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ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

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BRITISH TELEVISION RELAY NETWORK

PRODUCTION–LINE SOLDERING AND BRAZING

NETHERLANDS–DENMARK COAXIAL-CABLE SYSTEM

SOME APPLICATIONS OF COLD-CATHODE TUBES TO SWITCHING SYSTEMS

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INTERNATIONAL · TELEPHONE AND TELEGRAPH ·

Volume 29 **SEPTEMBER, 1952** **Number 3**

ELECTRICAL COMMUNICATION

Technical Journal of the
INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION
and Associate Companies

H. P. WESTMAN, Editor
J. E. SCHLAIKJER, Assistant Editor

Published Quarterly by the
INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION
67 BROAD STREET, NEW YORK 4, NEW YORK, U.S.A.

Sosthenes Behn, Chairman William H. Harrison, President
Geoffrey A. Ogilvie, Vice President and Secretary

Subscription, $2.00 per year; single copies, 50 cents
Electrical Communication is indexed in Industrial Arts Index

Volume 29 SEPTEMBER, 1952 Number 3

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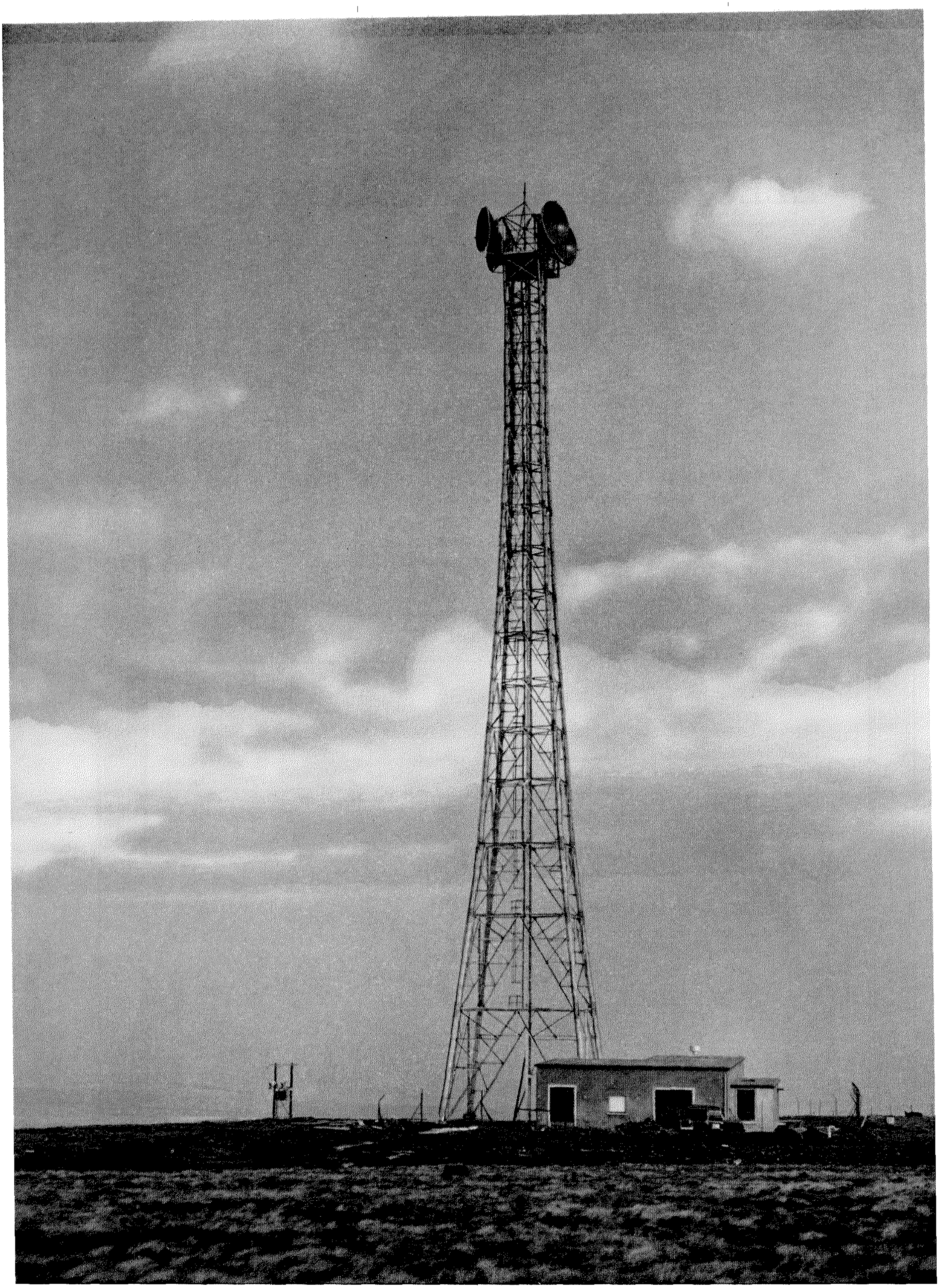


The 200-foot steel tower of the repeater station at Blackcastle Hill near Dunbar, Scotland. This is the tallest tower along the relay chain, the height being necessary due to the hilly terrain to be traversed to the southward.

British Television Relay Network

THE TELEVISION development programme of the British Broadcasting Corporation envisions the availability of television service to over 75 per cent of the population of the United Kingdom by 1953. With the opening of the station at Kirk o'Shotts, Scotland, early this year, the greater part of the programme has been brought to reality with service now available to about 68 per cent of the nation.

Since the post-war re-opening in 1946 of the station at Alexandra Palace, near London, the British television network has been extended to include first, a station at Sutton Coldfield, near Birmingham (1949), another at Holme Moss, near Manchester (1951), and now into Scotland with a transmitter at Kirk o'Shotts, near Edinburgh. The accompanying map illustrates the extent of the network, which is now more than 400 miles (644 kilometres) in length.

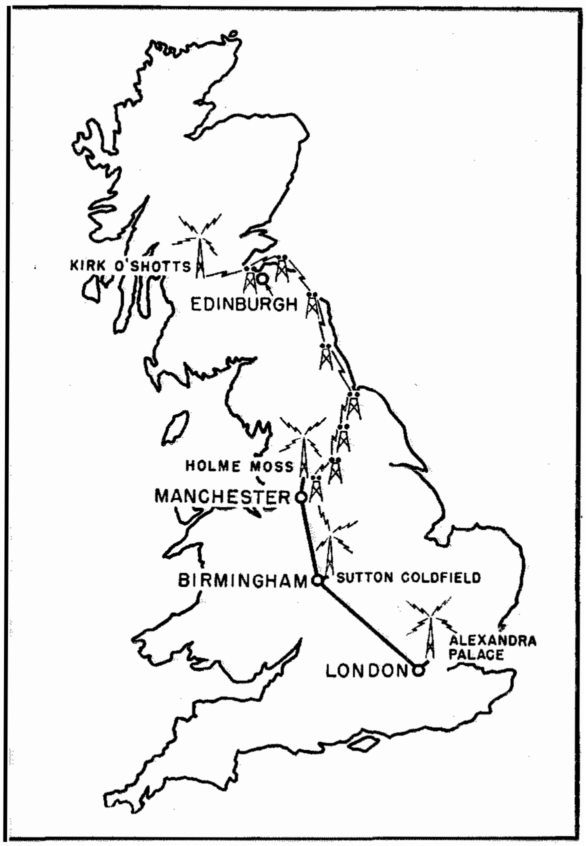
Standard Telephones and Cables, Limited, have contributed considerably to the network development, including the manufacture and installation for the British Post Office of the coaxial cables between London, Birmingham, and Manchester, the terminal equipment for the Birmingham–Manchester cable, and lately, the microwave radio relay system linking the Manchester and Edinburgh stations.

It is proposed in this paper to give a short description of the design and characteristics of the various links in this television network, with special emphasis on the features of the new radio relay system.

1. London—Birmingham Coaxial Cable

The first step in the post-war development plan of the British Broadcasting Corporation was the provision of a high-powered transmitting station at Birmingham to supply service in the Midland area.

Two links between London and the Sutton Coldfield station were projected and ordered by the Post Office for purposes of comparison—one link by means of a coaxial cable and the other by a radio relay system. At the time of writing, the television service is being provided by the coaxial cable, which was constructed and laid by Standard Telephones and Cables, Limited, while the repeaters for the cable were designed, made, and installed by the Post Office.[1]



The above map illustrates the British television relay network at its present stage of completion.

The unique feature of this cable is that it contains two tubes of 0·975-inch (2·48-centimetre) diameter that are capable of transmitting an extremely wide band of frequencies to cater for television signals of very high definition, for colour television, or for any future development requiring the transmission of a very wide bandwidth.

At present, the repeaters are spaced at 12-mile (19-kilometre) intervals but an ultimate repeater spacing of 3 miles (4·8 kilometres) is envisaged.

[1] H. Stanesby and W. K. Weston, "London–Birmingham Television Cable," *Electrical Communication*, v. 26, pp. 186–200; September, 1949.

In addition to the two large tubes—one for each direction of transmission—there are four smaller tubes of the standard 0·375-inch (0·953-centimetre) dimension, with repeaters spaced at 6-mile (9·6-kilometre) intervals. These tubes are satisfactory for the transmission of 405-line television signals but are actually in use for telephone channels.

The length of the main portion of the cable is 121·1 miles (194·9 kilometres), but at each end are shorter lengths providing connection to the television transmitters and to the switching centres.

2. *Birmingham—Manchester Coaxial Cable*

The second stage of the programme was provided by a coaxial cable laid from Birmingham to Manchester, a distance of 71 miles (114·2 kilometres), with a spur cable between Manchester and the transmitter at Holme Moss.

The main cable contains six 0·375-inch (0·953-centimetre) tubes, of which two are at present used for transmission of television signals, 16 paper-insulated quads for supervision and control of the repeater stations, and 172 paper-insulated quads for local telephone traffic. The repeater stations, with equipment designed and installed by the Post Office, are spaced at about 6-mile (9·7-kilometre) intervals.

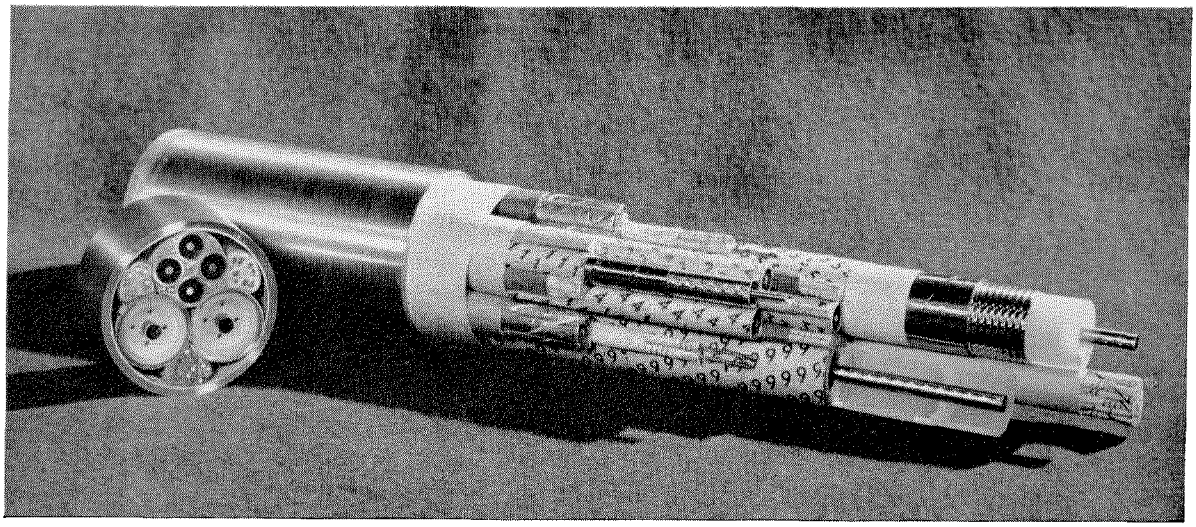
The spur cable between Manchester and the transmitter contains 2 coaxial tubes with 4 screened pairs for sound transmission and 24 paper-insulated quads for repeater control.

Vestigial-sideband terminal equipment is used with this cable to translate the video signal into a form suitable for transmission over the tubes. The principle consists in the modulation of a carrier with the video band, and the transmission over the cable of one complete sideband, the carrier, and only a vestige of the other sideband. This arrangement enables the video signal to be shifted to an appropriate part of the frequency spectrum without increasing the bandwidth to twice the video bandwidth as would be required by double-sideband practice. Transmission of television signals in a strictly single-sideband fashion presents very great technical difficulties that are obviated with vestigial-sideband transmission at the expense of only a small increase in bandwidth.

3. *Manchester—Edinburgh Radio Link*

The latest link of the network to be completed is the microwave radio relay chain between Manchester and Kirk o'Shotts, near Edinburgh. This system was brought into operation on March 14, 1952, but was actually used nearly a month earlier on an experimental basis to permit Scottish viewers to witness the funeral services of the late King George VI.

On this route the Post Office did not foresee a need for additional trunk telephone facilities in



The cable between London and Birmingham contains two large coaxial tubes and a group of four smaller tubes interspersed with five quads. At present, the large tubes are used for television programmes. Screened pairs for broadcasting and quads for supervisory circuits make up the other three sets of wires.

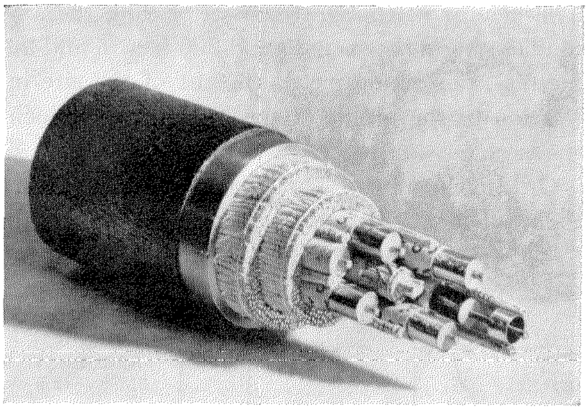
the near future; furthermore, the terrain lends itself well to the planning of a radio relay system. It was therefore decided to adopt the latter alternative rather than to lay a 2-tube cable for television alone. A two-way radio relay system employing 7 repeater stations was routed from Manchester towards the east coast, north to the Firth of Forth, and thence inland to Edinburgh and Kirk o'Shotts. The length of the route is about 250 miles (400 kilometres). Provision has been made at repeater stations to inject into the circuit programmes originating at major cities along the route. The repeaters normally operate unattended, with control and supervision of repeaters performed from each of the terminals.

The system operates on a frequency of 4000 megacycles, a wavelength of 7·5 centimetres. The line-of-sight paths between the repeaters were surveyed before construction was started by the use of Standard Telephones and Cables' portable microwave links[2] for determination of the heights required for each antenna at sites that in many cases could not be chosen solely from considerations of propagation effects. The average spacing between repeaters over the route is about 31 miles (50 kilometres).

Equipment at terminal and repeater stations affords one uni-directional channel in each direction. Each channel is designed to handle the 0-to-3-megacycle bandwidth of a 405-line, 50-frame, interlaced television signal.

To ensure a high degree of reliability, all transmission equipment with the exception of paraboloidal reflectors and main waveguide feeds is provided in duplicate; the change-over from one set of equipment to the other being fully automatic in the event of a breakdown. In the same way, automatic stand-by sources of power supply have been provided in case of a breakdown in local public mains supply.

The terminal transmitting equipment at each end of the link is designed to accept the video-frequency bandwidth of 0–3 megacycles at a peak-to-peak voltage of 1 volt in an impedance of 75 ohms. This video signal is applied to the reflector–cathode circuit of a velocity-modulation tube oscillating at a basic frequency of 4000



The cable between Birmingham and Manchester contains 6 coaxial tubes, 16 quads for supervisory control, and 172 quads for local telephone traffic.

megacycles. The resulting frequency excursion extends to 6 megacycles for a change in the video input from the bottom of a synchronising pulse to the peak white-picture signal.[3] The basic oscillator-tube frequency is stabilised by a precision crystal oscillator operating at a frequency of the order of 20 megacycles.

A travelling-wave amplifier provides the final amplification at all stations before the modulated signals are passed by waveguide to the transmitting antenna. The travelling-wave tube is a most important and powerful new tool in the development of centimetre-wave technique. It provides substantial amplification over a very wide frequency band, its operation depending on the interaction between the electric field of a stream of electrons beamed through the centre of a wire spiral and the electric field of a signal travelling along the spiral. This wartime invention has been developed to the commercial stage by Standard Telephones and Cables, Limited, and so far as is known the Manchester–Kirk o'Shotts link is its first application in public service.

The antennae for both the terminals and repeaters of the link consist of 10-foot (3-metre) paraboloids fed by waveguide horns. The width of the beam is 1·5 degrees between half-power points, and the transmitted power is 1 watt.

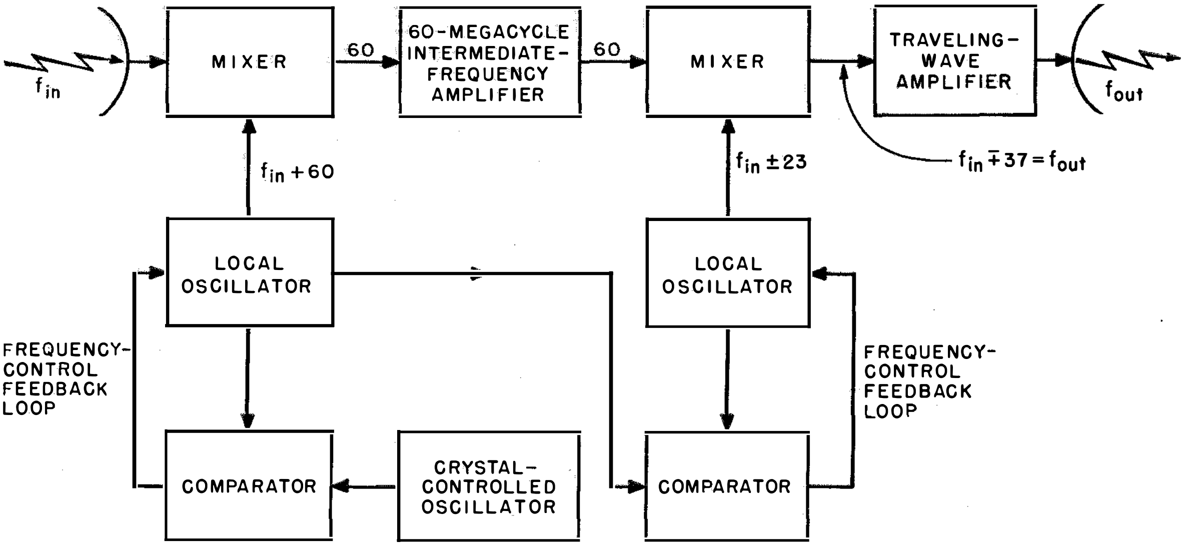
[2] "Portable Microwave Television Link," *Electrical Communication*, v. 27, pp. 295–297; December, 1950.

[3] British television standard where peak white = 70 per cent of signal acting positively and synchronizing pulse = 30 per cent of signal acting negatively, both with respect to the black level.

The heights of the steel towers bearing the antennae ranges between 20 and 200 feet (6 and 60 metres), depending on the height of the site in relation to the profile of the terrain over which the transmission passes.

On arrival at the first repeater in the chain, the wave is passed from the receiving antenna to for amplification, and then detected to provide a 1-volt peak-to-peak signal for application to a television transmitter or to a coaxial cable system.

The waveguides and their circuit elements that carry the 4000-megacycle signals are fabricated out of brass or copper, and have a rec-



Block diagram of the frequency-transforming and amplifying equipment at a repeater for one direction of transmission. The stand-by equipment is not indicated here.

a superheterodyne receiver. This receiver employs a crystal-controlled local oscillator operating at a frequency 60 megacycles above the mid-frequency of the incoming frequency-modulated transmission. The 60-megacycle intermediate-frequency signals are applied to an amplifier fitted with automatic gain control, and then again translated to a frequency 37 megacycles above or below the incoming signal.

Unwanted products of the last process are rejected in a filter, and the desired sideband is passed to a travelling-wave amplifier and thence to the transmitting antenna. The frequencies transmitted from successive repeaters are alternated by multiples of 37 megacycles to prevent possible interference between repeaters.

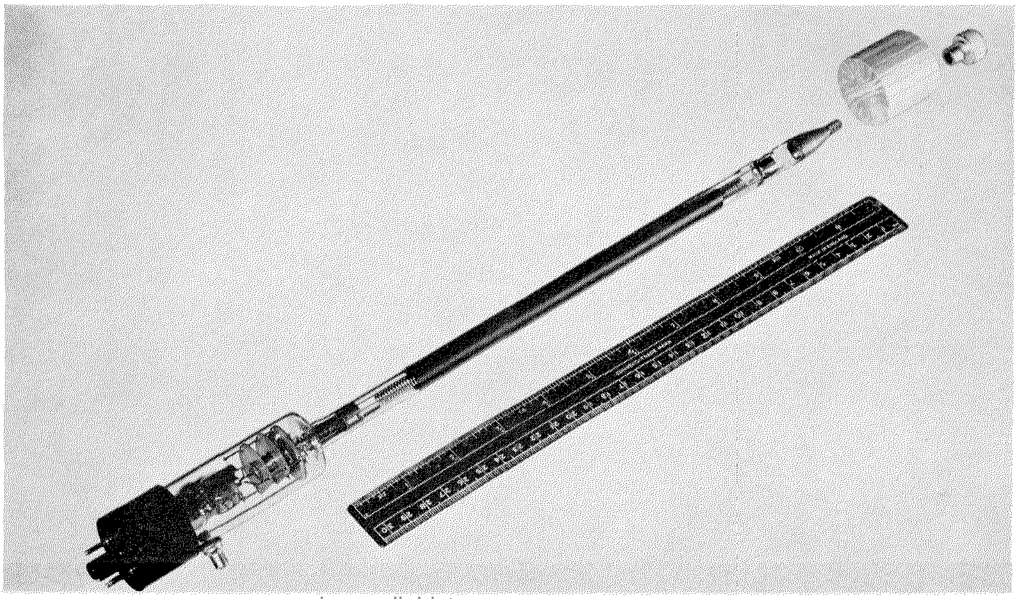
This process occurs at each successive repeater until the farther terminal of the link is reached, where the incoming 4000-megacycle signal is applied to a receiver, converted to 60 megacycles tangular cross-section of 2 by $\frac{2}{3}$ inches (5·1 by 1·7 centimetres). The waveguides from the equipment racks up to the antennas are filled with dry nitrogen under slight pressure to prevent corrosion of the inside surfaces. Inside the equipment racks, it would be impractical to fill the guides with nitrogen, and therefore the waveguide surfaces here are gold plated.

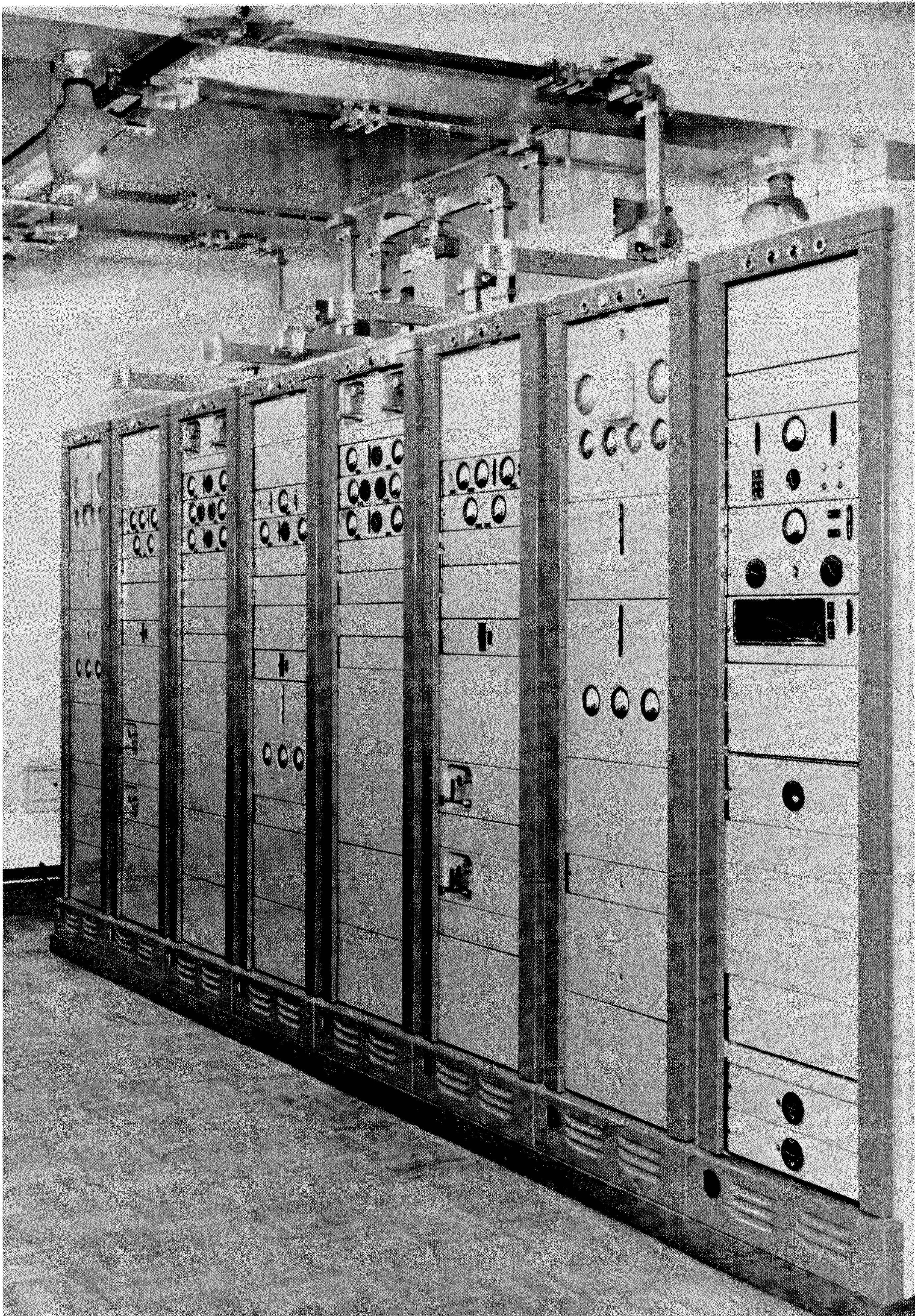
A 4-wire land-line telephone circuit interconnects all repeaters and terminals of the system to provide a party-line circuit and a 2-tone voice-frequency-telegraph circuit that carries the supervisory and remote-control signals. The remote-control desks at each terminal control the apparatus at repeaters which transmit television signals in the direction toward the terminal in which the desk is installed.

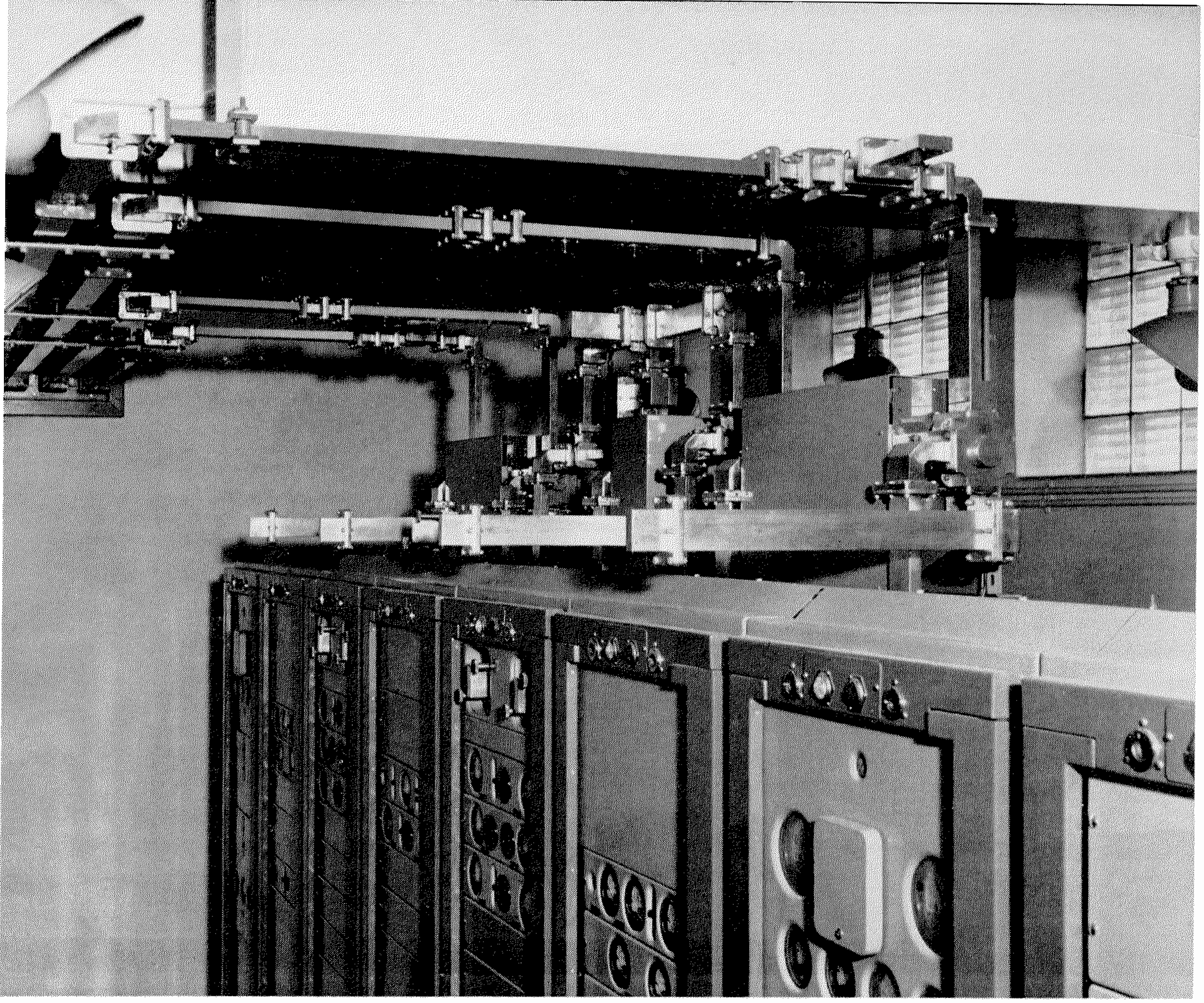
On the following few pages are presented a series of photographs illustrating some of the equipment comprising this new radio relay link.

The steel tower of a repeater station built on Blackford Hill, near Edinburgh.

The travelling-wave amplifier has for the first time been put into public service on the Manchester—Edinburgh radio relay link. The tube is cooled by a stream of air directed through the radiator (shown dismounted at the right).







The waveguide switching arrangement above a suite of cubicles at a repeater. The four lines at the left run up the tower to the antennae.

On the facing page is a suite of apparatus cubicles at one of the repeater stations. A second set mounted at the back of this one contains the stand-by apparatus.

At the right, this waveguide terminates in a horn feed at the focal point of the 10-foot paraboloid. The small tube carries nitrogen for pressurizing the waveguide.

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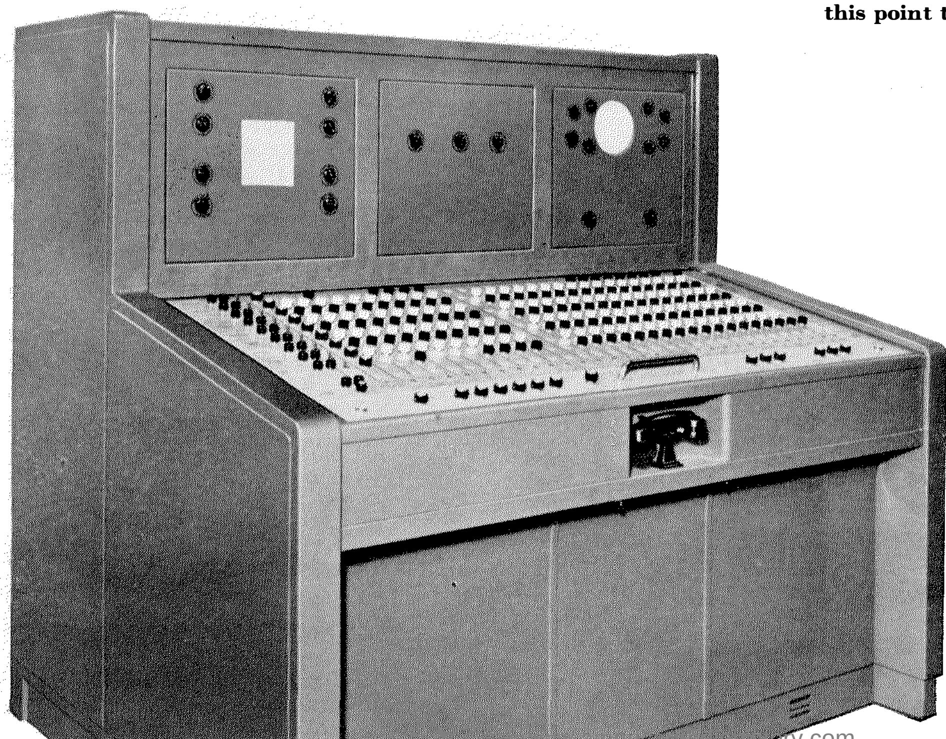
The terminal relay station at Kirk o'Shotts. short coaxial cable carries the video signals fro this point to the adjacent broadcast transmitt



Full supervision of the incom ing signals may be exercise from the control desk at Kir o'Shotts. The remote control for the apparatus at the vari ous repeaters terminate her

Production-Line Soldering and Brazing*

By C. E. EADON–CLARKE

Standard Telephones and Cables, Limited; Footscray, England

IN LIGHT as well as in heavy engineering, it has become increasingly common to fabricate parts by brazing or soldering relatively simple components together. By far the most widely employed tools for these operations are the soldering iron and the blowpipe, but in recent years the need for brazing and soldering on the production line has led to the development of three methods that can be push-button operated and process controlled for use by semi-skilled personnel.

• • •

Soldering and brazing are processes for joining metals in which molten filler material is drawn by *capillary attraction* into the space between the adjacent surfaces of the parts to be joined. In general, the term soldering is applied to the process when the filler material flows below 500 degrees centigrade and brazing when above this value. To ensure sound mechanical as well as electrical joints, the surfaces must have the correct spacing, be clean, and a suitable flux employed to assist wetting.

These processes differ from welding, with which brazing is often confused, in that welding is an operation in which the metallic parts are joined by heat and pressure fusion or by the addition of filler metal *direct* to the joint.

1. *Processes*

Of all the various possible methods of applying the necessary heat to the parts to be fastened together, the oldest and most widely known are the soldering iron and the gas blowpipe. It is felt that no further discussion of these systems is necessary here, but Table 1 has been prepared to compare them with three new systems that have come into widespread use in the last decade. These new methods, the electro-gas, electrical-resistance, and radio-frequency heating systems, all are characterized by the fact that the operation is almost completely automatic, requiring relatively little skill for production use. It might be surmised that one of these systems would be most effective for any particular job under consideration, and this is indeed true, as is illustrated

* Portions of this paper have been reprinted from an article by C. E. Eadon-Clarke on "'Push-Button' Soldering and Brazing" published in *Welding and Metal Fabrication*, v. 19, pp. 247–252; July, 1951.

TABLE 1

COMPARISON OF VARIOUS SOLDERING AND BRAZING METHODS

	Soldering Iron	Blowpipe	Electro-Gas Equipment	Resistance Machine	Radio-Frequency Equipment
Surface Pattern	Immaterial	Immaterial	Immaterial	Clean and smooth	Immaterial, but preferably not too irregular
Heat Generated	By conduction	By convection	By convection	Throughout body and by conduction from electrodes	By induction on outside surface only
Contact Required	Yes	No	No	Yes	No
Control of Heated Area	Only area in contact with bit	Moderately localized	Very localized	Very localized	Very localized
Process Control	Manual	Manual	Process timed or moving conveyor	Process timed	Process timed or moving conveyor
Suitability for Production Quantities	Flexible and ideal for operations that cannot be readily jigged	Flexible, suitable for small quantity batches	Flexible, suitable for small or large quantities of jobs that can be handled in production line	Medium or large quantities; moderately flexible	Large-scale production, particularly for long runs
Capital Cost	Very low	Very low	Low	Low	Moderately high
Type of Personnel Required for Maintenance	General factory engineer	General factory engineer	General factory engineer	General factory engineer	Routine by factory engineer, but visits by makers' engineers needed periodically

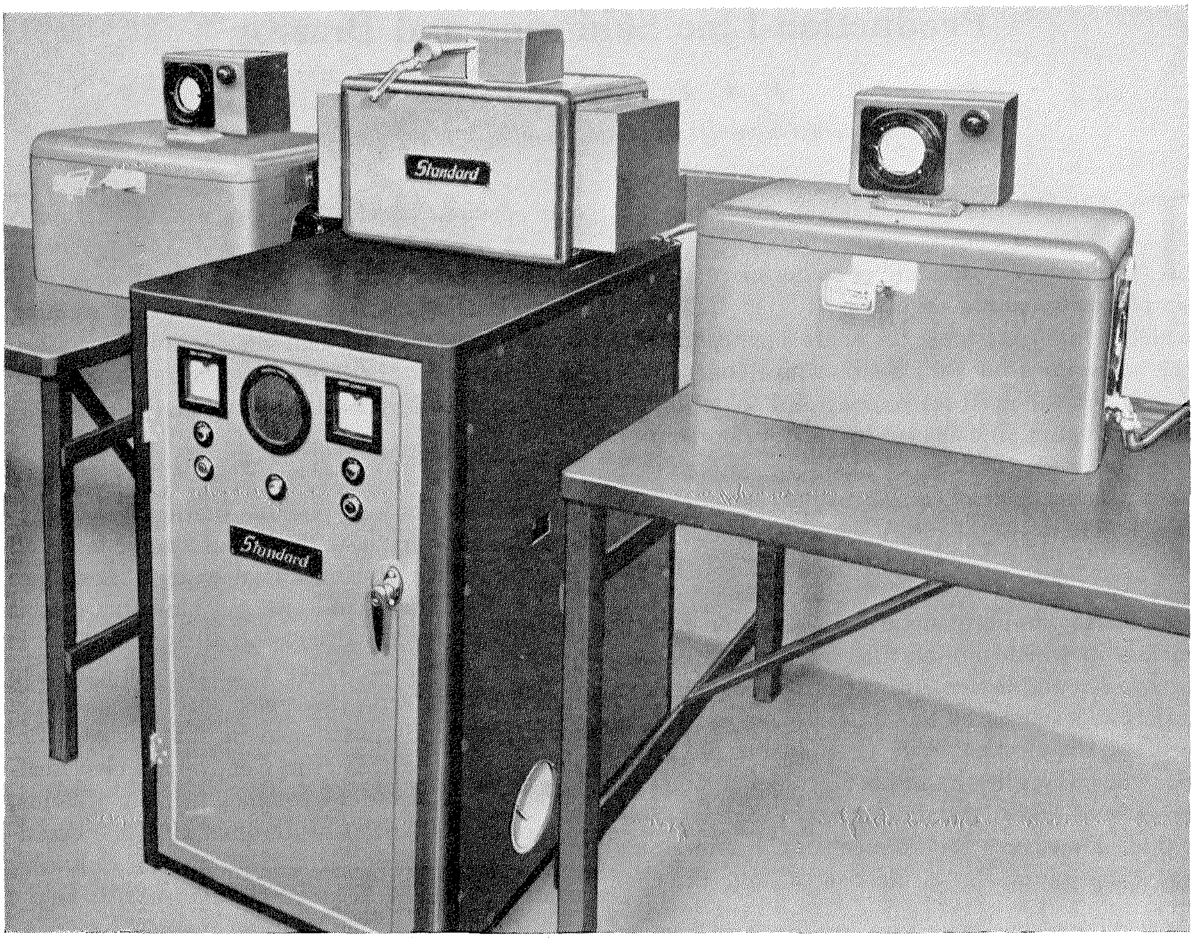


Figure 1—A 4-kilowatt induction-heating equipment with dual work stations. The stations are used alternately, but since each has individual controls, differing brazing or soldering work can be treated without changing the control settings.

TABLE 2

APPLICABILITY OF METHODS TO VARIOUS TYPES OF JOBS

Type of Work to be Handled	Soldering Iron	Blowpipe	Electro-Gas Equipment	Resistance Machine	Radio-Frequency Equipment
Soldering of Light Tin-Plate Box Work, Lids, Side Seams, Tagger Plates, Terminal and Wire Joining	Most suitable	Too violent unless flame is kept in motion	May be too violent since flame is stationary	Hand tools suitable in some cases, but component will often not stand pressure of electrodes for machine operation	Usually not possible
Soldering of Rigid Tin-Plate Structures, Terminals into Lids. Details where Solder is Pre-placed	Suitable in many cases	Too violent unless flame is kept in motion	May be too violent since flame is stationary	Suitable	Suitable for large quantities
Soldering of Heavy Section Details	Not suitable	Suitable for ferrous or non-ferrous materials	Suitable for ferrous or non-ferrous materials	Suitable for ferrous and even better for non-ferrous materials	More suitable for ferrous than non-ferrous loads
Brazing Medium and Heavy Section Details. Alloy Flowing Up to 650 Degrees Centigrade	Not possible	Suitable for all materials	Suitable for all materials	Suitable for all materials, but particularly for non-ferrous	More suitable for ferrous than non-ferrous loads. Speed limited by time for heat to conduct inwards
Brazing Medium and Heavy Section Details. Alloy Flowing Above 650 Degrees Centigrade	Not possible	Suitable for all materials	Suitable for all materials	Not suitable since carbon electrodes deteriorate rapidly	Suitable for ferrous but rate of heating decreases above Curie point (780 Degrees Centigrade). Not really suitable for non-ferrous materials

by Table 2. In the following pages, some of the more salient characteristics of these automatic processes will be pointed out, with emphasis on the newest of the three, the electro-gas method.

2. *Induction Heating*

Radio-frequency induction heating is the most widely known method, since it has received much attention during the recent war. An example of an induction heating equipment is given in Figure 1.

In this process, the alternating-current mains supply is transformed to an output frequency of from 400 kilocycles to 5 megacycles, and this radio-frequency current is passed through a water-cooled copper tube (known as the heating inductor) that is shaped to enclose the area to be treated. The object to be heated does not touch the inductor, but is separated from it by an air-gap of from $\frac{1}{16}$ to $\frac{1}{4}$ inch (1·5 to 6·4 millimetres). The smaller the air-gap, the more efficient the process, but this is often offset by the complexity of working with close tolerances. The radio-frequency current flowing through the heating inductor induces opposing currents in any metallic object nearby, and if the object to be heated is brought close to the inductor, these currents will generate heat in it by virtue of the electrical resistance of the metal. The heating efficiency is to some extent thus dependent on the resistance of the metal to be heated, the efficiency increasing with resistance for any given induced current.

Since the strength of the opposing currents induced in the object to be heated falls off very rapidly with the distance of the object from the heating inductor, very localized heating of small areas is possible on relatively large objects; this accounts for the rather strange shapes of some of the heating inductors shown in Figure 2. These coils have been shaped to fit around those parts of an object in which heat is desired.

The most striking feature of radio-frequency induction heating is that the induced currents flow only on the outside surface of the object and, therefore, this is the only area in which heat is generated. The remaining mass of material receives heat by conduction from this outer layer, and this may set a time limit on the brazing process since burning of the outside layer may occur before sufficient heat penetrates to the brazing alloy and the joint.

The overall power efficiency of induction heating varies considerably depending on the type of work, but it is of the order of 40 to 50 percent for ferrous objects, and may fall as low as 25 percent for nonferrous work. On the whole, the process is most suitable where very large numbers of similar components must be heated, particularly if they are ferrous.

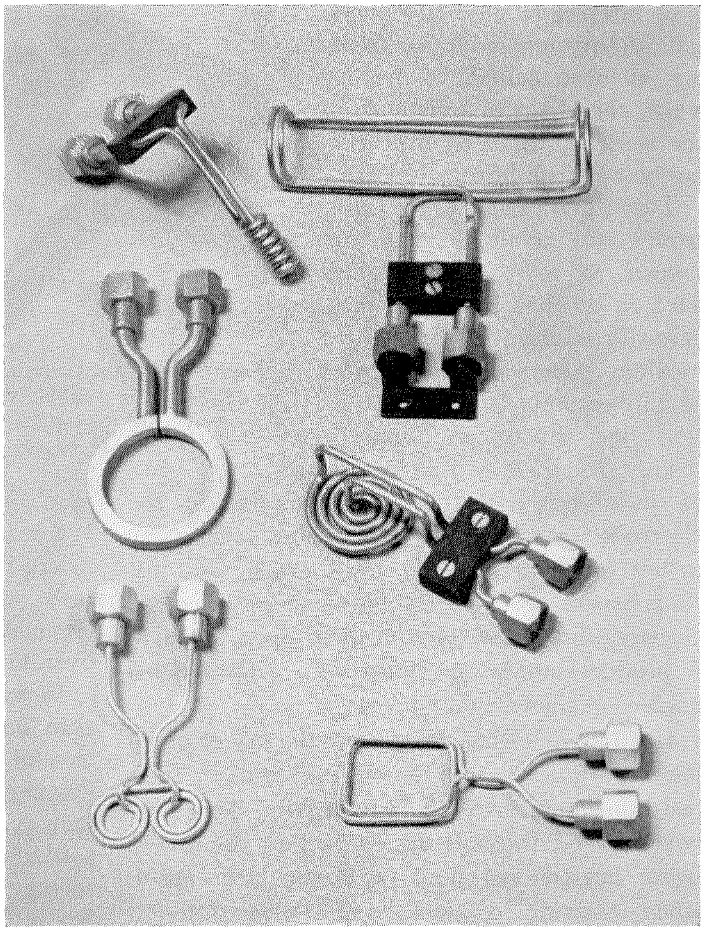


Figure 2—The heating inductors are shaped to place the coil near only those parts of an object in which heat is required.

3. Resistance Heating

The second process is not so well known, but has proved very successful; this is resistance or electrode heating. Here, the mains supply voltage is transformed down to about 2 to 12 volts, which is safe for the operating personnel without protection. This voltage is applied directly through carbon electrodes to the object to be heated (Figure 3). The resulting current passing through the resistance of the object generates the heat, and since this current may be of the order of hundreds or thousands of amperes, it will be seen that for a given voltage the resistance of the object has a marked effect on the amount of current and on the heating. It is of interest to note that while radio-frequency induction heating is best suited to ferrous work, the effect of resistance in the work makes resistance heating best for nonferrous objects.

Three grades of carbon electrodes are available and are known as soft, medium, and hard, in order of increasing electrical resistance. Soft and medium electrodes are most usually employed in the production line, since they can be readily shaped to fit the contour of the object, and the heat generated in the electrode itself is not excessive. They burn away rather quicker than the hard grade, and for that reason are often mounted in water-cooled electrode holders when the duty cycle is high. A production-type machine with water-cooled holders is shown in Figure 4.

Hard-grade electrodes permit the use of much lower currents than the soft, but due to their resistance they tend to heat rapidly. This heat is conducted through the contact to the object being heated and may be particularly useful when treating copper details. The different grades are useful in obtaining a uniform heat balance. For example, when joining copper to steel, a hard electrode would be used against the copper, and a medium or soft against the steel.

The correct grade and the shaping of the electrodes is a matter of experience and trial, but this is quickly obtained by practice. It is particularly important to keep the resistance between each electrode and the work to a minimum, or surface burning at these points will occur. When treating castings, the surface is smoothed where the electrodes are to be applied; iron castings are particularly troublesome to handle, especially where the final temperature has to be over 450 degrees centigrade.

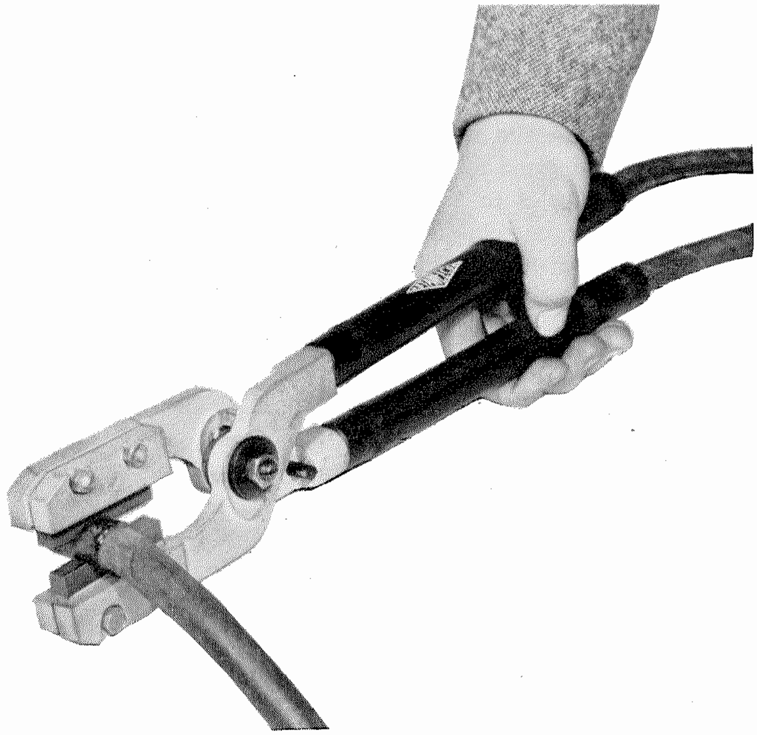


Figure 3—A resistance-heating hand tool is used to soft-solder a terminal lug on the end of a cable. Note that the carbon electrodes are shaped to fit partially around the lug.

In practice, the heating rate and final temperature in resistance heating is limited by deterioration of the electrodes, and brazing with alloys flowing above 650 degrees centigrade is not satisfactory. The process is particularly applicable to soft-soldering operations, being equally satisfactory for pre-placed solder rings or for hand-fed solder. The electrode pressure, held for the entire heating cycle and for at least part of the cooling cycle, ensures strong good-quality joints.

Figure 4—A production-line-type resistance brazing machine. On this 10-kilowatt equipment, the carbon electrodes are mounted in water-cooled holders. Pressure on the foot-pedal causes the electrodes to come together and grip the work firmly. The current is then applied until the brazing alloy melts; pressure is maintained while the work cools.

4. *Electro-Gas Equipment*

The most recently developed method of production-line soldering and brazing employs gas as the heating medium but relies for the control and timing on electricity. This has been named intense localized heating is so often required—by the development of specialized burners that operate with compressed air and pressurized gas, or with a pre-mixed and pressurized air-and-gas supply. The basic design of these burners is

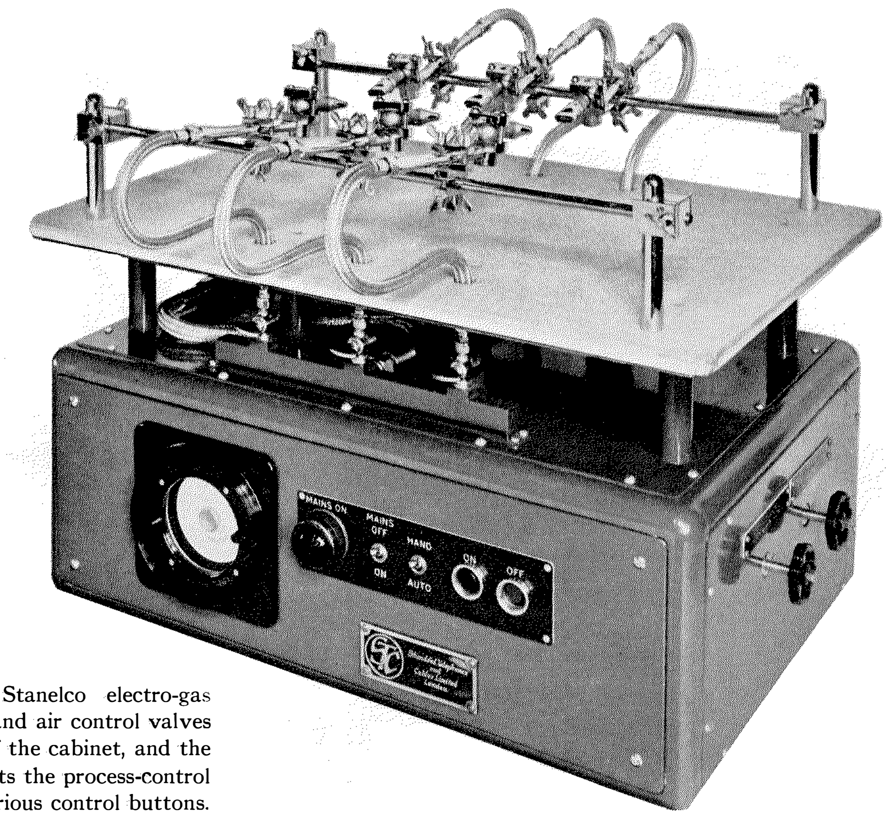


Figure 5—The Stanelco electro-gas equipment. Gas and air control valves are on the side of the cabinet, and the front panel mounts the process-control timer and the various control buttons.

the Stanelco electro-gas process and an example of this equipment is shown in Figure 5.

The effectiveness of the ordinary blowpipe depends entirely on the skill of the operator. Its operation is limited by the fact that, when the gas and air pressure are increased to obtain the highest heat, it is often found that the flame is unstable and tends to "blow off." This fault was overcome in the glass-working industry—where

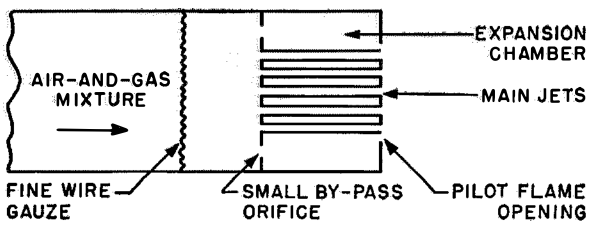


Figure 6—Design of a typical high-intensity burner.

shown in Figure 6. The air–gas mixture passes through a fine wire mesh that acts as a final mixer and a filter to remove extraneous particles, and then the bulk of the mixture emerges from the burner at high speed via the main jets. A small proportion, however, enters an expansion chamber on the burner and leaves this chamber at a relatively low speed. This latter gas burns as pilot flames that continuously ignite the mixture emerging from the main jets. The result is a steady, narrow, intense, ribbon flame.

Previous limitation to the use of these burners on production lines has been caused first, by the lack of knowledge of their existence outside the glass industry, and second, by the need for gas-pressure boosters or mixing plant, which would not be economically justified except where many heating stations are used.

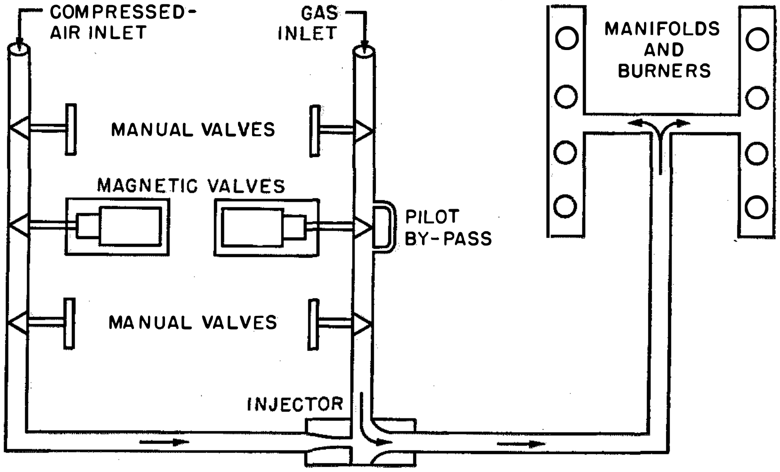


Figure 7—Diagram of the electro-gas equipment showing the control valves for gas and air and the injector unit. The electrical process timer that operates the magnetic valves is not indicated above.

The more recent development of efficient air–gas injectors has, however, removed the need for boosters, and hence the picture has been changed. Compressed air at any pressure over 10 pounds per square inch passes through a fine jet in the injector shown in Figure 7 and then across a small chamber containing gas; the mixture emerges through a specially shaped orifice. The diameter of the jet and of the orifice are carefully related to the number and type of burners in use, but there is sufficient tolerance for practical operation, and it is a simple matter to change the injector where necessary.

The air and gas supplies to the injector are regulated by valves until steady flames are produced. The flame width is about $\frac{1}{8}$ inch (3·1 millimetres), and many well-defined cones, similar to small bunsen flames, can be seen.

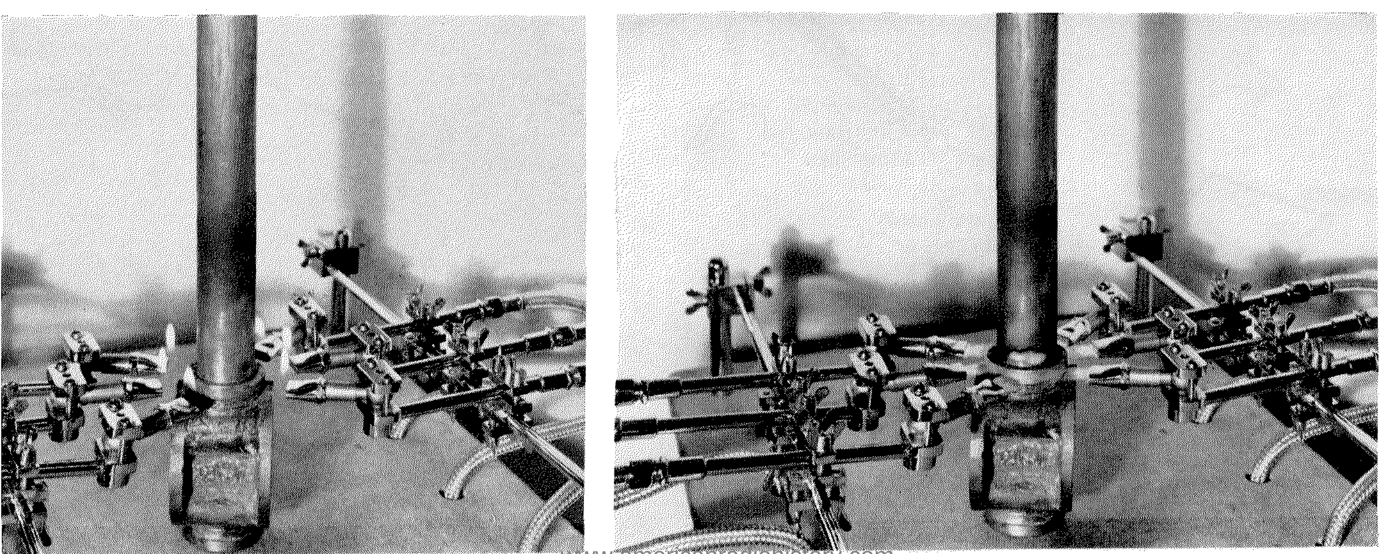
The next advance was to harness this form of heating with electrical control, and this was achieved by the introduction of electrically operated valves in the air and gas supplies. When in the "off" position, sufficient gas is by-passed to act as a pilot so that the burners instantly re-light when the "on" button is depressed (Figure 8). A process timer is included and this can be switched in when desired for production-line work.

The coal-gas-and-air flame burns at about 1300 degrees centigrade, which is a suitable temperature for most brazing operations and is also applicable to soft soldering and annealing. There is a wide range of suitable burners already available on the market, which is a great asset, and the flexibility of the mounting arrangement to give them the desired heating pattern appeals to the production engineer.

5. *Conclusion*

There will always be a place for use of each of the five methods of soldering and brazing that have been presented. The production engineer must consider carefully the range of work to be handled before purchasing equipment. Resistance machines and electro-gas equipment are becoming increasingly more popular due to their flexibility, low capital cost, and very low maintenance.

Figure 8—In the photograph at the left, the pilot flames are alight, and the work has been placed in position When the "on" button is depressed (right), the burners produce steady and intense flames to give clean high-quality brazed joints.



Netherlands-Denmark Coaxial-Cable System

By

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BEFORE the second world war, practically all international telecommunications traffic between the Scandinavian countries and the rest of Europe was routed via Germany. After the war ended, everything was done to re-establish as fast as possible the communications facilities that had been interrupted by hostilities. The most important circuits were quickly established, but it soon became obvious that it was not possible, at least for several years, to reconstitute the communications network that had formerly existed in Germany. This was highly disorganised by the separation into military zones of occupation and by the destruction of much equipment only partially replaced by provisional systems of a military character. Even the former complete network would have been inadequate to meet the growing needs of countries deprived by the war of international communications, and the part that could be re-established had to carry a large proportion of military traffic.

These considerations caused the Netherlands and the Danish administrations to review in August, 1945, the possibility, which had been considered long before the war, of a direct connection by submarine cable between the two countries. The considerable distance involved, about 260 kilometres (140 nautical miles), had previously made such a project impracticable

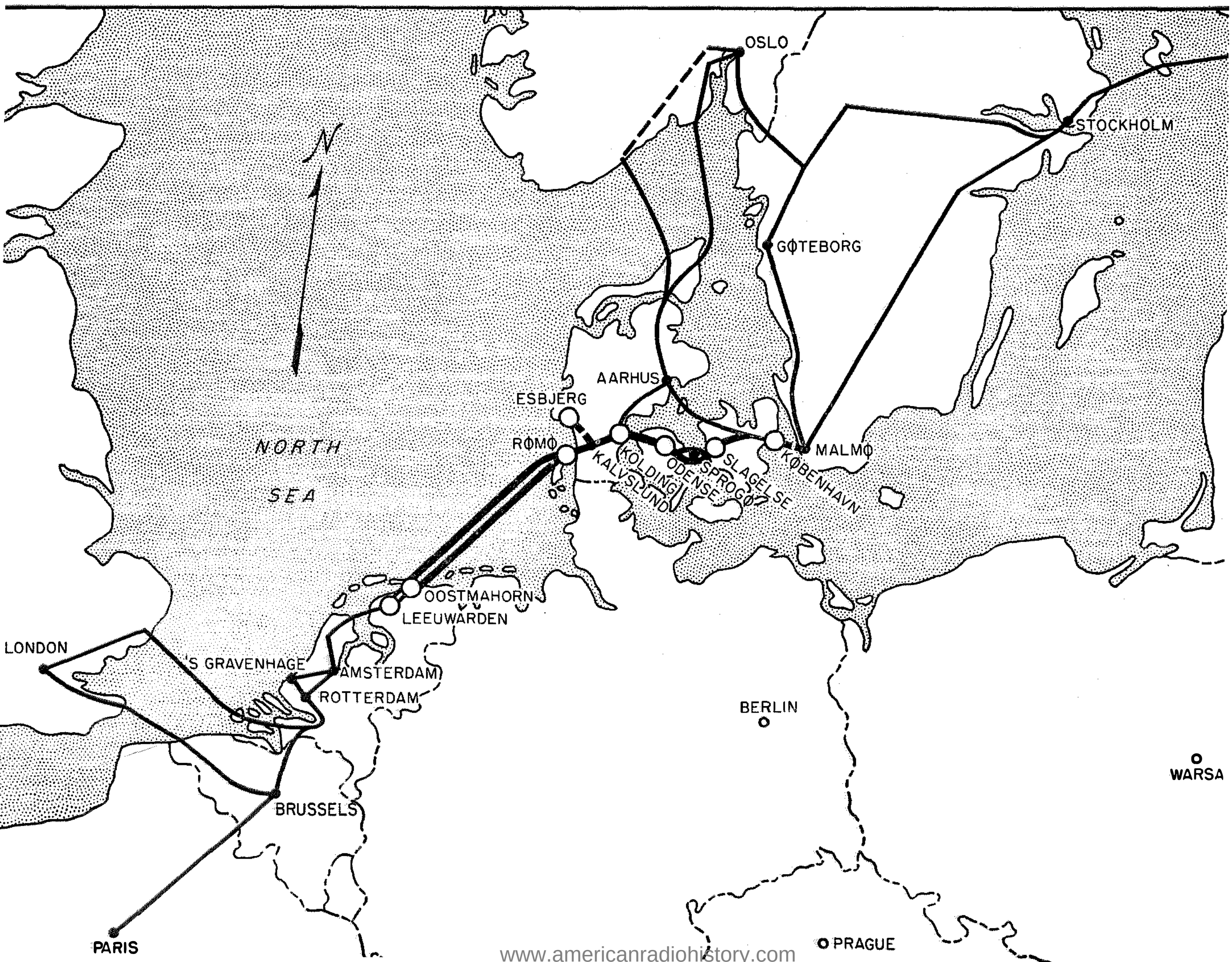


Figure 1—Main routes involved in the cable scheme.

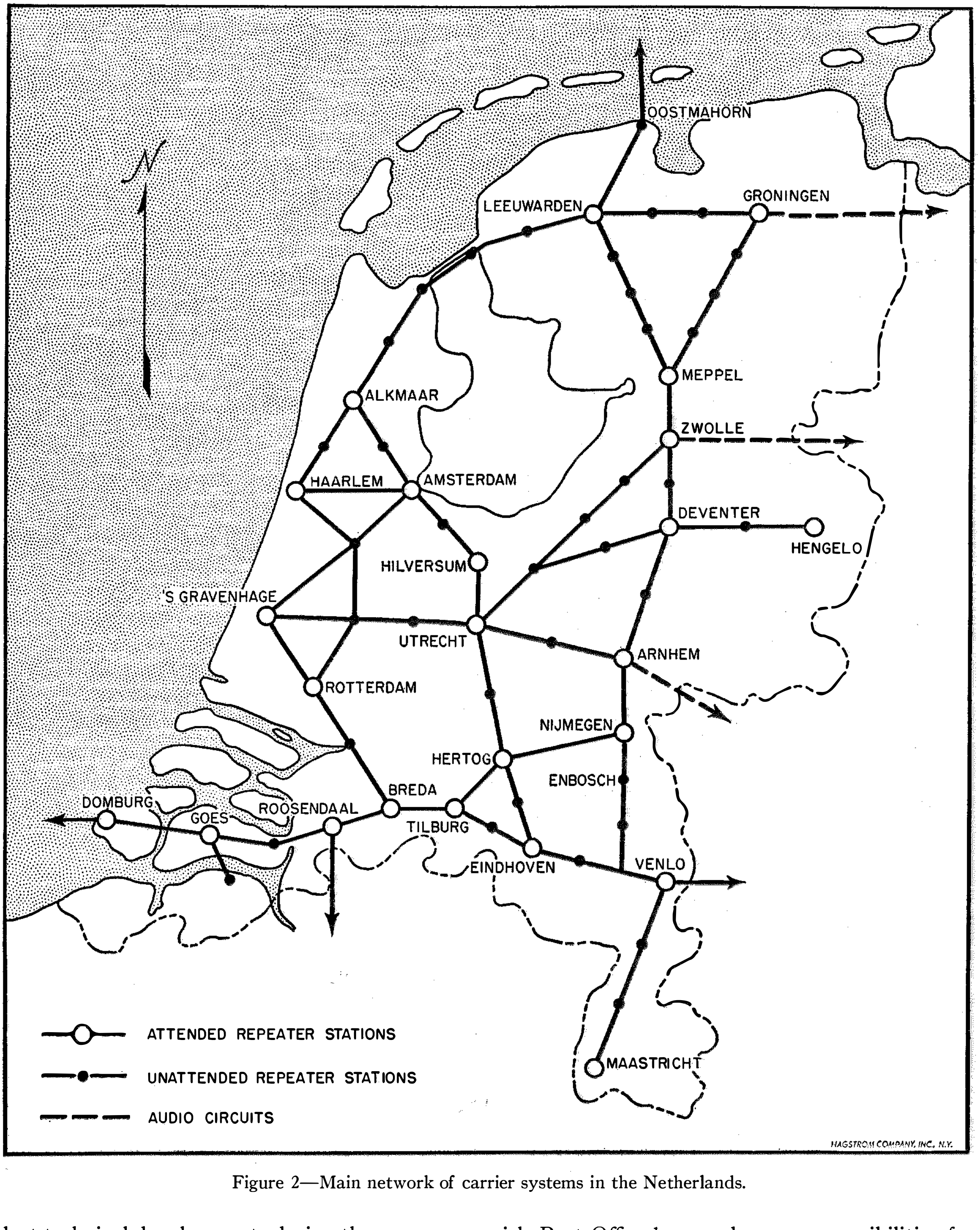


Figure 2—Main network of carrier systems in the Netherlands.

but technical developments during the war years, such as the introduction of improved dielectrics together with coaxial construction for submarine cables and the possible use of submerged repeaters, which had been introduced by the British Post Office,[1] opened up new possibilities for

[1] R. J. Halsey and W. T. Duerdoth, "Modern Submarine Cable Telephony and Use of Submerged Repeaters," *Post Office Electrical Engineer's Journal*, v. 37, pp. 33–39; July, 1944.

long submarine cables. Discussions between the two administrations showed that such a cable would not only form a main artery for international traffic, but in its passage across Denmark could be planned to provide a large number of much needed circuits for inland use. Enquiries of the nearest interested European administrations showed that Norway (linked with Denmark during the war by cables Hirtshald–Arendal and Frederikshavn–Sandefjord), Sweden (with the Baltic cables now terminating in Eastern Germany), Belgium, France, and Great Britain would welcome this proposal. Accordingly, the two administrations proposed this cable as part of the proposals of the Comité Consultatif International Téléphonique for rebuilding the European long-distance cable network, and it was included in the 'Programme Générale d'Interconnexion en Europe' agreed to at the meeting at Montreaux in 1946. Figure 1 shows a map of the main routes of interest in connection with this scheme. It was planned that all these routes would consist of 4-wire, high-velocity circuits using coaxial or quadded carrier cables of the highest quality, fully conforming to the recommendations of the Comité Consultatif International Téléphonique. Figures 2 and 3 show the main routes of the internal cable networks of the Netherlands and of Denmark, with connections to other countries.

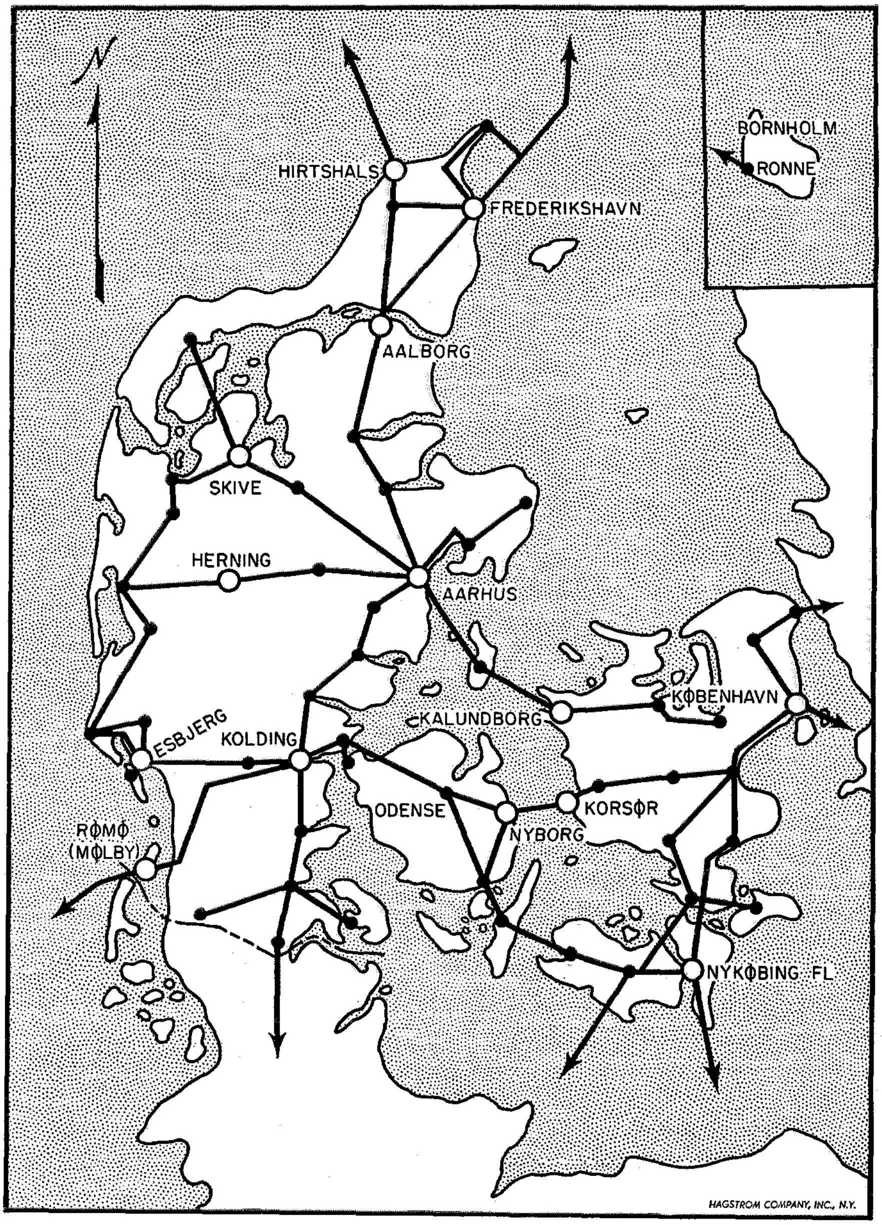


Figure 3—Long-distance telephone network in Denmark.

1. Netherlands–Denmark Link

The first project for the North Sea crossing involved a single air-spaced polythene coaxial submarine cable of the same construction as that of the Anglo–Dutch cable *6* that was laid in 1947. This somewhat expensive cable, which would have provided 48 circuits without submerged repeaters or 120 circuits with one submerged repeater using different frequency

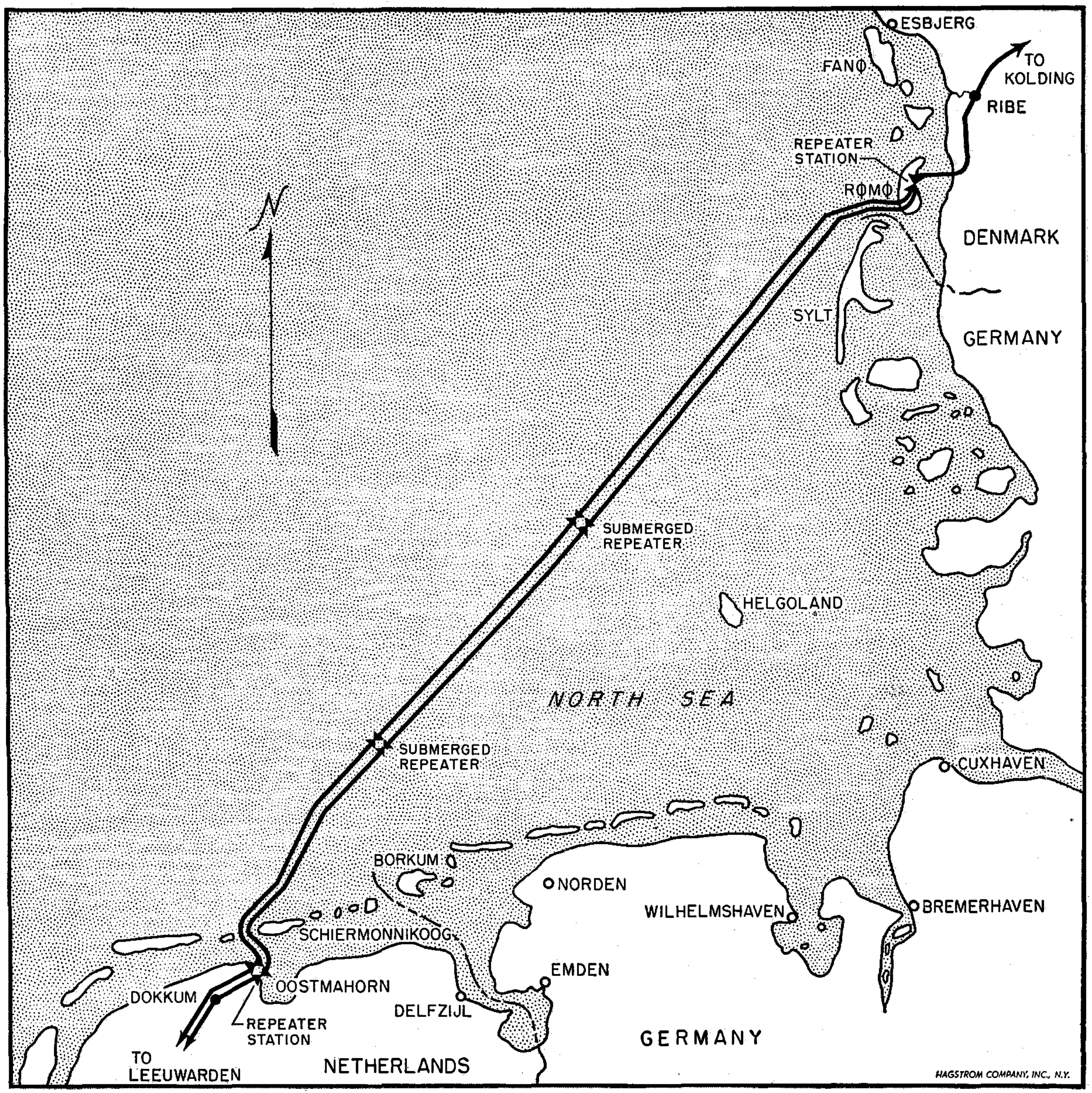


Figure 4—Route of the submarine cable between Oostmahorn and Rømø.

bands for each direction of transmission, was abandoned in favour of two solid-polythene coaxial cables of smaller diameter and of cheaper construction, each of which, using different frequency bands for the two directions, would provide initially 36 circuits with two submerged repeaters. The proposal to use two independent cables had the great advantage that in the event of interruption of either cable only half the circuits would be lost. Advantage can also be taken of two cables to allow the circuit capacity to be increased by replacement in each cable in turn of the initial repeaters and the addition of further repeaters without complete loss of service. It was proposed that a normal land coaxial cable with one tube for each direction of transmission should be used for the route across Denmark.

After many discussions regarding the best route, bearing in mind that only a few narrow shipping lanes had been cleared of mines, it was decided to land the submarine cables at Oostmahorn on the Frisian coast of the Netherlands and on the Island of Rømø connected by causeway to the Jutland mainland of Denmark. From there, the route would pass to Copenhagen via Kolding and on the Netherlands side would be terminated made by Nordiske Kabel-og Traadfabriker, Copenhagen, under license from the International Standard Electric Corporation.

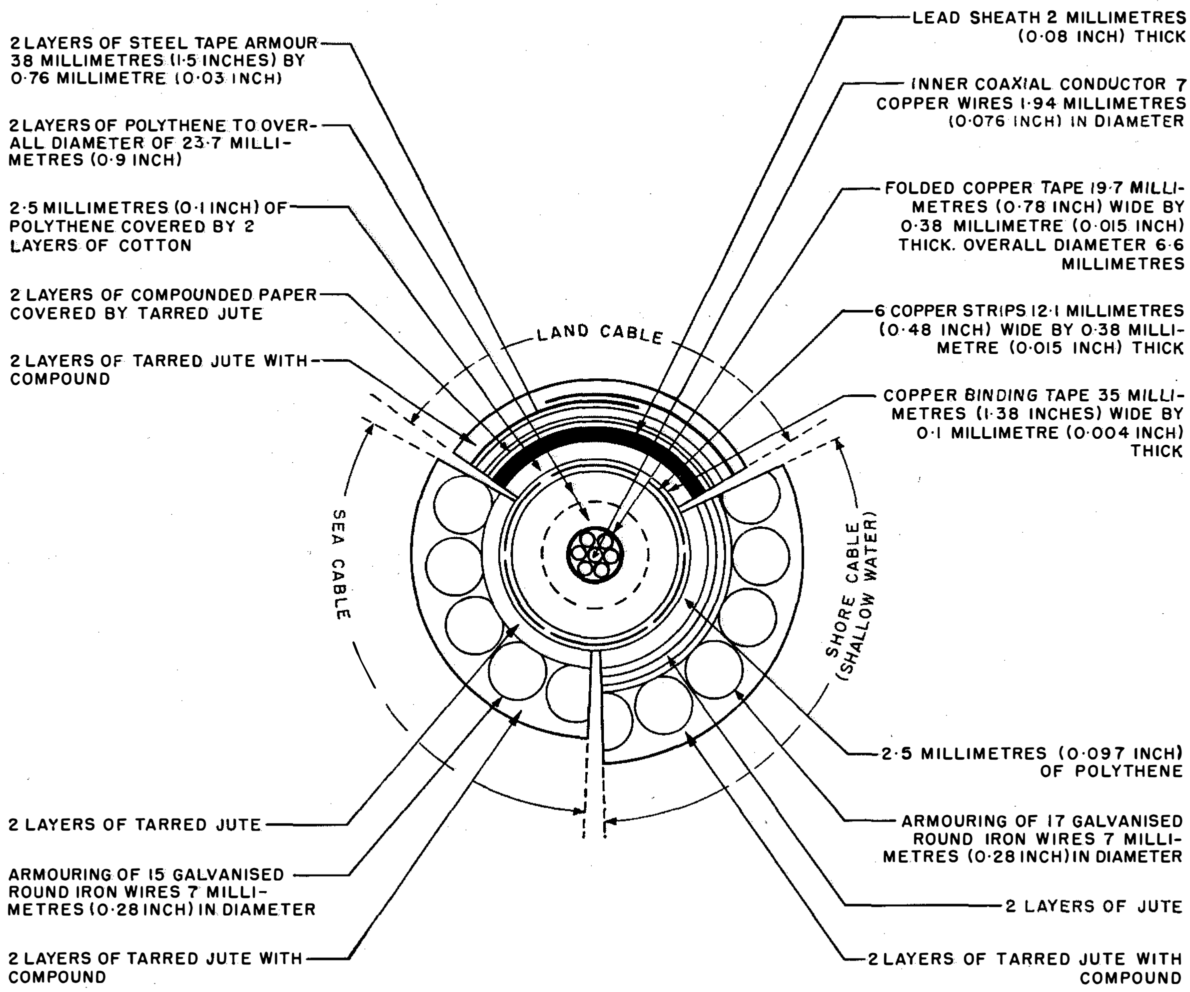


Figure 5—Cross section showing construction of the sea, land, and shore cables.

at the repeater station at Leeuwarden, from where it would join up with the existing Netherlands symmetrical-pair carrier cable network.

With the acceptance in principle of the scheme by the two administrations, the contract for the overall complete system and also for the equipment was awarded to Standard Telephones and Cables, Limited of London, in association with Standard Electric A/S of Copenhagen and Nederlandsche Standard Electric Mij N. V. of The Hague. The submarine cables across the North Sea and the short land ends were manufactured by Submarine Cables Limited, London. The land coaxial cable (including the submarine portions across the Great Belt in Denmark) was

2. *Land and Sea Cables*

The route taken by the submarine portion between Oostmahorn and Rømø is shown in Figure 4. The total length is approximately 263 kilometres (142 nautical miles). The sea cables consist basically of an inner coaxial copper conductor of overall diameter of 6·6 millimetres (0·26 inch) insulated with solid polythene bringing the diameter to 23·7 millimetres (0·935 inch) followed by an outer coaxial conductor with armouring and protection. The outer conductor is insulated with an additional layer of polythene at the shore ends of which about 0·93 kilometre (0·5 nautical mile) is in the sea in order that the contact with the sea water will be free from the disturbing influences of stray fields. This same insulation of the outer conductor, protected by an additional lead sheath and steel tape armouring, is followed out across the island of Rømø where the repeater station is about 5 kilometres (2·7 nautical miles) from the landing point. A cross section of the sea cables, shore ends, and land cables is shown in Figure 5. Between the coast repeater station at Oostmahorn and Leeuwarden where the system terminates, a distance of about 39 kilometres (21 nautical miles), solid polythene cables similar to the submarine cable but of somewhat smaller dimensions are used. The outer conductor of these cables is insulated from earth and protected by a lead sheath for a distance of about 3 kilometres (1·6 nautical miles) from Leeuwarden.

The submarine cables, each with its two submerged repeaters, were laid by the British Post Office cable ship *Monarch* with the assistance of the Netherlands cable ship *Poolster* and the Danish cable ship *C. E. Krarup*, which rendered particular assistance in buoying and in laying the shallow-water parts of the cable for distances of 26 and 20 kilometres (14 and 11 nautical miles) respectively. The whole of the work including loading, buoying, and jointing of the submerged repeaters in the open sea was carried out between August 8 and September 2, 1950. Although only a narrow channel about 1·9 kilometres (1 nautical mile) wide in parts had been swept of mines and particular navigational care was necessary, the work proceeded smoothly and was interrupted for only one day by storms. From the point of view of crosstalk between them and also in order to facilitate picking up, the two cables were laid as far apart as possible in the cleared channel, the spacing being of the order of 0·9 kilometre (0·5 nautical mile).

The section across Denmark, a distance of 323 kilometres (174 nautical miles), consists of a 2-tube coaxial cable between Rømø and Kalvslund with a 4-tube coaxial cable between Kalvslund and Copenhagen. Two of these 4 tubes will also serve to extend the Danish national circuits from Copenhagen to Esbjerg by means of a coaxial cable planned between Kalvslund and Esbjerg. These cables are all constructed to standard specifications of the Comité Consultatif International Téléphonique for coaxial cables (Livre jaune, Tome III, Specification A.VI), each tube of which has an inner copper conductor 2·64 millimetres (0·104 inch) in diameter insulated with polythene discs and an outer folded copper tape of 9·5 millimetres (0·375 inch) in diameter. The 2-tube section is provided with 8 star quads in the tube interstices and the 4-tube cable with 5 star quads, which are used for system maintenance purposes. The cable is lead covered and for the most part it is steel-tape armoured and laid in trenches at a greater depth (80 centimetres (31·5 inches) as against 60 centimetres (23·6 inches)) than normal in Denmark to give better protection against frost damage. In some sections, particularly around Copenhagen, the armouring is omitted and the cable is drawn into ducts.

The normal coaxial land cable is laid on the causeway joining Rømø to Jutland and on the Little Belt Bridge between Jutland and the island of Fyn. The only part of this route where special technical problems have been encountered is in the Great Belt crossing between Fyn and Zeeland. The distance is about 20 kilometres (10·8 nautical miles) and the island of Sprøgø being conveniently situated in the middle would allow repeater stations to be located at each side of the crossing and on Sprøgø thus maintaining the normal spacing of approximately 10 kilometres (5·4 nautical miles) for unattended coaxial repeaters. This assumes that coaxial tubes with characteristics similar to those of the land cables could be used for the sea crossings. Whilst it would, of course, have been possible to use solid polythene coaxial cables of the submarine type, as these normally consist of a single tube, either a special system of frequency translation would be necessary for the sea crossing or several cables with specially designed repeaters would be required. Both these solutions would be somewhat clumsy and expensive whereas, if a normal air-spaced 4-tube land cable could be used, the uniformity of the system between Copenhagen and Rømø could be preserved. It was considered that the shallow water, smooth sea bed, and comparative freedom from anchorages in the Great Belt would be acceptable conditions for a land-type cable provided it could be laid.

As a precaution against damage by shipping, each submarine section is duplicated and provision is made at the 3 repeater stations involved to change the repeaters from the working to the spare cable by manual switching. In addition, to prevent water penetration along the whole cable lengths in the event of damage, paper plugs were wound round the inner coaxial conductors at distances of 100 metres (328 feet), and a 13-millimetre (0·5-inch) wide plastic tape was applied longitudinally along the joint in the outer coaxial tube. The tubes were covered with a layer of paper wrapping and the interstice quads were wrapped with 3 layers of loose crepe paper. This does not produce a completely watertight cable but, due to the swelling of the paper plugs and wrapping on water penetration, it is considered that only a few hundred metres of cable would be damaged before repairs could be put in hand. Figure 6 shows the coaxial tubes with this protection. Impedance and pulse echo tests show that, although the presence of the plugs can be detected at frequencies above 1·2 megacycles, the deviations in the impedance curve are in all cases less than half the maximum value (3 per cent) prescribed by the Comité Consultatif International Téléphonique. The first of the Great Belt cables, all of which were made by the Nordiske Kabel-og Traadfabriker and laid by *C. E. Krarup*, was spliced in the cable ship, but subsequent cables were jointed as one length in the factory and laid in the normal manner for submarine cables. The whole cable section in Denmark was laid between June, 1949, and December, 1950.

hagen, the transmission is the usual 4-wire arrangement over two coaxial tubes in a single cable.

The equipment at Leeuwarden is provided to translate the band of frequencies associated with each channel to frequencies suitable for transmission over the submarine cables. These bands have been chosen as:—

24–168 kilocycles per second* in the Netherlands-to-Denmark direction.

208–352 kilocycles in the Denmark-to-Netherlands direction.

Provision is not made, however, to translate each channel from its audio frequencies 300-to-3400 cycles as it is planned to switch the circuits in blocks of twelve at Leeuwarden to the Netherlands network of carrier cables. The input to the translating equipment consists, therefore, of three basic 12-channel groups in the 60-to-108-kilocycle band.

3.1 Cable 1 (East Cable)

The three basic groups, 60–108 kilocycles, are modulated with carrier frequencies of 516, 564, and 612 kilocycles respectively, producing a continuous band of frequencies of 408–552 kilocycles. This band is then modulated with a carrier of 576 kilocycles, producing the band 24–168 kilocycles, which is transmitted to line.

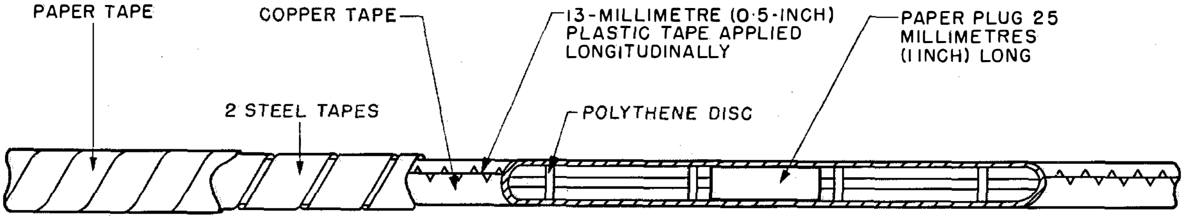


Figure 6—Coaxial tube for Great Belt cable.

3. *Overall System*

The system is worked on a grouped frequency basis on each of the two cables between Leeuwarden and Rømø, each cable carrying both directions of 36 speech circuits. Provision is also made in the system for a wideband two-way broadcast channel, which displaces three speech channels when radio programme transmission is taking place. From Rømø to Kolding and Copenhagen, the transmission is the usual 4-wire arrangement over two coaxial tubes in a single cable.

The incoming frequency band 208–352 kilocycles is first demodulated with a carrier of 760 kilocycles producing the band 408–552 kilocycles; this band being broken down to the three basic groups by a demodulation process that is the reverse of the modulation process described above.

* "Per second" has been omitted from the frequency designations that follow.

3.2 Cable 2 (West Cable)

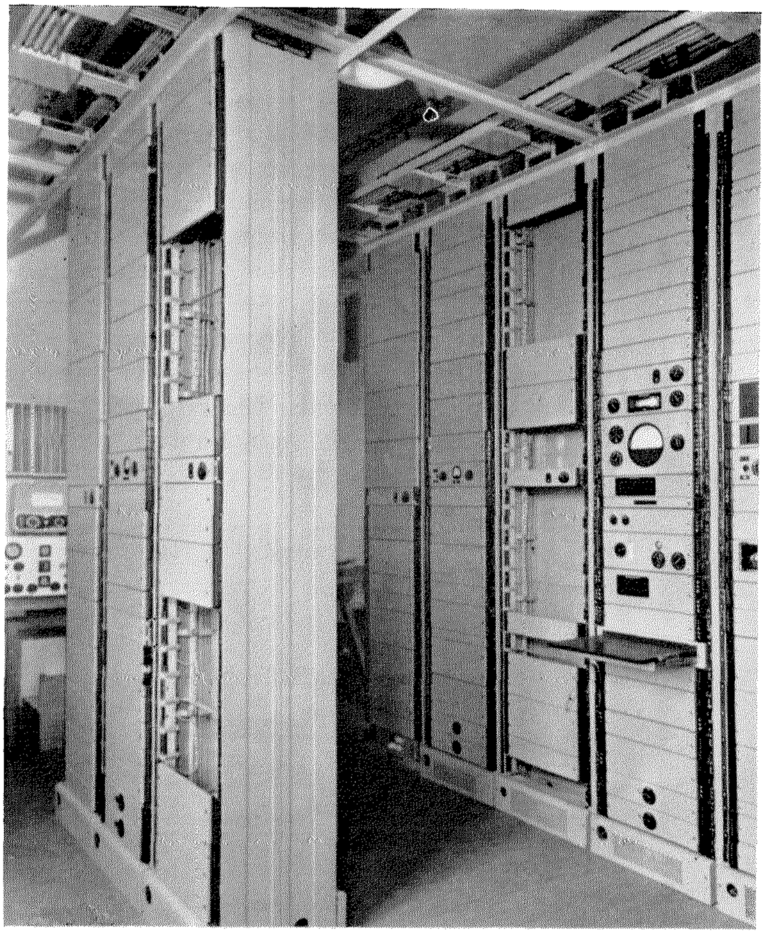
In this case, the three basic groups are modulated with carriers of 420, 468, and 516 kilocycles, respectively, producing a band of frequencies from 312 to 456 kilocycles, which is then modulated with a carrier frequency of 1364 kilocycles giving the band 908–1052 kilocycles finally with a carrier of 1076 kilocycles producing the same line frequencies as for Cable *1*, that is 24–168 kilocycles.

In the reverse direction, the incoming frequency band 208–352 kilocycles is first demodulated with a carrier of 1260 kilocycles producing the band 908–1052 kilocycles, which is then broken down to the three constituent groups by double-stage demodulation, first with a carrier of 1364 kilocycles and then with individual group carriers of 420, 468, and 516 kilocycles as above. The arrangement of frequency translations of the submarine system and of the Netherlands carrier system extensions at Leeuwarden is shown in Figure 7.

After the modulation processes described above, the signal band is amplified to a level suitable for transmission to line, first passing through an equaliser, which results in the higher frequencies being transmitted at a higher level than the lower, thus simplifying the equalization at the remote end and easing the amplifier loading problem. The signals then pass through the low-pass section of the directional filters, these filters being provided to separate the frequency bands used for the two directions of transmission. Finally the transmission is through the high-pass section of the line filters, this section passing all frequencies above 20 kilocycles, the low-pass section passing frequencies from zero up to a little over 10 kilocycles. This low-frequency band is used to provide an engineering speaker circuit operated on a party-line basis with two-wire repeaters at Oostmahorn and Rømø.

In the reverse direction, the transmission is via the high-pass section of the line filters, the high-pass section of the directional filters, and then through an equaliser that corrects the slope introduced by the line. The signal is then amplified and demodulated as described previously.

At Rømø, the frequency-translation processes are similar to those described for Leeuwarden, but in this case the band is not broken down to the basic twelve channel groups but its position in the frequency spectrum is changed to suit transmission as a 4-wire circuit over the coaxial land-cable system into Kolding and Copenhagen.



A suite of bays comprising the main frequency-translation equipment at Leeuwarden. Two of the bay sides are being installed: it only remains to slide the appropriate panels into their positions in the rack frameworks and to complete the circuit by means of plug-in connectors.

3.3 Cable *1* (East)

The incoming frequency band from the submarine cable, 24–168 kilocycles, is modulated with a carrier of 576 kilocycles to produce a band, 408–552 kilocycles, which is then transmitted over one of the coaxial tubes to Kolding.

In the reverse direction, the band received from the other coaxial tube from Kolding, namely 408–552 kilocycles is modulated with a carrier of 760 kilocycles, thus producing the band of 208–352 kilocycles for transmission to the Netherlands.

3.4 Cable *2* (West)

In this case, the band, 24–168 kilocycles incoming from the submarine cable is modulated with a carrier of 1076 kilocycles, producing a band, 908–1052 kilocycles, which is then transmitted to Kolding together with the 408–552-kilocycle band derived from cable *1*.

In the reverse direction, the band received from Kolding, namely 908–1052 kilocycles, is modulated with 1260 kilocycles to produce the required band, 208–352 kilocycles, for transmission to the Netherlands. The frequency translations at Rømø are shown in Figure 8.

The amplification, filtering, and equalisation is similar to that described for Leeuwarden with two major differences. One is that as the transmission from Rømø to Kolding and Copenhagen is by means of a 2-tube coaxial system, special filters have been provided to separate the two bands, 408–552 kilocycles and 908–1052 kilocycles, which are transmitted together over each of the two coaxial tubes and to route these bands to the equipment associated with cables *1* and *2* respectively. Secondly, as the attenuation of the submarine cables is very much higher at the frequencies transmitted from Denmark to the Netherlands (208–352 kilocycles), the signals are

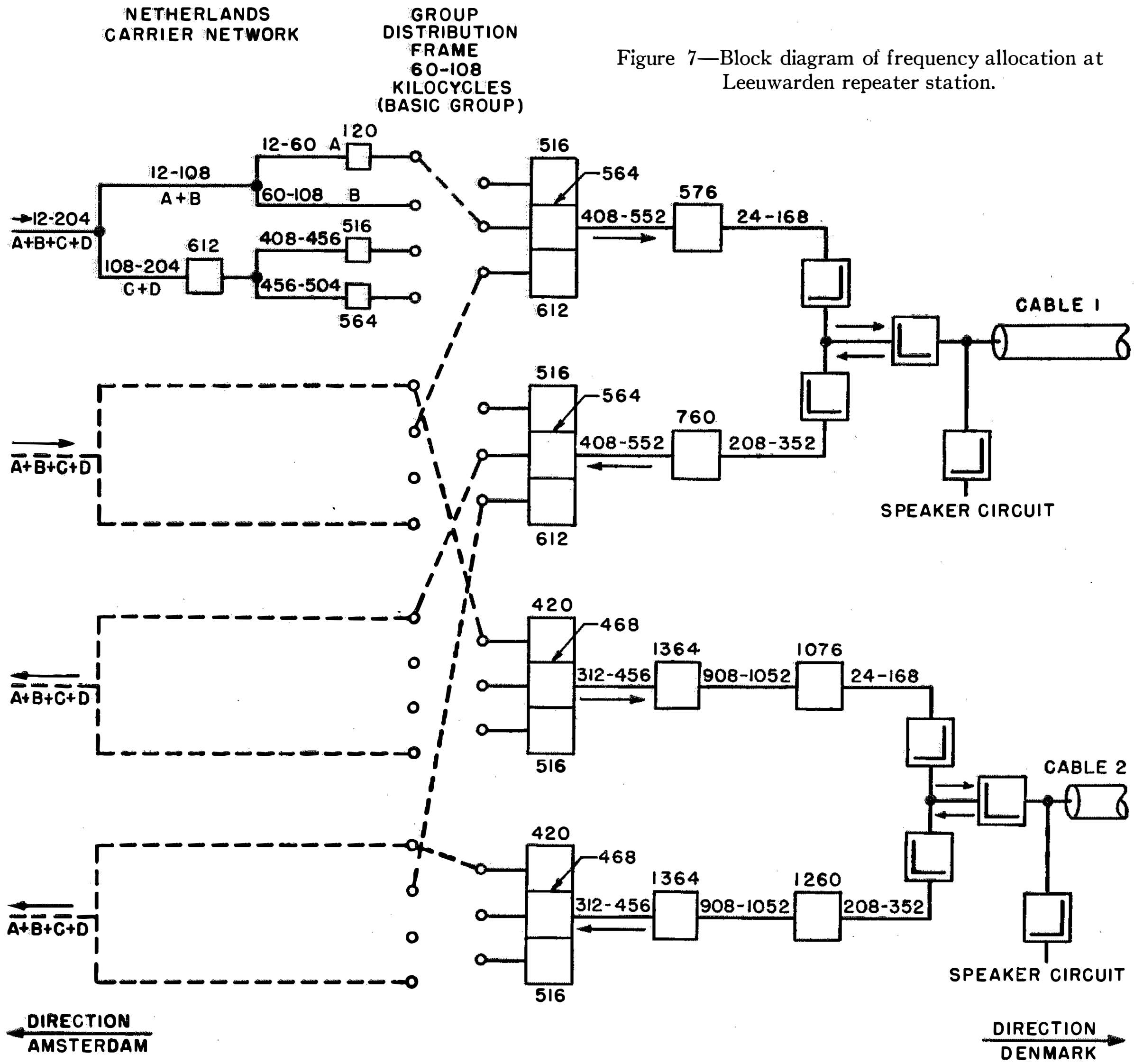


Figure 7—Block diagram of frequency allocation at Leeuwarden repeater station.

transmitted at a much higher level (at + 30 decibels referred to 1 milliwatt at 352 kilocycles). This means that a special amplifier is necessary to handle these high powers. The amplifiers provided have a power-handling capacity of 100 watts.

has a frequency of 124 kilocycles and which feeds the harmonic generators.

The master oscillator, frequency dividers, and harmonic generators are duplicated, and provision is made for automatic change-over within a time of 4 milliseconds from one set to the other

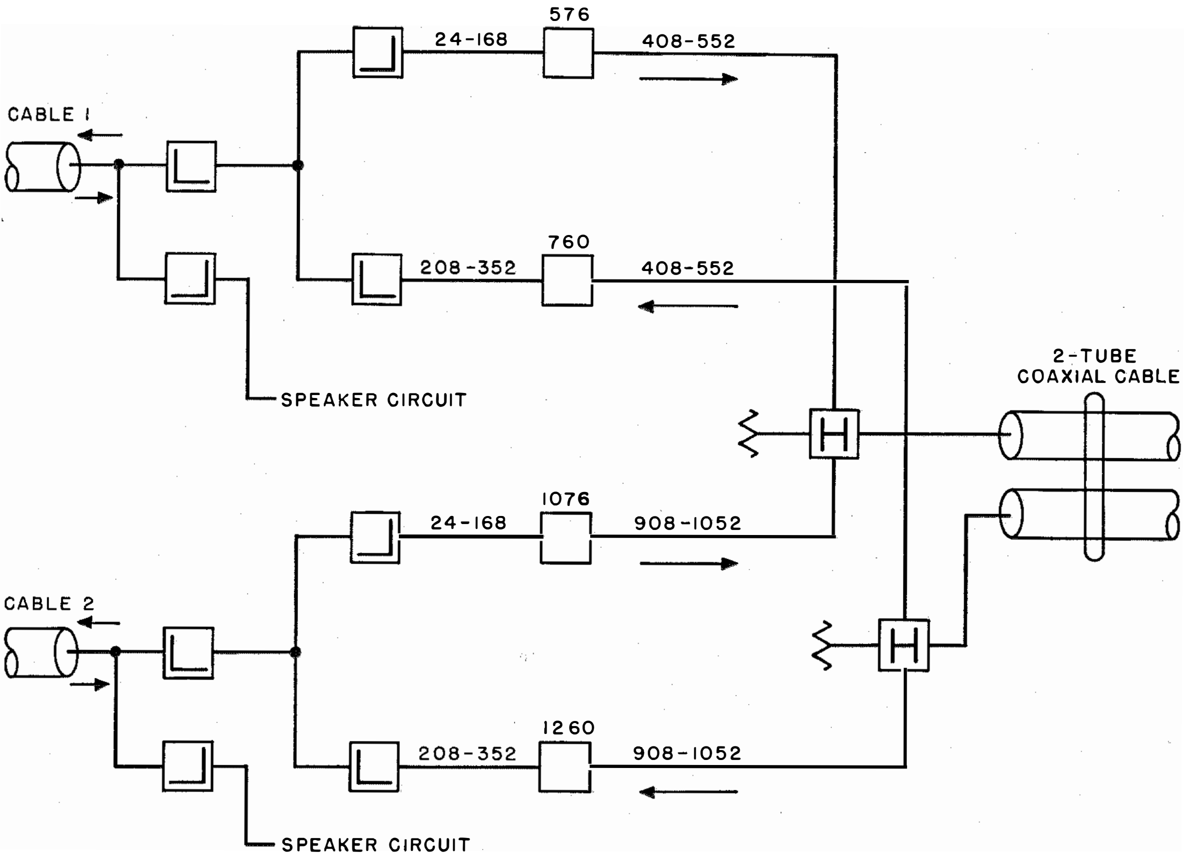


Figure 8—Frequency translations at Rømø repeater station.

The level diagrams for the limit frequencies of each direction of transmission are shown in Figure 9.

All the carrier modulating frequencies used both at Leeuwarden and Rømø are multiples of 4 kilocycles and are derived from a number of harmonic generators, the fundamental frequencies of which are 4 kilocycles or multiples thereof. Certain of the frequencies derived from these harmonic generators are modulated together to produce the wanted modulating frequencies.

A high degree of carrier-frequency accuracy and stability is achieved by the provision of an oven-controlled master crystal oscillator, which

in the event of failure of the working set. The frequency-generating equipment is similar in general principle to that used in Kolding and Copenhagen on the normal coaxial system.

At Kolding, the frequency bands, 408–552 and 908–1052 kilocycles (parts of supergroups *2* and *4*), are translated by the normal coaxial terminal equipment (partially equipped) to the basic supergroup range of 312–552 kilocycles. The 36 circuits of supergroup *2* are connected by through-supergroup filters at the basic supergroup frequency range to the normal coaxial system between Kolding and Copenhagen. Of the remaining 36 circuits of supergroup *4*, these are first frequency-translated to the range 60–108 kilocycles and 12 (alternatively, 9 and 1 programme channel) are translated to audio frequencies and terminated at Kolding. The remaining 24 circuits are connected by through-group filters to the coaxial system to Copenhagen. The same frequency translations are used in the reverse direction for transmission between Kolding and Rømø.

Between Kolding and Copenhagen, in addition to the through international circuits described above, 96 inland circuits together with one 2-way music channel are carried over one pair of tubes. Each pair of tubes between Copenhagen and Kolding is at present equipped with repeaters to take 600 circuits and may ultimately be equipped to take 960 circuits. The Kolding–Rømø section is similarly equipped, and the Rømø–Leeuwarden submarine system is planned to take up to 120 circuits on each submarine cable by changing the submerged repeaters and increasing their number. Kolding is arranged to provide full flexibility of growth between the following networks:—A) the Danish cable network, with extension to Norway radiating from Kolding, B) circuits to Copenhagen, with extension to Sweden, C) circuits to Esbjerg over the coaxial cable via Kalvslund, and D) the international circuits to the Netherlands. By equipping through-supergroup filters, through-group filters, or channel equipment as necessary in Kolding, any section can be connected to any other section at will as growth proceeds.

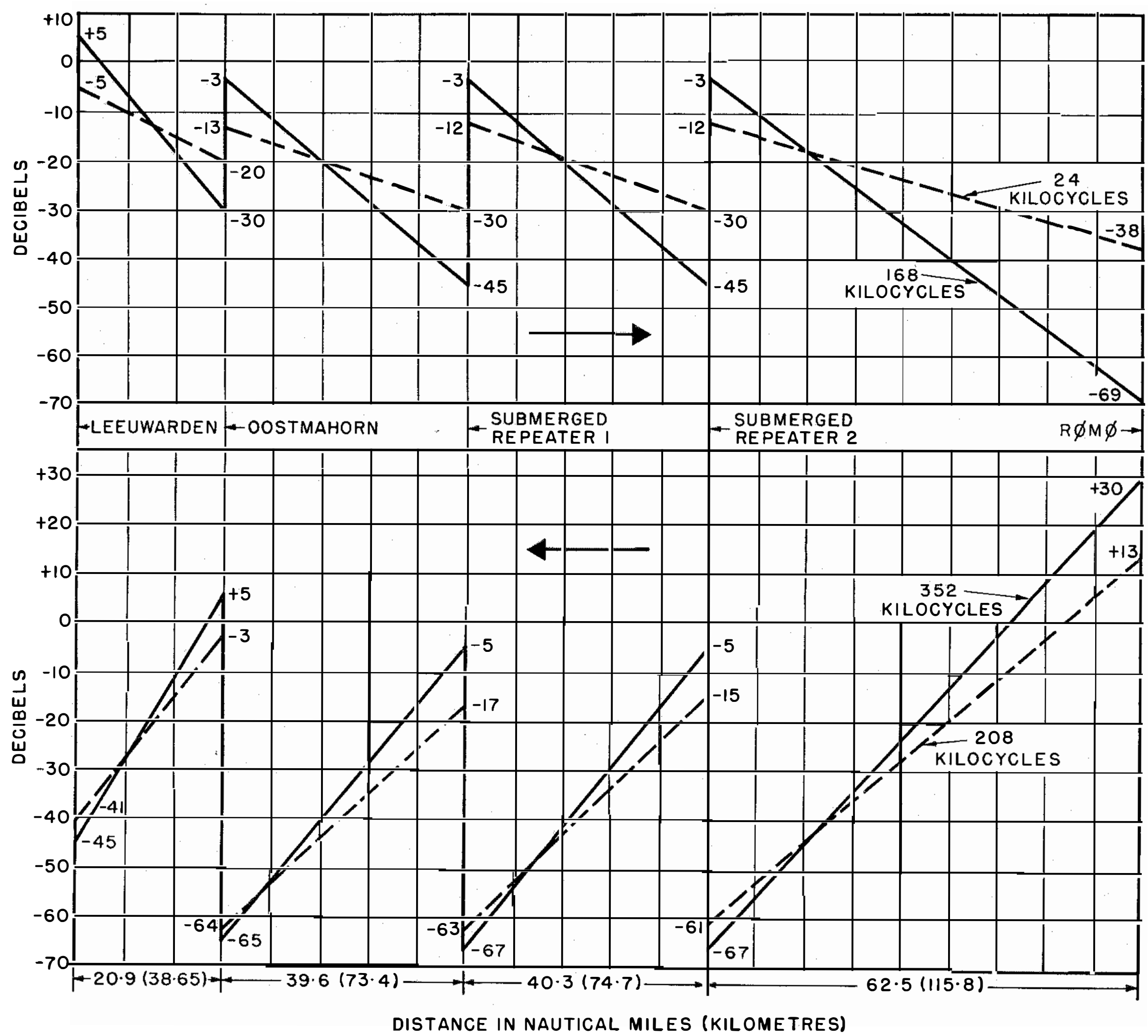
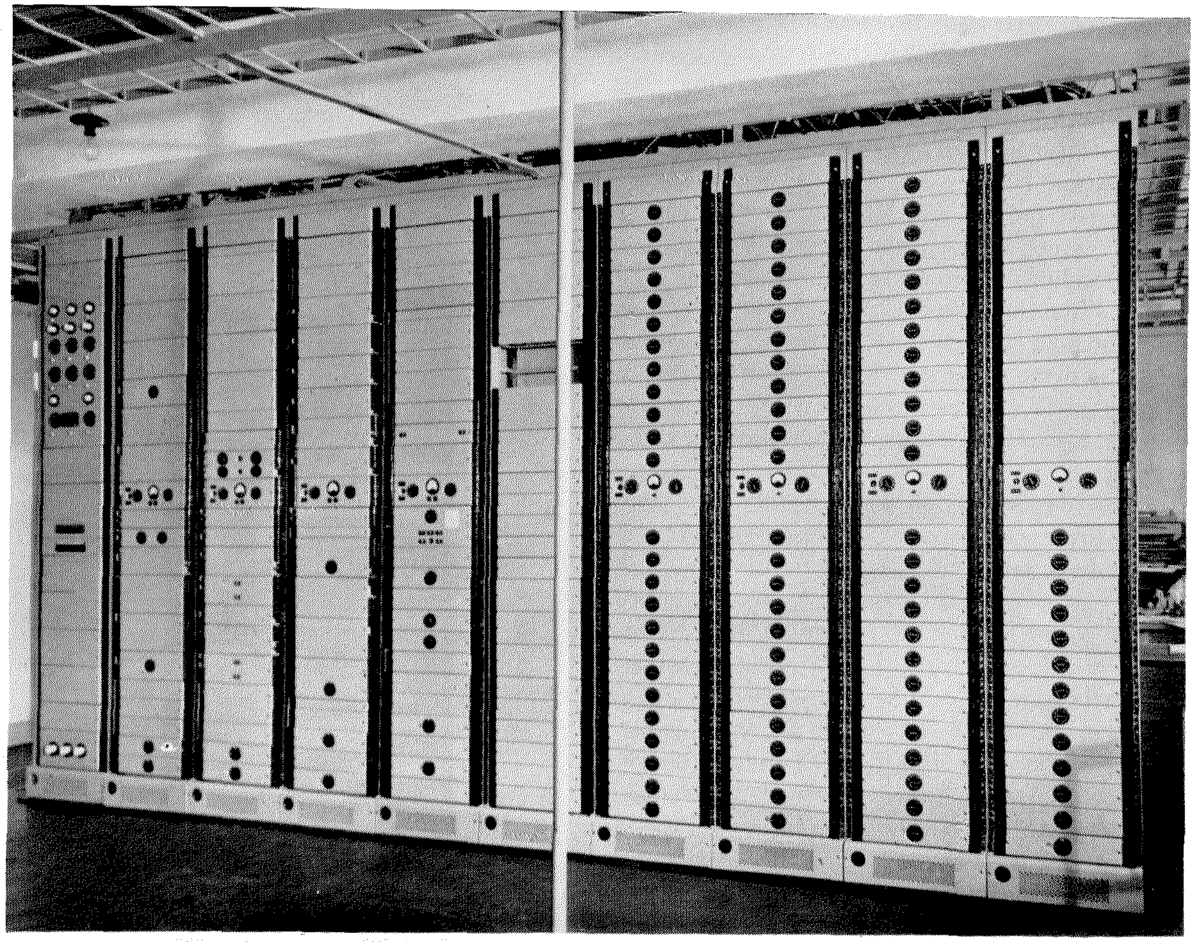


Figure 9—Level diagrams between Leeuwarden and Rømø.



A suite of bays in the Copenhagen carrier repeater station. From left to right are the following bays:—power supply supergroup carrier generator, group carrier generator, two channel carrier generators, and four channel bays.

4. *Repeater Equipment*

4. Submerged Repeaters

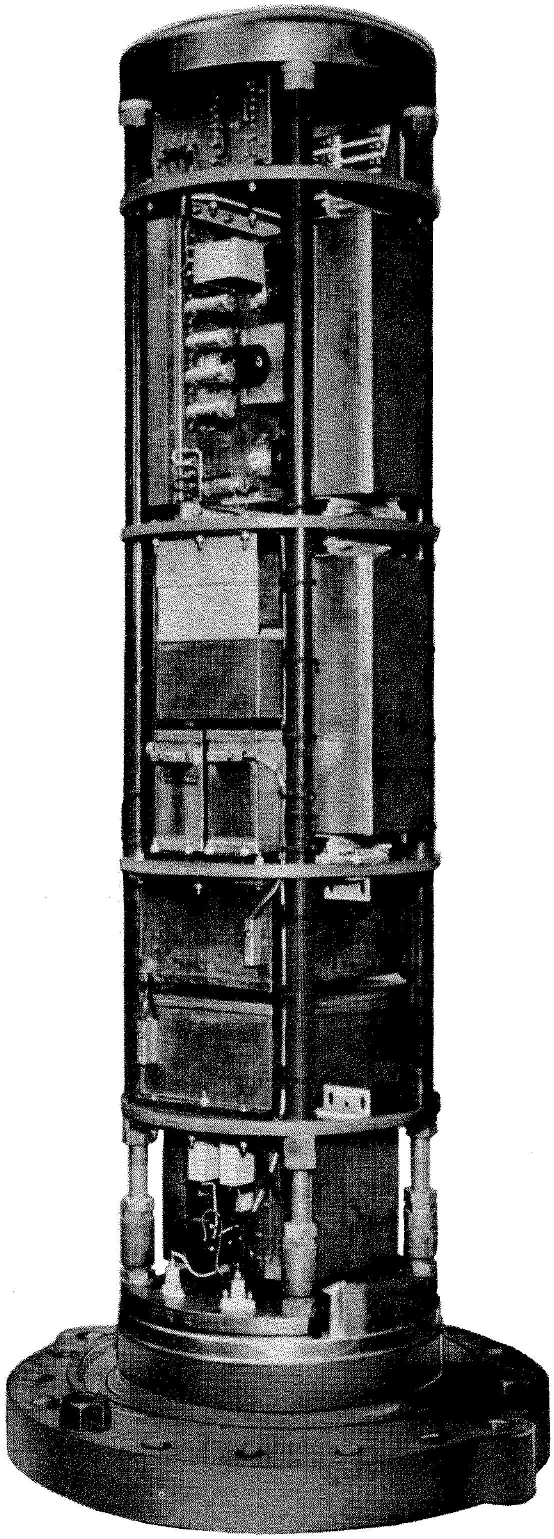
Each submarine cable is equipped with two submerged repeaters.

Many factors must obviously influence the design of repeaters, their size being chiefly determined by whether they are required to operate on a 2-wire or 4-wire basis. In this instance, 2-wire repeaters are necessitated by the fact that it is desired to use the same coaxial tube for both directions of transmission. In turn, the size of the repeater and the depth of water in which it is required to be laid dictate to a large extent whether it is mounted in a flexible housing constructed as an integral part of the cable or as a separate entity to which the cable is connected. In the case of the submarine cables between Rømø and Oostmahorn, 2-wire operation necessitated the use of a repeater that, because of its size, could not be designed as an integral part of the cable, while the relatively shallow depth of water (40 metres (131 feet) at its deepest point) did not unduly complicate the process of laying the cable to which it is connected. The type of repeater employed is mounted in a cylindrically shaped steel casing. The weight of the complete repeater is 919 kilograms (2026 pounds) and the outer casing is designed to withstand a pressure of about 70 atmospheres (72·3 kilograms per square centimetre or 1029 pounds per square inch). It is considered suitable for sea depths up to 400 metres (1312 feet).

The components of the repeater are mounted on a framework that is secured to the sealing plate of the repeater housing, and this framework is then enclosed in a hermetically sealed brass case, the case being filled with nitrogen to exclude the effects of oxidation on the components. The assembly thus secured to the sealing plate is then enclosed in a heavy steel case, the sealing plate being secured to this housing by a number of heavy steel bolts, a gutta percha sealing gasket being clamped between the sealing plate and the lip on the housing, to exclude the sea water. The cables enter the repeater through special glands in the sealing plate and the joints between these cables and the main sea cables are made within a chamber that gives mechanical protection but does not exclude the sea water, this of course being unnecessary.

The power supply to the repeaters is by constant direct current of 0·445 ampere. At the Oostmahorn end, one 224-volt battery for both cables is provided with its positive pole connected to the centre conductor of the cables and its negative pole connected via the insulated copper conductor of a cable to a large steel plate lying on the sea bed some distance off shore. The voltage drop on this conductor is continuously monitored. At the Rømø end, a metal rectifier cubicle for each cable, operating from the station alternating-current supply and producing 278 volts, is provided with its negative pole connected to the centre conductor of the cable and its positive pole connected to a similar earth plate to that at Oostmahorn. Thus the two power supplies at the ends of each of the cables are in series and produce a current along the centre conductor of the cable that returns through the sea. The valve heaters are connected in series with the centre conductor and the screen and anode voltages are produced by dropping the appropriate number of volts in the heaters and in a series resistance. The total voltage drop through the repeater is 220 volts. The rectifier set at Rømø is automatically regulated so that the current is maintained to ± 1 percent of its nominal value, this degree of control being desirable to get the maximum life out of the repeater valves. Provision is also made for automatic shut down if the current varies suddenly by more than ± 7 per cent, thus preventing surges in the primary supply or breaks in the cable from damaging the repeaters.

A block schematic of the repeater is shown in Figure 10. In principle, the repeater is essentially



Submerged amplifier showing assembly and wiring. The amplifier unit is mounted on the heavy sea plate forming, as it were, the bulkhead between the amplifier section and the 'jointing chamber.'

the same as that used with land coaxial cables, being a 3-stage amplifier with negative feedback, except that only one amplifier is used for both directions of transmission.

A by-pass low-pass filter across the repeater circuit is arranged to give a low-frequency speaker circuit between the terminals that is independent of the repeater.

An important aspect in the construction of any submerged repeater is that all components shall be capable of continuous service without maintenance. This means that not only must all components be manufactured with the greatest precision but that, as far as possible, they must be tested to see if they are likely to make good their predictable life.

This is particularly the case with valves, all of which are specially designed for long life and subjected to an endurance test over a period of 500 hours during which a very careful observation is made of heater, anode, and grid currents. If there are no signs of deterioration during this time, past experience suggests that it is reasonable to assume that they will continue to give the same standard of performance for several years. Similar tests are applied to other vulnerable components such as coils and capacitors.

4.2 Land Repeater at Oostmahorn

The repeater in each cable transmits the same frequency bands as the submerged repeaters and operates in a similar way on a 2-wire basis, but is provided with separate amplifiers for each direction of transmission and does not receive its power supply over the cables. Access is provided to the speaker circuit at this station, which although normally unattended can be visited for maintenance purposes. A block diagram of the repeater is shown in Figure 11.

4.3 Coaxial–Line Repeaters in Denmark

The land repeaters in Denmark are located at distances of up to 10 kilometres (5·4 nautical miles) apart, and between Copenhagen and Rømø there are 34 repeater sections. Figure 12 shows the general layout of these repeaters and the power supply arrangements. The only stations normally attended are Copenhagen, Kolding, and Rømø. The interstice wires of the coaxial cables are arranged to provide alarms from unattended to attended stations and also service speaker circuits.

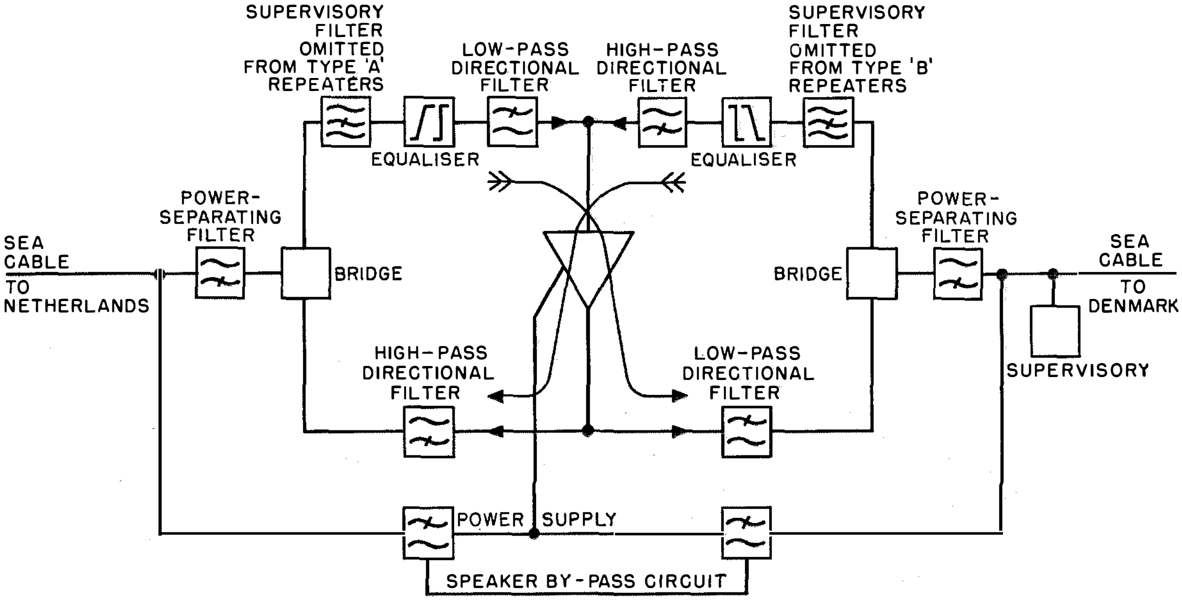


Figure 10—Simplified block diagram of the submerged repeaters.

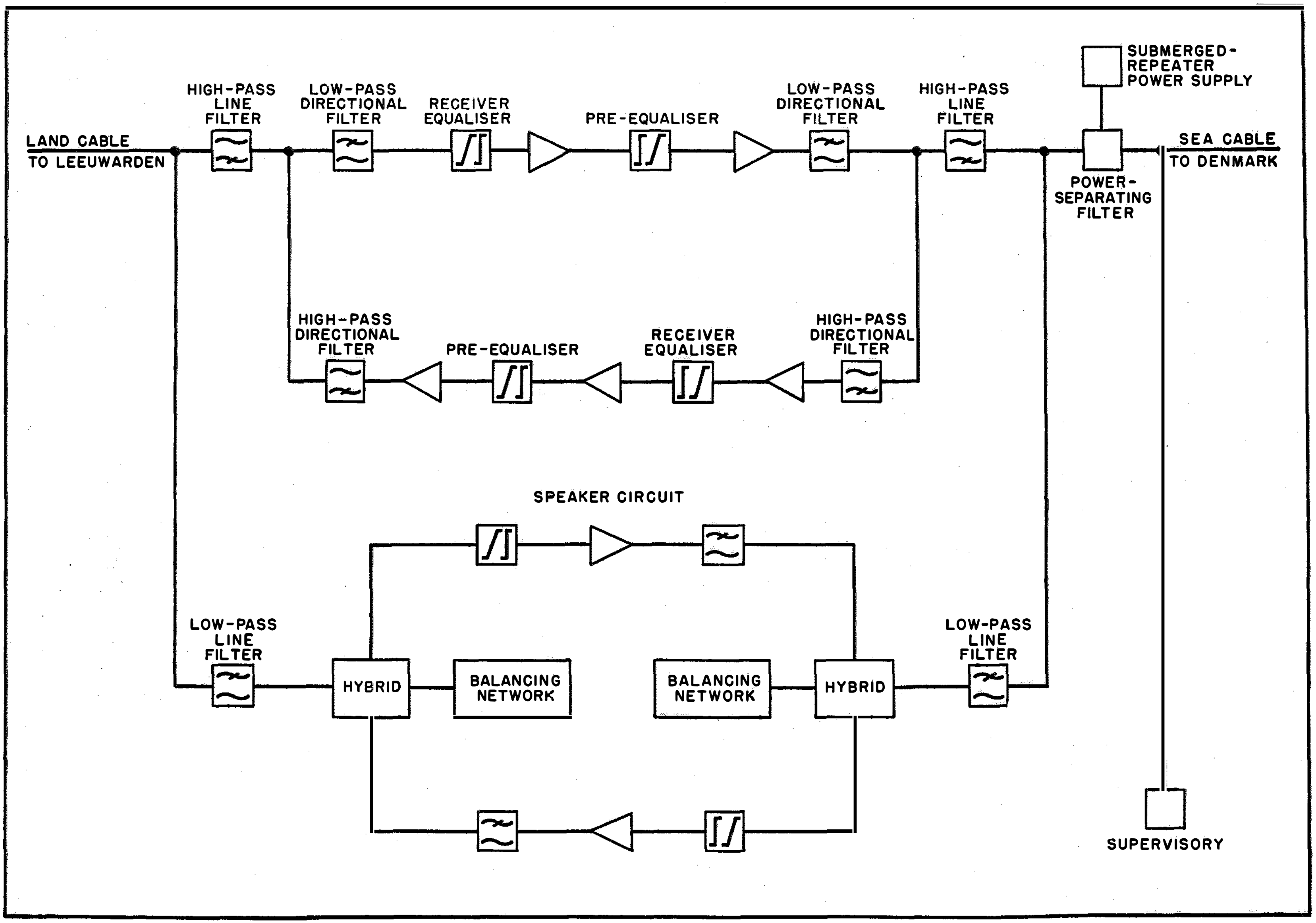


Figure 11—Shore repeater station at Oostmahorn. It is equipped for a single cable only.

5. *Pilot and Test Frequencies*

Between Leeuwarden and Rømø, the small temperature changes of the cable in the sea do not warrant automatic pilot regulation; minor adjustments of equalisation on the terminal equipment, only required at long intervals, are facilitated by the transmission over the cable of test frequencies at each edge of the frequency bands. The frequencies transmitted from Leeuwarden are 20 and 172 kilocycles and from Rømø, 204 and 356 kilocycles. These test frequencies may be applied to either of the two cables and, by the use of selective measuring equipment routine checks may be made without interruption of service.

A test frequency of 60 kilocycles may also be transmitted from Leeuwarden over each cable, and at Rømø may be connected to a frequency-comparison panel where it is compared with a frequency of 120 kilocycles derived from the local carrier-generation equipment. In this way, the master oscillators in the Netherlands and in Denmark can readily be synchronised.

It is important, of course, to be able to test the submerged repeaters while they are in service so that preparations can be made in good time for replacement before actual failure takes place. Two testing facilities are provided, one of which measures the loop gain and the other intermodulation.

To measure loop gain, a virtual carrier frequency in the band, 24–168 kilocycles, is sent from Leeuwarden and in the repeater this is picked off at the high-level output by a narrow-band filter, passed to a frequency doubler, and then twice the frequency, which lies in the band, 208–352 kilocycles, is put back into the repeater at the same point at a suitable level. This doubled frequency then passes through the repeater in the reverse direction and back to Leeuwarden, where its level can be measured by means of a selective measuring set. Thus a measure of the loop gain is obtained. Because Oostmahorn is an unattended station, a similar facility is provided. Thus by arranging that different frequencies are selected by each repeater, measurements of loop gain can be made from Leeuwarden to Oostmahorn, to the first submarine repeater end, and to the second submarine repeater. The transmitted frequencies from Leeuwarden are 152, 160, and 168 kilocycles respectively.

To measure intermodulation, which gives an

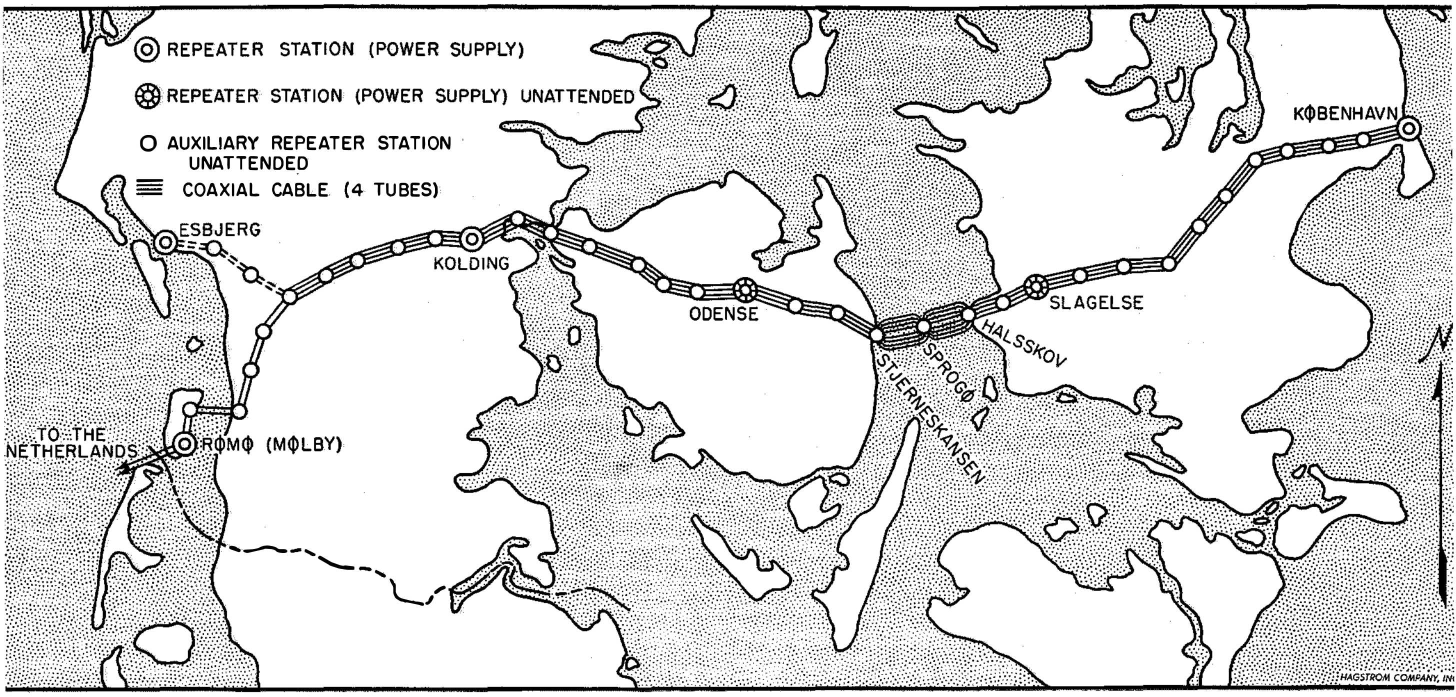


Figure 12—Coaxial cable and repeaters between Copenhagen and Rømø via Kolding.

indication of how the valves are ageing, two high-level test tones are sent simultaneously, one from Leeuwarden and one from Rømø. Again these test tones are virtual carrier frequencies so that measurements can be made while the system is in traffic. The frequencies are chosen so that the third-order intermodulation product is 312 kilocycles, which then travels back to Leeuwarden where its level can be measured by means of the selective-measuring set mentioned above. Certain band-elimination filters have been included in the repeaters to ensure that, for one particular pair of test tones, intermodulation is produced in only one repeater and the second repeater is checked by a different pair of tones. The frequencies chosen are:—

A. 236 kilocycles from Rømø.
   160 kilocycles from Leeuwarden, as for the loop gain test.
   Product is $(2 \times 236) - 160 = 312$ kilocycles.
B. 240 kilocycles from Rømø.
   168 kilocycles from Leeuwarden, as for the loop gain test.
   Product is $(2 \times 240) - 168 = 312$ kilocycles.

It should be noted that 312 kilocycles is a virtual carrier frequency, and at Rømø there will be carrier leak at this frequency starting back in Kolding or Copenhagen. A 312-kilocycle band-elimination filter is therefore equipped at Rømø so that the intermodulation measured at Leeuwarden will not be obscured by carrier leak.

Over the land coaxial system Rømø–Kolding–Copenhagen, pilots of 2604 kilocycles and 60 kilocycles in each direction are used to control automatically the gain of the repeaters to compensate for temperature changes on the system.

Intersupergroup test frequencies of 308, 808, 1056, 1304, 1552, 1800, 2048, and 2296 kilocycles in each direction may be applied when required at Copenhagen and Kolding and by the use of selective-measuring equipment allow routine measurements of the line frequency characteristic to be checked. Between Kolding and Rømø, intersupergroup frequencies of 308, 808, and 1056 kilocycles may be used for the same purpose. Frequencies of the master oscillators may also be checked by the transmission over the cable of a test frequency of 60 kilocycles, where it is compared on an oscilloscope with a corresponding frequency at the receiving end.

6. Cable Supervision

The line filters of the speaker circuit in the submarine repeaters and at Oostmahorn, Leeuwarden, and Rømø effectively transmit a frequency band 0·3 to 10 kilocycles with little distortion. This band may be used for detecting submarine-cable faults by means of impedance measurements or pulse test gear.

The land cable in Denmark is kept under gas pressure using nitrogen at 0·5 atmosphere (0·5 kilogram per square centimetre or 7·4 pounds per square inch). To detect failure in the lead sheath of the cable, contact manometers, designed by the Royal Swedish Telegraph Administration, are inserted in the cables at each joint, about 300 metres (984 feet) apart. The contact manometer is shown in Figure 13. It consists of a glass tube that is closed at one end and that is provided with a mercury column that under influence of the pressure moves as a piston in the glass tube. At a given pressure, a short-circuit will appear between the platinum electrodes. The manometers are adjusted to give this contact at 0·3 atmosphere (0·3 kilogram per square centimetre or 4·4 pounds per square inch) and as the position of the mercury column is decided by the relative pressures in the closed air space in the manometer and in the cable, the adjustment is independent of the temperature. The extra mercury column in the manometer serves as protection against contamination of the actual contact column. If the pressure falls in the cable due to a leakage in the lead sheath, the contact manometer nearest the leak will close first and give an alarm over the service pair in the cable to the adjacent repeater station, from which, by a

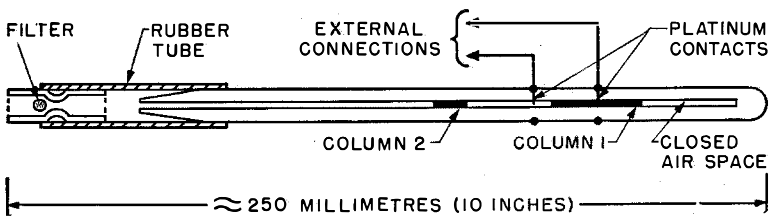
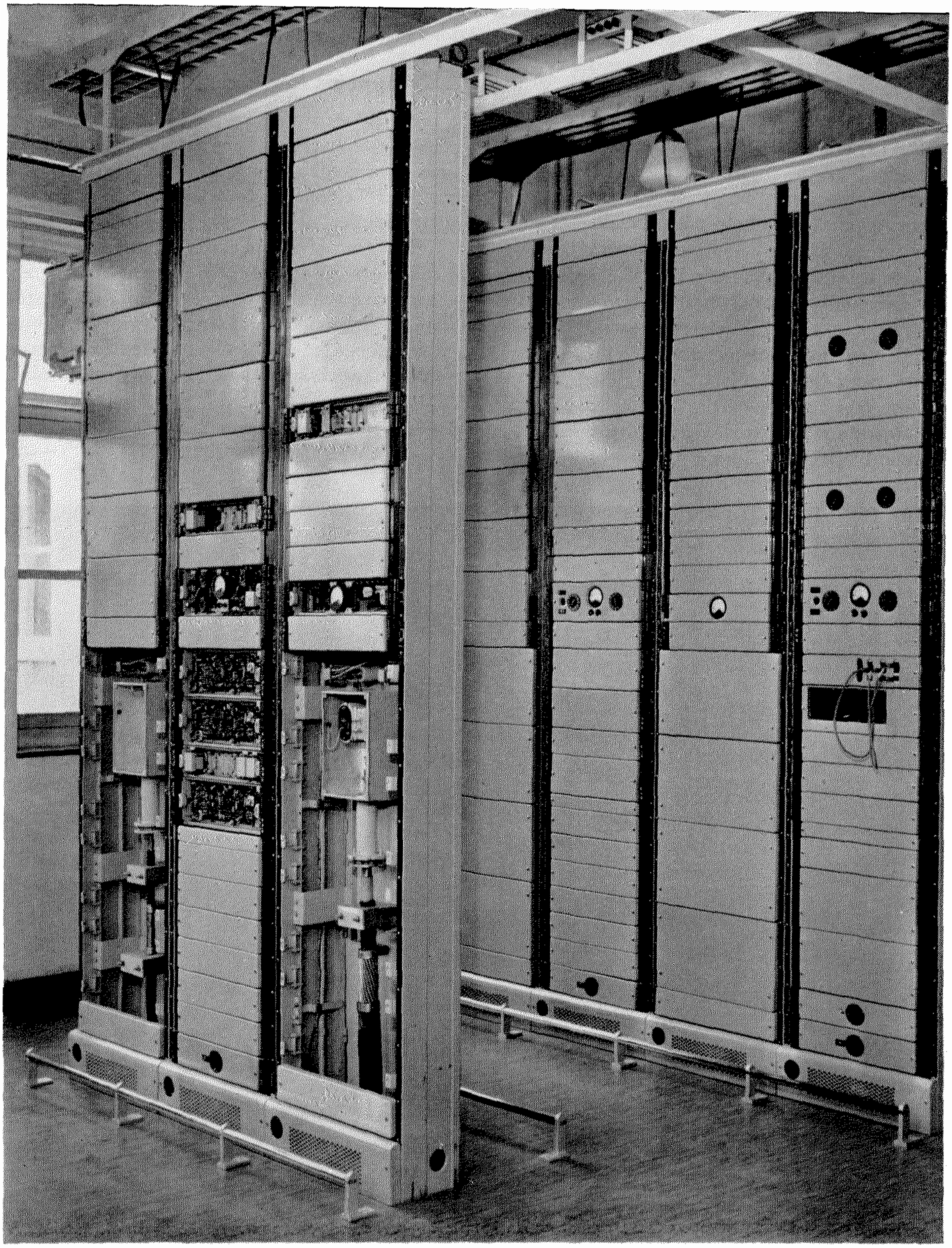


Figure 13—Contact manometer.



Suite of bays at Oostmahorn. Covers have been removed to show one of the submarine coaxial-cable terminations, the amplifier, and on the left the termination of one of the two-band coaxial cables to Leeuwarden.

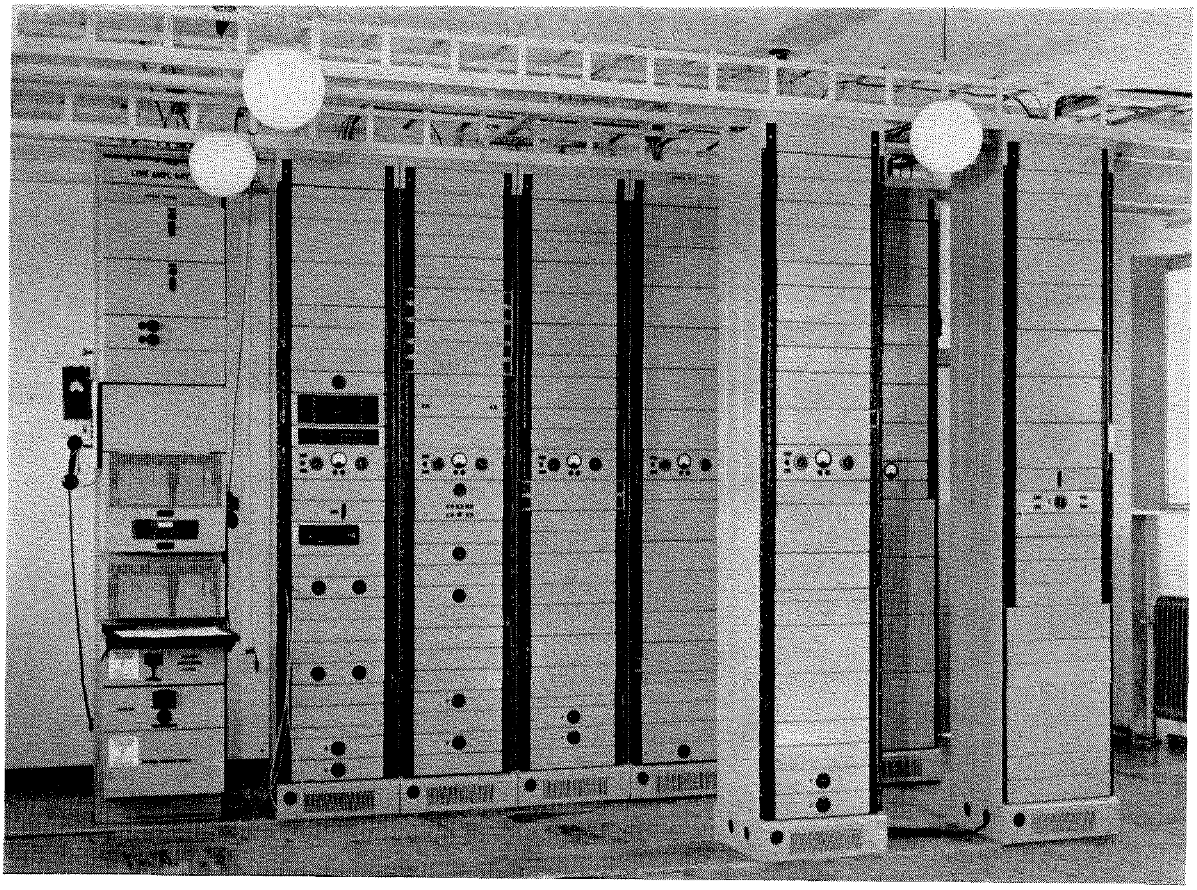
resistance measurement, it can be determined which manometer has operated. In this way the first rough determination of the position of the fault is given. The exact position is then found by putting into the cable a small amount of radium emanation (10 millicuries corresponding to under 0·01 millimetre (0·0004 inch)) from one terminal. By regulation of the pressure from the two ends of the cable, a constant flow of gas is maintained. The radio-active gas will now slowly flow to the faulty place and from here out into the surrounding earth, where it can be easily detected by means of a geiger counter. The radio activity cannot be detected through both cable and surrounding earth but only where the gas has escaped into the earth. This method was, as far as is known, first developed by the Danish Post and Telegraph Administration's laboratory, and it has been shown to give excellent determinations of faults within 1 metre (3·28 feet).

7. *Equipment Features*

With the exception of the submerged and coaxial-line repeaters, where special arrangements have been used, all the equipment is arranged in Standard Telephones and Cables 'New Equipment Practice,' described earlier in this journal.[2]

Connection of the panels with the bay wiring is made by means of plug-in-type connectors, which serve the dual purpose of completing the circuit and providing jack-in points for test leads. All components likely to be affected by climatic conditions are housed in hermetically sealed cans.

[2] F. Fairley, R. J. M. Andrews, and A. C. Delamare, "Improved Equipment Practice Reduces Size of Telephone Transmission Systems," *Electrical Communication,* v. 27, pp. 21–38; March, 1950.



General view of the submarine cable terminating station at Rømø. The equipment serves the essential functions of translating the two groups of 2-wire circuits over the submarine cable to normal 4-wire operation over the inland Danish network. Associate functions such as testing and providing power to the submarine repeaters and to the auxiliary land repeaters along the route to Kolding are also catered for at Rømø.

Rack frameworks and panels are specially designed to afford easy access to all components; panels calling for detailed examination may be removed from the bay and replaced by serviceable panels with a minimum loss in operating time.

Power supplies are arranged on each bay side with appropriate power-unit panels mounted at the bottom. These bay-side power units, of which there are a number of standardised types, are designed for connection directly to the alternating-current mains supply.

Photographs show the equipment installed at various stations.

Standby power supplies are provided at all stations. At Oostmahorn, the floated battery with reserve of 80 hours that supplies the submerged repeaters also supplies the anode voltage of the repeater at that station and the heaters are normally supplied directly from the mains by a transformer. In the case of mains failure, an automatic-starting motor-alternator operated from the same battery supplies the heaters with a break of two seconds, which is insufficient to allow them to cool appreciably. This arrangement ensures that no traffic break occurs on failure of the mains supply. At Rømø, where no mains supply exists, continuously running diesel alternators, with standby, allow the equipment to operate without break in the event of failure of the normal arrangement. In other terminal and power supply stations, standby alternators automatically switched into service in the event of failure of the mains supply are used. The power supplied to the equipment bays is automatically

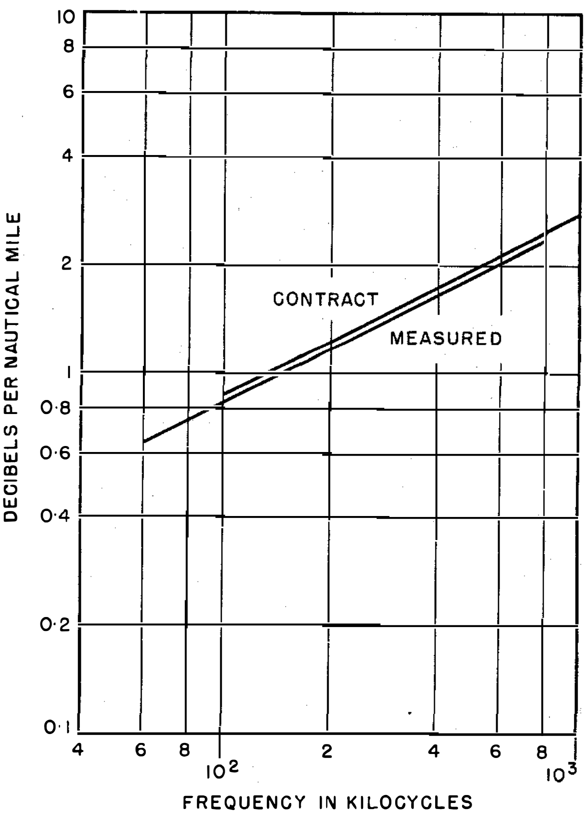


Figure 14—Attenuation versus frequency for the Netherlands–Denmark submarine cable.

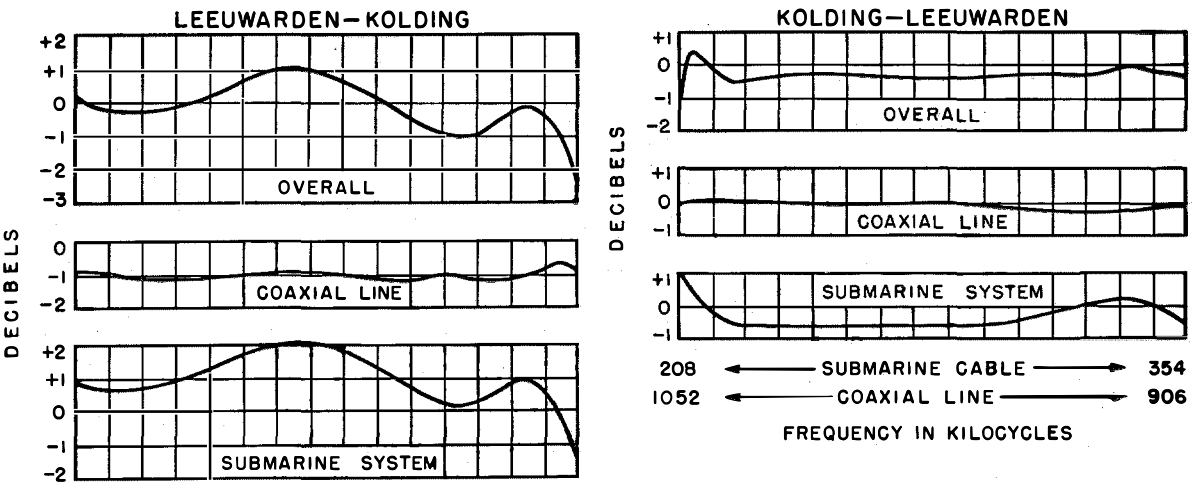


Figure 15—Attenuation characteristics of the Leeuwarden-to-Kolding section including terminal and repeater equipment (of cable *2*).

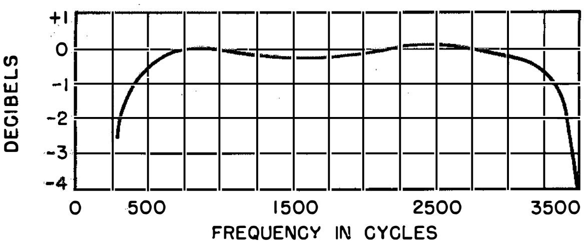


Figure 16—Overall frequency characteristic of a typical speech circuit between Copenhagen and Rotterdam.

regulated to ±1 per cent by station regulators that operate on both normal and standby supplies.

8. Performance Data

Figure 14 shows the attenuation–frequency characteristic of the submarine cable and Figure 15 shows the corresponding overall frequency characteristics including terminal and repeater equipments of the whole section from Leeuwarden to Kolding.

Figures 16 and 17 show overall frequency characteristics typical of the speech circuits between Copenhagen and Rotterdam and of the broadcast programme channel.

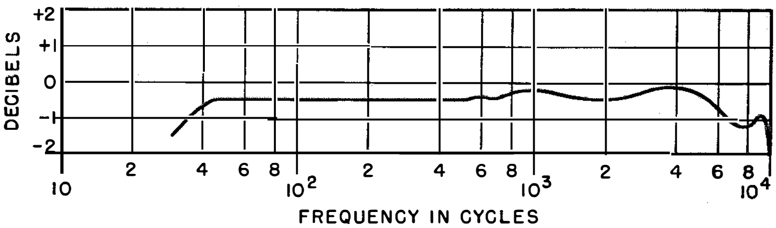


Figure 17—Overall frequency characteristic of the broadcast programme circuit between Copenhagen and Rotterdam.

9. Conclusion

The Netherlands–Denmark link, providing long-distance circuits of the highest quality and planned for extension as traffic develops, will play a major part in the improvement of communications over the whole of the western part of Europe. As an example of this, it may be mentioned that direct groups of 12 channels are operating between London and Copenhagen over this link without demodulation to audio frequency at any intermediate point, and that such groups are amplified together en route about 60 times with a total amplification of about 3300 decibels or 10^{330}.

Some Applications of Cold-Cathode Tubes to Switching Systems

By S. SIMON

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AMONG the wide diversity of problems with which the telephone switching engineer is daily faced are some that can be solved most readily by resorting to electronic devices. This will be the case, for instance, when the energy available is too small to operate directly an ordinary relay or when signals of very short duration have to be detected.

Given the quality of performance expected from the device, it is left to the designer to solve the particular problem in the most economical way by a judicious choice among the available different types of electronic tubes, such as multi-electrode vacuum tubes, thyratrons, and gas-filled cold-cathode tubes. Although the performance of thyratrons is in many respects superior to that of cold-cathode tubes, there are many instances where the limitations of the latter are more than outweighed by their unique advantages; they require no heating power, are subjected to no wear during idle periods, have a long service life, and need little or no maintenance.

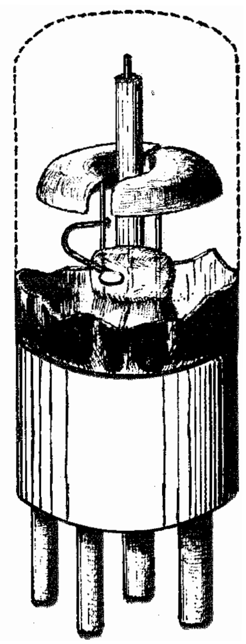


Figure 1—Cold-cathode relay tube.

It is the purpose of this paper to give a short description of some of the most characteristic applications of cold-cathode tubes to telephone switching and in particular to the rotary system.

Before starting with this description, a brief account will be given of the structure and main characteristics of the *2313* type of cold-cathode tube, which is used throughout in the applications described later.

1. *Structure and Characteristics of the 2313 Tube*

This tube is similar to the Western Electric type *313* tube, which was developed by Bell Telephone Laboratories. As may be seen in Figure 1, it consists of three elements, a cathode, anode, and control anode, enclosed in a glass envelope filled at low pressure with a monatomic-gas mixture consisting of 99 percent of neon and 1 percent of argon.

The cathode is a semicircular nickel electrode having a relatively large surface that is coated with a thin layer of a mixture of barium and strontium oxides.

The anode proper is just the end of a bare nickel wire separated some 12 millimeters (0.46 inch) from the other electrodes. The control anode is placed very near the cathodes, has the same shape as the latter, and is similarly located with respect to the anode. In fact, the control anode in the *2313* tube is interchangeable with the cathode. Although this disposition is not essential to cold-cathode tubes in general (the control anode might as well be reduced to a small surface near the cathode), it may prove useful in certain applications as will be seen later.

The space between anode and cathode will be referred to as the main gap. Similarly, the space between control anode and cathode will be defined as the control gap.

Assuming now that no current is flowing in the control gap, if a source of direct voltage is connected across the main gap with the positive side to the anode, it will be observed that the current flowing in the main gap remains negligible as long as the potential difference remains below a critical value called the main-gap breakdown voltage.

When this voltage is reached, however, cumulative ionization takes place in the gas and the current increases suddenly to a very high value that would lead to the instantaneous destruction of the tube if no limiting resistance had been placed in series with the tube. At the same time, the potential difference between anode and cathode drops to a lower value, called the main-gap sustaining voltage, which is practically constant (at least within the current range for which the tube is rated) and independent of the current, the magnitude of which is determined by the limiting resistance and the voltage of the direct-current source. The current can now be interrupted only by reducing the potential difference between anode and cathode below the sustaining voltage.

Similar observations can be made if the direct-current source is connected across the control gap, the anode being disconnected: the gas ionizes when the potential difference reaches the so-called control-gap breakdown voltage and deionizes when it drops below the control-gap sustaining voltage.

Assuming now that the control gap is brought into the conducting condition, it will be observed that the potential difference between

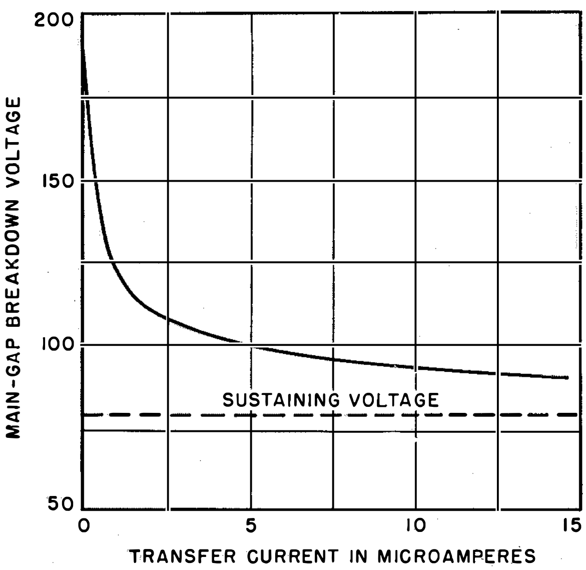


Figure 2—Breakdown voltage plotted against transfer current.

cathode and anode required to initiate the breakdown between these two electrodes is smaller than without current in the control gap: the main-gap breakdown voltage is a function of the current flowing through the control gap.

Given the cathode–anode voltage, the control-gap current needed to initiate breakdown in the main gap is called the transfer current.

The interesting fact is that only a very small transfer current is needed to initiate the main glow discharge. Once the latter has started, it becomes self-sustaining and is quite independent of the control anode.

Figure 2 gives the relation between transfer current and main-gap breakdown voltage. It will be seen from this curve that no transfer current is required to initiate breakdown in the main gap when the anode–cathode voltage is higher than 200 volts, while about 1 microampere transfer current is sufficient with 130–150 volts between anode and cathode.

The transfer process requires only a very short time, of the order of 100 microseconds, depending in some measure on the amplitude of the signal applied to the control anode and on the main-anode voltage.

To restore the tube to its deionized condition, it is necessary to interrupt the main discharge, as already mentioned, either by opening the anode circuit or by causing the main-gap potential to drop below the sustaining voltage.

This interruption must be maintained for a time sufficient to allow for the deionization of the gas. The minimum deionization time varies with the nature and the pressure of the gas, with the intensity of the current in the main gap, with the nature of the anode load impedance, and with the relative potentials on the electrodes during the deionization process. For the *2313* tube, this time does not exceed 10 milliseconds and by a suitable choice of the above-mentioned factors can be reduced considerably below this value.

As already noted, the main-gap sustaining voltage is fairly constant and independent of the main-gap current. It is primarily dependent on the cathode fall of potential, which in turn is dependent on the nature of the cathode material. The slope of the current–voltage characteristic depends not only on the nature of the cathode surface but also on its area: a small cathode surface would produce a steep characteristic. It can thus be readily understood that if the relative polarity between anode and cathode is reversed, i.e., anode negative and cathod positive, a steep current–voltage characteristic will be obtained, owing to the fact that the anode with its small noncoated area is now used as a cathode. Thanks to this asymmetrical current–voltage characteristic, the cold-cathode tube can be used as a rectifier and as a polarity discriminator. This

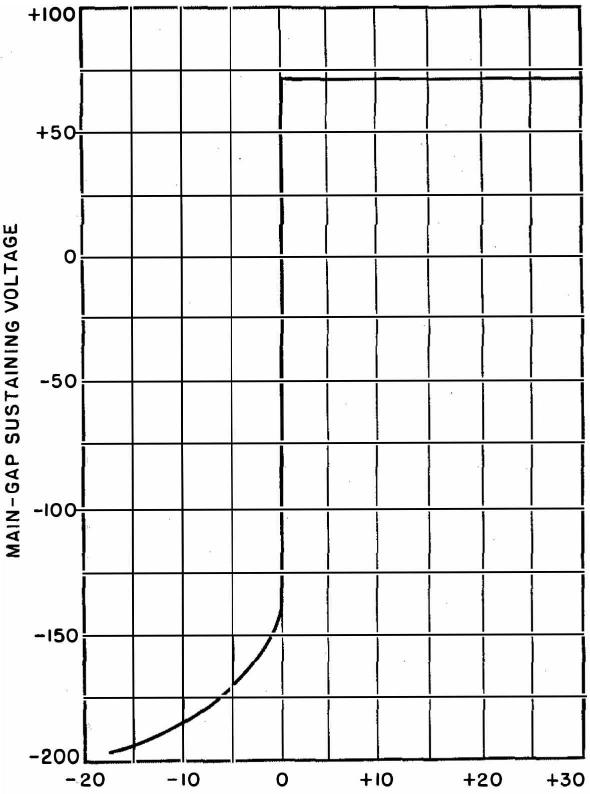


Figure 3—Asymmetrical main-gap current–voltage characteristic. The sustaining voltage is maintained constant across the main gap over a wide range of current.

characteristic is shown in Figure 3 for the *2313* tube. Table 1 gives the electrical characteristics of the tube.

2. *Life of Tube*

The useful life of such a cold-cathode tube is limited by the destruction of the cathode coating. The cathode is submitted to a positive ion bombardment that causes sputtering of the coating. Life decreases rapidly with increasing current as will appear from Table 2. This table is based on previously published data.[1]

[1] G. H. Rockwood, "Current Rating and Life of Cold-Cathode Tubes," *Transactions of the American Institute of Electrical Engineers*, v. 60, pp. 901–903; September, 1941.

It is important for the designer to keep well in mind this relationship between life and current drain. As there is no wear whatever when the tube is at rest, such tubes will be especially adapted to devices that are maintained for most of the time in the idle or waiting condition and have to work for only brief periods. Under such conditions, and with a high working current corresponding to a life of, say, a few tens of hours of

TABLE 1

ELECTRICAL CHARACTERISTICS OF THE *2313* TUBE

Nominal Control-Gap Breakdown Voltage	70 Volts
Nominal Control-Gap Sustaining Voltage	60 Volts
Minimum Main-Gap Breakdown Voltage	200 Volts
Nominal Main-Gap Sustaining Voltage	80 Volts
Maximum Transfer Current (130 Volts on Anode)	5 Microamperes
Rated Current for Expected Life of 20,000 Hours	10 Milliamperes

actual duty, the practical service life may amount to several years or tens of years.

Let T be the life expressed in hours of operation at a given current i and let k be the duty factor that is obtained by dividing the time the current is effectively passing through the tube during a given period (one day for instance) by the length of that period. Then the useful life would be T/k. In other words, it may be said that each hour a fraction k/T of the cathode coating would be sputtered away. If, during an operating cycle, the current assumes different values $i_1, i_2, \cdots i_n$, with duty factors $k_1, k_2, \cdots k_n$, respectively, let $T_1, T_2, \cdots T_n$ be the corresponding life times as given in the table above.

Thus, each hour a fraction

$$\frac{k_1}{T_1}+\frac{k_2}{T_2}+\cdots+\frac{k_n}{T_n}$$

TABLE 2

LIFE RELATED TO TUBE CURRENT

Tube Current in Milliamperes	Life in Hours
10	20,000
20	1,000
30	150
50	40
100	9
150	4

of the cathode coating would be destroyed. The useful life is readily obtained by the expression

$$\frac{1}{\frac{k_1}{T_1}+\frac{k_2}{T_2}+\cdots+\frac{k_n}{T_n}}.$$

If for instance for each operating cycle, the tube has to pass 50 milliamperes i_1 during 0.01 second and 15 milliamperes i_2 during 0.2 second, and if as an average of 100 such operating cycles occur each hour, the duty factor corresponding to i_1 is

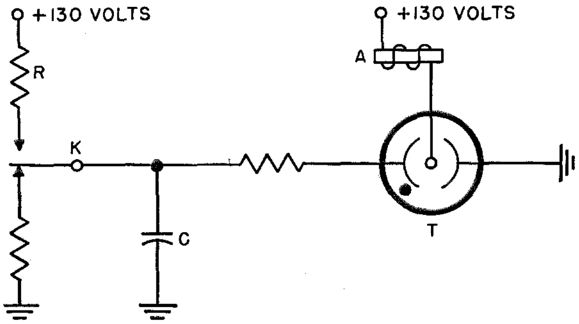


Figure 4—Delay device. The time interval between the operation of K and the energizing of relay A is controlled by the values of R and C.

$$k_1=(0.01\times 100)/3600=1/3600$$

and that corresponding to i_2 is

$$k_2=(0.2\times 100)/3600=1/180.$$

The life T_1 according to the table would be 40 hours and T_2 would be 2000 hours. The useful service is thus

$$\frac{1}{\frac{1}{3600\times 40}+\frac{1}{180\times 2000}}=100{,}000 \text{ hours}$$

or 11.5 years.

This example has been given to illustrate the fact that for a given service life expected from the tube, the lighter the duty cycle, the higher the current that may be allowed to pass through the tube.

As suggested in Rockwood's article,[1] the most useful type of rating for cold-cathode tubes would be to state the currents that will give a life of 100, 1000, and 10,000 hours. One can only regret that this method has not been adopted yet by cold-cathode-tube manufacturers.

3. Applications to Rotary Switching Systems

The particular properties exhibited by cold-cathode tubes make them adaptable to various specific uses, such as; highly sensitive relay, short-impulse-detecting device, self-locking relay, snap-action relay, marginal relay, polarized relay, rectifier, stabilizer, and potential limiter. Several of these specific uses are combined in some of the applications to be described.

Two delay devices are shown in Figures 4 and 5. In Figure 4, when contact K is operated, Capacitor C charges via resistance R and eventually reaches the breakdown potential of tube T after a delay determined by the time constant of R and C. Delays from a few milliseconds to several seconds or tens of seconds can be obtained using only small paper capacitors. The maximum delay attainable is set by the internal leakage resistance of the capacitors and the insulation resistance of the wiring.

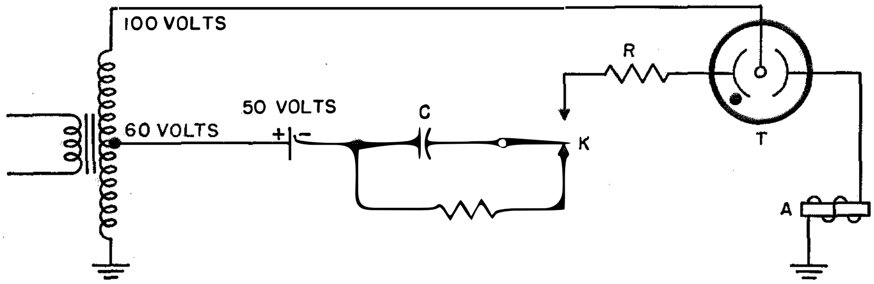


Figure 5—Alternating-current-operated time-delay circuit.

The delay device of Figure 5 can work with alternating currents exclusively as the 50 volts of direct current required in the control-anode circuit can be derived from the alternating-current mains by means of a small rectifier. When contact K is operated, capacitor C charges through resistance R on the negative half cycles through a reverse breakdown in the control gap until the charge is sufficient to break down the control gap on the positive half cycles. The main gap thus ionizes on every positive half cycle and the relay operates. The relay has been placed in the cathode circuit to avoid capacitor C loosing an appreciable part of its charge through the control gap on the positive half cycles. In fact, each time current flows in the main gap, the voltage drop through the relay coil is sufficient to interrupt the control-gap current. It will be noted that in this scheme, the tube *T* acts simultaneously as a rectifier and as a sensitive relay.

The periodic interrupter shown in Figure 6 may be used, for instance, to interrupt periodically a tone (busy tone, dead-level tone, etc.) in any required cadence. It is composed of two delay devices, such as the one described in Figure 4, mutually controlling each other. The pulse speed and ratio can be easily adjusted to any required values by a suitable choice of *R1*, *C1*, *R2*, and *C2*.

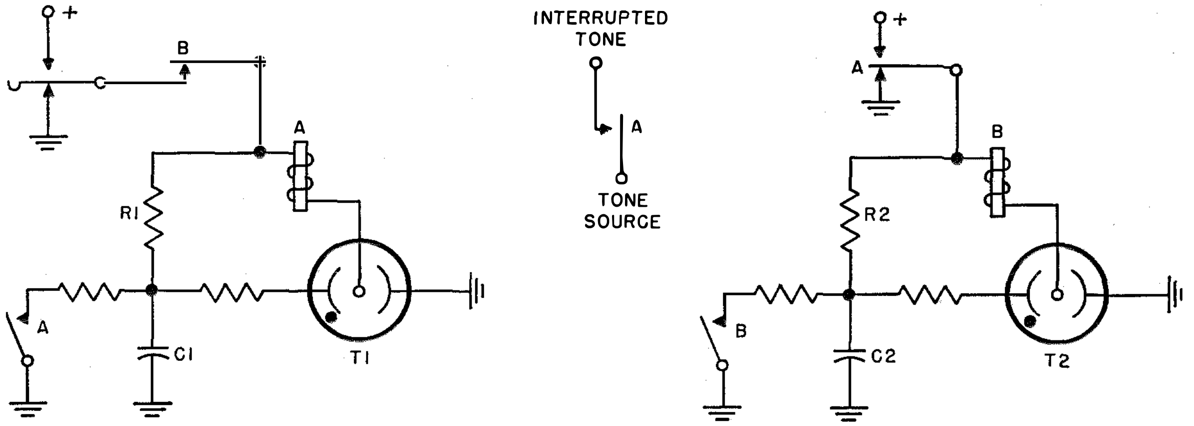


Figure 6—Periodic interrupter consisting of two interconnected timing circuits. The contacts are associated with the similarly lettered relays.

Figure 7 shows a device that is used to totalize the number of calls in a group of circuits. Each circuit of the group is provided with a contact *K* and associated resistance *R*. All these resistances are commoned to battery via the primary winding of transformer *TR*. Whenever a circuit is engaged, its contact *K* closes, causing an abrupt current increase in the primary winding of the transformer and a short impulse in the secondary winding. This impulse has a positive polarity and causes tube *T* to ionize. Relays *A* and *B* and the traffic meter operate. Relay *B* opens the anode circuit. Relays *A* and *B* and the traffic meter fall off and the device is ready for an eventual next impulse. Tube *T* does not respond to the negative impulses that are induced when-

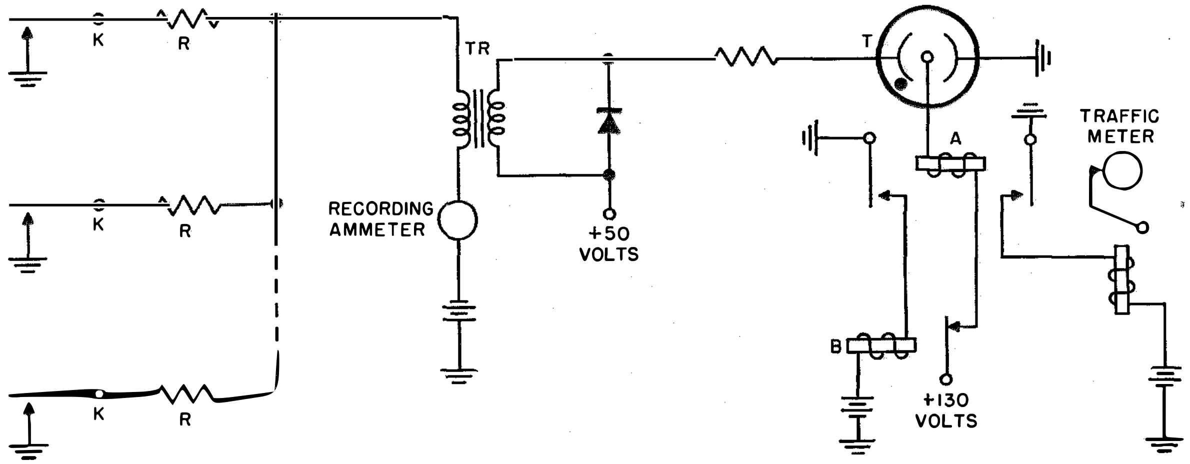


Figure 7—Traffic totalizer. The closing of any contact *K* will advance the traffic meter total by one unit. The recording ammeter will permit average holding time to be computed.

ever one of the contacts K opens as rectifier S acts as a short circuit for such negative impulses.

If used in conjunction with a recording ammeter calibrated to indicate traffic expressed in call–minute units, the average holding time for each circuit of the group can be readily obtained by dividing the total holding time of all calls as indicated on the ammeter by the total number of calls registered by the traffic meter.

The devices shown in Figures 8 to 12 were developed to convert low-energy alternating-current signals into direct-current signals without appreciable distortion. In Figure 8, the alternating-current impulses are rectified by the bridge rectifier S. The rectified signal added to the signal should be kept within reasonable limits by means of a signal limiter if necessary to avoid a reverse breakdown in the control gap on a signal of large amplitude.

In the scheme of Figure 9, two tubes $T1$ and $T2$ have been provided, mutually controlling each other by means of relays A and B. Tube $T1$ ionizes and relay A operates at the beginning of each alternating-current signal. Relay A prepares the circuit for relay B in series with tube $T2$ which, however, cannot ionize as long as the signal is present. In fact, $T3$, which acts both as a limiter and as a full-wave rectifier, keeps the potential on capacitor C sufficiently low to prevent the control gap of $T2$ from ionizing. As

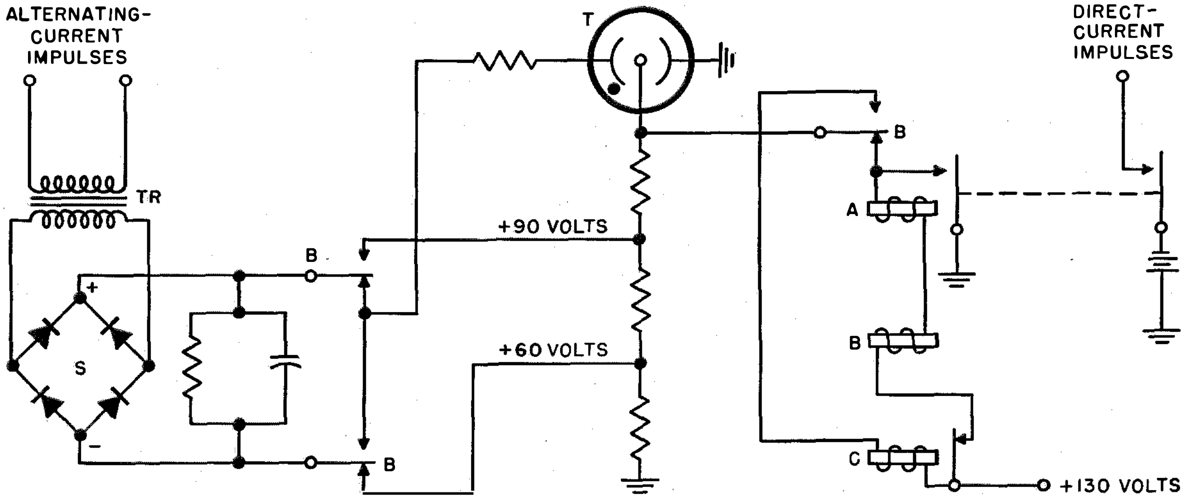


Figure 8—Method of converting alternating-current impulses into direct-current impulses without timing distortion.

the 60-volt bias causes tube T to ionize so that relays A and B operate and lock to ground through one contact of relay A. Relay B now raises the control-anode biasing voltage to +90 volts and reverses the polarity of the rectifier bridge, so that the rectified voltage now prevents the ionization of the tube as long as the signal is present. When the signal is over, the tube ionizes again, relay C operates and opens the circuit of relays A and B, which fall off. The latter in turn opens the anode circuit of the tube, which deionizes, while relay C releases. At the same time the control-anode biasing voltage is set again to +60 volts.

The same cycle of operations takes place for each of the following alternating-current impulses. For reliable operation, the amplitude of soon as the signal disappears, $T3$ deionizes, capacitor C charges to +130 volts through resistance R, tube $T2$ ionizes, and relay B operates. Tube $T1$ deionizes and relays A and B are de-energized.

The same sequence of operations takes place for each of the following alternating-current impulses. Relay A thus follows closely and repeats the incoming alternating-current signals. Owing to the limiting action of $T3$, the amplitude of the signals may vary within very wide limits. It will also be noted that advantage has been taken of the interchangeability of cathode and control anode, which characterizes the *2313* type of tube, to allow $T3$ at act simultaneously as a rectifier and as a limiter.

In the device shown in Figure 10, the sending relay SR sends alternating-current signals of different frequencies $F1$ and $F2$ to the line. At the receiving part of the equipment, the filter directs frequency $F1$ to the tube $T1$ and the frequency $F2$ to the tube $T2$. Both tubes mutually control each other by means of relays A and B. step-by-step switch $S2$. When the starting contact is closed, relay B operates on the first impulse and locks in series with relay C, which energizes in turn at the end of the first impulse. Relay C on the one hand places $S1$ under the

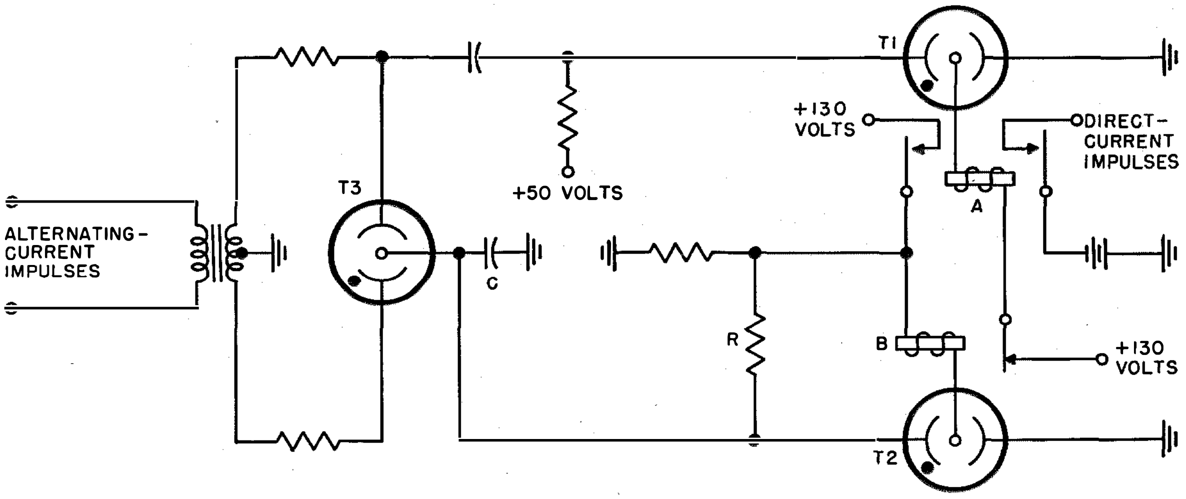


Figure 9—Another method of converting alternating-current impulses into direct-current impulses.

It can easily be seen that relay A repeats exactly the impulses from the sending relay SR.

When impulses have to be transmitted by means of alternating current from one circuit to another located in the same office, the simple scheme shown in Figure 11 may prove useful.

Relay D, which is stepping continuously, is common to both sending and receiving parts of the circuit and controls simultaneously the circuits connected to the step-by-step switch $S1$ and the receiving relay E with its associated control of relay D and, on the other hand, sends an alternating current to the receiving tube T that ionizes its control gap as long as relay C is operated. Relay E is now also under the control of the common relay D in series with the main gap of the tube. Transformer TR and its associated tube T thus act as a remotely controlled switch. Both switches $S1$ and $S2$ step in unison until $S1$ closes the "stop" contact. Relay A then energizes, causing relays B and C to release. The latter removes the alternating current so

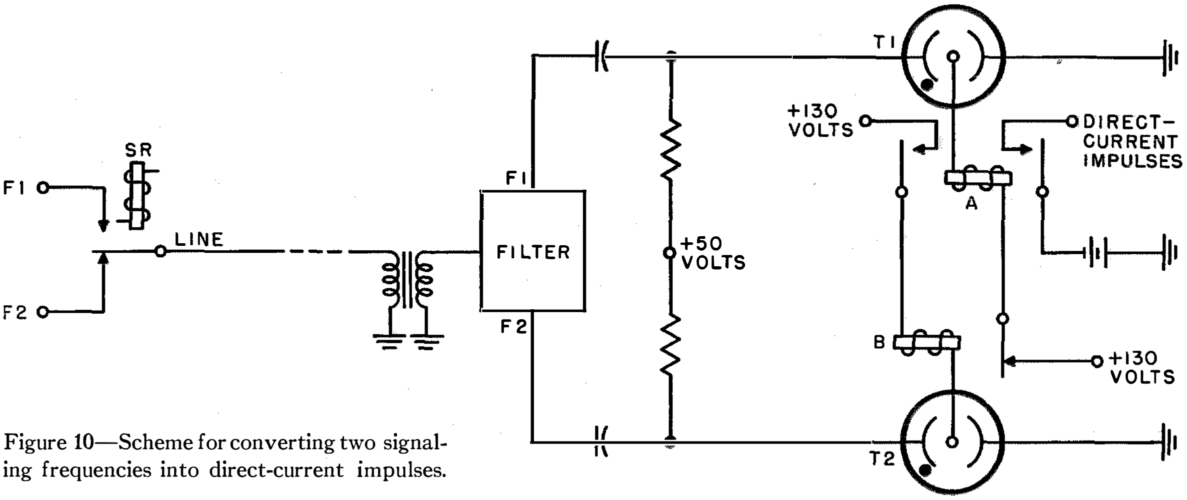


Figure 10—Scheme for converting two signaling frequencies into direct-current impulses.

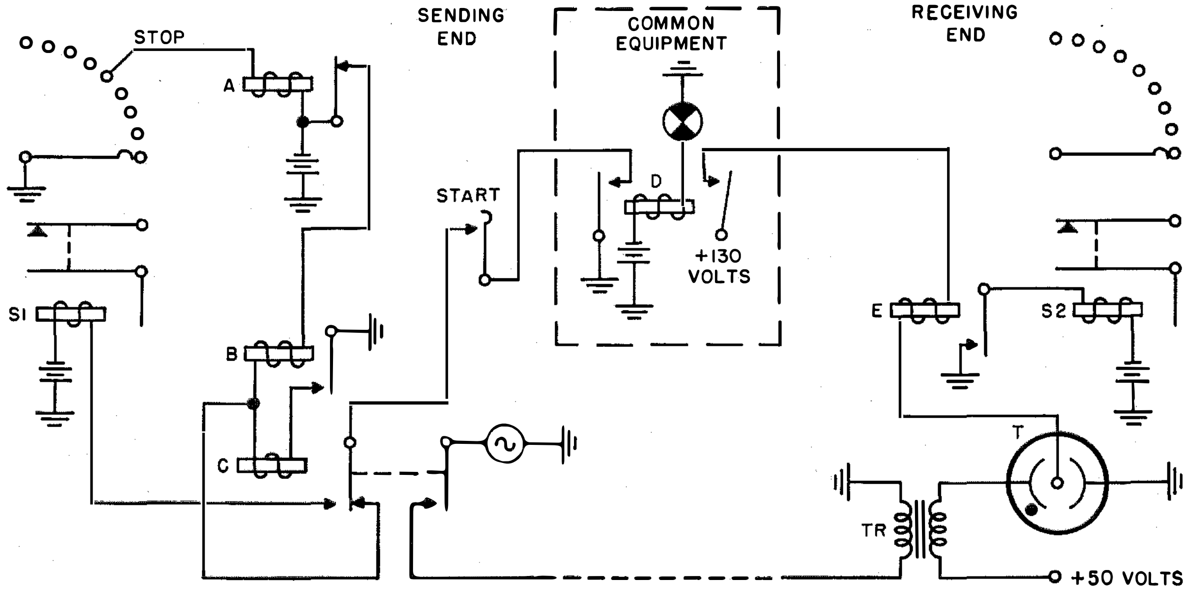
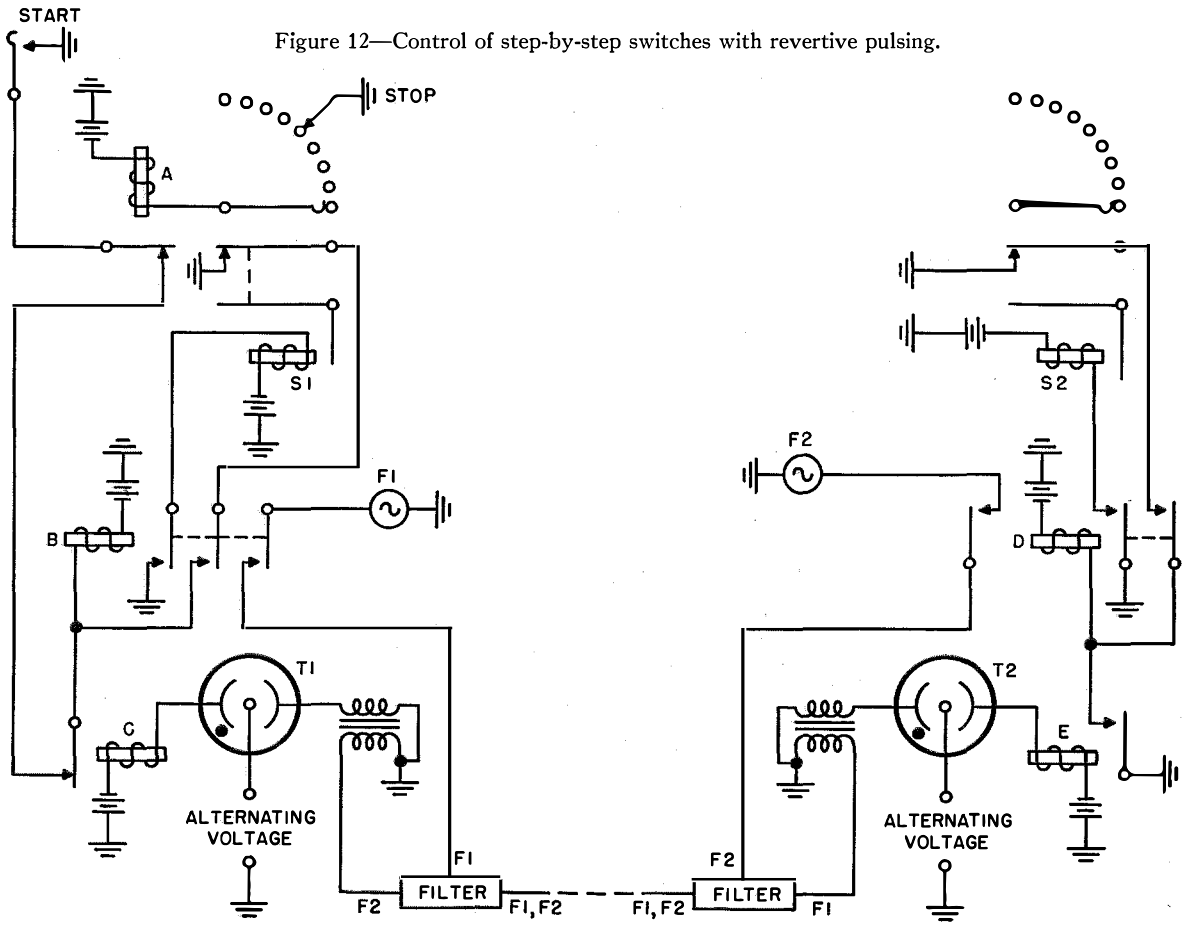


Figure 11—Control of step-by-step switches.

Figure 12—Control of step-by-step switches with revertive pulsing.



that tube T stops conducting and opens the circuit of $S1$. Both switches $S1$ and $S2$ come to rest in identical positions.

In the impulse-transmission schemes shown in Figures 8 through 11, if for any reason the receiving circuit fails to operate correctly, the send-

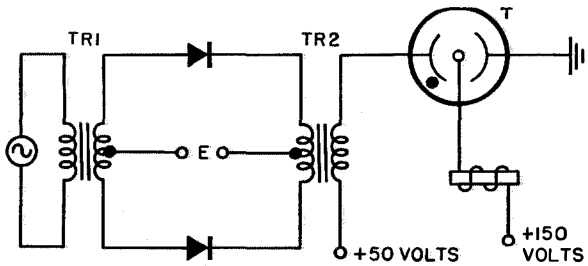


Figure 13—Arrangement equivalent to a highly sensitive polarized relay.

ing circuit goes on transmitting impulses, quite unaware of the trouble. In applications where this performance would be unacceptable, the arrangement shown in Figure 12 might be used. In this scheme, both sending and receiving circuits are under the permanent control of each other.

When the start contact is closed, relay B operates and connects an alternating current of frequency $F1$ to the line. Tube $T2$ thereupon ionizes, relays E and D being energized. An alternating current of frequency $F2$ is now sent back over the line. Tube $T1$ ionizes and relay C operates, causing the release of relay B and of the step-by-step switch $S1$, which makes one step. The alternating current $F1$ is removed from the line, tube $T2$ deionizes, relays D and E are deenergized as well as switch $S2$, which in turn makes one step. The alternating current $F2$ is at the same time disconnected from the line, causing the deionization of tube $T1$. Relay C releases, relay B and switch $S1$ are energized again, $F1$ is connected to the line, and the same sequence of operations is repeated. Both step-by-step switches advance alternately until switch $S1$ reaches the "stop" contact, whereupon relay A operates and opens the starting ground potential.

The arrangement shown in Figure 13 may be considered to be a substitute for a highly sensitive quick-acting polarized relay. If the terminals E are left open, an alternating voltage (450 cycles per second, for instance) applied to transformer $TR1$ would be unable to produce a significant current in the primary of $TR2$ because one or the other of the two rectifiers would offer a high reverse resistance to such a current. If a negative reference voltage were applied to the E terminal connected to $TR1$, the same condition would prevail. If, however, the polarity of this reference voltage were to be reversed, the two rectifiers would be operated in their forward low-resistance condition and would offer a low-resistance path to the alternating current. The alternating voltage produced across the secondary of $TR2$ from $TR1$ would ionize the gas tube T and operate the relay. In a practical setup, a reference current of only 100 microamperes is sufficient to insure the ionization of the tube.

Figure 14 illustrates an application of this scheme to stop a hunting selector on a free outlet that is characterized by a ground potential via a resistance $R1$ and by applying a negative potential to it make this outlet immediately inaccessible to other selectors hunting over the same bank of outlets.

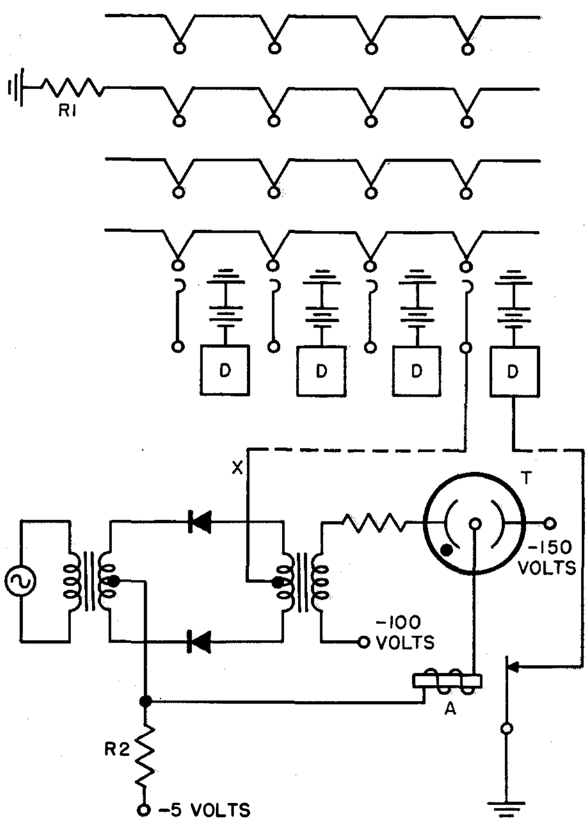


Figure 14—Method of selecting an outlet and busying it to other selectors.

As soon as the selector reaches the free outlet, a direct current flows through the rectifiers from the −5-volt lead to ground through resistances *R1* and *R2* in series. Tube *T* then ionizes, operating relay *A* and opening the circuit to the driving magnet *D* of the selector, which stops. The anode current flowing through resistances *R1* and *R2* brings the potential at point *X* to some −10 volts effectively busying that outlet for any other hunting selector. This busy condition stems from the fact that the potential prevailing at the outlet is now more negative than the −5-volt point, thus preventing operation of the rectifiers associated with the other selectors that contact the outlet.

Another application of this system is illustrated in Figure 15, which represents a relayless subscriber's line circuit combined with a static call-detector circuit common to a group of 50 subscribers. So long as none of the lines is in the calling condition, the polarity of the potential difference prevailing between points *X* and *Y* is such that the rectifiers are blocked. When a subscriber lifts his receiver, a current starts flowing in the voltage divider *R1*, *R2*, *R3* via the calling subscriber's loop. As a consequence, the potential of point *Y* is raised in the positive direction so that the polarity between *X* and *Y* is reversed and the rectifiers become conducting. Tube *T* ionizes, relay *A* operates, and the line finders start their hunt for the calling line.

Relay *A* thus acts as a common line relay for 50 subscribers. It will be noted that the anode of the tube is supplied with alternating current to allow the tube to deionize and release relay *A* as soon as the calling line has been found by the line finder. The application of −48 volts to the *C* wire from the cord circuit prevents the recti-

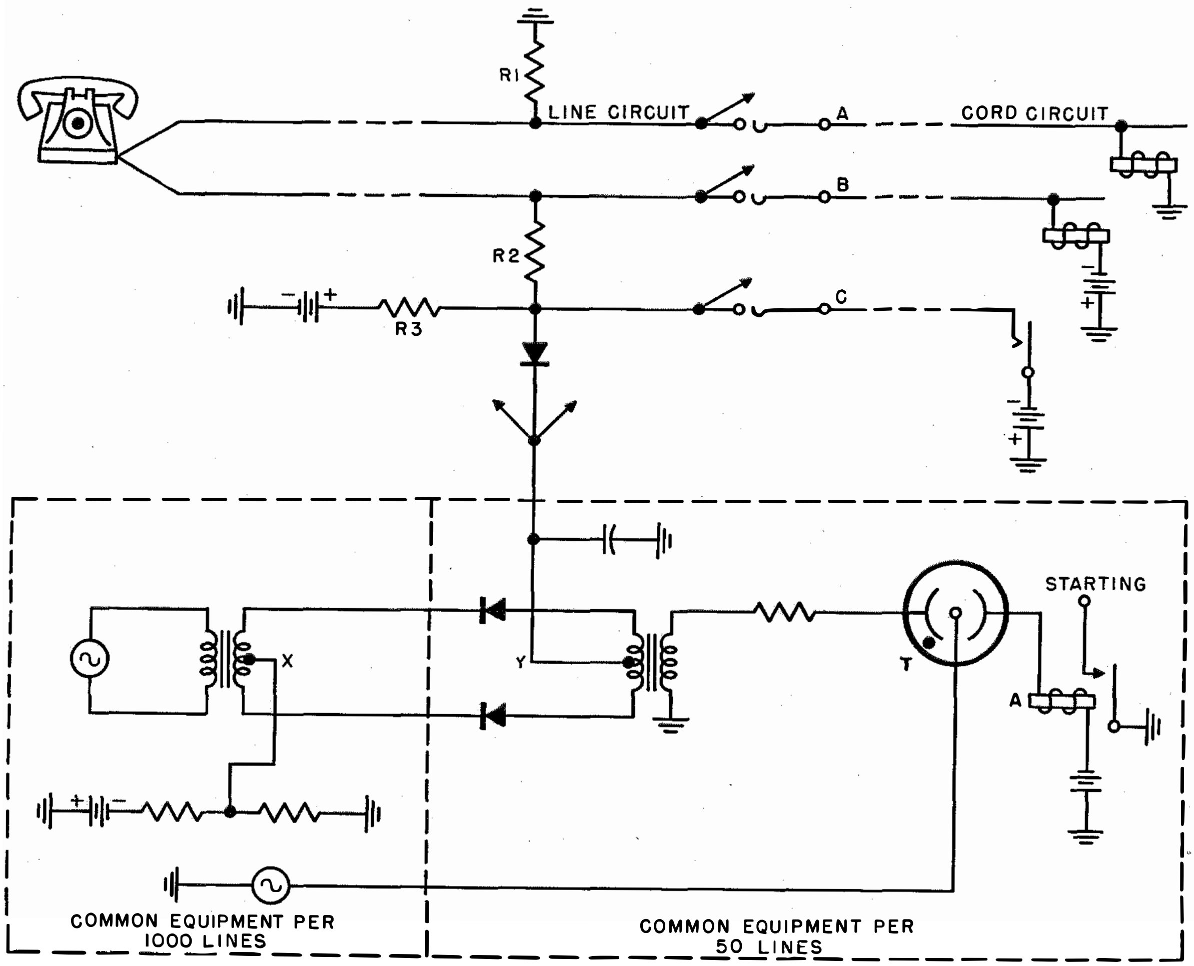


Figure 15—Relayless subscriber's line with common call-detector circuit.

fiers from working and removes the ionizing potential from the gas tube. This is a typical case in which the use of a cold-cathode tube is advantageous, because the circuit should stand ready to operate on any call during 24 hours of the day, whereas the total time during which it has to respond to calls is of the order of only a few minutes per day.

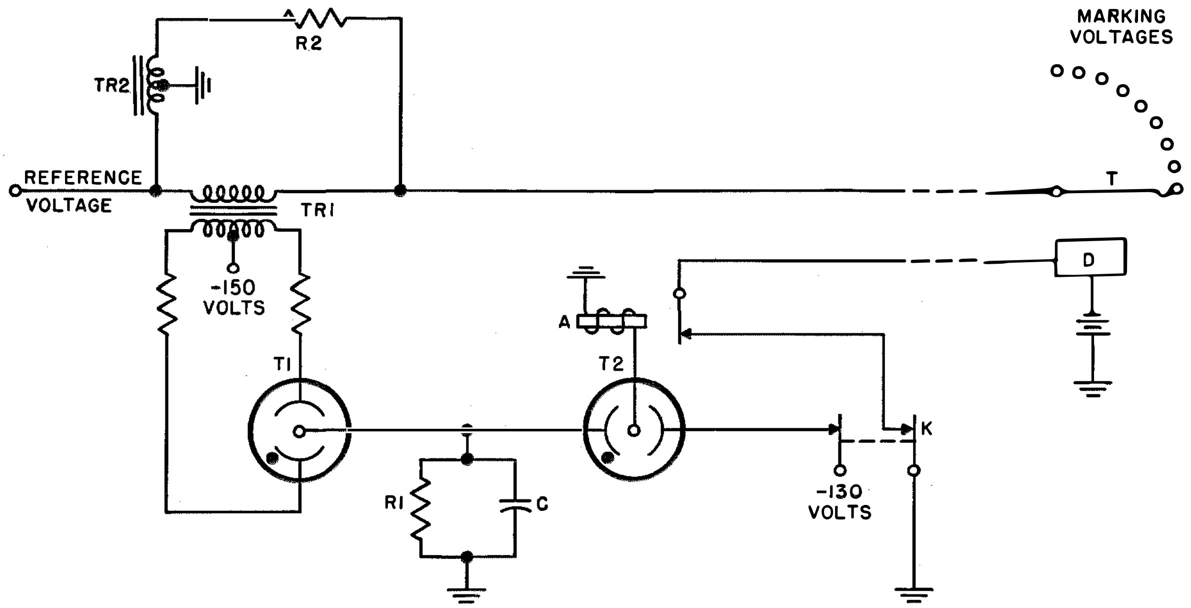


Figure 16—Phase discriminator used in *7E* rotary system.

The phase discriminator shown in Figure 16 is used in the *7E* rotary system, which is based on the use of 100-point single-motion selectors with multiphase marking. Each outlet in the selector arc is characterized by 1 of 12 alternating voltages, identical in frequency (450 cycles) and amplitude, but differing in phase angle by units of 30 degrees. The marking alternating voltages encountered by the test brush *T* of a hunting selector are compared in the register circuit with a reference alternating voltage characterizing the wanted group of outlets. This comparison is performed by a "phase discriminator," which will stop the hunting selector as soon as the test brush reaches a terminal marked with an alternating voltage in every respect identical (voltage, phase-angle, and frequency) to the reference voltage.

The operation is as follows. One end of the primary winding of transformer *TR1* is connected at the register to the reference alternating voltage while the other end is connected to the test brush of the hunting selector. So long as the compared alternating voltages are not identical, the difference voltage that appears across transformer *TR1* keeps the control gap of tube *T1* ionized. The tube acts simultaneously as a potential limiter and as a 2-phase rectifier, the limiting action taking place in the control gap and the rectifying action taking place in the main gap. The rectified voltage that appears on capacitor *C* is sufficient to keep the control anode of tube *T2* below its breakdown voltage. As soon, however, as the selector reaches the wanted outlet, the compared alternating voltages become identical, the potential difference across transformer *TR1* drops to zero, tube *T1* deionizes; capacitor *C* discharges to ground through resistance *R1*, tube *T2* ionizes, and relay *A* is energized and opens the circuit of the driving magnet *D* of the selector, which stops on the selected outlet. Autotransformer *TR2* and resistance *R2* are provided to maintain a voltage across transformer *TR1* during the time the circuit via the test brush *T* is open while it passes from one terminal to the next or passes a terminal on which no potential is present. During this time, the test brush is at ground potential, thereby ensuring the continuous existence of a potential between the brush and a terminal on which it arrives, so that the high-resistance film of the contact may be broken down.

In the final selector, which must be able to select one particular outlet out of a hundred, the arc is divided in 10 groups of 10 terminals. The first terminal of each group is marked by 1 of 10 alternating voltages with a potential of 10 volts, while the other terminals of each group of 10 are each marked by 1 of 9 alternating voltages with a potential of 6 volts. (Of the 12 voltages available in each group, 10 and 9, respectively, are used only for the tens and units selections.) The wanted outlet is selected in two successive steps: first a 10-volt reference alternating voltage is connected to the comparator, according to the tens figure of the wanted subscriber. When the selector has been set on the first terminal of the group of 10 to which the wanted subscriber belongs, the discriminator is connected to a 6-volt reference alternating voltage corresponding to the units figure of the wanted number, contact K is momentarily opened to allow tube $T2$ to deionize, after which the units selection takes place.

By a suitable choice of the various components in the discriminator, the operating time, i.e., the time from the moment the test brush connects the wanted potential until the moment relay A opens its break contact, is approximately 3 milliseconds.

The specific applications of gas tubes to telephone switching outlined in this paper are based primarily on the performance of the tubes as high-speed sensitive relays. However, it must be noted that these are not the only properties possessed by the tubes that make them of great interest to the telephone engineer.

Glide-Slope Receiver

By R. C. DAVIS

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INSTRUMENT landing systems have been used for many years to guide aircraft to a position directly above the landing strip of an airport from which the pilot can make a safe landing despite limitations in visibility. The principles of operation require the setting up of radio-frequency radiated fields that at points of equal intensity define surfaces above the ground. In the case of the localizer, the surface is a vertical plane aligned directly along the center of the landing strip of the airport. For the glide slope, it is approximately horizontal but is tipped sufficiently to touch the landing strip at a point allowing adequate additional runway to bring the plane to a safe stop. The intersection of the two surfaces is the approach path the aircraft is to travel. Many papers have been written on these fundamentals so reference will be made here to only one[1] published recently in this journal.

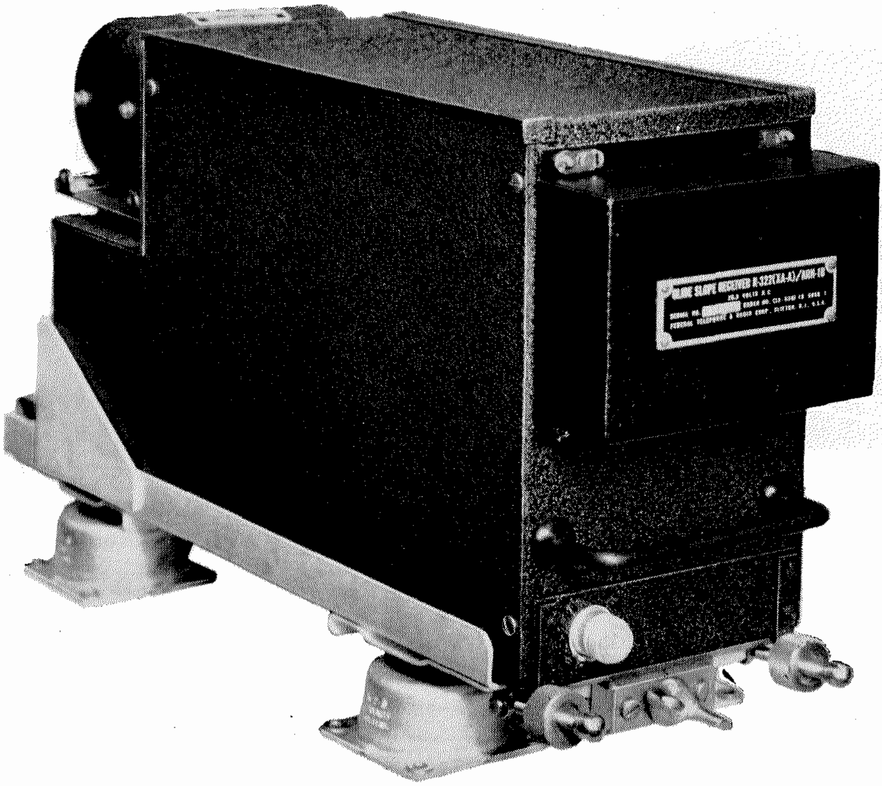


Figure 1—*154A* glide-slope receiver.

1. Requirements for New Receiver

The need for an improved glide-slope receiver to provide increased performance and for operation at several additional frequencies was widely recognized during the war. Consideration was given to the development of such a receiver at that time and during the immediate post-war period. However, the problem was of such a nature and the number of interested parties was so great that sound progress could result only by preparing a target specification. This essential step was initiated by the air lines of the United States through their mutually owned Aeronautical Radio, Incorporated. Representatives of the individual air lines, the United States Air Force, Bureau of Aeronautics, Civil Aeronautics Administration, and interested manufacturers were invited to prepare jointly a set of characteristics for a new receiver. The first draft was distributed in 1948 and a final version was issued in March of 1950.

In the preparation of this report, much consideration was given to systems for remotely controlling the receiver and to the relation

[1] R. A. Hampshire and B. V. Thompson, "*ILS-2* Instrument Landing Equipment," *Electrical Communication*, v. 27, pp. 112–122; June, 1950.

between the apparent width of a course and the intensity of the field producing it.

It was finally agreed that the remote control of the receiver would be done through the grounding of any one of 10 wires that would select a pair of frequency-determining crystals, the single desired crystal of that pair being chosen by the grounding of one of two additional wires.

The course-width problem centered chiefly about the question of whether course softening should be employed. Course softening refers to a condition whereby the angular width of the course as indicated by the instrument on the aircraft becomes greater as the touchdown point is approached. Without softening, a constant angular deflection from a course provides the same indicator reading regardless of the distance from the touchdown point. It was agreed that the receiver should not be concerned with course softening but that this property should be controlled by the transmitting equipment. Table 1 indicates the difference in performance requirements as reflected in the characteristics of the currently used receivers, type *R-89B/ARN-5*, and a new receiver *154A* described in this paper.

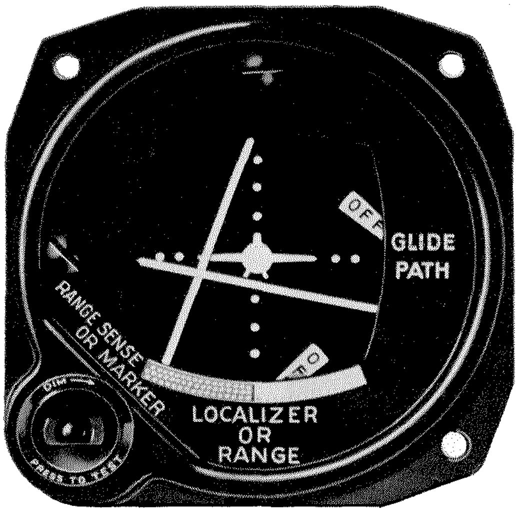


Figure 2—Flight-deviation indicator with OFF flags visible to warn pilot that the system is not operating. The flags are automatically retracted from view when localizer and glide-slope signals are being received.

A photograph of the *154A* receiver is shown in Figure 1, and the flight-deviation indicator used with it may be seen in Figure 2. Right- and left-side views of the receiver with the side plates removed may be seen in Figures 3 and 4.

2. *Circuit Features*

Figure 5 is a block diagram of the receiver. The selection of the operating channel being controlled entirely by the choice of a crystal to produce the required heterodyne frequency, the remaining circuits are all of the fixed-tuned type.

TABLE 1

PERFORMANCE REQUIREMENTS OF GLIDE-SLOPE RECEIVERS

Characteristic	*R89B/ARN-5*	*154A*
Number of Channels	3	20
Frequency Range in Megacycles	332.6–335.0	329.3–335.0
Size* in Inches (Centimeters)		
Width	$5\frac{3}{16}$ (13)	$4\frac{7}{8}$ (12)
Length	15 (38)	$15\frac{1}{2}$ (39)
Height	$6\frac{1}{2}$ (17)	$7\frac{5}{8}$ (19)
Weight in Pounds (Kilograms)	12.5 (5.7)	15 (6.8)
Plate Voltage in Volts	26.5	>150
Remote-Control Circuit in Number of Wires Plus Ground	3	11
Sensitivity in Microvolts	≈ 100	20
Selectivity. Over-all Attenuation at 300 Kilocycles Off Resonance in Decibels	28	90
Harmonic Distortion in Percent for Inputs from 200 to 10,000 Microvolts	High	$\leqslant 5$
Antenna Input		
Type	Balanced	Coaxial
Impedance in Ohms	95	52

* Includes front and rear space for cables and plugs.

2.1 Radio-Frequency Amplifier

The radio-frequency amplifier was designed to cover the entire band over which the receiver must operate, from 329.3 to 335 megacycles per second. The attenuation outside this pass band was made as great as possible by suitable compromises between sensitivity and tube loading for the four radio-frequency stages. Figure 6 shows the selectivity of this amplifier and the ranges over which spurious responses may occur as a result of undesired outputs from the oscillator and frequency-multiplier stages.

The 6th harmonic of a crystal oscillator operating near 50 megacycles is used to produce the intermediate-frequency signal. It falls below the frequency of the received channel as is indicated in Figure 6. No trouble from spurious responses will result from the 5th harmonic as it will be near 253 megacycles and so far removed from the pass band of the amplifier as to produce no evident effect.

The 7th harmonic falls above the channel frequency and for operation at 329.3 megacycles would produce a spurious response that would be only 45 decibels below full output based on the selectivity of the radio-frequency amplifier. This is indicated by point A in Figure 6. As a protection against spurious responses of at least 75 decibels was considered necessary, the frequency multipliers producing the heterodyning wave were designed to attenuate the 7th harmonic by at least 40 decibels with respect to the 6th harmonic.

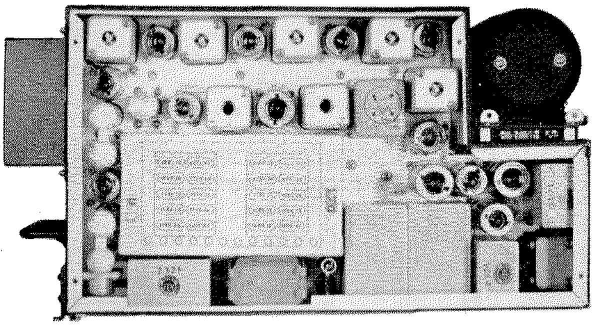


Figure 3—Right side of *154A* receiver.

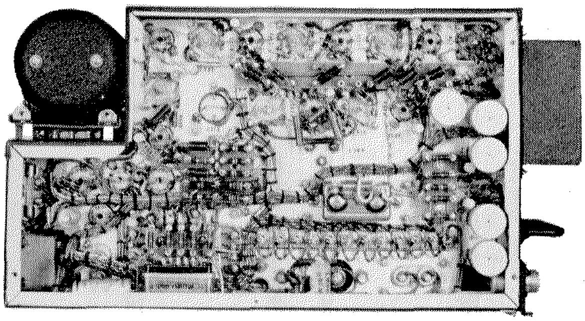
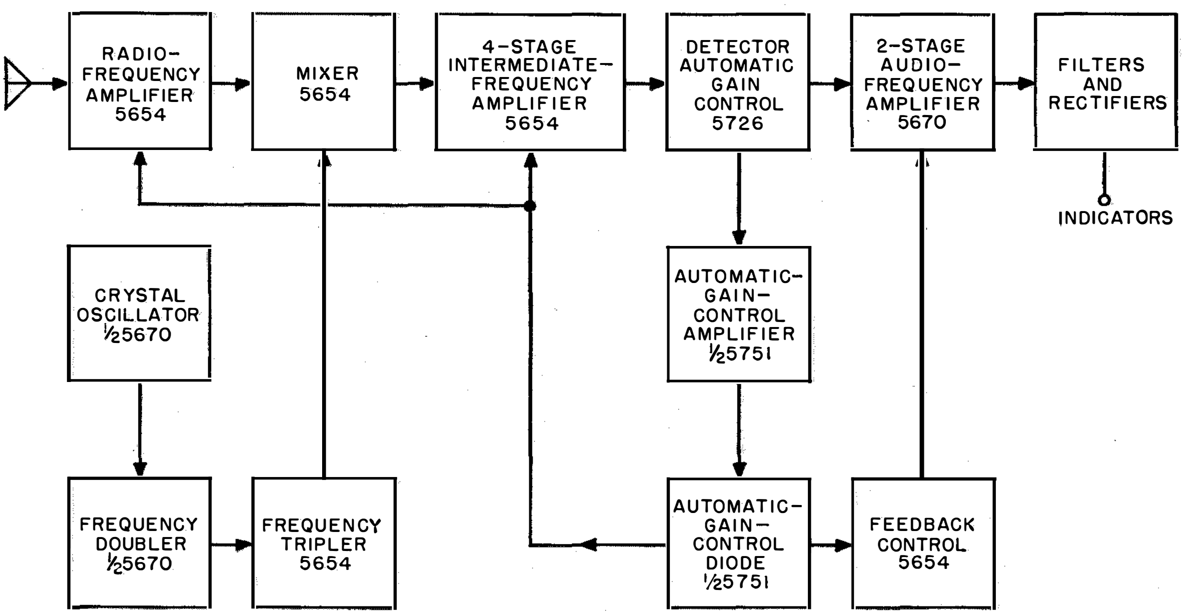


Figure 4—Left side of *154A* receiver.

Figure 5—Block diagram of receiver.



Coaxial transmission-line sections are employed for tuning the radio-frequency amplifier stages. The central conductor of each line is provided with an end capacitance adjustment for tuning. By using invar for the central conductor and brass for the outer tube, the end capacitance changes as a function of temperature and, when adjusted to have the proper negative coefficient, will compensate for the effects of temperature on the tuning of the circuits. This arrangement is shown in Figure 7.

C1 represents the input capacitance of the vacuum tube in series with a coupling capacitance. *C2* is a line-shortening capacitance that is not particularly affected by changes in temperature. *C3* provides the temperature-sensitive variation in capacitance, having a negative temperature coefficient because the distance from the end cap to the central conductor will increase with temperature as the brass shield expands more rapidly with temperature than the invar rod. The proper division of capacitance between *C2* and *C3* is made by adjusting the position of the cap along the threaded end of the brass tubing. The position is determined experimentally by measuring the shift in the radio-frequency selectivity of the amplifier for a known change in temperature and then calculating the dimensional changes required to correct for the shift.

2.2 Oscillator

In general, the higher the frequency of an oscillator that must be multiplied to a particular injection frequency, the less trouble will result from spurious responses. As the oscillator frequency is made higher, the unwanted harmonics of the mixer will be separated from the desired harmonic by a larger percentage of frequency, and fewer multiplications of frequency will be required. These factors reduce the requirements on selectivity of the radio-frequency amplifier,

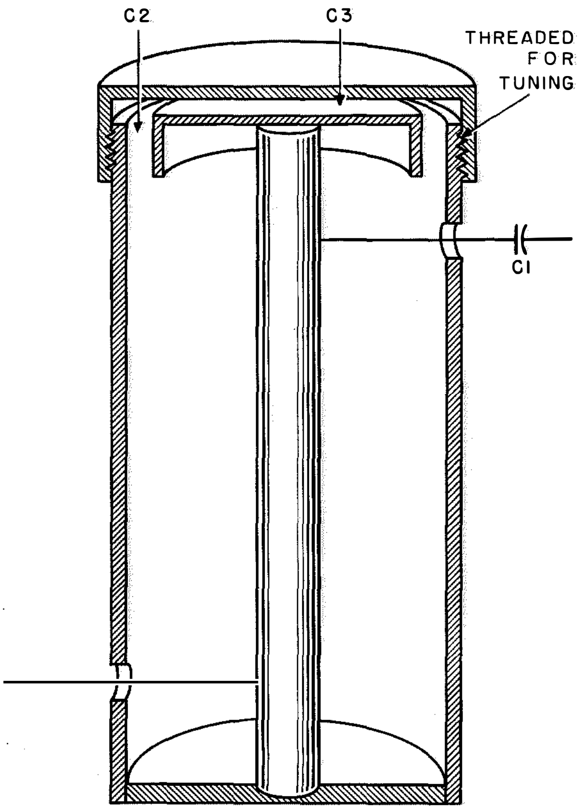


Figure 7—Coaxial-line tuning elements used in the radio-frequency amplifier. The end cap is threaded to permit adjustment. By using an invar rod for the central conductor and brass for the outer tube, a negative coefficient of capacitance to temperature is obtained to stabilize operation over a wide temperature range.

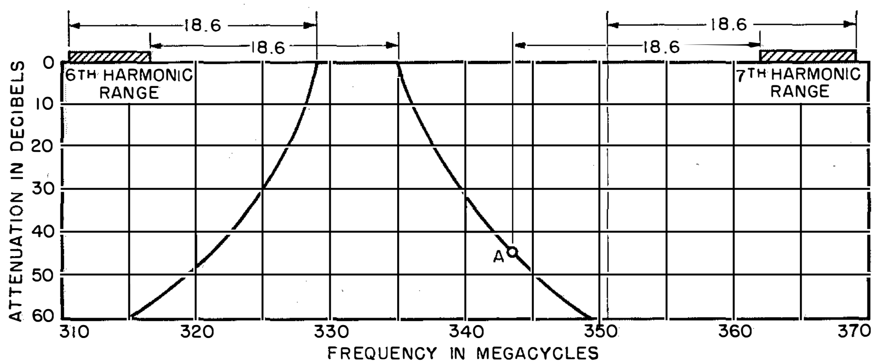


Figure 6—Selectivity characteristic of the fixed-tuned radio-frequency amplifier. The 6th harmonic of the local crystal oscillator produces the 18.6-megacycle intermediate frequency. The 7th harmonic range is shown and would produce spurious responses if it were not suitably attenuated.

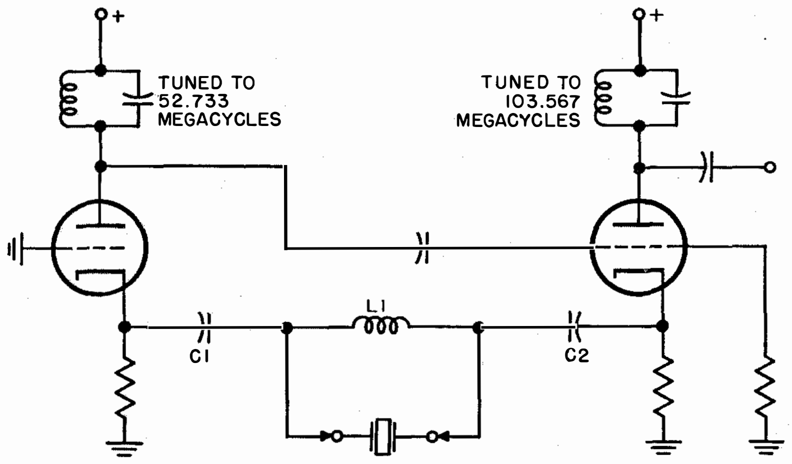


Figure 8—Crystal oscillator circuit. *L1*, *C1*, and *C2* reduce the effects of the inductance and capacitance of the leads through series tuning.

oscillator–frequency-multiplier selectivity, and shielding.

As the oscillator frequency is increased, however, less power is available for driving the multiplier stages and the problem of avoiding "free running," when the crystal is not in control, increases. As a compromise, it was decided to use an oscillator frequency in the vicinity of 50 megacycles. In the required range from 51.833 to 52.733 megacycles, reliable fifth-harmonic-mode crystals may be obtained.

Twenty crystals are provided and may be selected individually through the operation of a bank of 11 relays. The selected crystal is connected between the cathodes of a pair of triodes in a cathode-coupled-oscillator circuit as may be seen in Figure 8. The crystal operates at or near its series-resonant mode. It was found necessary to tune out approximately the series lead inductance of the wiring and to neutralize stray capacitance. The components required for this are *C1*, *C2*, and *L1* in Figure 8. The channel-switching operation does not require adjustment of these elements or of the oscillator-multiplier tuned circuits as the band over which the frequency may be varied is less than 2 percent of the frequency.

2.3 Intermediate-Frequency Amplifier

The 4-stage intermediate-frequency amplifier provides most of the gain before detection. Its selectivity characteristic is shown in Figure 9. As regeneration can be very damaging to stability in such an amplifier, the placement of components and the filtering of leads was given careful attention. The transformers employed ceramic insulation throughout to obtain a high degree of stability with variation in temperature. The center frequency of 18.6 megacycles holds to within ±9 kilocycles over the full range of ambient conditions prescribed in the specification.

2.4 Audio-Frequency Amplifier

The audio-frequency amplifier required compensation to maintain the course width over the specified temperature range. The increasing attenuation in the bandpass filters and rectifiers supplying signals to the indicator as the temperature

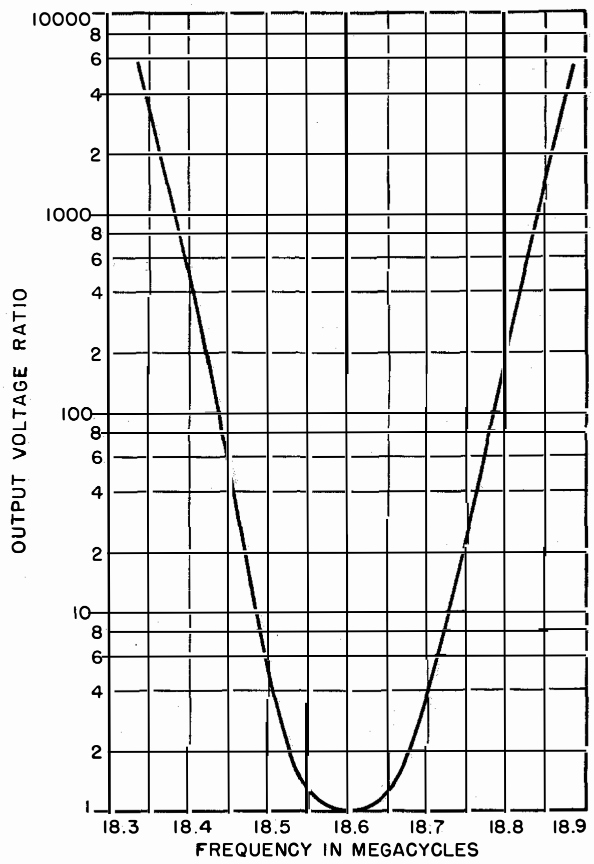


Figure 9—Selectivity curve for the four-stage intermediate-frequency amplifier.

increased is compensated for by the use of negative-temperature-coefficient resistors. Disk-type thermistors responsive to the variations in ambient temperature are used as shown in Figure 10 to accomplish this compensation.

TABLE 2

VACUUM TUBES USED IN *154A* RECEIVER

Aerinc Type	Commercial Equivalent
5654	*6AK5*
5626	*6AL5*
5751	*12AX7* (approximately)
5670	*2C51*

2.5 AUTOMATIC GAIN CONTROL

The receiver is provided with an automatic-gain-controlled amplifier such that it will provide "fail-safe" operation of the indicator. Any tube failure in the automatic-gain-control system will produce a flag-alarm current of less than 125 microamperes, which is insufficient to move the warning flag from its OFF position. This protection is provided over a signal input range from 200 to 100,000 microvolts. The automatic-gain-control characteristics are shown in Figure 11.

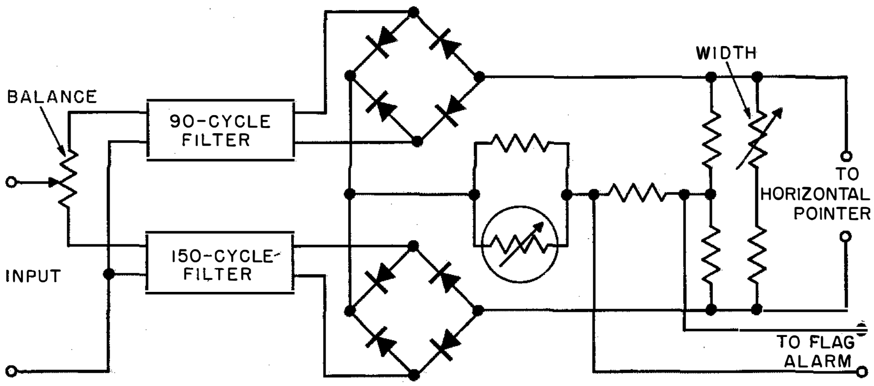


Figure 10—Indicator circuit. The negative-temperature-coefficient resistors compensate for increases in attenuation with temperature in the filters and rectifiers.

The very wide range of input voltage over which the receiver must operate and maintain substantially constant output placed great emphasis on the design of the automatic-gain-control system. The glide-slope pointer must not indicate any increase in course sharpness as the touchdown point and, consequently, transmitter are approached. The receiver output must be maintained within +0, −1 decibel from the time the signal would first be used to the touchdown point, which corresponds to a range of open-circuit voltage from 200 to 100,000 microvolts.

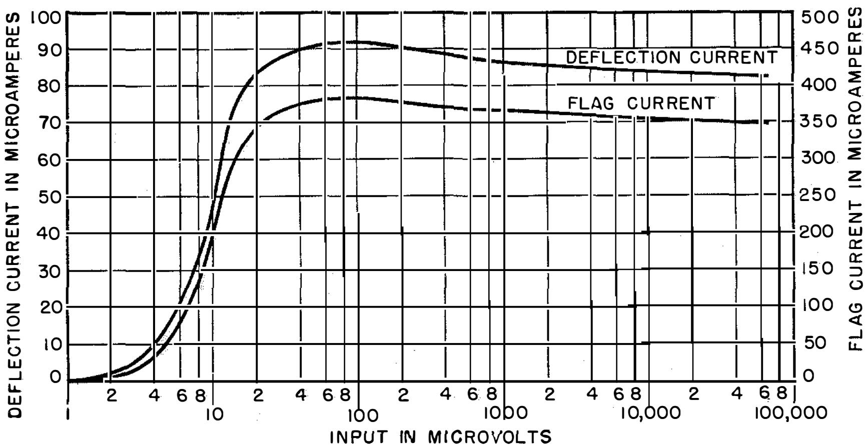


Figure 11—Automatic-gain-control characteristic. The pointer-deflecting current and flag-operating current are plotted against receiver input applied through a resistance of 50 ohms and expressed in microvolts for a 2-decibel ratio of 90- and 150-cycle side frequencies at 90-percent modulation.

In addition to the usual action of the automatic gain control on the radio- and intermediate-frequency amplifiers, an inverse-feedback circuit is used in the audio-frequency amplifier. As will be seen in Figure 12, *V1* is a resistance-coupled audio-frequency amplifier supplying signals to a shunt-fed output stage *V2*. Between the plate of *V2* and the cathode of *V1*, an inverse-feedback circuit is connected. This circuit forms an attenuator that transfers a portion of the output signal to the input. The series arms of the attenuator are *R1* and *R2*. The plate resistance of *V3* forms the shunt arm. Since the grid of *V3* is connected to the automatic-gain-control circuit, the shunt resistance will vary with the input signal level. The voltage on the grid of *V3* becomes more negative as the input signal increases, thus increasing the shunt resistance for higher signal levels. This increases the amount of inverse feedback and reduces the gain of the audio-frequency amplifier. The values chosen for this circuit are so related to the rest of the system as to provide a slight reduction of output as the input signal varies from 200 microvolts to the maximum value.

2.6 Tubes

With the exception of the *OA2* voltage-regulator tube, the 12 tubes used in the receiver are of the reliable types developed for air-line use under the auspices of Aeronautical Radio, Incorporated (Aerinc). The types used and their commercial equivalents are shown in Table 2.

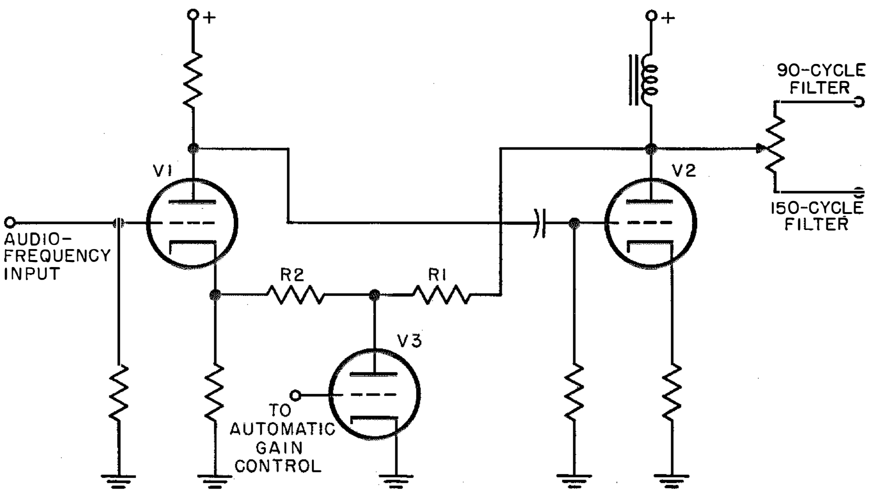


Figure 12—Inverse-feedback circuit in the audio-frequency amplifier. The automatic-gain control voltage is applied to the control grid of *V3* increasing the inverse feedback and reducing the gain of the amplifier as the input signal increases.

3. *Acknowledgments*

The assistance provided by E. B. Moore and B. L. Smith and the encouragement of E. W. Butler and D. P. Wilson during the development of the receiver are deeply appreciated.

Dynamic Aspects of Errors in Radio Navigational Systems Particularly in Cases of Fast-Moving Receivers and Transmitters*

By HENRI BUSIGNIES

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In free space, the relative movement of a radio transmitter and receiver makes the latter observe a different frequency from that transmitted. According to the simple law of the Doppler effect,

$$F_{rec} = F_0(1 + V/c), \quad (1)$$

where

F_{rec} = received frequency
F_0 = transmitted frequency
c = speed of light
V = relative speed of approach or recession.

If transmission is not in free space, reflections will occur at all boundaries. However, to make this paper of practical use, consideration will be given only to reflections from the surface of the actual earth.

The spectrum of received frequencies resulting from the relative movement of the transmitter and receiver, complicated by reradiation from reflecting surfaces, can be calculated by using (1). If both the transmitter and receiver are moving, the relative speed of each reflector with respect to them must also be considered. As a function of time, the received frequency spectrum changes in accordance with the variation in the shape of the triangular path connecting the transmitter, receiver, and reflectors.

1. *Values of Doppler Frequencies*

The highest Doppler frequency will result from the greatest relative speed among the elements in the system: the transmitter, receiver, and reflector. The lowest frequency will approach direct current.

At a transmission frequency of 10,000 megacycles per second, the Doppler frequency spectrum may extend from zero to ±6600 cycles from the frequency of the originally transmitted wave for speeds of 720 kilometers (447 miles) per hour, and at 100 megacycles this spectrum may extend from zero to ±66 cycles.

The phase of each Doppler component depends on the number of wavelengths in each path and its amplitude on the efficiency of the reflector. The reflections would normally be weak compared with the direct wave but shielding of the direct wave or some equivalent directivity effect may result in the reflected wave being stronger than the direct wave. Reflections are, therefore, a major factor in the design of many radio systems, particularly those dealing with directive and timing features for navigational purposes.

2. *Integration*

In 1937 and 1938, a study was made of the dynamic aspect of errors due to mountains[1] and the ionosphere[2] introduced into a medium-frequency radio compass carried by an airplane. It was shown that these errors could be reduced by integrating the variations and by wide-band transmissions. Reflection effects were studied soon after in direction finders and omnirange systems operating in the 100-to-150-megacycle region.

In directive systems such as direction finders, instrument landing, and omnidirectional ranges, the Doppler frequencies caused the indicated bearing to oscillate around the true value. Where

* Reprinted from *Navigation*, v. 3, pp. 60–62; September–December, 1951. Summary of a paper presented at the national convention of the Institute of Radio Engineers in New York, New York, on March 9, 1950; and at the national convention of the Institute of Navigation in San Diego, California, on June 30, 1950.

[1] H. Busignies, "Mountain Effects and the Use of Radio Compasses and Radio Beacons for Piloting Aircraft," *Electrical Communication*, v. 19, n. 3, pp. 44–70; 1941.

[2] H. Busignies, "Evaluation of Night Errors in Aircraft Direction Finding, 150–1500 Kilocycles," *Electrical Communication*, v. 23, pp. 42–62; March, 1946.

there was but one main reflection, the variations were quite regular; they may seem to be sinusoidal but are generally not. The sinusoidal oscillation is a particular case. While averaging reduces the error by a substantial ratio, it will not eliminate it except in special cases.

As a function of the number of reflections, the received spectrum looks more and more like random noise. The lowest-frequency components often are most disturbing as they persist too long for satisfactory integration. Some beats can occasionally equal the scanning rate of an automatic direction finder or omnidirectional range (30 cycles) or the modulation characteristics (90 and 150 cycles) of the instrument landing system resulting in large deviations for short periods of time. Variations of bearings up to 10 degrees have often been recorded, as have deviations of 1 to 2 degrees in instrument landing courses. Increased accuracy through averaging carries the penalty of slower response.

An illustration of trading speed for accuracy is the case of a continuous-wave split localizer in which ground reflections made accurate readings with portable ground equipment next to impossible but with which a straight course could be flown over the same area when flying at a suitable integrating speed.

These improvements are obvious but it must be remembered that a delay in receiving navigational data is a form of error that is directly proportional to speed. It cannot always be accepted but, fortunately, a delay of a few seconds provides a substantial amount of integration. Delay errors must be analyzed carefully, particularly in the case of automatic flying.

For instrument landing, considerable integration may be used for during a proper approach, the speed of the aircraft across the course is quite small with respect to that along the course. In the case of distance measurement, a delay of a few seconds it not acceptable unless proper accounting is made for it.

3. Polarization and Directivity

In directive systems, both directivity and polarization affect the results strongly. Directivity is generally used at only one terminal of the link. The usual figure-of-eight or cardioid patterns will restrict the energy to half the area of a circular pattern, whereas a sharply directive system may distribute it over, say, 10 degrees or $\frac{1}{36}$th of a circular pattern.

The figure-of-eight and cardioid patterns are practically the worst systems with regard to reflections but are widely used because of their convenience. With antennas giving sharp multilobed patterns, the reflecting surfaces are still scanned or illuminated but at different times and in most cases this permits satisfactory discrimination against reflections. The Doppler frequencies resulting are low frequencies and will cause more modulation or fading of the received signal than directivity error. Improvement may be considered to be proportional to the increase in sharpness of pattern.

If the transmitting and receiving antennas deal with pure polarized waves, directional errors will be diminished as changes in polarization occurring at a reflecting surface will reduce the effect of the reflected wave on the receiving antenna. Thus, a lack of polarization purity accentuates the effect of reflections and reduces directivity information.

Many flight recordings have been made on instrument landing localizers at 110, 320, and 1000 megacycles and around omnidirectional ranges at 117 megacycles. Very useful data can be derived from these recordings on the best design of radio nagivational aids of the directive type for use under dynamic conditions.

4. Time-Based Systems

Navigational systems such as loran, gee, shoran, and distance-measuring equipment depend on timing for their performance and are independent of directivity factors. While all Doppler frequencies due to reflections are present, they produce beats only when pulses due to echoes overlap. When they do not overlap, they can be observed separately and as the echoes trail behind the direct pulse, the leading edge of the direct pulse is free from these dynamic errors.

5. Bandwidth Versus Integrating Time

The above suggests that accuracy may be effectively improved by trading increased bandwidth for a reduction in time.

Errors due to reflections can be averaged out by radiating a frequency spectrum such that the phases of the various reflected components are received at random. Integration can then be applied over the frequency spectrum and will not require motion to be effective. Andrew Alford proposed and Sidney Pickles developed a localizer in which the carrier consisted of a large number of frequencies generated by beats among quartz crystals. A sinusoidal distribution of frequencies would be best but this condition was only approached in the equipment built; the spectrum was approximately 4 megacycles wide. Many flight recordings made at a poor site, which was deliberately chosen, showed course bends as large as 13 degrees with a standard instrument landing localizer and were reduced to 1 degree with the frequency-spectrum system.

Spectrum radiation would be particularly effective in the medium-frequency band for marine beacons. Frequencies around 300 kilocycles are subject to the so-called night effect due to reflections from the E layer of the ionosphere, which limits the range of loop-type shipboard direction finders to about 50 miles (80 kilometers) at night.

In the night effect, the error goes through a complete cycle of variation for a 360-degree phase shift of the reflected wave with respect to the direct wave. If a sufficient number of different phases are present over a complete cycle of reflection errors, an integrating direction finder (automatic direction finder or equivalent) will show considerable improvement. An improvement in accuracy of 5 to 1 may be expected using a relatively narrow band of frequencies. A loop direction finder of the null type will also demonstrate the same improvement but will lose sharpness of null. Such a system, however, would permit the continued use of the present marine direction finders and also encourage the procurement of better automatic direction finders, both of which could be used with the present beacons without modification.

P. Aigrain has calculated that using a 3.3-kilocycle band in the medium-frequency spectrum should reduce the error by a factor of 5 and a 5.5-kilocycle band by 20 to 30. Experimentation has not yet been started.

6. *Acknowledgment*

Messrs. P. Adams, L. DeRosa, S. Pickles, and P. Aigrain have contributed experimental data, recordings, and calculations used in the preparation of this paper.

Some Recent Developments in Traveling-Wave Tubes for Communication Purposes*

By J. H. BRYANT, T. J. MARCHESE, and H. W. COLE

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BASIC features of traveling-wave tubes are reviewed together with their typical performance characteristics and their applications. Specific data on a commercial type of glass-envelope tube is presented to illustrate some of the characteristics. Factors leading to a new mechanical design are outlined, and a tube is described that embodies features of ruggedness and ease of handling. One object of the development was to use the lowest voltage consistent with obtaining the required output power. The low voltage leads to a structure that is physically short, resulting in minimum size and weight of magnet.

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It is the purpose of this paper to describe a traveling-wave tube of new mechanical design that is far more rugged than those previously developed. First, however, the basic features of traveling-wave tubes will be reviewed. Typical performance characteristics will be discussed and consideration will be given to applications for traveling-wave tubes. The characteristics of a glass-envelope tube of conventional design will then be reviewed, and finally some factors leading to a new mechanical design will be outlined.

1. *Important Characteristics of Traveling-Wave Tubes*

Traveling-wave tubes are broadband amplifiers for microwave frequencies. These tubes have the unique characteristic that the gain is constant and independent of frequency over a very wide bandwidth. A sketch showing the principal parts of a traveling-wave tube is given in Figure 1.

Here, a helical radio-frequency transmission line serves to carry the radio-frequency wave to be amplified. The electromagnetic wave thus propagated travels along the wire at approximately the speed of light. This means that the wave front progresses down the axis of the tube at a rate that is approximately proportional to the ratio of pitch of the helix to the length of a single turn. For example, the velocity of 1200-volt electrons corresponds to one-fourteenth of the velocity of light. As shown, an electron beam

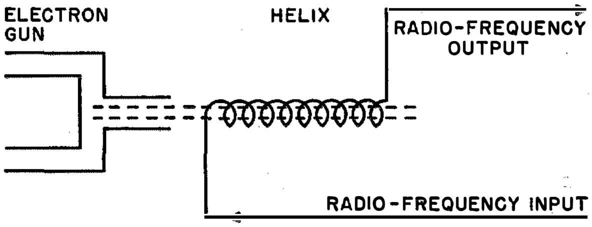


Figure 1—Elementary traveling-wave tube.

is made to traverse the region inside the helix. The moving electron beam interacts with the axial component of the electromagnetic field in such a manner as to produce three forward waves. One of these waves, the one of interest, has a negative attenuation component, which is to say that it is a gaining wave. It extracts energy from the direct-current beam and grows in magnitude so that an amplified signal is obtained at the output. One reason that such a tube has extremely large bandwidth is that the velocity of propagation of the electromagnetic wave is constant over a very large frequency range. Also, as the frequency is changed, the degree of interaction between the beam and the wave is such as to make the total gain practically constant. That is, as frequency is decreased and therefore the number of total wavelengths on the helix is decreased, the coupling between the field and the beam increases, giving a greater gain per wavelength in such a way as to make the gain tend to stay constant.

A typical glass-envelope traveling-wave tube is illustrated in Figure 2. Typical characteristics of gain and output power as functions of input power are shown in Figure 3. As can be seen, the

* Presented at the National Electronics Conference in Chicago, Illinois on October 23, 1951. Reprinted from *Proceedings of the National Electronics Conference—1951*, v. 7, pp. 299–303.

output power is a fairly linear function of input power to well above the rated power of 10 watts. Note the break in the power output curve, which is referred to as the point of initial limiting, and then the subsequent leveling off, which is effectively power limiting. This characteristic of power limiting can be used to obtain certain desired system characteristics.

Also of interest, when the tubes are to be used in a communication system, is the signal-to-noise performance. A typical plot of signal-to-noise ratio as a function of driving power is given in Figure 4. As shown here, the signal-to-noise ratio is of the order of 60 decibels and is reasonably constant as drive power is varied.

2. *Applications of Traveling-Wave Tubes*

2.1 Amplifier

As previously stated, traveling-wave tubes are broadband amplifiers for microwave frequencies and have the unique characteristic that gain is practically constant over an extremely wide frequency band. This means that a tube may be used to amplify several frequencies spaced hundreds of megacycles apart. Alternatively, the tube may be used at any number of frequencies at will without the necessity of employing tuning controls. Another unique characteristic of traveling-wave tubes is that they are insensitive to load, as contrasted with tubes requiring resonant circuits for their operation.

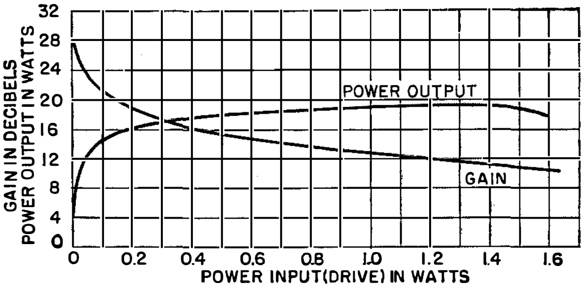
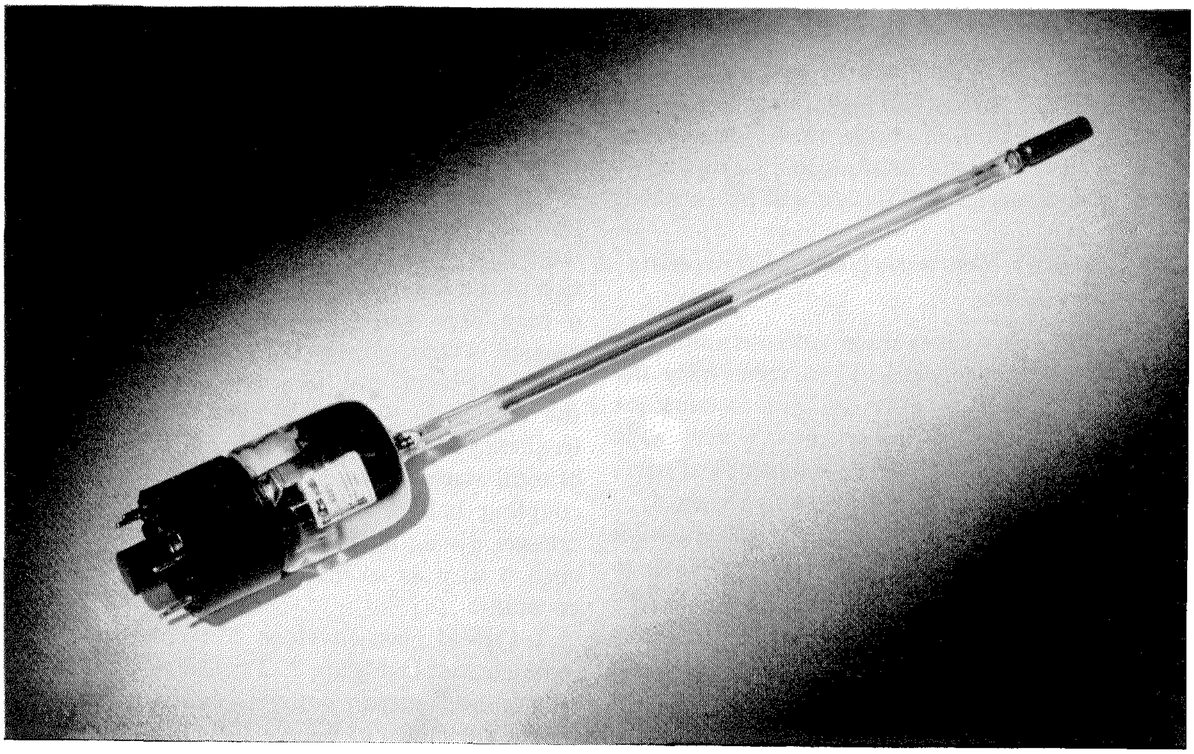


Figure 3—Gain and output power as functions of drive power.

Thus far only 10-watt, medium-power tubes have been considered. Traveling-wave tubes at other power levels in other frequency bands have also been developed. Their characteristics are similar to those presented previously. An important application of traveling-wave tubes is

Figure 2—Type *5929* traveling-wave tube.



that of low-noise signal amplification for the input stage of radio receiving equipment. The most important characteristic of these tubes is that they present a low noise figure. Recent developments have led to reproducible noise-figure values of 10 decibels or better.[1]

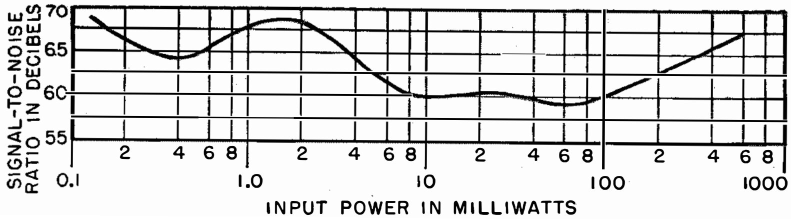


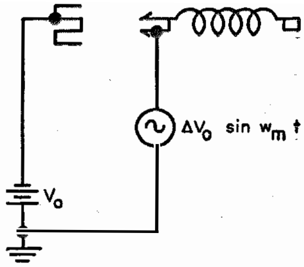
Figure 4—Signal-to-noise ratio as a function of drive power.

2.2 Phase Modulation of Traveling-Wave Tubes

Although traveling-wave tubes are characteristically amplifiers, they have other interesting applications, such as for high-level mixing. A typical tube contains about 40 electrical wavelengths in the wave-propagating structure at the midfrequency of the band. With this large number of electrical wavelengths, frequency modulation is relatively easy to achieve, as illustrated in Figure 5. Although this tube was not designed for frequency modulation, by application of audio or video voltages through electrodes in the tube one may realize phase modulations that result in frequency modulation of the output signal. As illustrated here, the modulating signal may be applied in series with the anode supply. Alternatively, it may be applied between cathode and focus electrode in the electron gun. A typical spectrum is shown in Figure 6. Here the modulating voltage is adjusted for maximum output power in the first sideband, which means that each of the first-order sidebands contains 30 percent of the total power of the unmodulated carrier. The modulating signal is a 20-megacycle sine wave so that the frequency of this first sideband is 20 megacycles higher than the carrier frequency. This property opens several possible applications. One application is to use a traveling-wave tube to modulate a carrier signal with any desired modulation such as voice or other intelligence. Another application is that of realizing a frequency shift such as might be desired in a straight-through repeater station in a cross-country radio-relay link.

$$\frac{\Delta\phi}{\phi} \approx \frac{1}{2}\frac{\Delta V_0}{V_0}$$

$$\Delta\phi = \frac{2^{\frac{1}{2}}\phi V_{\mathrm{mod}}}{V_0}$$



	Frequency
$\frac{e}{E} = J_0(\Delta\phi)\cos\omega_0 t$	ω_0
$+ J_1(\Delta\phi)\cos(\omega_0 + \omega_m)t$	$\omega_0 + \omega_s, \quad \omega_0 - \omega_s$
$+ \cdots$	

For the type *5929*,

$$V = 3000 \text{ volts}$$
$$N = 40\lambda$$
$$\phi = 80\pi,$$

$V_{\mathrm{mod}} = 15$ root-mean-square volts for maximum power in the first-order sideband.

Figure 5—Frequency modulation of a traveling-wave tube.

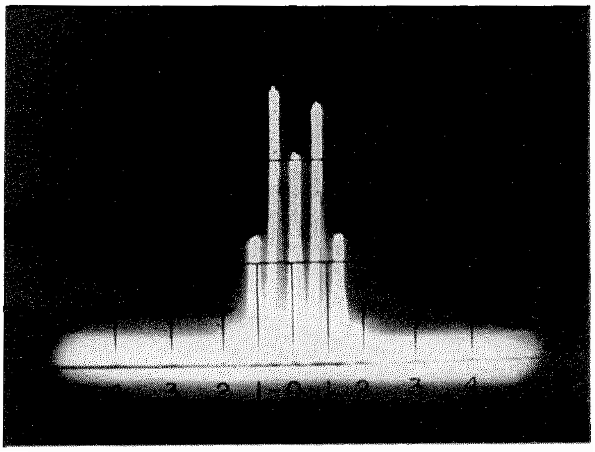


Figure 6—Output spectrum of frequency-modulated traveling-wave tube.

[1] A. G. Peifer, Philip Parzen, and J. H. Bryant, "Low-Noise Traveling-Wave Tube," *Electrical Communication*, v. 29, pp. 234–237; September, 1952: also *Proceedings of the National Electronics Conference—1951*, v. 7, pp. 314–317.

3. *Traveling-Wave Tube of Improved Mechanical Design*

No practical difficulties have so far been encountered in laboratory or field tests of the *5929*, which may be described as conventional in general concept. It is recognized, however, that from the point of view of strength an all-metal envelope would be preferable to a slender, elongated glass envelope. Consequently, a new tube has been developed which uses an all-metal envelope, as seen from Figure 7. The metal shell comprises the vacuum envelope and also provides complete shielding from stray radio-frequency fields. The operating frequency range is is 5900 to 7100 megacycles. One object of the development was to use the lowest voltage consistent with required output power. The rated output power is 10 watts with 25 decibels gain, and the tube operates at 1200 volts. This reduced voltage leads to a structure that is physically short, resulting in minimum size and weight of magnet. It will be noted that the input and output radio-frequency circuit connections are coaxial lines that are brought out at one end of the tube.

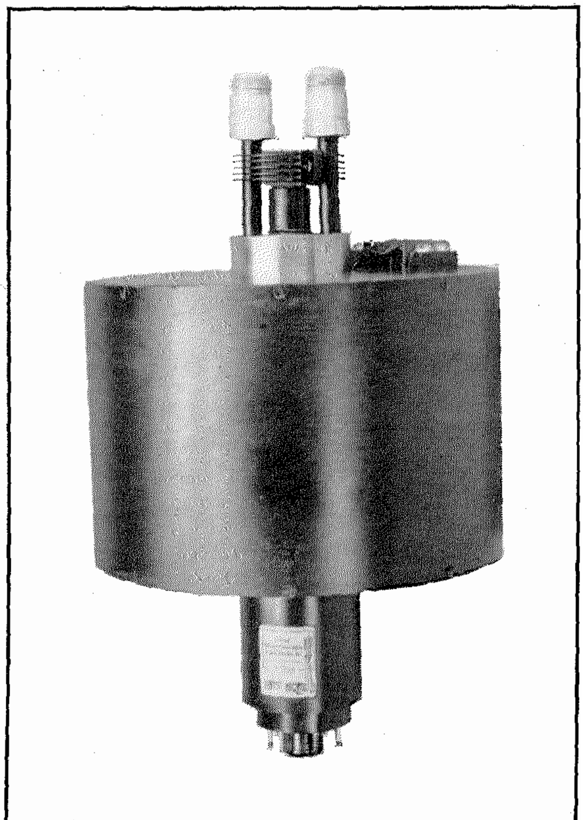
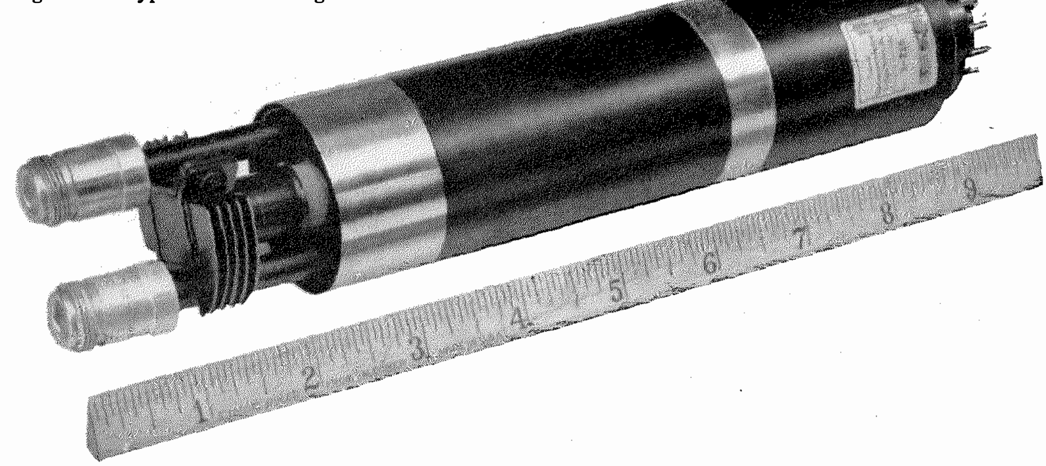


Figure 8—Type *X216* tube and electromagnet.

Figure 7—Type *X216* traveling-wave tube.



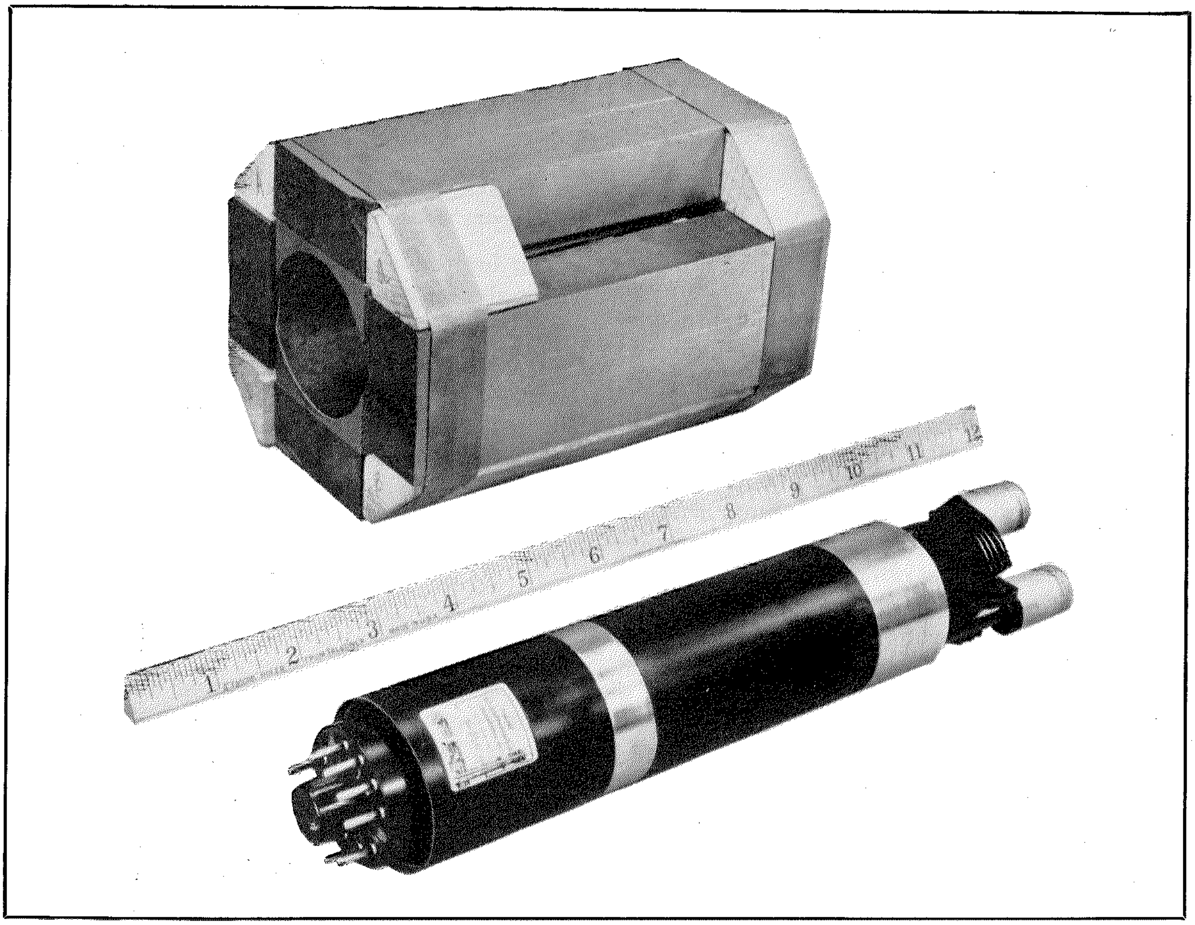


Figure 9—Experimental permanent-magnet arrangement.

For operation, a single solenoid four inches in total length is required, as shown in Figure 8. The use of coaxial radio-frequency lines has simplified the magnet design, allowing the use of a single continuous solenoid without the breaks in the field that are a consequence of using waveguide terminations. Because of the realization that permanent magnets should eventually replace electromagnets, a considerable amount of development has been carried out on the operation of these tubes in permanent magnets. An experimental permament-magnet arrangement is illustrated in Figure 9. Thus far, operation in permanent magnets has not been as satisfactory as operation in electromagnets, but it is believed that the difficulty is not fundamental, and it is expected that good permanent-magnet operation will be realized in the near future.

As illustrated in the foregoing diagrams, the tube is packaged so far as the radio-frequency connections are concerned, but it is removable from the magnet. When the tube is placed in the magnet, the metal envelope serves to align the tube with the magnetic field and also to index the tube in the proper longitudinal position. To place the tube in operation, it is necessary only to connect the radio-frequency lines and the socket.

4. *Acknowledgment*

The authors wish to express thanks to all their colleagues for their participation in the work described here and, in particular, to Mr. E. Reynoso for the fabrication of the *X216* tubes.

Low-Noise Traveling-Wave Tube*

By A. G. PEIFER, PHILIP PARZEN, and J. H. BRYANT

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THE LOW-NOISE traveling-wave tube to be described has been designed for use as an input stage of a microwave receiver. For such service, low noise and moderately high gain are desired. The traveling-wave type of tube offers a considerable bandwidth with no tuning adjustments. The tube has a gain of 15 decibels with an output power of one-half milliwatt at frequencies from 4200 to 5200 megacycles per second. With a velocity-jump electron gun and appropriate drift-tube lengths, a noise figure of 10 decibels is obtained. The operating voltage and current are 600 volts and one-half milliampere, respectively. A new version of the tube is now under construction in which the gain will be increased from 15 to 30 decibels with a comparable noise figure.

• • •

A simplified diagram of a helix-type traveling-wave tube is shown in Figure 1. The three essential parts consist of the matching sections, the helix, and the electron gun. The matching sections couple the input and output radio-frequency lines to the helix. To obtain energy transfer between the electron beam and the electromagnetic wave propagating down the helix, the axial velocity of the beam must be nearly equal to the axial velocity of the electromagnetic wave. The axial velocity of the electromagnetic wave is determined by the pitch and diameter of the helix and, therefore, the helix determines the operating voltage of the electron gun.

A brief outline will be given of means for noise reduction in traveling-wave tubes, and consideration given to the method of measurement of noise figure. Finally, some results of measurements will be given, and a low-noise tube will be described.

* Presented at the National Electronics Conference in Chicago, Illinois on October 23, 1951. Reprinted from *Proceedings of the National Electronics Conference—1951*, v. 7, pp. 314–317. Development of this tube is being sponsored by the United States Army Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey.

1. Theory of Noise Reduction

The limiting noise in a traveling-wave tube, which determines the ultimate noise figure, is generated by the velocity and current fluctuations of the electron beam. The theory used for design of the tube described here considers the velocity and current fluctuations in three distinct regions, namely, the diode region, the field-free drift region, and the helix.

The final results can be simply stated by means of the equation

$$F = 1 + \frac{q^2}{2c}\frac{T_0}{T_a} f \sec^2 a_0 \tag{1}$$

where q^2 is the magnification of the velocity fluctuations from the cathode to the drift region preceding the helix, c is the gain parameter, T_0 is the cathode temperature in degrees Kelvin, T_a is the ambient temperature in degrees Kelvin, f is a factor that gives the variation of noise figure with length of drift region preceding the helix, and a_0 is the ratio of the current to velocity fluctuations at the entrance to the drift region

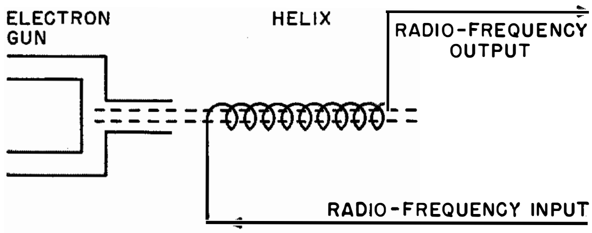


Figure 1—Elementary traveling-wave tube.

preceding the helix. When these factors are known, the length of the drift region necessary to produce a minimum noise figure may be computed.

Another noise-reducing scheme derived by an extension of the theory requires two drift regions and a velocity jump, which is a sudden acceleration or deceleration of the electron beam. A velocity jump upward is made to occur at a velocity-modulation maximum.

The velocity fluctuations are reduced by the square root of the ratio of the potentials of the drift regions while the current fluctuations remain constant. Calculations for the tube described here, which incorporates two drift regions with a velocity jump, predict a noise figure of eight decibels. However, measurements thus far have shown a noise figure of 10 decibels.

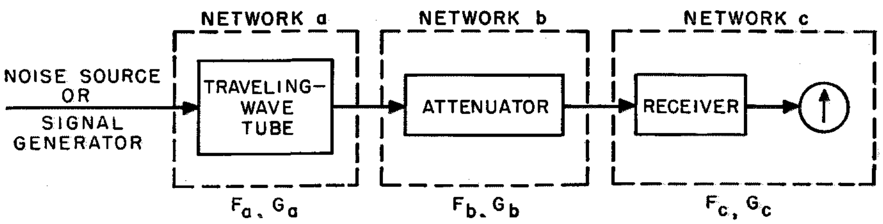


Figure 2—Noise figure of three cascaded networks.

2. Measurement

The noise figure of a network is defined as the ratio of the input available signal to available noise power divided by the ratio of the output available signal to available noise power. The over-all noise figure of the traveling-wave tube a, attenuator b, and receiver c shown in Figure 2 is

$$F_{abc} = F_a + \frac{F_b - 1}{G_a} + \frac{F_c - 1}{G_a G_b} \tag{2}$$

where F and G represent the noise figure and gain of the network designated by the subscripts. The noise figure of the attenuator, network b, is one and its gain is the reciprocal of the loss. Upon rearranging (2) and substituting into it, the noise figure of the tube becomes

$$F_a = F_{abc} - \frac{F_c - 1}{G_a \times 1/L_b} \tag{3}$$

Therefore, to find the noise figure of the tube, it is necessary to know the over-all noise figure of the cascaded networks, the gain of the tube, and the noise figure of the receiver.

The noise-figure measurements were made by using the apparatus illustrated in Figure 3. The input power to the three networks may be obtained from a noise source or from a signal generator. The initial level of the receiver indicator is set by attenuator b with no input power applied to the networks, but with the tube in operation. The over-all noise figure may be calculated by measuring the power necessary to double the receiver output power.

Since the bandwidth of the traveling-wave tube is quite large, the limiting bandwidth of the cascaded networks is the bandwidth of the receiver c. The amplitude-modulation microwave receiver used was constructed in the laboratory and has a bandwidth of 40 megacycles. A gas-discharge-lamp-type noise source was used to measure the noise figure of the receiver.

To eliminate the necessity of changing radio-frequency circuit connections during an experiment, a waveguide switch was constructed. The two positions of the switch in Figure 3 correspond to the positions for measuring over-all gain and noise, respectively. For gain measurements, the input power bypasses the tube and the input and

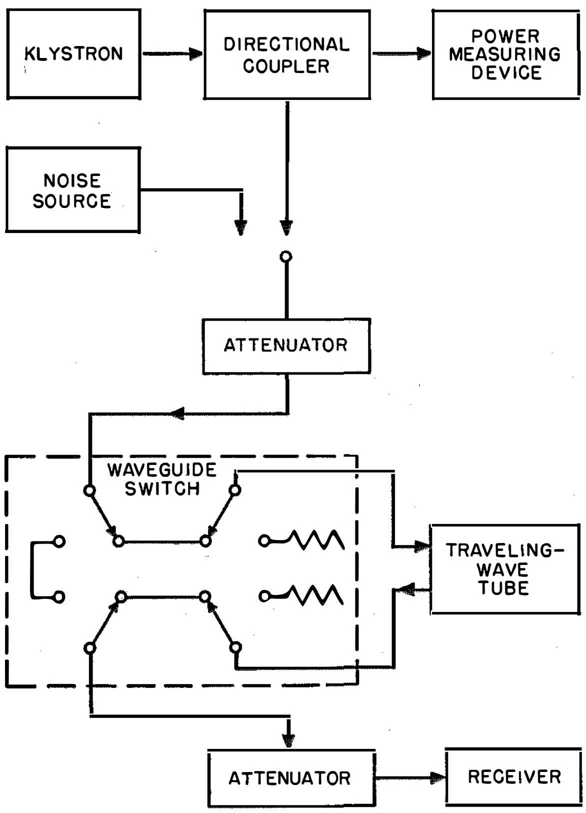


Figure 3—Schematic of noise-measuring equipment.

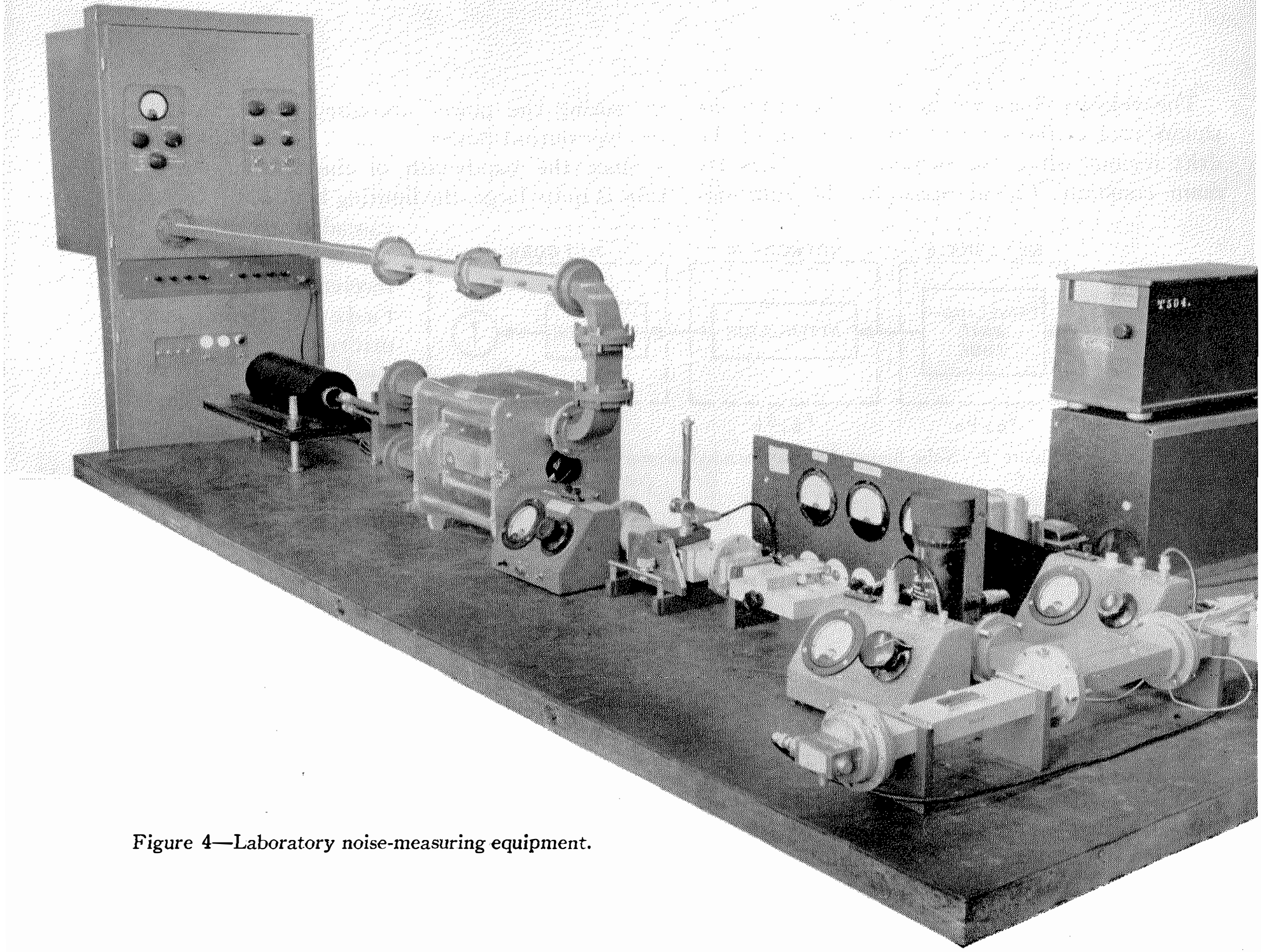


Figure 4—Laboratory noise-measuring equipment.

output lines of the tube are terminated with waveguide loads. In Figure 4 is shown the laboratory equipment for making noise measurements.

3. Preliminary Investigations

At the beginning of the development, a demountable traveling-wave tube, whose cathode-to-helix spacing could be varied, was set up. The electron gun was a single drift gun with no velocity jump. The correlation between the calculated and experimental minima was good; however, the absolute value of the experimental noise figure was two decibels higher than the theoretical value. This may be explained by the fact that the partition current was higher than normal in this experimental tube.

4. Magnetic-Field Requirements

The uniform axial magnetic field required for low-noise operation is about 275 gausses over a length of $11\frac{3}{16}$ inches (28.4 centimeters). A solenoid to provide this field requires approximately 50 watts of power. In Figure 5 is shown a solenoid with the tube in place. The solenoid is 4 inches (10 centimeters) in diameter, $11\frac{3}{16}$ inches long, and weighs about 30 pounds (13.5 kilograms). Plots of the magnetic field of permanent-magnet structures available to date have shown the fields to be too nonuniform in the axial direction. However, the permanent magnets measured were only preliminary samples and better performance may be expected with further development and design.

5. Mechanical Design

The mechanical design was aimed at a rugged tube that would require few adjustments when being set up for operation. Figure 6 is a photograph of the tube. To ruggedize the tube, a metal envelope was used to enclose all of the components. The envelope, besides being rugged, serves as a radio-frequency shield preventing stray radiation from interfering with the operation of the tube.

Bearing surfaces on the metal envelope align the tube with the magnetic field and adjustment is not necessary for the intial setup or when tubes are interchanged. The matching sections do not require adjustment when the tube is operated in the frequency band of 4200 to 5200 megacycles per second. The coaxial radio-frequency lines are terminated with coaxial type-N connectors. Connections to the electron gun are made with a 12-pin medium-shell diheptal base located at the opposite end from the radio-frequency connectors.

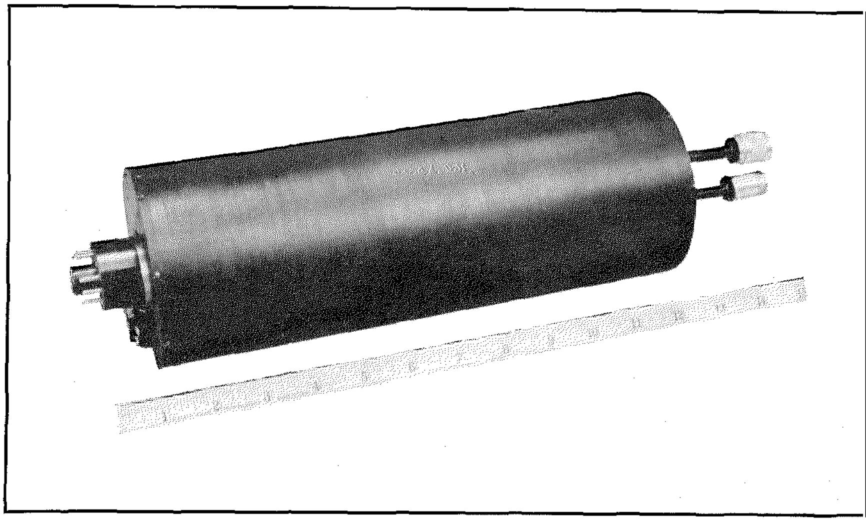


Figure 5—Low-noise traveling-wave tube in solenoid.

TABLE 1
TYPICAL OPERATING CONDITIONS

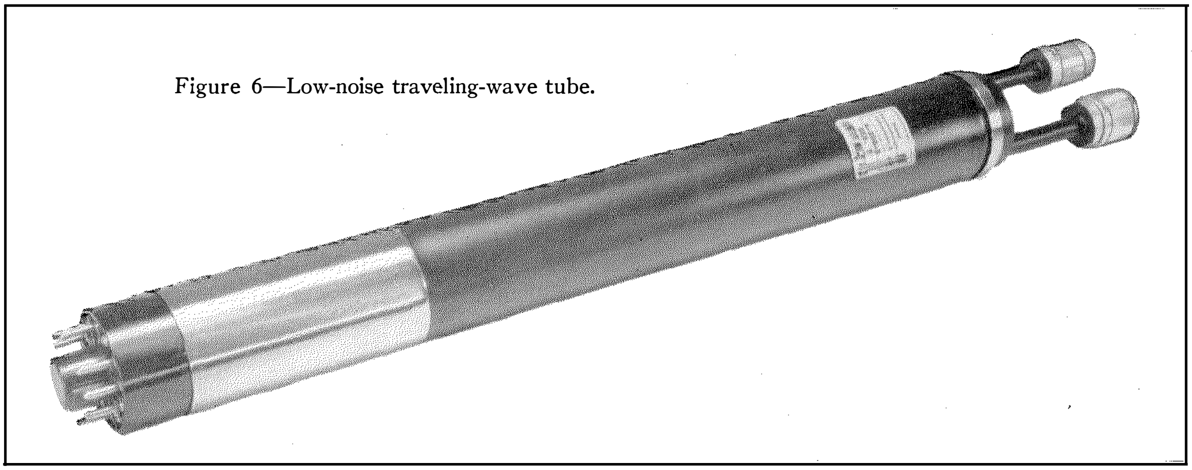
Heater voltage	6.3 volts
Heater current	0.6 ampere
Cathode potential	0 volts
Cathode current	0.5 milliampere
Focus voltage	0 volts
Anode *1* voltage	150 volts
Anode *2* voltage	600 volts
Helix voltage	600 volts
Collector voltage	650 volts
Collector current	0.5 milliampere
Solenoid voltage	100 volts
Solenoid current	0.5 ampere
Frequency range	4200–5200 megacycles
Gain	15 decibels
Noise Figure	10 decibels

6. Operating Data

Typical operating conditions are given in Table 1.

Beam transmission to the collector is better than 99 percent of the cathode current. The helix, tube envelope, and solenoid are operated at ground potential.

Figure 6—Low-noise traveling-wave tube.



Space-Charge-Wave Propagation in a Cylindrical Electron Beam of Finite Lateral Extension*

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A METHOD is outlined for obtaining the effect of geometry on the propagation of space-charge waves in a cylindrical electron beam under arbitrary direct-current conditions. In particular, the case of the complete-space-charge diode is discussed in detail. The results of this theory are verified by the experiments of Cutler and Quate.

• • •

1. Introduction

The nature of the propagation of space-charge waves in cylindrical electron beams of finite lateral extension has been discussed by Hahn[1] and Ramo[2] for the case of an electron beam drifting with a constant velocity v_0 in a tube. According to the latter paper, the field quantities propagate with a wave number Γ given by

$$\Gamma=\frac{\omega}{v_0}\pm\frac{\omega_p}{v_0[1+(T^2/\gamma_0^2)]^{1/2}} \tag{1}$$

where ω is the angular frequency, $\gamma_0=\omega/v_0$, the plasma frequency

$$\omega_p=(e\rho_0/m\epsilon_0)^{1/2},$$

ρ_0 is the direct-current charge density, the direct-current density $J_0=\rho_0 v_0$, and T/γ_0 is a function of $R=a/b$ and $\gamma_0 b$. T is the radial propagation number in the beam, a is the radius of the drift tube, and b is the radius of the beam.

A plot of T/γ_0 versus $\gamma_0 b$ for various values of R is given in Figure 1. For $b=\infty$ (infinite electron beam), T equals zero, and for $R=1$, $Tb=2.4$. For all other conditions, Tb varies between 0 and 2.4.

Equation (1) may be interpreted physically as follows: The first term denotes the phase shift due to the direct-current transit angle. The second term denotes the phase shift due to space charge and is a function of geometry. The latter term is proportional to $[J_0/1+(T^2/\gamma_0^2)]^{1/2}$. Thus, it appears that the behavior of this phase shift is governed by an effective direct-current density J_{0e}, which is given by

$$J_{0e}=J_0/1+(T^2/\gamma_0^2). \tag{2}$$

Thus, one may obtain the nature of the propagation of space-charge waves in those problems where the direct-current parameters and geometry may vary along the direction of propagation by replacing J_0 by J_{0e} in the differential equations governing such propagation in an infinite beam. A more-rigorous discussion of this replacement process is given in the appendix. This method assumes that any lateral variations of direct-current parameters are small. The equations for an infinite beam will now be given.

1. *Space-Charge-Wave Propagation in an Infinite Electron Beam*

The Maxwell equation for small-signal theory and for a one-dimensional propagating process in an electron beam are

$$\left(j\omega+\frac{\partial v_0}{\partial Z}\right)v+v_0\frac{\partial v_0}{\partial Z}=\frac{e}{m}E, \tag{3}$$

$$\frac{\partial J}{\partial Z}+j\omega\rho=0, \tag{4}$$

$$J+j\omega\epsilon_0 E=0, \tag{5}$$

$$J=\rho_0 v+\rho v_0. \tag{6}$$

Here all field quantities vary as $\exp j\omega t$ and the alternating-current quantity is assumed small with respect to the corresponding direct-current quantities.

E, J, v, ρ are the alternating-current electric field, current density, velocity, and charge density, respectively, and E_0, J_0, v_0, ρ_0 are the cor-

* Reprinted from *Journal of Applied Physics*, v. 23, pp. 215–219; February, 1952. This work was sponsored by the Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey.

[1] W. C. Hahn, "Small Signal Theory of Velocity-Modulated Electron Beams," *General Electric Review*, v. 42, pp. 258–270; June, 1939.

[2] S. Ramo, "Electronic-Wave Theory of Velocity-Modulation Tubes," *Proceedings of the IRE*, v. 27, pp. 757–762; December, 1939.

responding direct-current quantities. e is the charge of the electron and is negative, and m is the mass of the electron.

After solving for all field quantities in terms of J, one obtains the following differential equation for J.

$$v_0^3\frac{\partial^2 J}{\partial Z^2}+\left[3v_0^2\frac{\partial v_0}{\partial Z}+j2\omega v_0^2\right]\frac{\partial J}{\partial Z}+\left[\frac{eJ_0}{m\epsilon_0}+j2\omega v_0\frac{\partial v_0}{\partial Z}-\omega^2 v_0\right]J=0. \quad (7)$$

$$v=\frac{v_0}{J_0}\left[J+\frac{v_0}{j\omega}\frac{\partial J}{\partial Z}\right]. \quad (8)$$

Equation (7) can be simplified by replacing Z by the transit time τ. Thus let

$$J=(Y/v_0)\exp -j\omega\tau,$$

$$\tau=\int_0^Z (dZ/v_0). \quad (9)$$

Then Y satisfies the equations

$$(\partial^2 Y/\partial\tau^2)-K^2Y=0, \quad (10)$$

$$K^2=\frac{1}{v_0}\left[\frac{\partial^2 v_0}{\partial\tau^2}-\frac{eJ_0}{m\epsilon_0}\right], \quad (11)$$

and

$$v=\frac{v_0}{J_0}\left[J+\frac{1}{j\omega}\frac{\partial J}{\partial\tau}\right]. \quad (12)$$

To account for a finite geometry, one may now apply the replacement of J_0 by J_{0e} discussed previously. Thus, (11) now becomes

$$K^2=\frac{1}{v_0}\left[\frac{\partial^2 v_0}{\partial\tau^2}-\frac{eJ_0}{m\epsilon_0}\frac{1}{1+(T^2/\gamma_0^2)}\right]. \quad (13)$$

Thus, given K as a function of τ, one obtains a solution of (10) with two arbitrary constants. These constants are then determined by the initial conditions for J and v. Particular consideration will now be given to the case of the complete-space-charge diode.

2. *Space-Charge Waves in a Complete-Space-Charge Diode*

The case to be considered in Figure 2 is that of a cylindrical beam of radius b surrounded by a

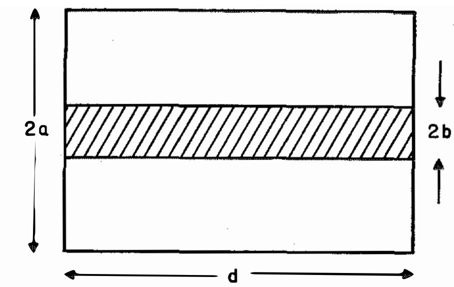

Figure 2—Complete-space-charge diode.

cylinder of radius a in a complete-space-charge diode. Denoting conditions at the cathode and anode by subscripts 1 and 2, respectively,

$$v_0=v_{02}(Z/d)^{2/3},$$

$$\tau=(3d^{2/3}/v_{02})Z^{1/3}.$$

Hence

$$v_0=\frac{v_{02}^3}{d^2}\left(\frac{\tau}{3}\right)^2. \quad (14)$$

Also

$$v_{02}^3=\frac{9eJ_0}{2m\epsilon_0}d^2.$$

Now let

$$\phi=g^{1/4}\frac{v_{02}}{d}\frac{\tau}{3};\ 0<\phi<g^{1/4},$$

$$g=5.76/(\gamma_{02}b)^2,$$

$$G=(Tb/2.4)^2,$$

$$v_0=(v_{02}/g^{1/2})\phi^2. \quad (15)$$

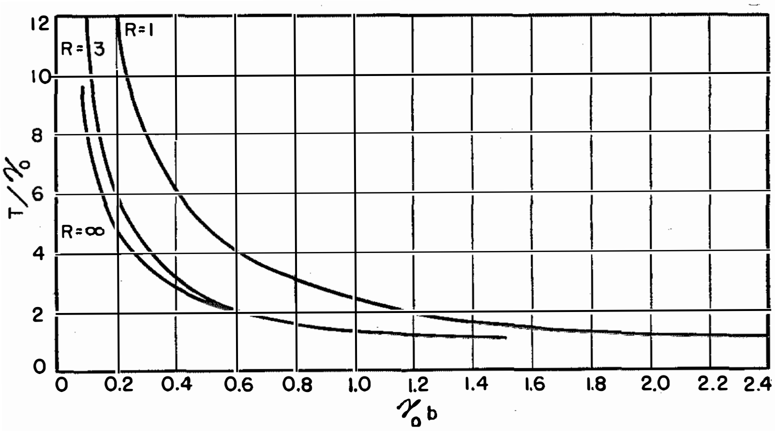

Figure 1—Correction to plasma frequency in drift tube for beam and drift-tube geometry. R is the (radius of the drift tube)/(radius of the beam), $\gamma_0 = \omega/v_{0d} = (2\pi f)$/(beam velocity in the drift tube), b is the beam radius, ω_d is the plasma frequency in the drift tube, $\omega_{d'}$ is the plasma frequency corrected for beam and drift-tube geometry, and $\omega_{d'}/\omega_d = 1/[1+(T/\gamma_0)^2]^{1/2}$.

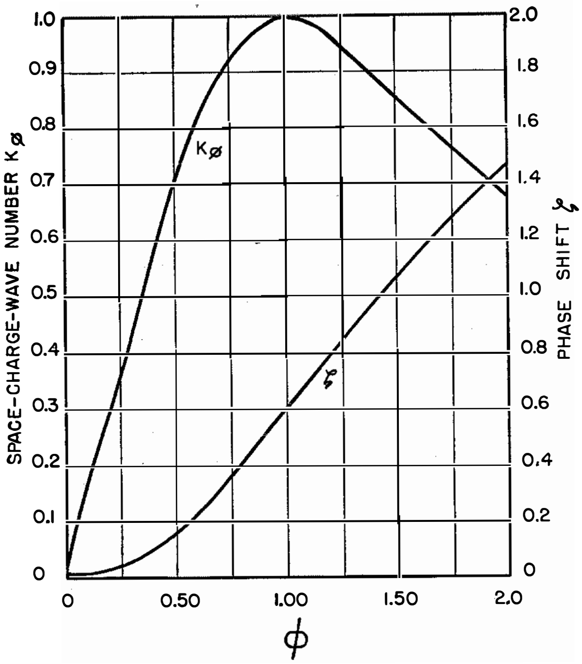

Figure 3—Phase shift ζ and space-charge-wave number K_ϕ plotted against ϕ.

Thus (10) becomes

$$(\partial^2 Y/\partial\phi^2) - K_\phi{}^2 Y = 0,$$
$$K_\phi{}^2 = 2G\phi^2/(1+G\phi^4),$$
$$v = \frac{v_0}{J_0}\left(J + \frac{g^{1/4}}{j3\gamma_{02}d}\frac{\partial J}{\partial\phi}\right). \tag{16}$$

Now applying the WKB method,[3] an approximate solution is

$$Y = C_1 P + C_2 Q,$$
$$P = \left(\frac{\zeta}{K_\phi}\right)^{1/2} I_{1/4}(\zeta),$$
$$Q = \left(\frac{\zeta}{K_\phi}\right)^{1/2} I_{-1/4}(\zeta),$$
$$\zeta = \int_0^\phi K_\phi d\phi. \tag{17}$$

Here $I_{1/4}$ is the modified Bessel function of order $\frac{1}{4}$. C_1 and C_2 are arbitrary constants to be determined from the initial conditions. The case of $G=1$ for $a=b$ and $G\neq 1$ will now be discussed separately.

[3] L. Schiff, *Quantum Mechanics*, McGraw-Hill Book Company, New York, New York; 1949: p. 181.

2.1 $G=1$ for $a=b$

A plot of K_ϕ and ζ versus ϕ is given in Figure 3. From (9) and (17)

$$J = \exp\left(-j\frac{3\omega}{g^{1/4}}\frac{d}{v_{02}}\phi\right)\frac{C_1P+C_2Q}{v_0}. \tag{18}$$

For small ϕ

$$K_\phi = 1.414\phi\left(1-\frac{1}{2}\phi^4\right),$$
$$\zeta = 0.707\phi^2\left(1-\frac{1}{6}\phi^4\right),$$
$$P = 0.60\phi\left(1+\frac{27}{120}\phi^4\right), \tag{19}$$

and Q approaches a value not equal to zero. It follows readily from this that $C_2=0$. Otherwise, v would not be finite at the cathode. Thus

$$J = C_1(P/v_0)\exp -j\omega\tau \tag{20}$$

and

$$v = \frac{C_1}{J_0}\left(\frac{-g^{1/4}}{j\omega}\frac{v_{02}}{3d}\right)\left(\frac{\partial\log v_0}{\partial\phi} - \frac{\partial\log P}{\partial\phi}\right)P\exp -j\omega\tau. \tag{21}$$

Thus from the initial conditon for v and using (19)

$$C_1 = -\frac{1}{0.60}\left(\frac{j3\omega}{g^{1/4}}\frac{d}{v_{02}}\right)J_0 v_1. \tag{22†}$$

Now let J_{2p} and v_{2p} be the current and velocity modulations, respectively, at the anode that are obtained by the method, neglecting geometry.[4] Thus

$$J_2 = J_{2p}G_j,$$
$$v_2 = v_{2p}G_v, \tag{23}$$
$$G_j = \frac{P_2}{0.60\phi_2},$$
$$\frac{G_v}{G_j} = \phi_2\left(\frac{\partial\log v_{02}}{\partial\phi} - \frac{\partial\log P_2}{\partial\phi}\right). \tag{24}$$

Before undertaking any extensive calculations, it is useful to obtain values of G_j and G_v for small and large ϕ_2.

† This solution will satisfy the small-signal conditions everywhere except near the cathode. A rigorous solution can be obtained only by including the effect of thermal velocities.

[4] C. C. Cutler and C. F. Quate, "Experimental Verification of Space Charge and Transit Time Reduction of Noise in Electron Beams," *Physical Review*, v. 80, pp. 875–878; December, 1950.

2.1.1 *Small* ϕ_2

Using (19), one obtains

$$G_j = 1 + (27/120)\phi^4, \qquad G_v/G_j = 1 - (27/30)\phi^4. \tag{25}$$

2.1.2 *Large* ϕ_2

For $\zeta_2 > 0.75$, one may use the asymptotic formulas for the Bessel function. Thus

$$P_2 = \frac{1}{2.5}\frac{\exp \zeta_2}{K_2^{1/2}}, \tag{26}$$

$$G_j = \frac{P_2}{0.60\phi_2}, \tag{27}$$

$$\frac{G_v}{G_j} = 2.5 - \frac{\phi_2^4}{1+\phi_2^4} - \phi_2 K_2. \tag{28}$$

Accurate values may be obtained by using (25) for $\phi_2 < 0.75$ and (26)–(28) for $\phi_2 \geqslant 1.25$. A curve of G_j and G_v/G_j versus ϕ_2 is plotted in Figure 4. It is to be noted that the value of G_v/G_j for large ϕ_2 is independent of ζ_2.

2.2 $G<1$ OR $a>b$

Figure 4 may also be interpreted as a plot of G_j and G_v/G_j versus the value of $(T/\gamma_{02})^{1/2}$ at the anode. The simplest way to compute these quantities for this case would be to use Figure 4 for the corresponding values at the anode. This would be rigorously true if G were a constant equal to $G(\phi_2)$. Physically this is plausible since the greatest effect of geomertry would be at the anode where T/γ_0 attains its greatest value. G itself is a decreasing function of ϕ, which equals 1 for small values of ϕ and increases monotonically from 0 to 1 for large values of ϕ_2 as a/b goes from ∞ to 1. Now for large values of ζ_2, where the value of G_v/G_j is independent of ζ, only $G(\phi_2)$ is important. In this case and for small values of ϕ_2, this interpretation of Figure 4 will be approximately correct. The curve of G_j will be somewhat in error for cases other than $G=1$. By the same reasoning, this result will also be approximately valid for a conical converging beam.

An experimental verification of this theory may be found in the experiments of Cutler and Quate[4] wherein the position of the current-modulation minimum in a drift tube following the diode did not agree with that computed from their theory. Using their notation,

$$\left[1+\left(\frac{T}{\gamma_{02}}\right)^2\right]^{1/2} = \frac{\omega_{p2}}{\omega(4QC^3)^{1/2}}. \tag{29}$$

According to their theory,

$$X_{\min}{}^{(1)} \text{ is proportional to } \tan^{-1}[\theta_1(4QC^3)^{1/2}]. \tag{30}$$

According to the theory presented here,

$$X_{\min}{}^{(2)} = \tan^{-1}\left[\theta_1(4QC^3)^{1/2}\frac{G_j}{G_v}\right]. \tag{31}$$

The numerical values in their experiment were $(4QC^3)^{1/2} = 2.8\times10^{-2}$, $\omega_{p2}/\omega = 0.052$, $T_2/\gamma_{02} = 1.56$, $\phi_2 \doteqdot 1.25$, $\theta_1 = 32$. From Figure 4, $G_v/G_j = 0.55$. Hence,

$$X_{\min}{}^{(2)}/X_{\min}{}^{(1)} = 1.02/0.73 = 1.40. \tag{32}$$

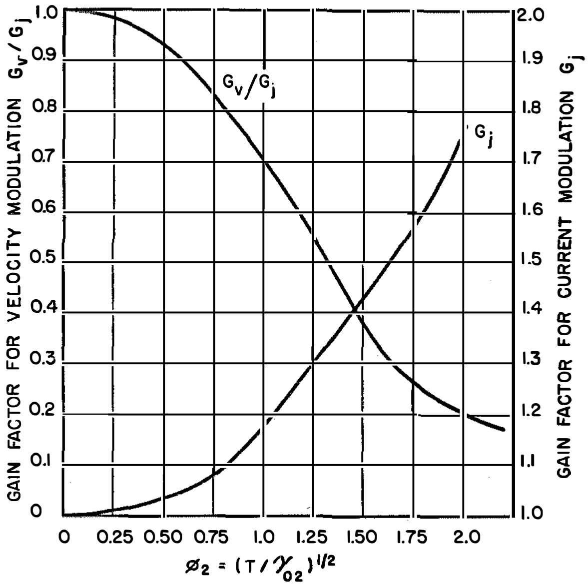

Figure 4—Gain factors for current and velocity modulations G_j and G_v/G_j plotted against $\phi_2 = (T/\gamma_{02})^{1/2}$.

By their measured values

$$X_{\min}{}^{(2)}/X_{\min}{}^{(1)} = 1.1/0.7 = 1.57. \tag{33}$$

Thus the agreement is seen to be rather good. Furthermore, since G_v/G_j is always less than 1, the position of the minimum will always be farther from the anode than that from the theory neglecting geometry.

A better agreement in this case can be obtained by considering in greater detail the effect of the variation of G.

Thus in this case, (28) becomes, letting $\psi = G^{1/4}\phi$.

$$\frac{G_v}{G_j}=2.5+\frac{\phi_2}{4G}\frac{\partial G}{\partial \phi}-\frac{\psi_2{}^4}{1+\psi_2{}^4}\left(1+\frac{\phi_2}{4G}\frac{\partial G}{\partial \phi}\right)-\psi_2 K(\psi_2). \quad (34)$$

Thus in the case $(\phi_2/4G)(\partial G/\partial\phi)\ll 1$, the previous discussion of the effect of G is rigorously correct. For the case of $a/b=\infty$, which is so in these experiments, an approximate formula for G for large ϕ is

$$G=\frac{0.087}{0.09+\log \phi}. \quad (35)$$

In this case, $\phi_2=1.6$ from Figure 1. Taking these values, $G_v/G_j=0.45$, which results in a computed value of $X_{\min}{}^{(2)}/X_{\min}{}^{(1)}$ equal to 1.52.

3. Conclusions

A general method has been given to obtain the effect of geometry on the propagation of space-charge waves in electron beams. The resulting differential equations, while complicated, may be approximately solved by WKB methods. The complete-space-charge diode has been discussed in detail, and the results have been checked by the experiments of Cutler and Quate. This method can be utilized for any direct-current conditions.

4. Appendix—Field-Theory Analysis of Propagation of Space-Charge Waves in Finite Beams

An attempt will now be made to derive the method of evaluating the effect of geometry from field-theory considerations. In this case, the ballistic equation is

$$v_0{}^3\frac{\partial^2 J}{\partial Z^2}+\left[3v_0{}^2\frac{\partial v_0}{\partial Z}+2j\omega v_0{}^2\right]\frac{\partial J}{\partial Z}+\left[2j\omega v_0\frac{\partial v_0}{\partial Z}-\omega^2 v_0\right]J=\frac{eJ_0}{m}j\omega E. \quad (36)$$

Here E is the component of the electric field in the direction of propagation. All transverse motions are neglected. The derivation of this is similar to that of (7).

The circuit equation, which formerly was given by (5), is now

$$\nabla_t{}^2E+\frac{\partial^2 E}{\partial Z^2}+k^2E=j\frac{\zeta}{k}\left(k^2J+\frac{\partial^2 J}{\partial Z^2}\right), \quad (37)$$

where

$$\nabla_t{}^2=\frac{\partial^2}{\partial X^2}+\frac{\partial^2}{\partial Y^2},$$

$k=2\pi/\lambda$, the free-space wave number, and $\zeta=(\mu_0/\epsilon_0)^{1/2}$.

This is derived directly from the wave equation. The case of $a=b$ will be considered first. Here let

$$\begin{aligned} E&=E_1\exp(-j\omega\tau)J_0(Tr),\\ J&=J_1\exp(-j\omega\tau)J_0(Tr),\\ \mathrm{Ta}&=2.4. \end{aligned} \quad (38)$$

This solution will satisfy the boundary condition that $E=0$ for $r=a$. E_1 is a function of Z only.

Equations (36) and (37) now become

$$v_0\frac{\partial^2 J_1}{\partial\tau^2}+2\frac{\partial v_0}{\partial\tau}\frac{\partial J_1}{\partial\tau}=\frac{e}{m}J_0 j\omega E_1, \quad (39)$$

$$v_0\frac{\partial^2 E_1}{\partial\tau^2}-\left(\frac{\partial v_0}{\partial\tau}+2j\omega v_0\right)\frac{\partial E_1}{\partial\tau}+\left[(-\gamma_0{}^2-T^2+k^2)v_0{}^3+j\omega\frac{\partial v_0}{\partial\tau}\right]E_1=j\frac{\zeta}{k}\left\{v_0\frac{\partial^2 J_1}{\partial\tau^2}-\left(\frac{\partial v_0}{\partial\tau}+2j\omega v_0\right)\frac{\partial J_1}{\partial\tau}+\left[(-\gamma_0{}^2+k^2)v_0{}^3+j\omega\frac{\partial v_0}{\partial\tau}\right]J_1\right\}. \quad (40)$$

Now it is known from the infinite-beam solution that E_1 varies as $\exp \pm j\omega_p\tau$. Hence if one assumes that:

$$\omega_p/\omega\ll 1, \quad (41)$$

$$\frac{1}{\omega v_0}\frac{\partial v_0}{\partial\tau}\ll 1, \quad (42)$$

$$k/\gamma_0\ll 1, \quad (43)$$

then the terms containing the derivatives may be neglected in (40). Equation (40) reduces to

$$E_1=-j\frac{\zeta}{k}\frac{J_1}{1+(T^2/\gamma_0{}^2)}. \quad (44)$$

Substituting this in (40), one has

$$v_0\frac{\partial^2 J_1}{\partial\tau^2}+2\frac{\partial v_0}{\partial\tau}\frac{\partial J_1}{\partial\tau}+\frac{e}{m\epsilon_0}\frac{J_0}{1+(T^2/\gamma_0{}^2)}J_1=0. \quad (45)$$

Thus it is seen that J_0 is reduced by the factor as stated in (2). This approximation will also be valid in the case $a>b$ provided T is a slowly varying function of Z. Conditions for (41) and (42) are obviously not satisfied near the cathode of a complete-space-charge diode. However, this does not matter as the effect of geometry is small there.

Determination of Electron Density and Collision Frequency in a Gaseous Discharge by Microwave Propagation Measurements*

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MICROWAVE measurement techniques have been adapted in the past few years to the determination of fundamental quantities in gaseous discharges. Rose et al.[1] measured electron densities and collision frequencies in pulsed radio-frequency discharges in a resonant cavity; Makinson et al.[2] measured electron densities in a direct-current discharge terminating a coaxial line.

A different microwave technique permitting the determination of electron densities and collision frequencies in gaseous discharges is reported here. This is a transmission method in which the gaseous discharge fills a section of waveguide several wavelengths long. Measurements are made of the absorption and phase shift of a low-power microwave signal passing through the discharge.

The electron gas produced in the discharge constitutes an absorbing dielectric medium. The resulting propagation constant γ for guided waves is given by

$$\gamma^2=\left(\frac{2\pi}{\lambda_{g0}}\right)^2\left\{\left[-1+\left(\frac{\lambda_{g0}}{\lambda}\right)^2\frac{c}{1+b^2}\right]+j\left(\frac{\lambda_{g0}}{\lambda}\right)^2\frac{cb}{1+b^2}\right\}, \quad (1)$$

where

λ = unbounded-space air wavelength,

λ_{g0} = air-filled guide wavelength,

$b = f_c/2\pi f$,

$c=\dfrac{f_p^2}{f^2}\approx\dfrac{0.8\times 10^{-4}\times N_0}{f^2_{\mathrm{mcps}}}$,

f = microwave signal frequency,

f_c = electron collision frequency,

f_p = electron "plasma" frequency, and

N_0 = number of electrons per cubic centimeter.

Let m be the measured absorption in decibels and δ the measured phase shift in degrees relative to air-filled-guide propagation, both per length λ_{g0} of discharge. Then, if

$$\left(\frac{m}{54.6}\right)^2-\frac{1}{4}\left(\frac{\delta}{180}\right)^2\ll\frac{\delta}{180}$$

(a condition realized in all but very-high-density discharges for microwave studies),

and

$$\left.\begin{aligned} b&\approx 6.6\,\frac{m}{\delta}\left(1-\frac{\delta}{360}\right)\\ c&\approx\left(\frac{\lambda}{\lambda_{g0}}\right)^2\frac{\delta}{180}\,(1+b^2).\end{aligned}\right\} \quad (2)$$

Thus, from measurement of absorption m and phase shift δ in the electron gas, b and c are determined, hence f_c and N_0. If b is small, N_0 is determined from phase-shift measurements alone.

A basic assumption underlying the above formulas is that the electron density be uniform over the whole volume of the discharge tube. Thus, only average electron density is determined. Corrections due to nonuniform density

* This work was sponsored by the United States Army Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey.

† Now with University of Illinois.

[1] D. J. Rose, D. E. Kerr, M. A. Biondi, E. Everhart, and S. C. Brown, "Methods of Measuring the Properties of Ionized Gases at Microwave Frequencies," Massachusetts Institute of Technology, Research Laboratory of Electronics, Technical Report 140; October 17, 1949.

[2] R. E. B. Makinson, P. C. Thonemann, R. B. King, and J. B. Ramsay, "Dielectric Constant and Electron Density in a Gas Discharge," *Proceeding of the Physical Society*, Section B, Part 8, v. 64, pp. 665–670; August, 1951.

distributions have not been studied, and may be substantial if the degree of nonuniformity is large in the region of large radio-frequency electric fields. Also, the effects of electron gas and glass boundaries on phase shift and absorption have been neglected, though this has been justified by calculations involving idealized boundaries.

Absorption in the electron gas is determined from measurements of total insertion loss and reflected power.

Phase-shift in the electron gas is measured by means of the circuit illustrated in Figure 1. The radio-frequency signal in arm *1* of the magic tee splits equally into two signals going down arms *2* and *3*, respectively. The signal in arm *3*, after being partially absorbed and phase shifted in the electron gas, interferes with the signal in arm *2* to form a standing-wave pattern. A direct measure of the phase shift is the shift of the position of the minimum of this pattern as detected by a radio-frequency probe in the slotted line. With large absorption in the electron gas, exceeding about 10 decibels, this method is limited by the difficulty of accurately locating the minimum of a shallow standing wave and also, possibly, by the presence of an interfering reflection of the signal from arm *2* at the gaseous discharge tube. This absorption limit can be extended by employing in arm *2* a compensating attenuator that is calibrated accurately for phase shift. An ultimate limit is set by the microwave noise generated by the discharge.

The experiments were performed with a standard X-band rectangular waveguide 1- by $\frac{1}{2}$- by 0.050-inch wall (25 by 13 by 1.3 millimeters) with a gaseous discharge tube approximately 9 inches (23 centimeters) long. Steady hot-cathode direct-current discharges in helium, neon, argon, and xenon were studied. Phase shifts were measured at signal frequencies of $f_1 = 8500$ megacycles per second and $f_2 = 12{,}000$ megacycles; absorptions, where measurable, only at 8500 megacycles. With subscripts 1 and 2 referring to measurements at f_1 and f_2, respectively, and with m and δ as defined above, the following relations, derived from (1), apply independent of gas and pressure.

$$\left.\begin{aligned} f_c &\approx 3.5\times 10^{11}\,\frac{m_1}{\delta_1},\\ N_{01} &\approx 2.0\times 10^{9}\delta_1(1+b_1^2),\\ N_{02} &\approx 7.0\times 10^{9}\delta_2(1+b_2^2), \end{aligned}\right\} \tag{3}$$

with $b_2 = (f_1/f_2)b_1$. N_{01} should equal N_{02} since both are, theoretically, equal to the electron density N_0. With M and Δ being total absorption and phase shift, respectively, $m_1/\delta_1 = M_1/\Delta_1$, $\delta_i = \Delta_i/n_i$ with n_i the number of air-filled-guide wavelengths along the discharge at frequency $f_i (i = 1, 2)$.

Table 1 gives the results obtained with argon. Parameter b_1 is reasonably constant with varying current at each pressure, as would be expected. Also, the average calculated densities at the two frequencies agree to within 20 per cent at 0.5 millimeter of mercury and to within 40 percent at 2 millimeters pressure. However, since a nonuniform density distribution may conceivably cause the same percentage error at the two frequencies, this agreement may not be very significant.

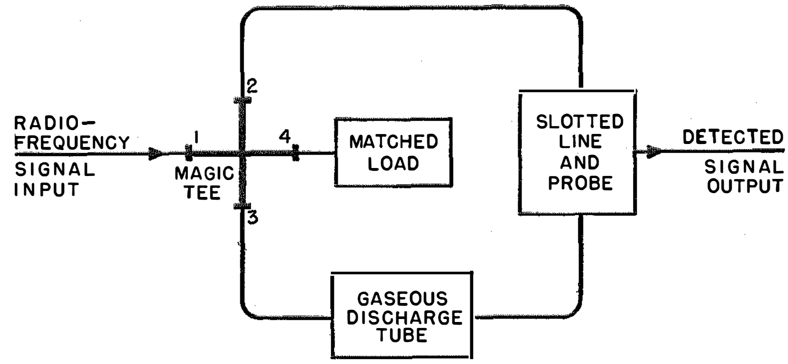


Figure 1—Phase shift in the radio-frequency signal is measured by the above apparatus.

In conclusion, the transmission method appears to be a convenient one for measuring average electron densities and collision frequencies in gaseous discharges. The accuracy is limited by the degree of nonuniformity of the electron density distribution. If the relative density distribution is known from other considerations or measurements, then the transmission method can possibly determine the absolute density values. This method, is, of course, applicable to the study of pulsed as well as steady discharges. By employing a narrowly defined beam in space, of say, submillimeter waves, and classical optical methods for measuring phase shift, the transmission technique may lend itself to direct exploration of electron density distribution in a discharge.

TABLE 1

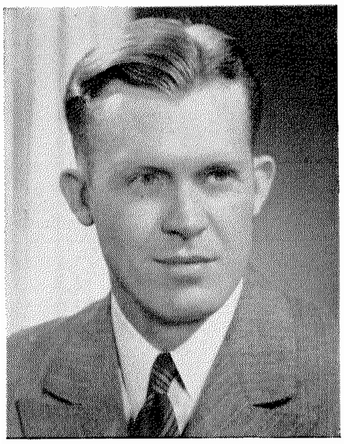
ARGON AT A PRESSURE OF 0.5 MILLIMETER OF MERCURY

Discharge Current in Milliamperes	Discharge Power in Watts	Δ_1 in Degrees	M_1 in Decibels	b_1	f_c in Cycles Per Second	N_{01} Per Cubic Centimeter	Δ_2 in Degrees	N_{02} Per Cubic Centimeter
2	0.43	1.9	—	—	—	1.0×10^9	1.2	1.1×10^9
4	0.82	5.2	—	—	—	2.6	2.4	2.3
6	1.16	7.8	—	—	—	3.9	4.8	4.6
8	1.49	9.7	—	—	—	4.9	6.0	5.7
10	1.83	14.3	—	—	—	7.2	8.4	8.0
12	2.14	18.2	—	—	—	9.1	10.8	10.3
14	2.44	20.7	0.4	0.13	6.9×10^9	10.4	13.2	12.5
16	2.72	26.6	0.5	0.12	6.4	13.3	15.6	14.8

ARGON AT A PRESSURE OF 2 MILLIMETERS OF MERCURY

Discharge Current in Milliamperes	Discharge Power in Watts	Δ_1 in Degrees	M_1 in Decibels	b_1	f_c in Cycles Per Second	N_{01} Per Cubic Centimeter	Δ_2 in Degrees	N_{02} Per Cubic Centimeter
4	1.00	5.8	—	—	—	3.3×10^9	3.6	3.7×10^9
8	1.76	19.4	1.0	0.34	1.8×10^{10}	11.2	13.2	13.5
10	2.10	25.2	1.5	0.39	2.1	14.5	19.2	19.7
12	2.40	33.6	2.0	0.39	2.1	19.3	24.0	24.6
16	2.96	53.7	3.6	0.44	2.3	30.9	36.0	36.9

Contributors to This Issue



JOHN H. BRYANT

JOHN H. BRYANT was born in Baird, Texas, on April 15, 1920. He received the B.S. degree in electrical engineering from the A&M College of Texas in 1942 and the Ph.D. degree from the University of Illinois in 1949.

From 1942 to 1946, he served as a commissioned officer in the Signal Corps of the United States Army.

For the summer of 1947, he was temporarily employed by the vacuum tube department of Federal Telecommunication Laboratories and in 1949 joined the staff as a project engineer. Dr. Bryant is now an assistant department head in the electron-tube laboratory.

He is a member of Sigma Xi, American Physical Society, and Institute of Radio Engineers.

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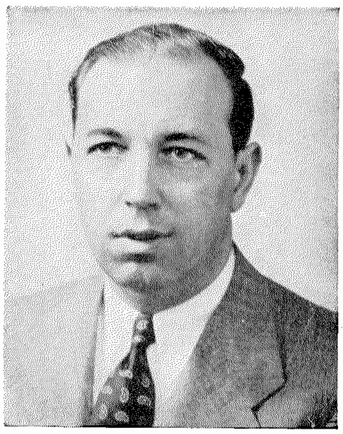

HENRI BUSIGNIES

HENRI BUSIGNIES was born at Sceaux, France, on December 29, 1905.

On completion of his military duties, he entered the Paris laboratories of the International Telephone and Telegraph Corporation in 1928. He has been active in the fields of direction finders, radio aids to navigation, and radar. He developed and demonstrated in Europe and the United States the first airborne automatic radio compass. The first of over a hundred patents was issued to him in 1926. He is now the technical director of Federal Telecommunication Laboratories.

Mr. Busignies received the Lakhovsky award of the Radio Club of France in 1926 for one of his inventions for an automatic radio compass. In 1948, he received the Presidential Certificate of Merit for his work with the National Defence Research Council during the second world war. He is a Fellow of the Institute of Radio Engineers.

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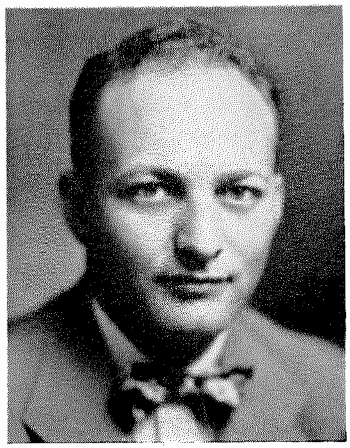

HERBERT W. COLE

HERBERT W. COLE was born in New York, New York, on April 30, 1923. He received a B.A. degree in physics from Gettysburg College in 1948. He then spent a year in the graduate school of Newark College of Engineering.

From 1940 to 1943, he was employed as a junior tool designer by Bendix Aviation Corporation. He then served as a commissioned officer and pilot in the United States Air Force from 1943 to 1946.

In 1948, Mr. Cole became a vacuum-tube engineer for Federal Telephone and Radio Corporation. In 1950, he was transferred to Federal Telecommunication Laboratories. He is now a senior engineer and is engaged in microwave tube research, specializing in traveling-wave tubes.

Mr. Cole is a Member of the Institute of Radio Engineers.

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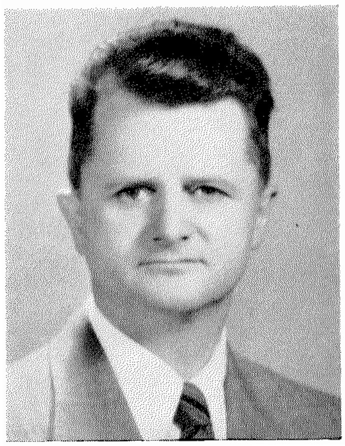


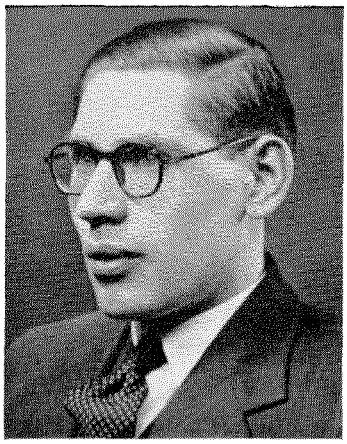
RICHARD C. DAVIS

RICHARD C. DAVIS was born on January 26, 1911 in Saugerties, New York. He received the M.E. degree from Stevens Institute of Technology in 1932.

From 1933 to 1938, he worked for Philco Radio and Television Corporation. He then served as a development engineer for Euthenics Products Corporation.

In 1939, he joined the engineering department of Federal Telegraph Company, where he worked on very-high-frequency radio links, receiver design, direction finders, and mobile radio equipment. He is now head of the engineering section on radio receivers and mobile equipment.

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



C. E. Eadon-Clarke

Frank Fairley was born in Preston, Lancashire, England in 1904. He received an honours degree in physics from Cambridge University in 1926.

Joining the European engineering department of International Standard Electric Corporation in 1927, he was transferred later to Standard Telephones and Cables, Limited.

From 1926 to 1936, he was engaged in planning and installing carrier systems, mostly in Malaya and India. Since then, he has developed transmission systems in London where he is now in charge of the planning and engineering of such systems.

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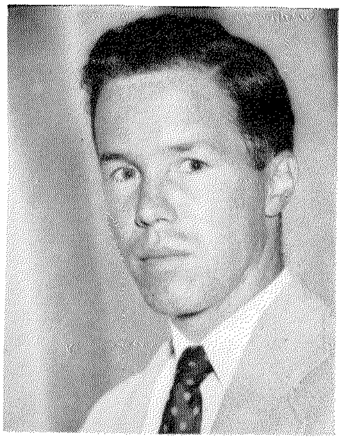

Richard H. Geiger

C. E. Eadon-Clarke was born in Hoddesdon, England, in September, 1919. He attended St. Lawrence College and Chelsea Polytechnic, receiving a special wartime diploma in radio with physics.

After working for some years in the chemical laboratories of Philips Electrical, Limited, he joined the radio division of Standard Telephones and Cables, Limited, in 1942. He worked on high-power radio transmitters during the war and was then transferred to the heat-treatment group of the industrial supplies division to develop and market all types of products associated with production-line heating systems.

Mr. Eadon-Clarke is an Associate of the Institution of Electrical Engineers.

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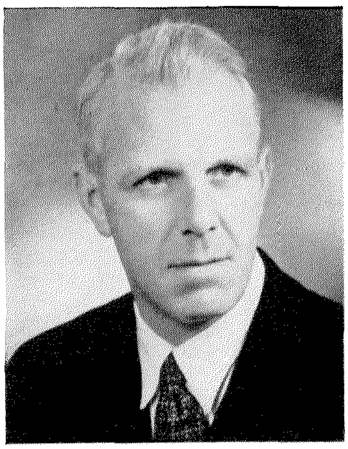

Frank Fairley

Richard H. Geiger was born in 1927 in New York, New York. He received the B.S. degree in engineering physics from Lehigh University in 1950.

Since 1950 Mr. Geiger has been with Federal Telecommunication Laboratories, where he is now working in the electron-tube laboratory on the application and development of gas-discharge devices at microwave frequencies.

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Ladislas Goldstein

Ladislas Goldstein. A photograph and biography of Doctor Goldstein appears on page 82 of the March, 1952, issue.

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Murray A. Lampert was born in New York City in 1921. He received the B.A. and M.A. degrees from Harvard University.

During the second world war, he taught in the Army-Navy Officers Electronics Training School at Harvard for two years and spent another year doing optical design work for the Harvard Observatory optical research project.

After the war, he worked for three years at the radiation laboratory of the University of California at Berkeley on the interaction of high-energy particles with nuclei.

Since 1949, Mr. Lampert has been in the vacuum tube department of Federal Telecommunication Laboratories, where he has worked on microwave amplifiers and microwave propagation through electron gases.

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K. L. Larsen was born at Ringsbolle, Denmark, on April 26, 1909. He received the M.Sc. degree in electrical engineering from the Royal Technical College in Copenhagen in 1934.

Joining the Danish Posts and Telegraphs administration in 1934, he was appointed assistant chief engineer in 1943. He is serving also as a lecturer in telecommunications to the Danish Army Officers School.

Knighthood in the Order of Danebrog was conferred on Mr. Larsen in 1950.

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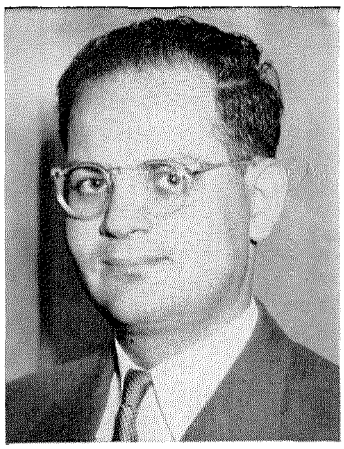

Murray A. Lampert

K. L. LARSEN

THEODORE J. MARCHESE was born on October 17, 1912 in Carlstadt, New Jersey. He received the B.S. degree in electrical engineering from the evening school of Newark College of Engineering in 1948.

He was employed by Federal Telegraph Company in 1932 as a technician and was transferred in 1941 to the engineering department of Federal Telephone and Radio Corporation as an engineer on large vacuum tubes. He was transferred again, in 1947, to Federal Telecommunication Laboratories, where he is doing development work on microwave power generators, negative grid tubes, and traveling-wave tubes.

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

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THEODORE J. MARCHESE

PHILIP PARZEN

PHILIP PARZEN. A photograph and biography of Mr. Parzen appears on page 83 of the March, 1952, issue.

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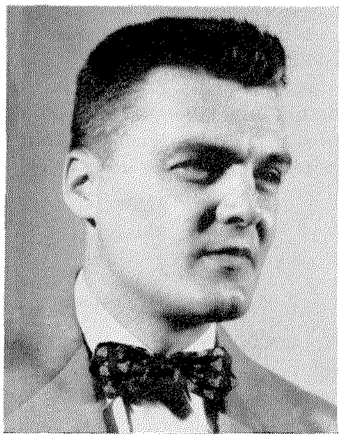
ALBERT PEIFER was born in 1922 in Mode, Illinois. He received the B.S. degree in electrical engineering in 1948 and the M.S. degree in 1950 from the University of Illinois. From 1947 to 1950 he was employed part time as a research assistant in the vacuum-tube laboratory at that university.

Since 1950, Mr. Peifer has been with Federal Telecommunication Laboratories doing research on traveling-wave tubes in the electron-tube laboratory.

Mr. Peifer is a Member of the Institute of Radio Engineers and of Eta Kappa Nu.



ALBERT G. PEIFER

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S. SIMON was born in Brussels, Belgium, on December 31, 1906. At the completion of his studies at Ecole Centrale des Arts et Métiers, Brussels, he was awarded the diploma of Ingénieur des Arts et Métiers of that school.

Mr. Simon joined the circuit laboratory of Bell Telephone Manufacturing Company in 1928, and has been in charge of the laboratory since 1942.

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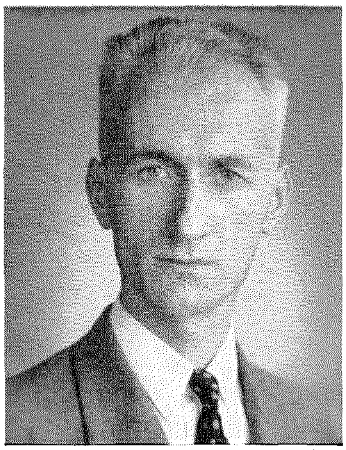
JAN TJEERD VISSER was born in the Netherlands in 1897. He graduated from the Institute of Technology (Technische Hoogeschool) at Delft in 1924.



S. SIMON

On graduation, he joined the Netherlands Posts, Telegraphs, and Telephones Administration as an engineer and has been active in the development of long-distance national and international telecommunication networks. Since 1925, he has been closely associated with the work of the Comité Consultatif International Téléphonique. The post-war restoration of damaged telephone network also occupied much of his time. In 1945, he was appointed head of the Central Cables and Repeaters Branch of the Posts, Telegraphs, and Telephones Administration.

Mr. Visser is a member of the Koninklijk Instituut van Ingenieurs and of the Vereeniging van Delftsche Ingenieurs.

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JAN TJEERD VISSER

INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION

Associate Manufacturing and Sales Companies

United States of America

Capehart-Farnsworth Corporation, Fort Wayne, Indiana
Flora Cabinet Company, Inc., Flora, Indiana
Thomasville Furniture Corporation, Thomasville, North Carolina
The Coolerator Company, Duluth, Minnesota
Federal Telephone and Radio Corporation, Clifton, New Jersey
Federal Electric Corporation, Clifton, New Jersey
International Standard Electric Corporation, New York, New York
International Standard Trading Corporation, New York, New York
I. T. & T. Distributing Corporation, New York, New York
Kellogg Switchboard and Supply Company, Chicago, Illinois

British Commonwealth of Nations

Standard Telephones and Cables, Limited, London, England
Creed and Company, Limited, Croydon, England
International Marine Radio Company Limited, Croydon, England
Kolster-Brandes Limited, Sidcup, England
Standard Telephones and Cables Pty. Limited, Sydney, Australia
Silovac Electrical Products Pty. Limited, Sydney, Australia
Austral Standard Cables Pty. Limited, Melbourne, Australia
New Zealand Electric Totalisators Limited, Wellington, New Zealand
Federal Electric Manufacturing Company, Ltd., Montreal, Canada

South America

Compañía Standard Electric Argentina, Sociedad Anónima, Industrial y Comercial, Buenos Aires, Argentina
Standard Electrica, S.A., Rio de Janeiro, Brazil
Compañía Standard Electric, S.A.C., Santiago, Chile

Continental Europe

Vereinigte Telephon- und Telegraphenfabriks Aktiengesellschaft Czeija, Nissl & Co., Vienna, Austria
Bell Telephone Manufacturing Company, Antwerp, Belgium
Standard Electric Aktieselskab, Copenhagen, Denmark
Compagnie Générale de Constructions Téléphoniques, Paris, France
Le Matériel Téléphonique, Paris, France
Les Téléimprimeurs, Paris, France
C. Lorenz, A.G. Stuttgart, Germany
Mix & Genest Aktiengesellschaft and Subsidiaries, Stuttgart, Germany
G. Schaub Apparatebau G.m.b.H., Pforzheim, Germany
Süddeutsche Apparatefabrik Gesellschaft m.b.H., Nuremberg, Germany
Fabbrica Apparecchiature per Comunicazioni Elettriche, Milan, Italy
Nederlandsche Standard Electric Maatschappij N.V., The Hague, Netherlands
Standard Telefon og Kabelfabrik A/S, Oslo, Norway
Standard Electrica, S.A.R.L., Lisbon, Portugal
Compañía Radio Aérea Marítima Española, Madrid, Spain
Standard Eléctrica, S.A., Madrid, Spain
Aktiebolaget Standard Radiofabrik, Stockholm, Sweden
Standard Telephone et Radio S.A., Zurich, Switzerland

Telephone Operating Systems

Companhia Telefônica Nacional, Rio de Janeiro, Brazil
Compañía de Teléfonos de Chile, Santiago, Chile
Cuban American Telephone and Telegraph Company, Havana, Cuba
Cuban Telephone Company, Havana, Cuba
Compañía Peruana de Teléfonos Limitada, Lima, Peru
Porto Rico Telephone Company, San Juan, Puerto Rico

Radiotelephone and Radiotelegraph Operating Companies

Compañía Internacional de Radio, Buenos Aires, Argentina
Compañía Internacional de Radio Boliviana, La Paz, Bolivia
Companhia Radio Internacional do Brasil, Rio de Janeiro, Brazil
Compañía Internacional de Radio, S.A., Santiago, Chile
Radio Corporation of Cuba, Havana, Cuba
Radio Corporation of Porto Rico, San Juan, Puerto Rico

Cable and Radiotelegraph Operating Companies

(Controlled by American Cable & Radio Corporation, New York, New York)

The Commercial Cable Company, New York, New York[1]
Mackay Radio and Telegraph Company, New York, New York[2]
All America Cables and Radio, Inc., New York, New York[3]
Sociedad Anónima Radio Argentina, Buenos Aires, Argentina[4]

[1] Cable service. [2] International and marine radiotelegraph services.
[3] Cable and radiotelegraph services. [4] Radiotelegraph service.

Laboratories

Federal Telecommunication Laboratories, Inc., Nutley, New Jersey
International Telecommunication Laboratories, Inc., New York, New York
Laboratoire Central de Télécommunications, Paris, France
Standard Telecommunication Laboratories, Limited, London, England