn/dn stands for the ratio of the two elliptic functions cn over dn, where the subscript v or f is the modulus of the function; and K_v and K_f are the complete elliptic integrals of the first kind, evaluated for the modulus value given by the subscript.

It is evident from Figure 8 that the modulus v in (15) is immediately set by *two voltage ratios*, i.e. the maximum allowable ripple in the pass band V_p/V_v and the minimum required attenuation in the reject band V_p/V_h. Similarly, the modulus f in (16) is immediately set by the ratio of *two bandwidths*, i.e. the valley bandwidth BW_v and the "hill" bandwidth BW_h, which are the two bandwidths on the attenuation skirts where the attenuation is equal to the peak-to-valley ratio V_p/V_v and peak-to-hill-ratio V_p/V_h, respectively.

Equation (14A) takes on two different forms depending on the part of the attenuation characteristic in which one is interested. From the middle of the pass band out to the valley bandwidth, i.e. when $\mathrm{BW}/\mathrm{BW}_v$ is less than unity, (14A) stays as given. In the cutoff region between the edge of the valley bandwidth and the edge of the hill bandwidth, i.e. where $\mathrm{BW}/\mathrm{BW}_v$ varies between unity and $1/f$, (14A) turns into (14B).

$$\left|\frac{V_p}{V}\right|^2 = 1 + \frac{(V_p/V_v)^2 - 1}{dn_{v'}[n(K_v/K_f)dn_{f'}^{-1}(\mathrm{BW}_v/\mathrm{BW})]}. \quad (14B)$$

The prime indicates the complimentary modulus, i.e. $v' = (1 - v^2)^{1/2}$ and $f' = (1 - f^2)^{1/2}$. (Note also the inversion of the bandwidth ratio in the bracketed expression.)

Finally outside the hill bandwidth, i.e. where BW/BW_v is greater than $1/f$, (14A) turns into (14C).

$$\left|\frac{V_p}{V}\right|^2 = 1 + \frac{(V_p/V_h)^2 - 1}{(cn/dn)_{v'}{}^2[n(K_v/K_f)(cn/dn)_{f'}{}^{-1}(BW_h/BW)]}, \qquad (14C)$$

and of course the relations of (15), (16), and (17) apply to (14A), (14B), and (14C).

In the above equations, the writer has used the (cn/dn) elliptic function rather than the (sn) elliptic function because with the former a single equation (14) can be written that holds for both n odd *and* n *even.* Similarly, the important zero- and pole-location equations that follow will apply to both n odd and n even. Insofar as numerical work is concerned, the fact that $(cn/dn)u = sn\ (K - u)$ enables the cn/dn values to be obtained from the more-common sn tables.

It should be mentioned that when an even number of arms n is to be used, (14) calls for a finite attenuation at infinite frequency. To produce this phenomenon, it is necessary to make the numeric P_p/P_m in Step *6* of the Darlington procedure equal to zero; this will then produce a short-circuited termination for a network ending in a series arm, or an open-circuited termination for a network ending in a shunt arm, which will in turn produce finite attenuation at infinite frequency.

For both the direct procedure and the Darlington procedure, it is necessary to solve (14) for its $j(BW/BW_v)$ zeros; the left-half-plane zeros are given in (18), (18A), and (18B).

$$j(BW_0/BW_v)_m = -\ r_m{}^c \pm j i_m{}^c \qquad (18A)$$

$$r_m{}^c = \frac{[(sn/cn)_{f'}B^c][f'^2(sn/dn^2)_f A_m{}^c]}{1+f^2[(sn/cn)_{f'}{}^2B^c][(cn/dn)_f{}^2A_m{}^c]} \qquad (18B)$$

$$i_m{}^c = \frac{[(dn/cn^2)_{f'}B^c][(cn/dn)_f A_m{}^c]}{1+f^2[(sn/cn)_{f'}B^c][(cn/dn)_f{}^2A_m{}^c]}, \qquad (18C)$$

where the angles $A_m{}^c$ and B^c are given by (19).

$$A_m{}^c = (2m - 1)(K_f/n) \qquad (19A)$$

$$B^c = \frac{1}{n}\frac{K_f}{K_v}\left(\frac{sn}{cn}\right)_{v'}^{-1}\left[\left(\frac{V_p}{V_v}\right)^2 - 1\right]^{-\frac{1}{2}}. \qquad (19B)$$

It should be realized that, for modulus values as close to unity as is v' in practical cases, one can use the approximation $(sn/cn)^{-1} = \sinh^{-1}$.

Thus for lossless elements, the numerator polynomial of (5) required by Step *3* of both the direct and Darlington procedures is obtained by multiplying together a total of n factors each of the form

$$[j(BW/BW_v) - (-\ r_m{}^c \pm j i_m{}^c)],$$

where $r_m{}^c$ and $i_m{}^c$ are given by (18) and (19). This procedure will now give us the U coefficients of (5).

Next needed for the Darlington procedure are the zeros of the equation formed when the numeric P_p/P_m is subtracted from (14); the two most-common cases are considered in the next two paragraphs.

For resistive loading in both sides, impedance-matched output is usually desired at the peaks, and for this case $P_p/P_m = 1.0$. The $j(BW/BW_v)$ zeros required in Step *7* of the Darlington procedure are then for this impedance-matched case given by (20).

$$\left(j\frac{BW_0}{BW_v}\right)_m = 0 \pm j\left(\frac{cn}{dn}\right)_f (2m - 1)\frac{K_f}{n}. \qquad (20)$$

Thus the numerator polynomial of (6) required by Step *8* of the procedure is obtained by multiplying together a total of n factors each of the form

$$[j(BW/BW_v) - (\pm\ ji)],$$

where i is value of the j term in (20). This procedure will now give us the V coefficients of (6).

The input-immittance equations (7) and (8) can now be formed.

For a reactive load on one side of the network, $P_p/P_m = 0$, and so the zeros of Step *7* will be identical to those used for Step *3*, and thus the V

coefficients of (6) will be identical to the U coefficients of (5). Thus for this case as soon as the U's are obtained, (7) can be formed.

In the continued-fraction expansion procedure, it is necessary to know the frequencies of infinite attenuation and to assign these frequencies to specific arms of the network; the infinite-attenuation frequencies of (14) are given by (21).

$$\left(j\frac{\mathrm{BW}_\infty}{\mathrm{BW}_v}\right)_m = 0 \pm j\frac{1}{f(cn/dn)_f A_m{}^c}. \qquad (21)$$

It has not yet been possible to recognize the law of formation that would give a closed-form solution for the network values.

10.2 Limiting Case of No Ripples in the Pass Band of *m*-Derived-Configuration Filters

When the peak-to-valley ratio V_p/V_v in (14B) is made to approach 0 decibels and the equation is expressed in terms of the hill bandwidth BW_h instead of valley bandwidth, then in the limit the response equation becomes (22) and the shape is that of Figure 8C.

$$\left|\frac{V_p}{V}\right|^2 = 1 + \frac{(V_p/V_h)^2 - 1}{\cosh^2[n\cosh^{-1}(\mathrm{BW}_h/\mathrm{BW})]}. \qquad (22)$$

It is of interest to note that if one applies (22) to two points on the skirt and allows the V_p/V_h ratio to become infinite, then the Butterworth equation (4) results. Similarly, (14) would turn into the Chebishev equation (3) if the V_p/V_h ratio is correctly made to approach infinity.

Paralleling the paragraphs of Section 10.1 to obtain the required element values, we must solve (22) for its $j(\mathrm{BW}/\mathrm{BW}_h)$ zeros; the n left-half-plane zeros are given by (23), (23A), and (23B).

$$j(\mathrm{BW}_0/\mathrm{BW}_h) = -r_m{}^b \pm ji_m{}^b \qquad (23)$$

where

$$r_m{}^b = \frac{\sinh B^b \sin A_m{}^b}{\sinh^2 B^b + \cos^2 A_m{}^b} \qquad (23\mathrm{A})$$

and

$$i_m{}^b = \frac{\cosh B^b \cos A_m{}^b}{\sinh^2 B^b + \cos^2 A_m{}^b} \qquad (23\mathrm{B})$$

where the angles $A_m{}^b$ and B^b are given by (24).

$$\left.\begin{aligned} A_m{}^b &= (2m-1)(90°/n) \\ B^b &= (1/n)\sinh^{-1}[(V_p/V_h)^2 - 1]^{1/2}. \end{aligned}\right\} \qquad (24)$$

Thus the numerator polynomial of (5) required by Step *3* of both the direct and Darlington procedures is obtained for lossless elements by multiplying together a total of n factors of the form

$$[j(\mathrm{BW}/\mathrm{BW}_h) - (-r_m{}^b \pm ji_m{}^b)],$$

where $r_m{}^b$ and $i_m{}^b$ are given by (23). This procedure will now give us the required U coefficients of (5).

Next needed are the zeros of the equation formed when the numeric P_p/P_m is subtracted from (22).

For the impedance-matched case where $P_p/P_m = 1.0$, the $j(\mathrm{BW}/\mathrm{BW}_h)$ zeros of Step *7* are all zero, so that the numerator polynomial (6) of Step *8* is simply the one term $j(\mathrm{BW}/\mathrm{BW}_h)^n$ and all the V coefficients are zero. Thus for this attenuation shape and for the impedance-matched case, the input-immittance equations (7) and (8) can be formed as soon as the U's are obtained.

For the reactive-load case, $P_p/P_m = 0$ and so the V coefficients of (6) are identical to the U coefficients of (5), and as soon as the U's are obtained the input immittance (7) can be written.

As in Section 10.1, it is necessary to know the frequencies of infinite attenuation and to assign these frequencies to specific arms of the network. These infinite-attenuation frequencies of (22) are given by (25).

$$\left(j\frac{\mathrm{BW}_\infty}{\mathrm{BW}_h}\right)_m = 0 \pm j\frac{1}{\cos A_m{}^b}. \qquad (25)$$

It has not yet been possible to recognize the law of formation that would give a closed-form solution for the network values.

11. Postscript—Recent Publications on this Subject

In Section 8, it was mentioned that in the 20 years that have passed since Norton and Bennett obtained their solutions, no one published the more-complicated Chebishev solutions. The attention of the writer has been directed to the following three papers whose existence emphasizes the often-repeated phenomenon that when the time is ripe, a problem is usually solved almost simultaneously in many parts of the world.

Ernest Green has apparently accomplished the most, for in his paper,[9] equations (2), (3), (5), and (6) give the general solutions for the Butterworth and Chebishev responses for *any ratio of terminations* instead of merely for equal loading on both sides or for loading on one side only.

The first paper to be published with the general solution for the case of equal loading on both sides is that of Vitold Belevitch.[10]

Finally, H. J. Orchard[11] using an "intuitive" process that was probably similar to that used in Section 8 of this paper, obtained the solution for the case of loading on one side only.

12. Acknowledgment

The author takes this opportunity to express his appreciation to Mr. Georges Deschamps of Federal Telecommunication Laboratories for many productive discussions with him on some of the material in this paper.

[9] E. Green, "Exact Amplitude Frequency Characteristics of Ladder Networks," *Marconi Review*, volume 16, number 108, pages 25–68; 1953.

[10] V. Belevitch, "Tchebyshev Filters and Amplifier Networks," *Wireless Engineer*, volume 29, pages 106–110; April, 1952.

[11] H. J. Orchard, "Formulae for Ladder Filters," *Wireless Engineer*, volume 30, pages 3–5; January, 1953.

United States Patents Issued to International Telephone and Telegraph System, May-July, 1953

FROM May 1st to July 31st 1953, 76 patents were issued by the United States Patent Office to the International Telephone and Telegraph System companies. These patents are listed below, followed by summaries of several that may be of more than usual interest.

International Telephone and Telegraph Corporation now owns or has licensing rights under some 3000 United States patents, of which approximately 10 percent are in the field of telephony including automatic telephone switching equipment and subscriber apparatus, 50 percent in the field of radio and television including radio navigation, pulse, and other telecommunication systems, the remaining 40 percent being in various fields such as selenium rectifiers, vacuum tubes and other components, cables, and miscellaneous devices.

P. R. Adams, "Radio Echo-Pulse System for Vehicles Following a Fixed Route," 2,641,688.

P. R. Aigrain, "Volume Control for Pulse-Code Modulation," 2,643,293.

P. R. Aigrain, see also J. B. Lair.

R. P. Arthur, "Multipoint Plug and Jack," 2,640,183.

R. S. Bailey, see E. D. Phinney.

J. M. Baxter, "Automatic Phonograph Mechanism," 2,645,496.

J. I. Bellamy, "Selecting and Actuating Mechanism for Crossbar Switches," 2,643,299.

E. M. Bradburd, "Electron-Discharge Device and Circuits for Neutralized Coaxial-Line Amplifiers," 2,643,302.

F. H. Bray, G. C. Hartley, and D. S. Ridler, "Electrical Storage of Information in Gas-Filled Tubes," 2,638,506.

W. C. Bruckman, M. A. Kreitchman, and W. J. Olson, "Washing Machine Provided with Resilient Vibration-Control Door-Sealing Means," 2,643,538.

F. X. Bucher, see F. J. Lundburg.

A. M. Casabona, "Pulse Navaglide System," 2,640,982.

R. Chapman and R. H. Trier, "Loudspeaker Baffle with Elongated Aperture for the Egress of Sound," 2,646,851.

R. F. Cleaver and T. D. Gray, "Electric Signal Modulator," 2,640,973.

R. F. Cleaver, see also C. W. Earp.

D. R. Clements, "Synchronizing-Signal Separator," 2,640,103.

N. Cohen, see L. Goldstein.

C. T. Daly, "Electric Amplifier with Automatic Gain Control," 2,647,176.

G. Deakin, "Electromagnetic Relay Armature Mounting and Biasing," 2,641,665.

M. den Hertog, "Voice-Frequency Code Signaling System," 2,647,164.

B. Derjavitch, see S. van Mierlo.

L. Devaux, "Fascimile Telegraph Transmitter," 2,640,875.

G. C. Dewey, "Traveling-Wave Tube," 2,643,353.

S. H. M. Dodington, "Antenna System," 2,641,756.

C. F. Drake, "Metallization of Nonmetallic Surfaces," 2,639,997.

C. Dumousseau, see E. Touraton.

C. C. Eaglesfield, "Electric Pulse-Code-Modulation System of Communication," 2,640,965.

C. W. Earp and R. F. Cleaver, "Combined Radio Direction and Distance," 2,640,191.

P. T. Farnsworth, "Television Image-Analyzing Tube," 2,641,723.

S. Frankel, "Pulse Converting System," 2,643,331.

S. Frankel, "Broadband Loop Antenna," 2,642,529.

S. Frankel, see also E. Labin.

J. H. Fremlin and R. N. Hall, "Electron-Discharge Apparatus," 2,645,739.

H. Galley, "Ring-Oscillator Pulse-Producing Circuit," 2,642,526.

W. L. Garfield, "Arrangement for Indicating the Rate of Change of a Physical Effect," 2,645,755.

F. P. Gohorel, "Route-Selecting Automatic Telephone System," 2,642,499.

L. Goldstein, N. Cohen, and W. Sichak, "Control of Wavelength in Waveguide and Coaxial Lines," 2,641,702.

L. Goldstein, N. Cohen, and W. Sichak, "Gas-Discharge Transmission Arrangement," 2,643,297.

J. M. Gomez-Diez, "Method of Testing Electric Cable Sheaths," 2,642,739.

T. D. Gray, see R. F. Cleaver.

H. C. Habegger, "Record-Changing Device," 2,643,129.

L. B. Haigh, see W. Hatton.

R. N. Hall, see J. H. Fremlin.

C. E. Hallmark, "Pulse Counter," 2,641,694.

R. Hardy, "Panoramic Receiver," 2,645,711.

G. C. Hartley, and W. J. Reynolds, "Binary-Multiplying Circuit," 2,638,267.

G. C. Hartley and D. A. Weir, "Telecommunication Exchange System," 2,640,872.

G. C. Hartley, see also F. H. Bray.

W. Hatton, L. B. Haigh, and L. Kozma, "Calculator Equipment Working with Teleprinter," 2,645,420.

A. G. Kandoian, "Circularly Polarized Broadband Antenna," 2,640,928.

R. Kelly, see V. J. Terry.

L. Kozma, see W. Hatton.

M. A. Kreitchman and W. J. Olson, "Cabinet Structure for Washing Machines," 2,645,548.

M. A. Kreitchman, see also W. C. Bruckman.

J. Kruithof, "Electric Multiswitch," 2,640,884.

J. Kruithof, "Two-Motion Switching System," 2,640,882.

J. B. H. Kuper, "Microwave Amplifier of the Magnetron Type," 2,640,951.

E. Labin, S. Frankel, and M. Silver, "Local Transmitter Frequency-Control Circuit," 2,641,693.

E. Labin, "Pulsed Radio Remote-Control System," 2,645,771.

J. B. Lair and P. R. Aigrain, "Speech-Communication System," 2,640,880.

G. X. Lens, "Switching Mechanism," 2,647,166.

F. J. Lundburg and F. X. Bucher, "Antenna Assembly," 2,640,930.

N. Marchand, "Shiftable Directional Antenna," 2,640,192.

E. M. S. McWhirter and F. G. Popp, "Intelligence Exchange System," 2,645,764.

T. Moore, "Timing Device," 2,641,701.

W. H. Myers, "Phonograph-Spindle Aperture Reducer," 2,645,499.

J. J. Nail, "Elliptically Polarized Antenna," 2,643,337.

S. Nevin, "Heated-Cathode Electron Multiplier," 2,640,169.

W. J. Olson, see W. C. Bruckman and M. A. Kreitchman.

L. W. Parker, "Color Kinescope," 2,643,352.

E. D. Phinney and R. S. Bailey, "Amplifier and Receiver System," 2,640,917.

A. Piquet, see E. Touraton.

L. C. Pocock, "Compliant Supports for Transducer Diaphragms," 2,646,853.

F. G. Popp, see E. M. S. McWhirter.

R. B. Reade, "Method of Controlling Susceptance of a Post-Type Obstacle," 2,645,679.

A. H. Reeves, "Electric-Discharge Device," 2,645,742.

H. R. Rensch, "Circuit Arrangement for Conference-Call Installations," 2,644,039.

W. J. Reynolds, see G. C. Hartley.

D. S. Ridler, see F. H. Bray

M. Rose, "Terminal Pin Block," 2,643,365.

L. Rosenberg, see L. Stashover.

T. R. Scott and W. K. Weston, "Tube and Its Manufacture," 2,640,501.

W. Sichak, "Circularly Polarized Antenna Arrangement for Radar," 2,640,915.

W. Sichak, see also L. Goldstein.

M. Silver, "Tracking System between Receiver and Transmitter," 2,643,329.

M. Silver, see also E. Labin.

L. Stashover and L. Rosenberg, "Phase Detector," 2,640,939.

V. J. Terry and R. Kelly, "Electric Power-Supply Equipment," 2,642,558.

E. Touraton and C. Dumousseau, "Hyperfrequency Vacuum Tube," 2,640,172.

E. Touraton and C. Dumousseau, "Catcher Circuits for Velocity-Modulation Tubes," 2,647,219.

E. Touraton, A. Piquet, and C. Dumousseau, "Velocity-Modulation-Amplifier Tube," 2,641,730.

E. Touraton and R. Zwobada, "Ultrahigh-Frequency Electron Tubes," 2,640,112.

E. Touraton and R. Weill, "Electronic-Discharge Device," 2,641,725.

M. C. Tournier, "Bandpass Filter," 2,640,879.

R. H. Trier, see R. Chapman.

S. van Mierlo and B. Derjavitch, "Pulse Electronic Switching System," 2,638,505.

H. M. Veaux, "Multichannel Electrical Pulse-Communication System," 2,640,881.

M. W. Wallace, "Cold-Cathode Discharge Tube," 2,642,548.

R. Weill, see E. Touraton.

D. A. Weir, see G. C. Hartley.

W. K. Weston, see T. R. Scott.

J. H. Wilson, "Means and Method for Testing Terminal Banks," 2,642,481.

N. H. Young, "Method and Means for Transmitting Intelligence," 2,645,677.

R. Zwobada, see E. Touraton.

Loudspeaker Baffle with Elongated Aperture for the Egress of Sound

R. Chapman and R. H. Trier
2,646,851—July 28, 1952

A loudspeaker arrangement in which a housing baffle is provided. In the wall of this housing, adjacent to the loudspeaker diaphragm, there is provided an elongated opening, the width of which is of the order of one wavelength at the highest frequency for which maximum diffusion is desired. By the use of this elongated slot, a better distribution of sound throughout the entire frequency range is obtained than with the conventional loudspeaker.

Antenna Assembly[1]

F. J. Lundburg and F. X. Bucher
2,640,930—June 2, 1953

A radio antenna assembly for producing omnidirectional-range information comprising a dipole antenna of small length relative to a half-wavelength of the operating frequency mounted within a resonator that is open to radiation around its periphery; and an omnidirectional antenna mounted symmetrically around the dipole radiator. The omnidirectional radiator comprises elements arranged symmetrically about the dipole

[1] F. J. Lundburg and F. X. Bucher, "Cage-Type Very-High-Frequency Phase-Comparison Omnidirectional Radio Range Antenna," *Electrical Communication*, volume 29, pages 108–116; June, 1952.

and energized in phase. The resonator comprises the cage with plates above and below the dipole radiator interconnected by vertical rods that tend to remove vertical-polarization errors that may occur from the dipole. In addition, an arrangement is made for extending this resonator above and below its normal area to attenuate horizontal-radiation errors that may be created through radiation from the top and bottom plates.

Electric-Discharge Device

A. H. Reeves
2,645,742—July 14, 1953

A pulse-counting tube and circuit arrangement for providing counting by a sequential discharge of gaps within the tube. In order that a single triggering pulse may cause the tube to operate, a first gap is provided having a lower striking voltage than the other gaps; ionization coupling being provided between adjacent gaps. A source of voltage sufficient to maintain discharge is applied between the anode electrode and the various cathode elements forming the sequential gaps. Accordingly, an applied pulse will start a sequential train of output pulses, the number of which will depend on the number of gaps in the tube.

Traveling-Wave Tube

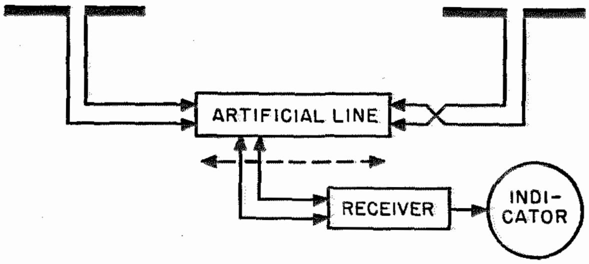
G. C. Dewey
2,643,353—June 23, 1953

A traveling-wave-tube amplifier arrangement using the normal form of electron-beam gun and a delay line for slowing down the phase velocity of applied energy so as to obtain the desired interaction between the beam and the energy to be amplified. The means for slowing the beam consists of conducting baffles positioned uniformly along either the outer wall or a central conductor of the coaxial line formed by the tube, the beam being transmitted along the edges of these baffles. To suppress angle-dependent modes of operation of the tube, a plurality of longitudinal slots are cut radially through the walls of the chamber defining the outer conductor of the tube.

Shiftable Directional Antenna

N. Marchand
2,640,192—May 26, 1953

A direction finder in which two spaced antennas are coupled in impedance-matching relation to an interconnecting transmission line. The line has a transposition so that when in-phase energy is received, a null will occur at the center of this line. The output tap points of a phase shifter provided in the transmission line are continuously moved in simple harmonic motion so that a sine-wave output is obtained. The zero of the output



will be displaced from the center of the line in accordance with the phase relation of energy from the spaced antennas. The output is applied to a cathode-ray-tube indicator.

Pulse Navaglide System

A. M. Casabona
2,640,982—June 2, 1953

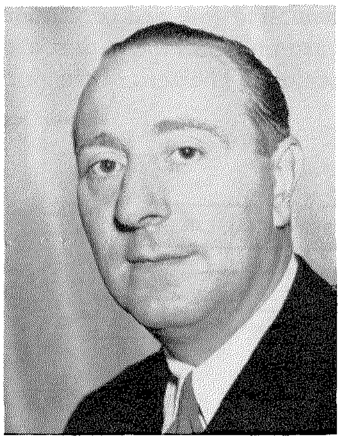
A radio-beacon arrangement that may simultaneously be used in distance measuring as well as in guiding a craft along a desired course; consisting essentially of a pair of antennas energized by a series of radio-frequency pulses. The antennas are spaced a desired fraction of a wavelength that will determine the subcarrier frequency for operation of the system. The complex pulses of radio-frequency energy produce the effective subcarrier-frequency energy. The antenna may be constructed for horizontal guiding of the craft and also for use in vertical glide-slope guidance.

Band-Pass Filter

M. C. Tournier
2,640,879—June 2, 1953

A band-pass filter consisting of a mean-frequency transformer having two similar circuits coupled by mutual induction and placed in the output or anode circuit of a tube and in the input or grid circuit of the following tube. These two circuits are associated with piezoelectric elements providing piezoelectric resonance transformers. One of these elements has a resonance that is slightly less than the lowest resonance of the mean-frequency transformer; the other resonant frequency being slightly higher than the resonant frequency of the mean-frequency transformer.

Contributors to this Issue



MAURICE ARDITI

MAURICE ARDITI was born on March 1, 1913, in Paris, France. In 1933, he received a degree in physics engineering at the Ecole de Physique et Chimie Industrielle of Paris, in 1934 a degree in electrical engineering at Ecole Supérieure d'Electricité, and in 1935 an M.S. degree in physics and chemistry from Sorbonne University, Paris.

From 1936 to 1939, Mr. Arditi did research work in physical chemistry at Paris University. In 1939, he joined the laboratories of Le Matériel Téléphonique, Paris, working on secondary emission, electron multipliers, and electron optics. He joined Federal Telecommunication Laboratories, New York, in 1944 and resumed work on special switching tubes. From 1947 to 1951, he was active in the field of microwave receivers for radio links and is presently engaged in the development of the microstrip line. Some of his studies on this subject are reported in the present issue.

Mr. Arditi is a Senior Member of the Institute of Radio Engineers.

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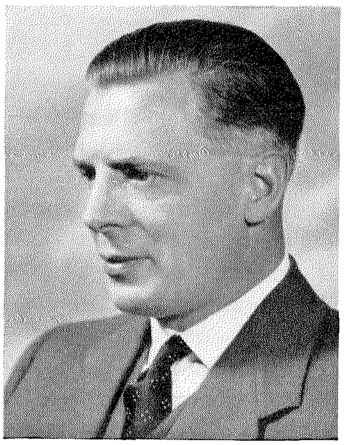

C. H. CHAMBERS

C. H. CHAMBERS received his technical training at the West Ham Technical College and later at the Northern Polytechnic.

He joined Standard Telephones and Cables in 1930 and has been actively engaged in the field of remote control and telemetering for power generation and transmission, on which subject some of his work is presented in this issue. Since 1946, he has been in charge of the planning and engineering of such systems.

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M. C. DETTMAN

M. C. DETTMAN was born in New York, New York, on March 5, 1919. He attended various courses at New York University, Polytechnic Institute of Brooklyn, College of the City of New York, and RCA Institutes.

From 1936 to 1937, Mr. Dettman was employed as supervisory engineer at National Recording Company in New York, New York. From 1937 to 1942, he was with Radiomarine Corporation of America working on all types of marine radio equipment. He was in charge of shipboard installations at Aruba, Dutch West Indies, and Venezuela, South America, in 1939.

In 1942, Mr. Dettman joined the staff of Federal Telecommunication Laboratories, where he is now department head and acting division head. He has been associated with work on shipboard and airborne pulse-time-modulation equipment, 500-megacycle color television, broadcast television transmitting equipment, navar beacon transmitters, and automatic-tuning shipboard communications transmitters. In this issue, Mr. Dettman discusses equipment of this last type. During 1944 and 1945, he also taught electronics at an industrial trade school.

Mr. Dettman is a member of the Institute of Radio Engineers.

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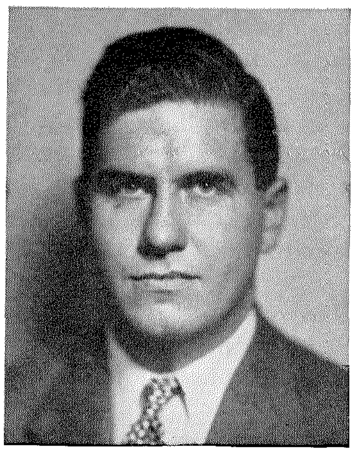

MILTON DISHAL

MILTON DISHAL was born on March 20, 1918, at Philadelphia, Pennsylvania. Temple University conferred on him the B.S. degree in 1939 and the M.A. degree in physics in 1941. He served as a teaching fellow in physics during the postgraduate years.

In 1941, he entered the employ of Federal Telecommunication Laboratories, where he is now a department head in charge of a group engaged in the development of radio receivers having special characteristics. Some of his studies on filters are presented in the present issue.

ROLAND H. DUNN

Mr. Dishal is a Senior Member of the Institute of Radio Engineers. He has been active in the work of the Radio Technical Committee on Aeronautics. He is also serving as a part-time instructor at Polytechnic Institute of Brooklyn.

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ROLAND HARRIS DUNN was born in Derbyshire, England, on December 22, 1906. He received the B.Sc. with Honours in electrical engineering in 1928 from the Faculty of Engineering of Manchester University.

He joined Standard Telephones and Cables in 1930 and has been associated with special systems in various capacities. In recent years, he has been in charge of special-systems development. In this issue he is a co-author of the paper on telemetering. Mr. Dunn became an Associate Member of the Institution of Electrical Engineers in 1935.

WILLIAM T. GIBSON

WILLIAM T. GIBSON was born in 1899 in Northhampton, England. Educated at Trinity College, Cambridge, he received the M.A. and B.Sc degrees. He returned to the university after a period of service during the first world war in the Royal Engineers, for which he received the O.B.E. award. In 1922, he did research work in Cavendish Laboratory under Professors Rutherford and J. J. Thompson.

His entire professional career was with the International System. He set up the valve laboratory of Le Matérial Téléphonique in 1928, and of Federal Telegraph Company in 1932, and was a consultant on similar activities to other associated companies of the International System.

At the time of his death on December 27, 1952, Mr. Gibson was chief valve engineer of Standard Telephones and Cables, Limited, and manager of the Ilminster laboratories and factory. One of his last papers, appearing in this issue, described the manufacture of thermistors.

Mr. Gibson was a Member of the Institution of Electrical Engineers.

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ARMIG G. KANDOIAN received the M.S. degree in electrical engineering from Harvard Graduate School of Engineering in 1935.

He has been associated with the International System since 1935 and has done extensive work on antennas, transmission lines, measurements, and on various problems connected with radar, communications, and navigation. In this issue, he is a co-author of the paper on helical antennas. He is head of the radio and radar components division of Federal Telecommunication Laboratories.

Mr. Kandoian received the honorable mention award of Eta Kappa Nu in 1943. He is a Fellow of the Institute of Radio Engineers and a member of the American Institute of Electrical Engineers and Harvard Engineering Society.

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WILLIAM SICHAK was born on January 7, 1916, in Lyndora, Pennsylvania. He received the B.A. degree in physics from Allegheny College in 1942.

From 1942 to 1945, he was engaged in developing microwave radar antennas at the Radiation Laboratory

ARMIG G. KANDOIAN

of Massachusetts Institute of Technology. Since then, he has been with Federal Telecommunication Laboratories, working on microwave antennas and allied equipment. In this issue, he is co-author of a paper on wideband helical antennas.

Mr. Sichak is a department head in the radio and radar components division of the laboratories. He is a Member of the Institute of Radio Engineers and of the American Physical Society.

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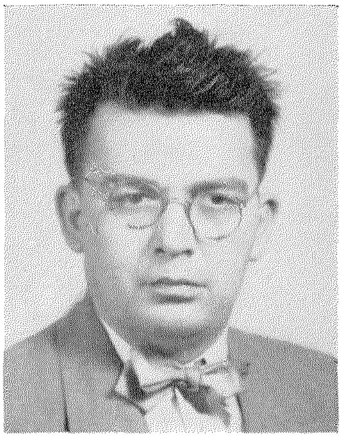
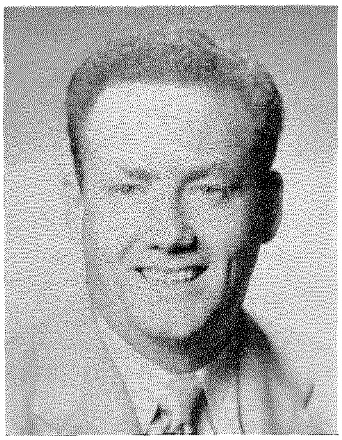
ROGER E. WHITE was born on November 2, 1920, in Lancaster, New Hampshire. He received the B.S. degree in engineering physics from the University of Maine in 1942. During 1944–1950, he did graduate work at the University of Maryland.



WILLIAM SICHAK



Roger E. White

Mr. White was with the Naval Research Laboratory from 1942 to 1952, working in the fields of aerial navigation, communication, countermeasures, and radar.

He has been with Federal Telecommunication Laboratories since 1952, serving as department head in the electron-tube laboratory, where he is engaged in traveling-wave-tube work. Some of his work on matching sections for these tubes is reported in the present issue.

Mr. White is a Senior Member of the Institute of Radio Engineers and a member of Tau Beta Pi and Phi Kappa Phi.

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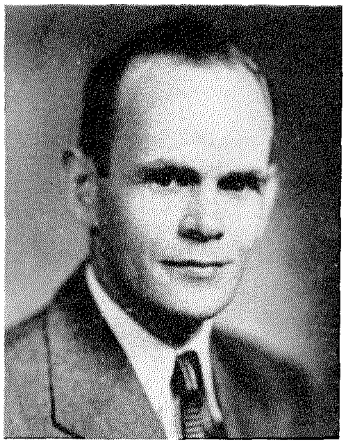

N. H. Young

Norman H. Young was born in Philadelphia, Pennsylvania, in 1913. He received the B.S. degree in electrical engineering from Pennsylvania State College in 1934 and the M.S. degree in 1935.

From 1935 to 1942, he was engaged in television engineering for the Philco Corporation and had charge of the transmitter of television station WPTZ.

In 1942, he became a department head in Federal Telecommunication Laboratories and during the war was largely concerned with the application of pulse-time modulation to military communication equipment. At the termination of the war, he was responsible for the engineering of the color-television transmitter for the Columbia Broadcasting System. He has done additional work on receivers and studio equipment for color television.

He is presently responsible for the development of automatically tuned transmitters of high accuracy for military service. A paper on the tuning system for transmitters of this type appears in this issue.

Mr. Young is a member of Eta Kappa Nu and a Senior Member of the Institute of Radio Engineers.

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Austral Standard Cables Pty. Limited, Melbourne

NEW ZEALAND — New Zealand Electric Totalisators Limited, Wellington

South America

ARGENTINA — Compañía Standard Electric Argentina, S.A.I.C., Buenos Aires

BRAZIL — Standard Electrica, S.A., Rio de Janeiro

CHILE — Compañía Standard Electric, S.A.C., Santiago

Europe

AUSTRIA — Vereinigte Telephon- und Telegraphenfabriks A. G., Czeija, Nissl & Co., Vienna

BELGIUM — Bell Telephone Manufacturing Company, Antwerp

DENMARK — Standard Electric Aktieselskab, Copenhagen

FRANCE —
Compagnie Générale de Constructions Téléphoniques, Paris
Le Matériel Téléphonique, Paris
Les Téléimprimeurs, Paris

GERMANY —
C. Lorenz, A.G., Stuttgart
Mix & Genest A. G., and Subsidiaries, Stuttgart
G. Schaub Apparatebau G.m.b.H., Pforzheim
Süddeutsche Apparatefabrik G.m.b.H., Nuremberg

ITALY — Fabbrica Apparecchiature per Comunicazioni Elettriche, Milan

NETHERLANDS — Nederlandsche Standard Electric Maatschappi N.V., The Hague

NORWAY — Standard Telefon og Kabelfabrik A/S, Oslo

PORTUGAL — Standard Eléctrica, S.A.R.L., Lisbon

SPAIN —
Compañía Radio Aérea Marítima Española, Madrid
Standard Eléctrica, S.A., Madrid

SWEDEN — Aktiebolaget Standard Radiofabrik, Stockholm

SWITZERLAND — Standard Telephone et Radio S.A., Zurich

TELEPHONE OPERATIONS

BRAZIL — Companhia Telefônica Nacional, Rio de Janeiro

CHILE — Compañía de Teléfonos de Chile, Santiago

CUBA — Cuban American Telephone and Telegraph Company, Havana

CUBA — Cuban Telephone Company, Havana

PERU — Compañía Peruana de Teléfonos Limitada, Lima

PUERTO RICO — Porto Rico Telephone Company, San Juan

CABLE AND RADIO OPERATIONS

UNITED STATES OF AMERICA —
American Cable & Radio Corporation; New York, New York
All America Cables and Radio, Inc.; New York, New York
The Commercial Cable Company; New York, New York
Mackay Radio and Telegraph Company; New York, New York

ARGENTINA —
Compañía Internacional de Radio, Buenos Aires
Sociedad Anónima Radio Argentina, Buenos Aires (*Subsidiary of American Cable and Radio Corporation*)

BOLIVIA — Compañía Internacional de Radio Boliviana, La Paz

BRAZIL — Companhia Radio Internacional do Brasil, Rio de Janeiro

CHILE — Compañía Internacional de Radio, S.A., Santiago

CUBA — Radio Corporation of Cuba, Havana

PUERTO RICO — Radio Corporation of Porto Rico, San Juan

RESEARCH

UNITED STATES OF AMERICA —
Federal Telecommunication Laboratories; Nutley, New Jersey; *a division of International Telephone and Telegraph Corporation*

International Telecommunication Laboratories, Inc.; New York, New York

ENGLAND — Standard Telecommunication Laboratories, Limited, London

FRANCE — Laboratoire Central de Télécommnnications, Paris

ASSOCIATE LICENSEES FOR MANUFACTURE AND SALES IN JAPAN

Nippon Electric Company, Limited, Tokyo

Sumitomo Electric Industries, Limited, Osaka