ELECTRICAL COMMUNICATION

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

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MANUFACTURE OF GERMANIUM POWER DIODES
POLYTHENE-INSULATED VIDEO-PAIR CABLES
LOUDSPEAKING INTERCOMMUNICATING SYSTEM AT LONDON AIRPORT
NEW TYPE OF DIFFUSION CATHODE
NOVEL GAS-GAP SPEECH SWITCHING VALVE
PRECISION RADAR RANGE CALIBRATOR
CATHODE-FOLLOWER PHASE-SHIFT OSCILLATOR
TELEPHONE STATISTICS OF THE WORLD
UNITED STATES PATENTS ISSUED TO THE INTERNATIONAL SYSTEM

INTERNATIONAL · TELEPHONE AND TELEGRAPH ·

Volume 32 SEPTEMBER, 1955 Number 3

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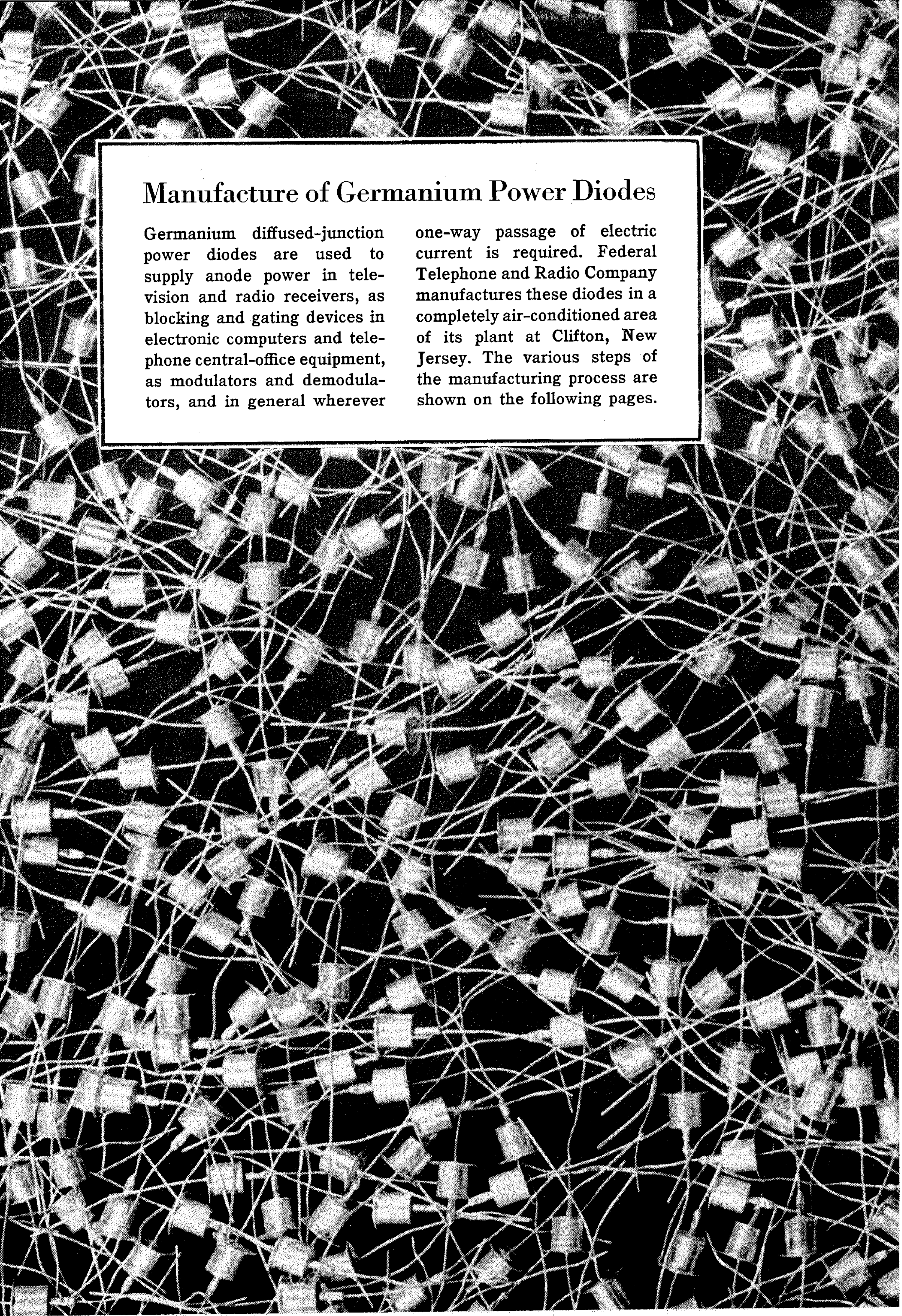
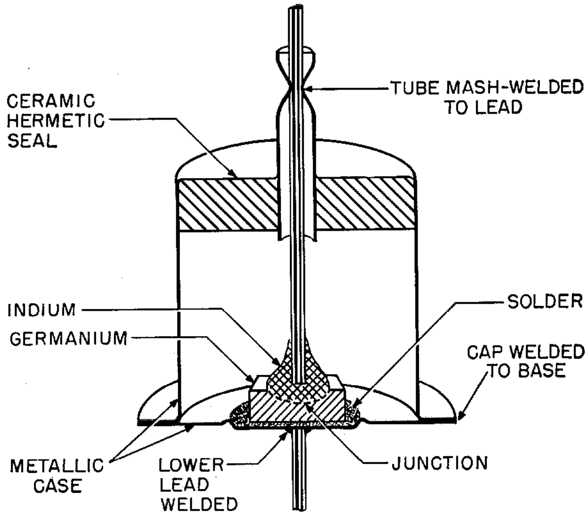
Germanium rectifier parts being assembled for heat treating.

Manufacture of Germanium Power Diodes

Germanium diffused-junction power diodes are used to supply anode power in television and radio receivers, as blocking and gating devices in electronic computers and telephone central-office equipment, as modulators and demodulators, and in general wherever one-way passage of electric current is required. Federal Telephone and Radio Company manufactures these diodes in a completely air-conditioned area of its plant at Clifton, New Jersey. The various steps of the manufacturing process are shown on the following pages.

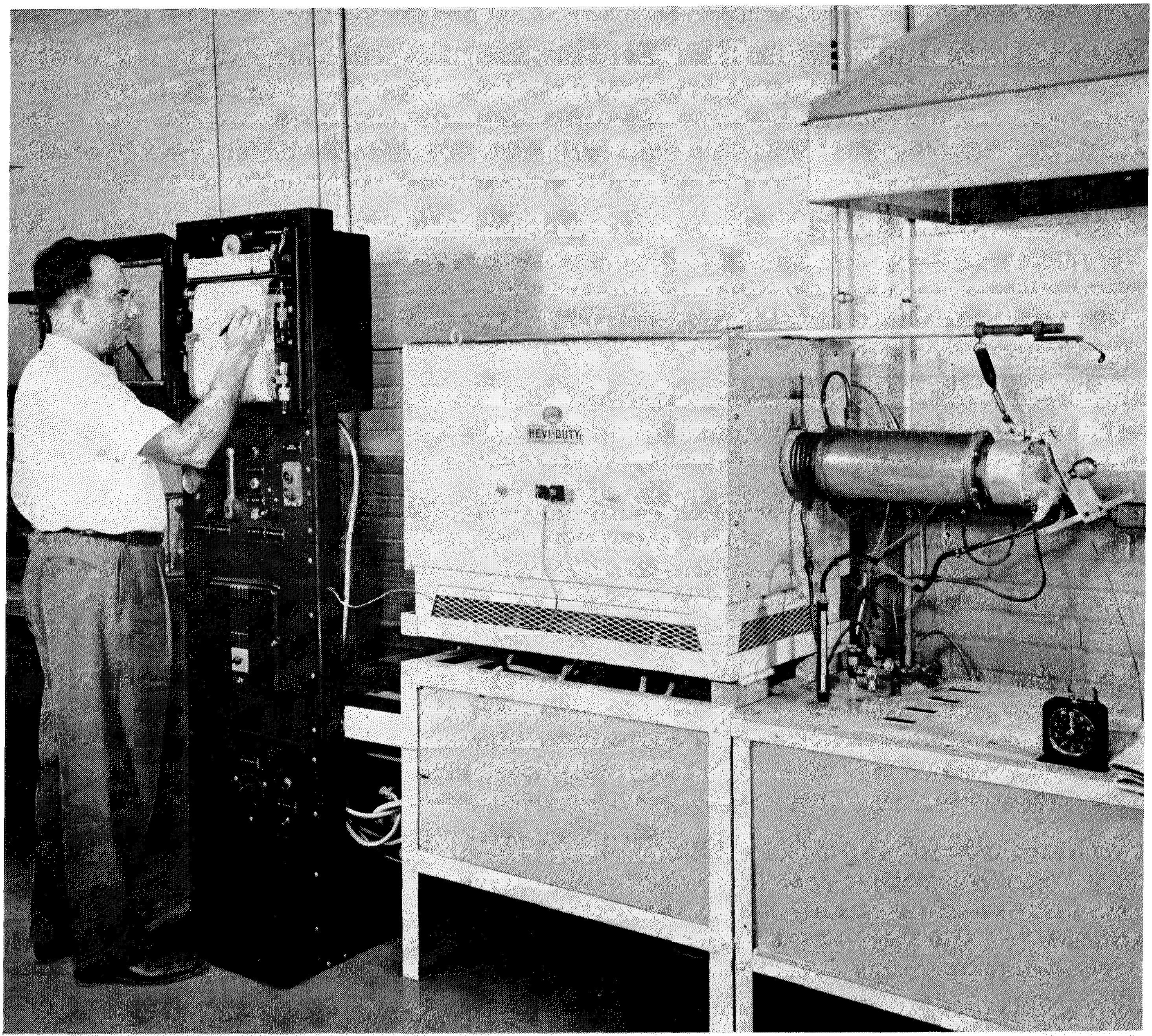




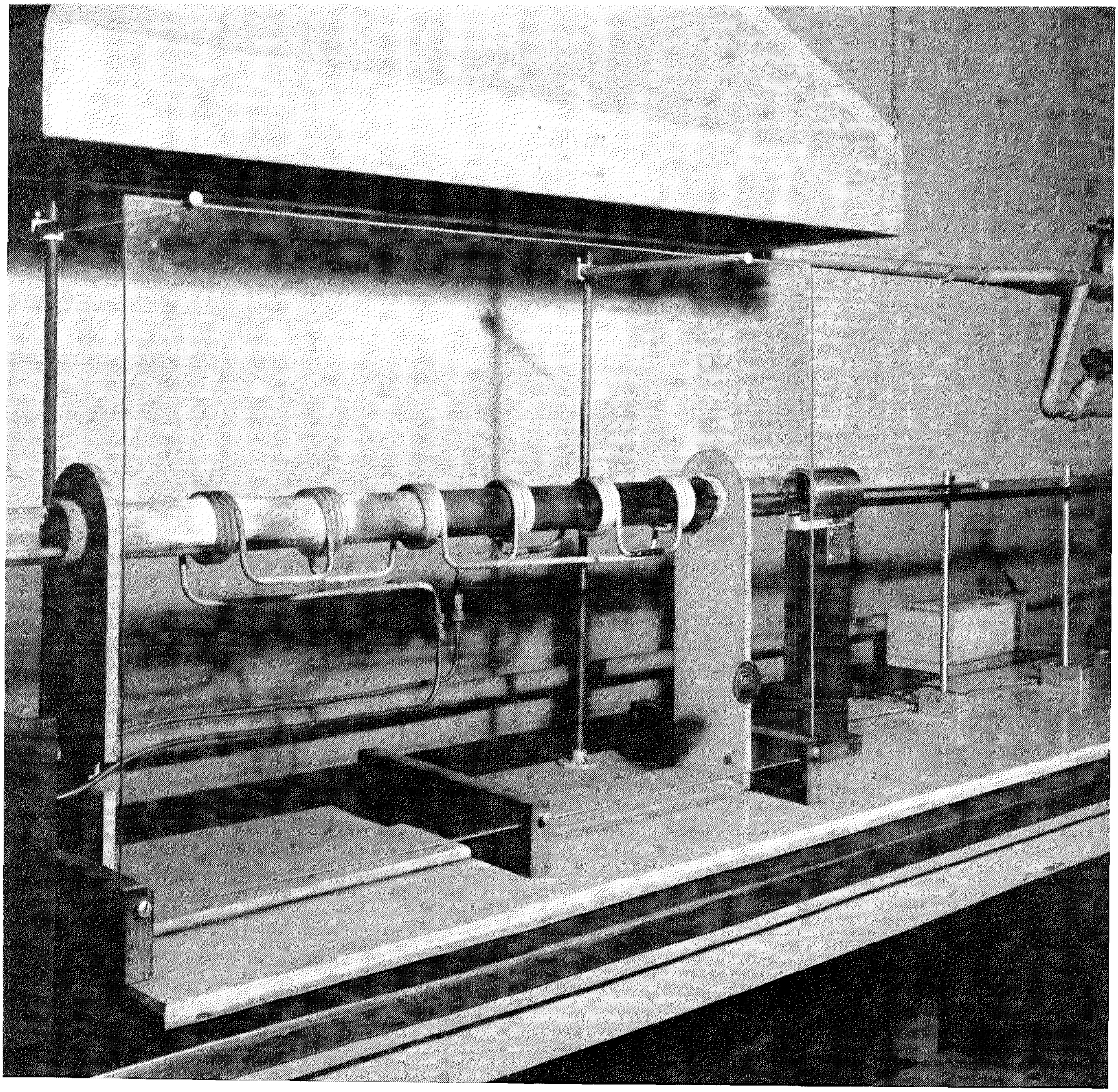
1. Cross-sectional drawing of a power diode is shown the left. The diodes pictured on the preceding page a approximately actual size.

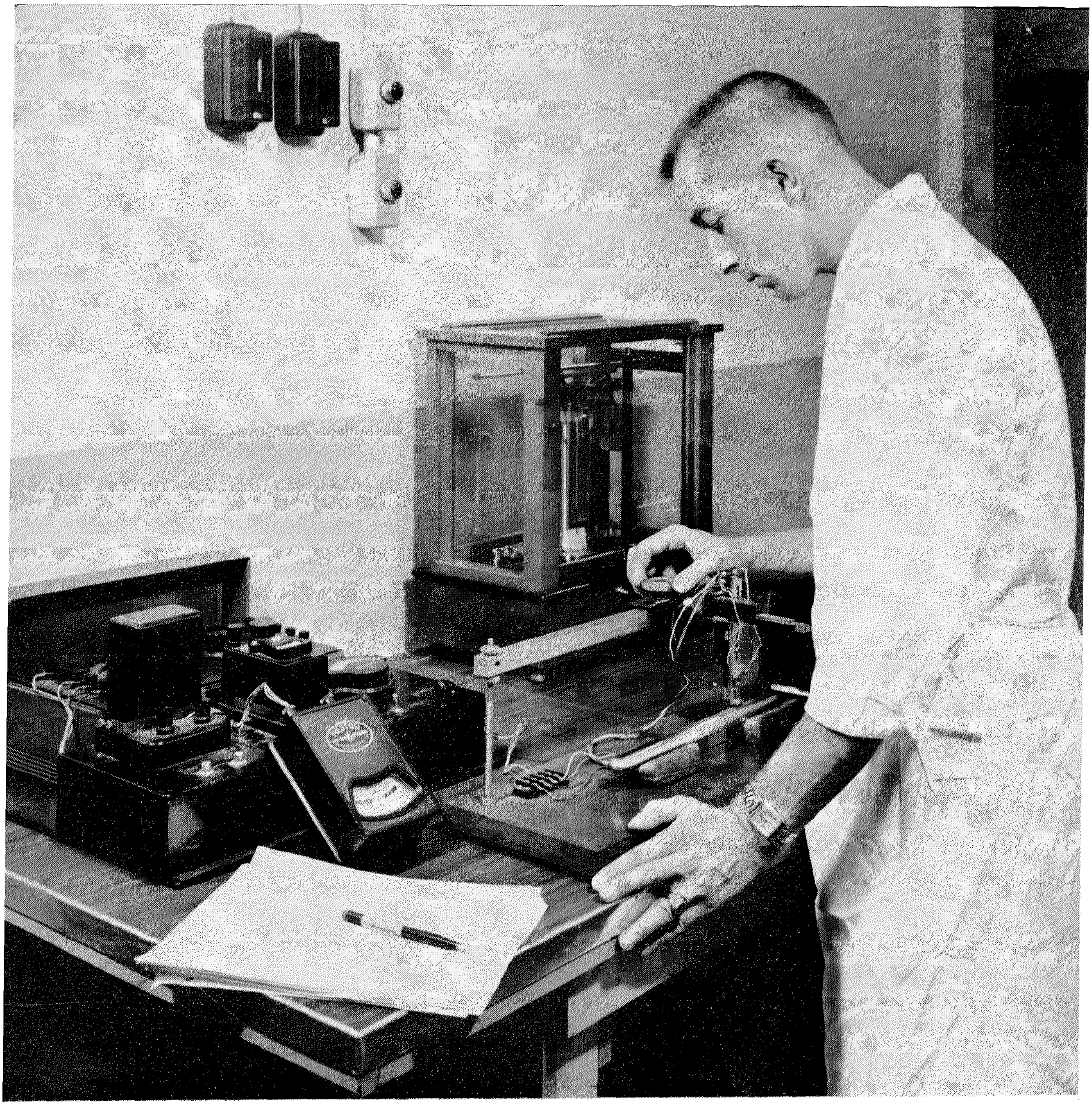
2. The first step in the production of the germaniu plates for the rectifiers is shown below. Extremely pu germanium dioxide is weighed in a graphite boat. Ter perature and humidity are very critically controlled and a monitored by the instrument in the left background.

3. In this electrically operated furnace, the germanium dioxide is heated in an atmosphere of hydrogen. The dioxide is reduced to molten germanium; the oxygen combines with the hydrogen and is carried out of the furnace as water vapor. The engineer checks the recorder on the automatic temperature-controlling instrument.

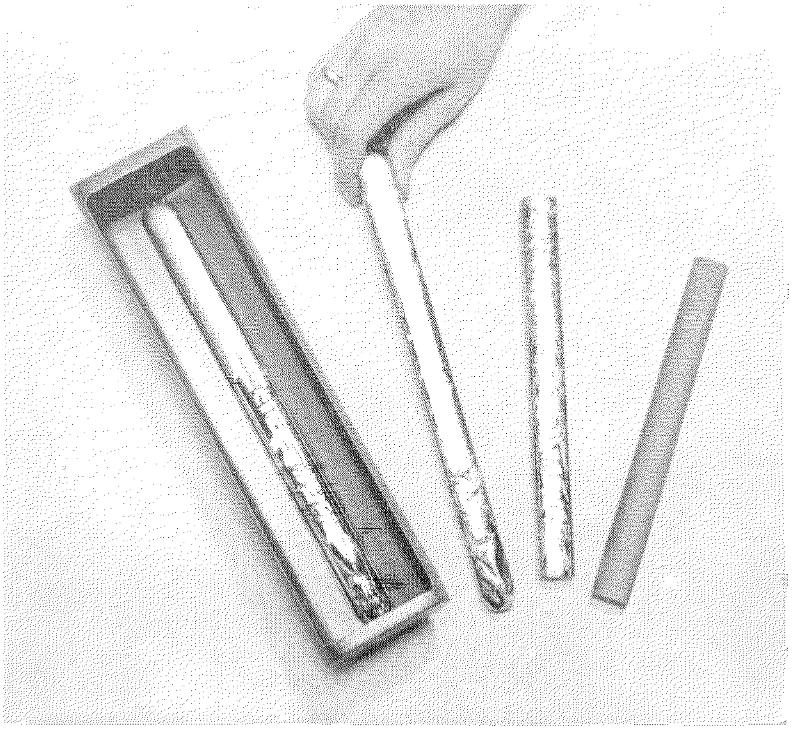


4. Successful production of germanium diodes requires very careful control of impurities in the germanium. Here is the zone-refining device. The germanium bar obtained after reduction in step 3 is placed in a graphite boat and very slowly drawn through the fused-quartz tube. The 6 coils, being energized by a high-frequency induction-heating generator, heat narrow zones of the boat to a red-hot temperature. As the boat is drawn through, each portion of the germanium bar in the boat is therefore melted and resolidified 6 times successively. Since impurities in the germanium are more soluble in molten germanium than in the solidified state, they are swept into the endmost portion of the bar, which is later cut off with a steel diamond-charged circular saw. The remaining bar consists of extremely pure germanium.



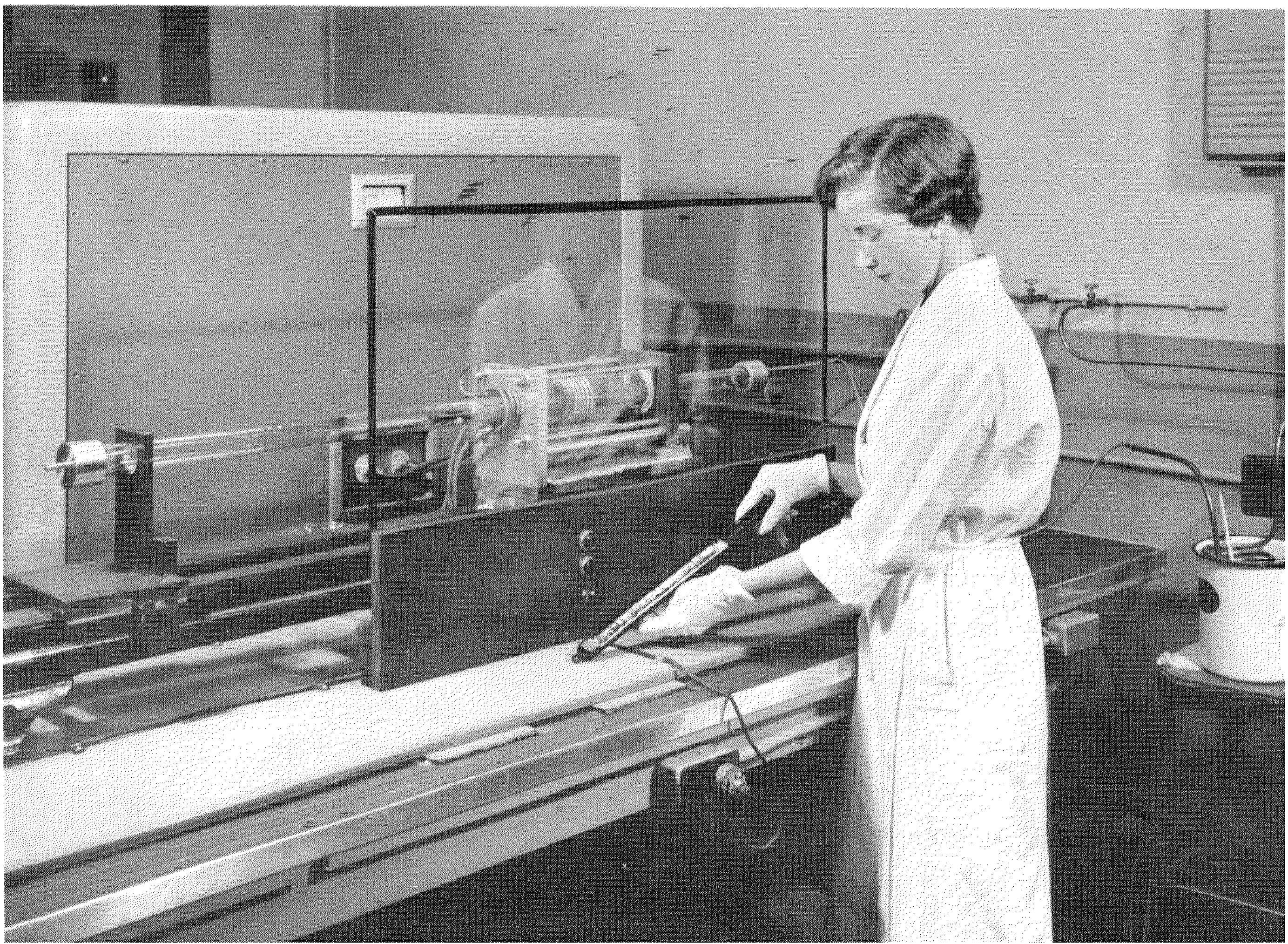


5. The electrical resistivity of germanium varies with the amount and type of impurities remaining after zone refining. Resistivity along the bar is measured accurately by means of the above device and portions having excess impurity can be removed and returned for further refining. This permits an accurate determination of how far from the end the bar must be cut off.



6. With an induction-heating device similar to that us
in zone refining, but having only one heating coil, the b
of germanium is melted in a narrow zone and as the bo
moves, it is allowed to recrystallize as a single large cryst
The worker below holds a boat used in the process. In t
front of the boat is loaded a small piece of monocrystalli
germanium. Between it and the long bar of pure ge
manium is placed a very small quantity of specially select
impurity that will give the germanium its rectifying pro
erty. Melting is started at this junction and as the melt
zone is swept along the length of the bar, the impurity
uniformly distributed throughout the bar. As the germa
ium resolidifies behind the molten zone, it conforms to t
crystalline pattern of the small piece of single-cryst
germanium; the whole bar thus becomes a single cryst

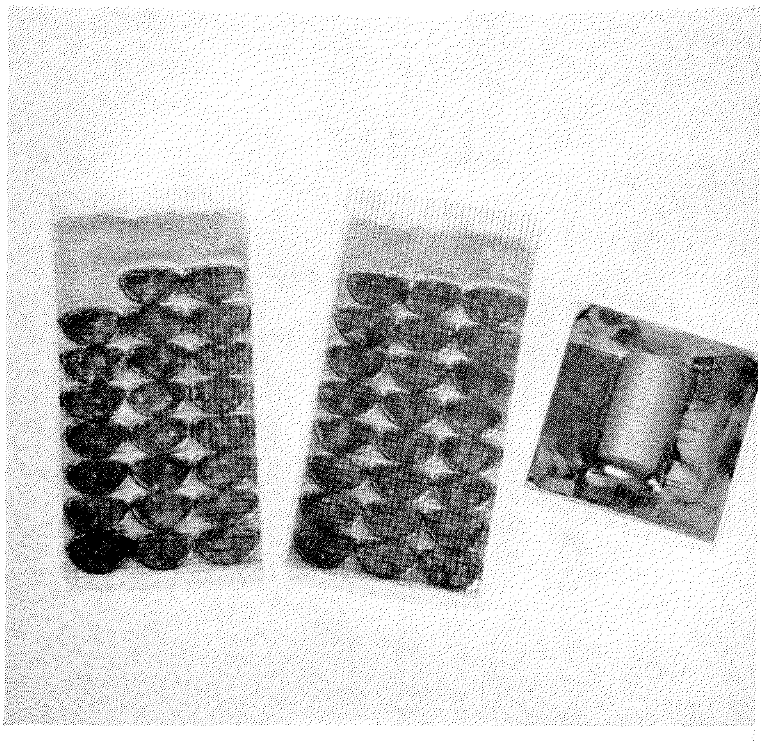
7. The stages in the production of a germanium bar a
shown at left. In the boat is a bar of germanium as reduc
from the oxide. The hand holds a reduced bar that h
been etched to bring out its polycrystalline state. Next

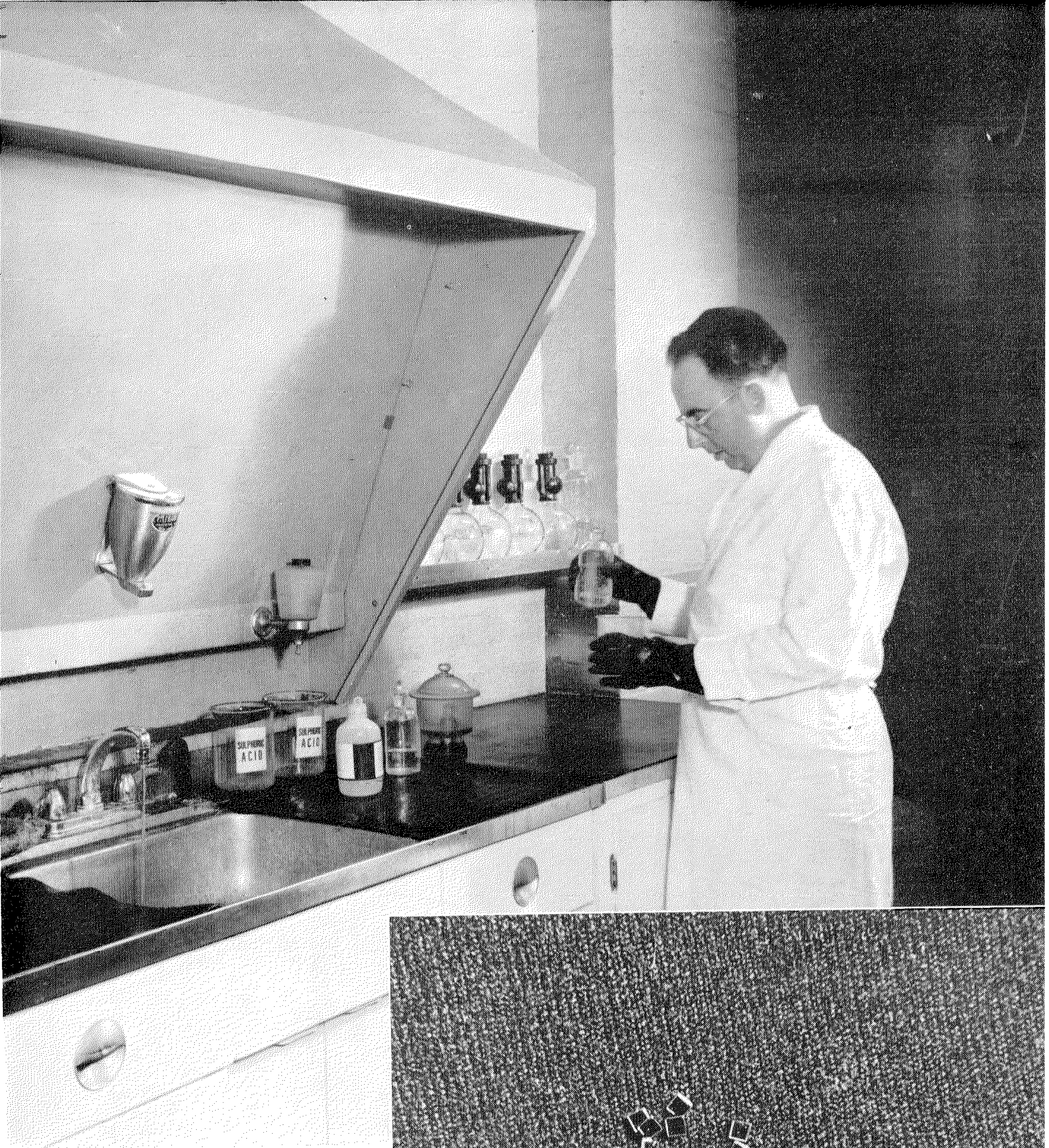


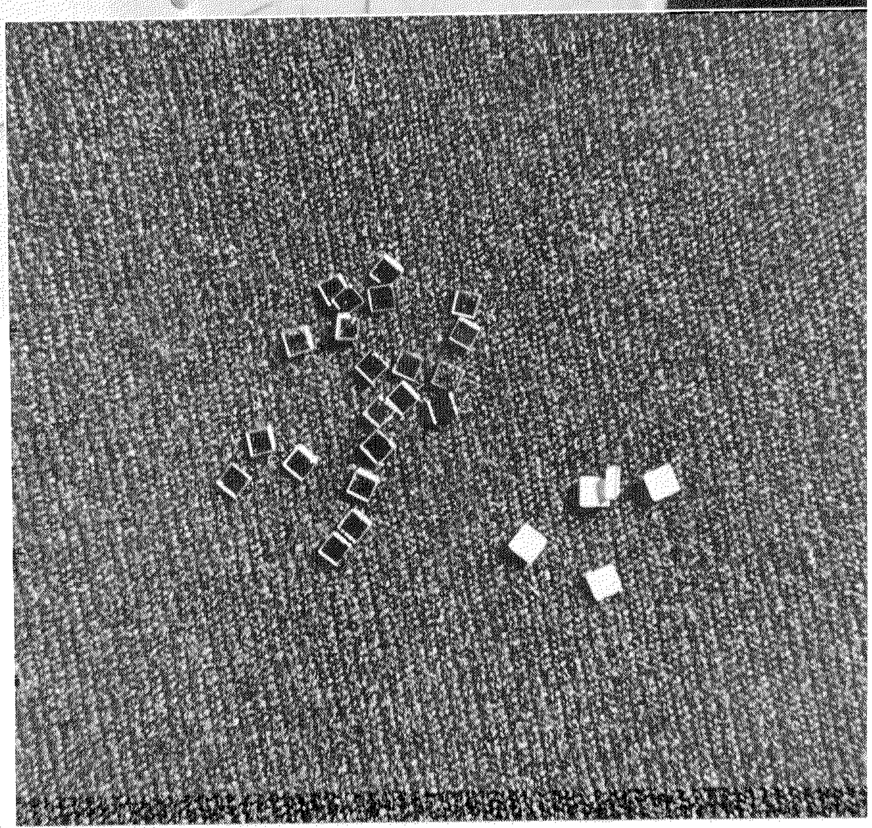
a bar after zone refining with the impure ends sawn off; this is also polycrystalline. Last is an etched bar of single-crystal germanium. After step 6, the bar is etched to remove all surface impurities and to prove its monocrystalline state; note the uniform matte surface.

8. Above are shown two machines used in sawing the germanium bar into plates that will be assembled into the diodes. At the operator's left is a saw that cuts the bar transversely into slices about 0.02-inch (0.5-millimeter) thick. These are then lapped to exact thickness and cemented with shellac onto a ceramic plate. In the machine in front of the operator, the slices are sawn into squares about 0.08-inch (2-millimeters) square. Both saws use diamond-impregnated steel disks.

9. At the right is a piece of bar after slicing and the two plates show the slices cut into squares.

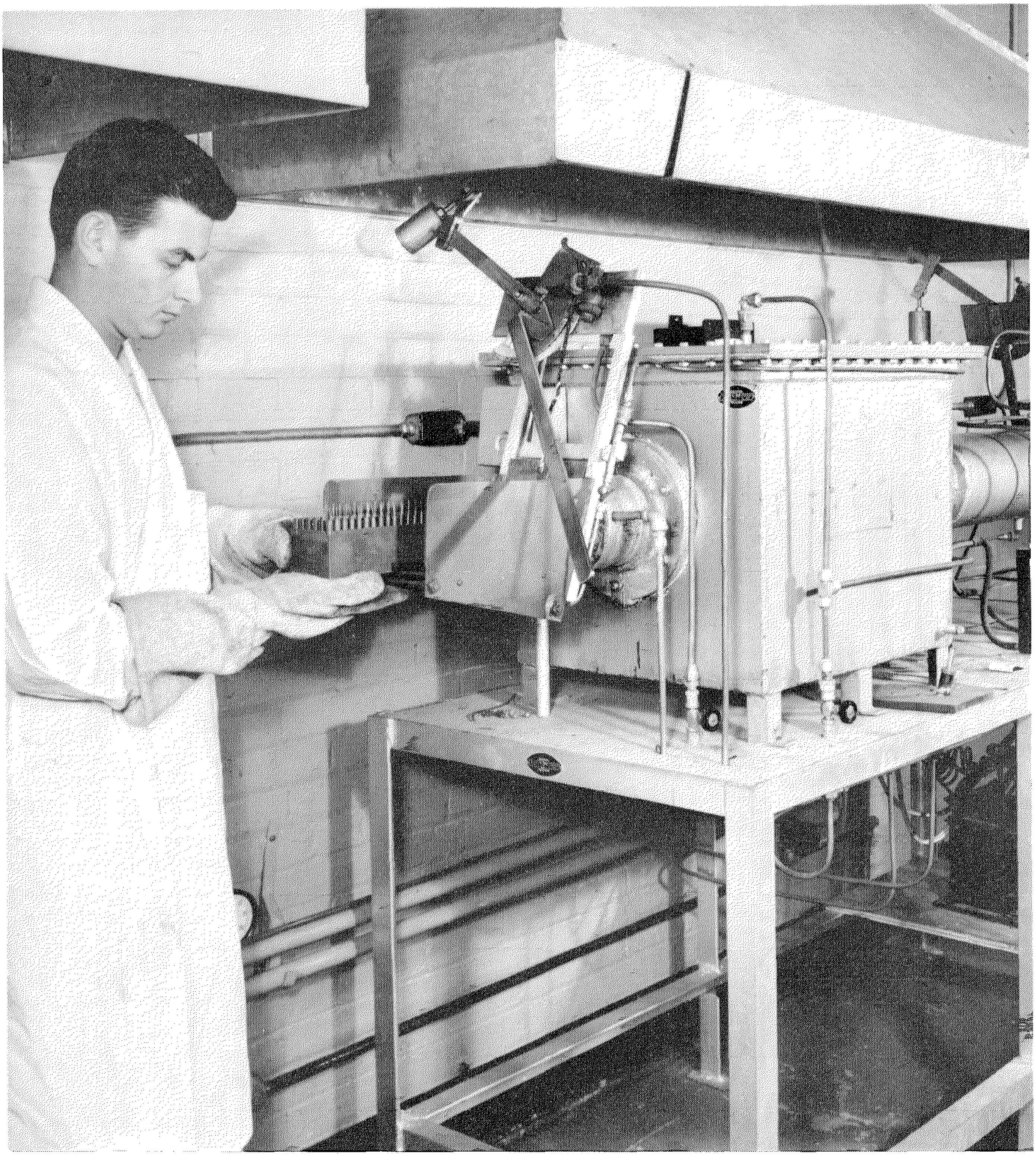




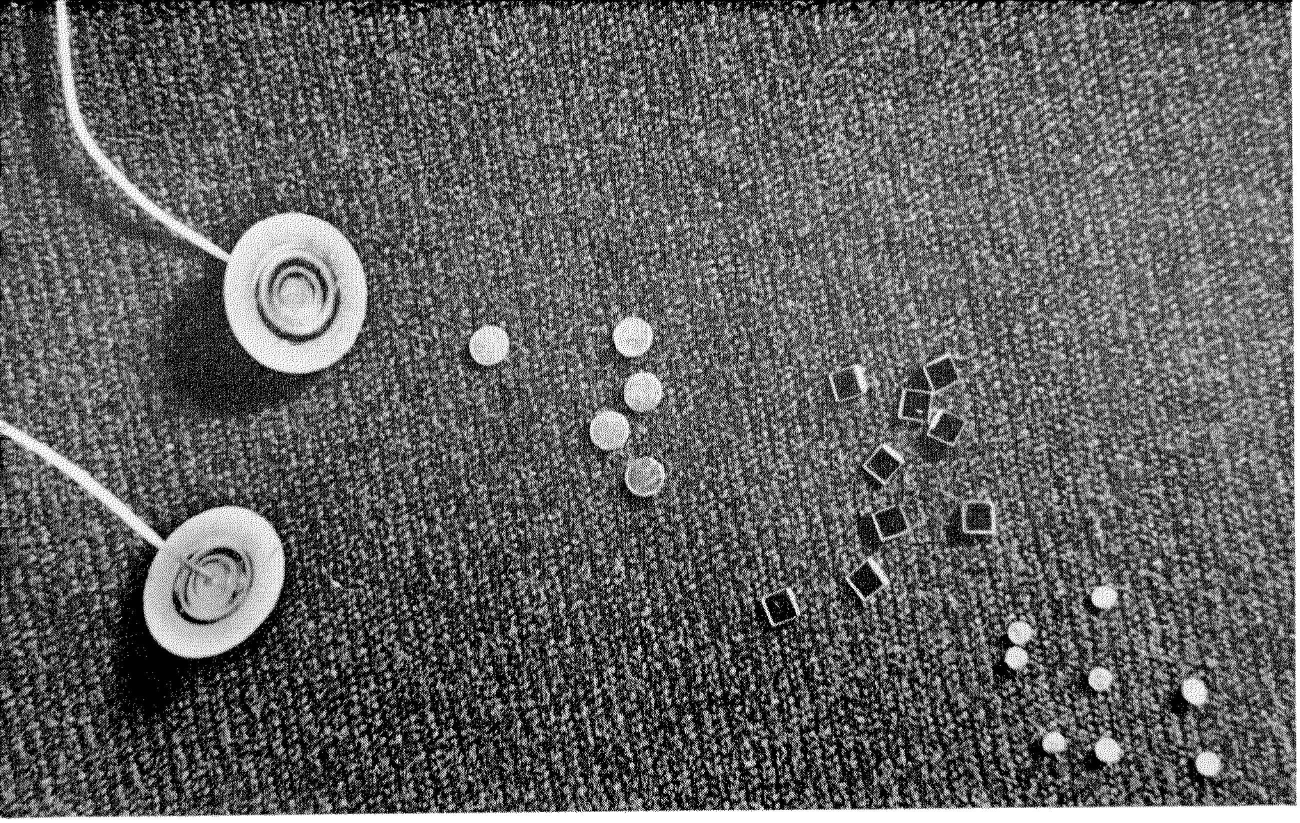


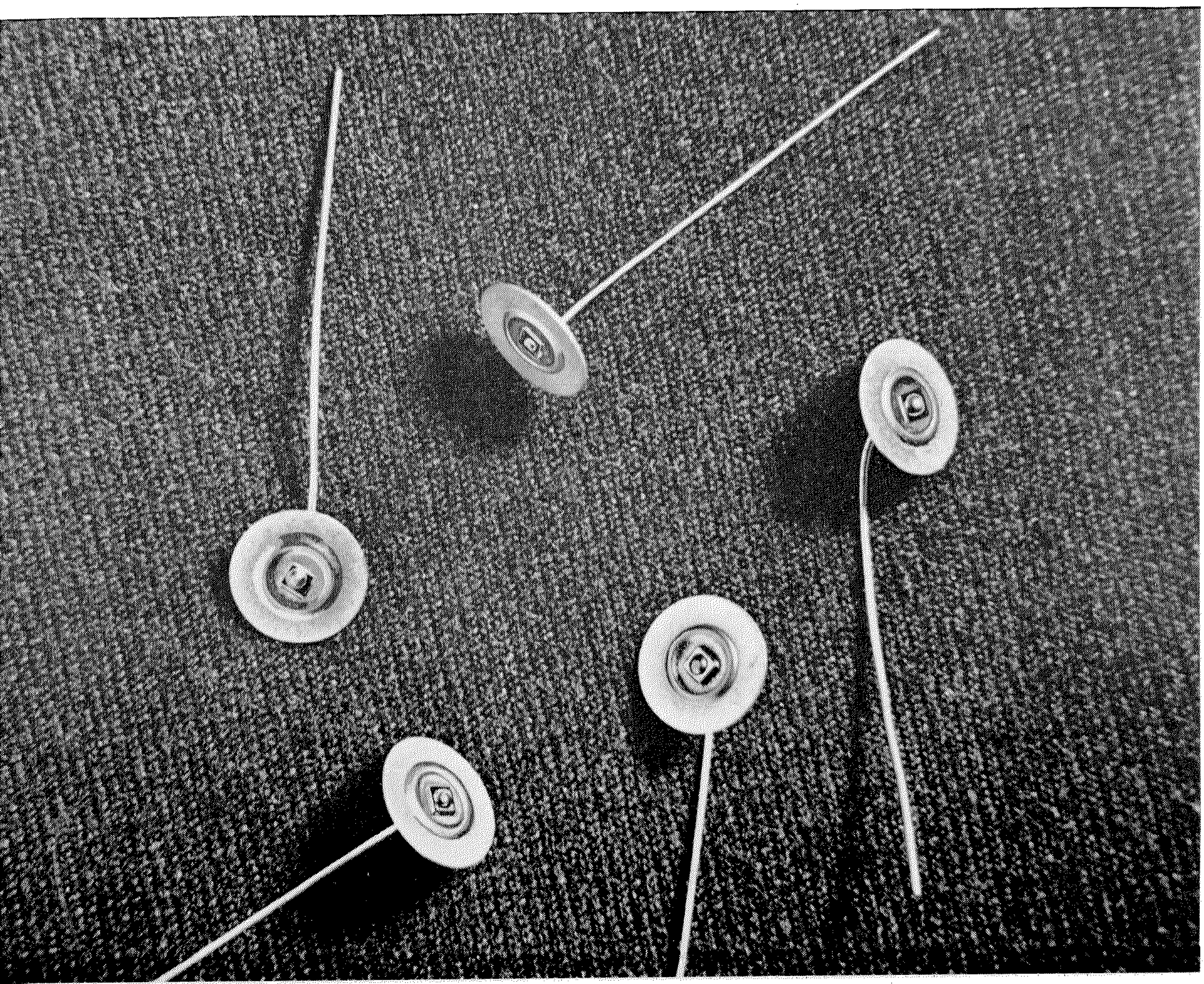
10. On the opposite page, lapping is done on a glass plate until the slices measure correct thickness on the gauge at the right.

11. After sawing, the plates are released from the ceramic plaque and are next cleaned by chemical etching as shown above. At the right, the rough bright surface of the plates are reduced to a smooth shining black by the etching.



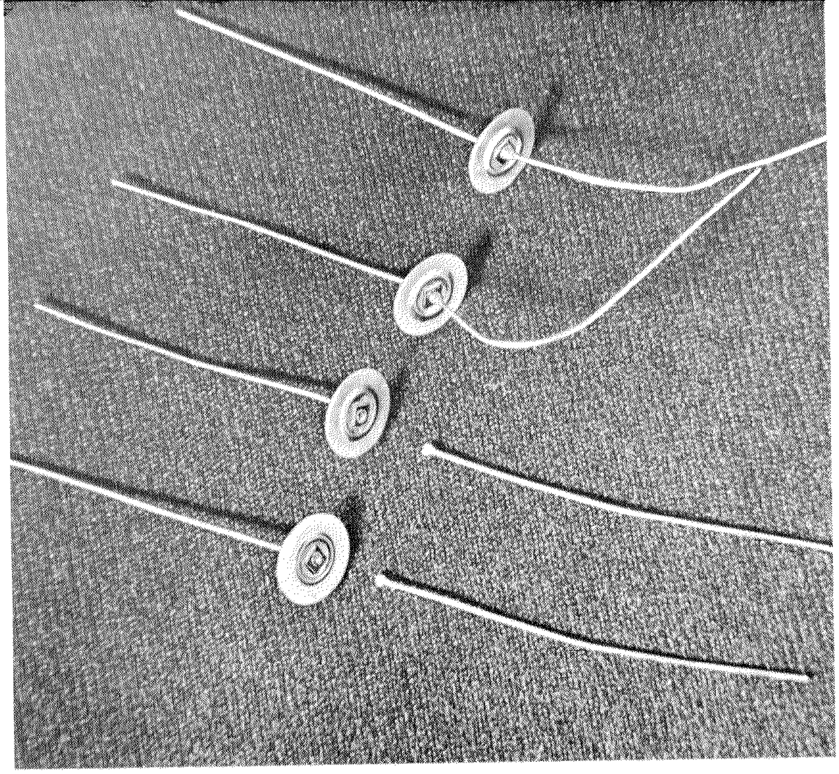
12. The frontispiece of this issue shows an operator loading a graphite boat with parts preparatory to the diffusing process. The parts are shown at the top of the facing page. From left to right are the bases, solder preforms, germanium plates, and indium pellets. Diffusion is done in a hydrogen-atmosphere electric furnace. The rectifying junction is formed at this stage, pictured above. At the bottom of the facing page are some diode bases after the diffusion stage. The small round indium pellets are partially diffused into the square germanium plates.





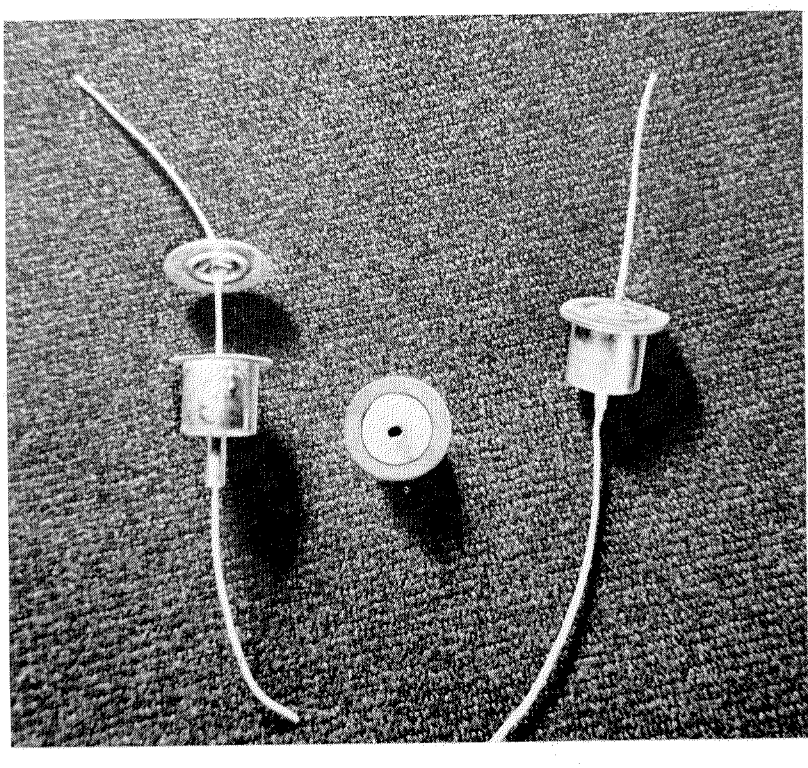
13. A pretinned nickel upper lead is fused in the indium electrode with the jig below. A preliminary measurement of forward resistance is next made to eliminate defective diodes before further operations. The process is illustrated below and at the top of the facing page.

14. In the apparatus shown below, the diodes are given their final cleaning by electrolytic etching. Noxious fumes are carried off by the hood and flues.

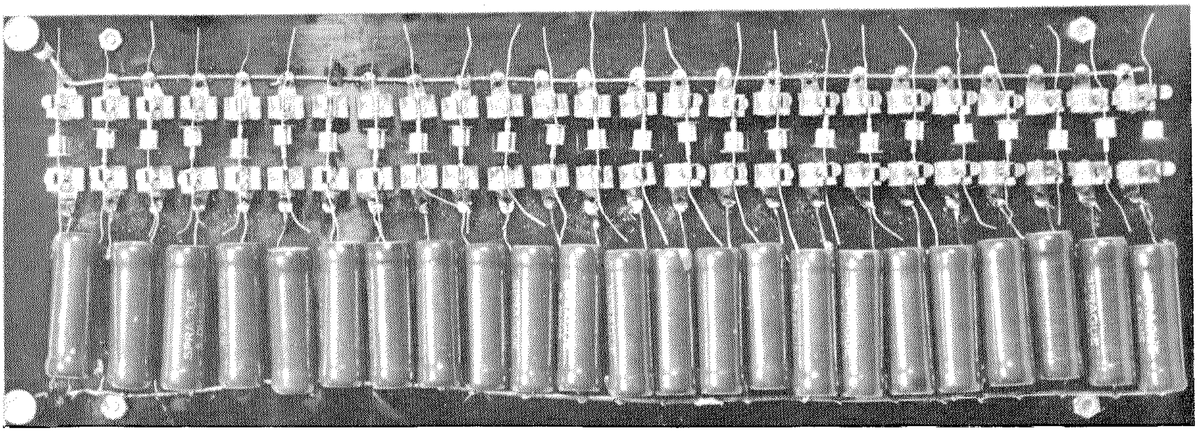


15. Characteristics of the diodes in the forward- and reverse-current directions are made at the two positions shown here.

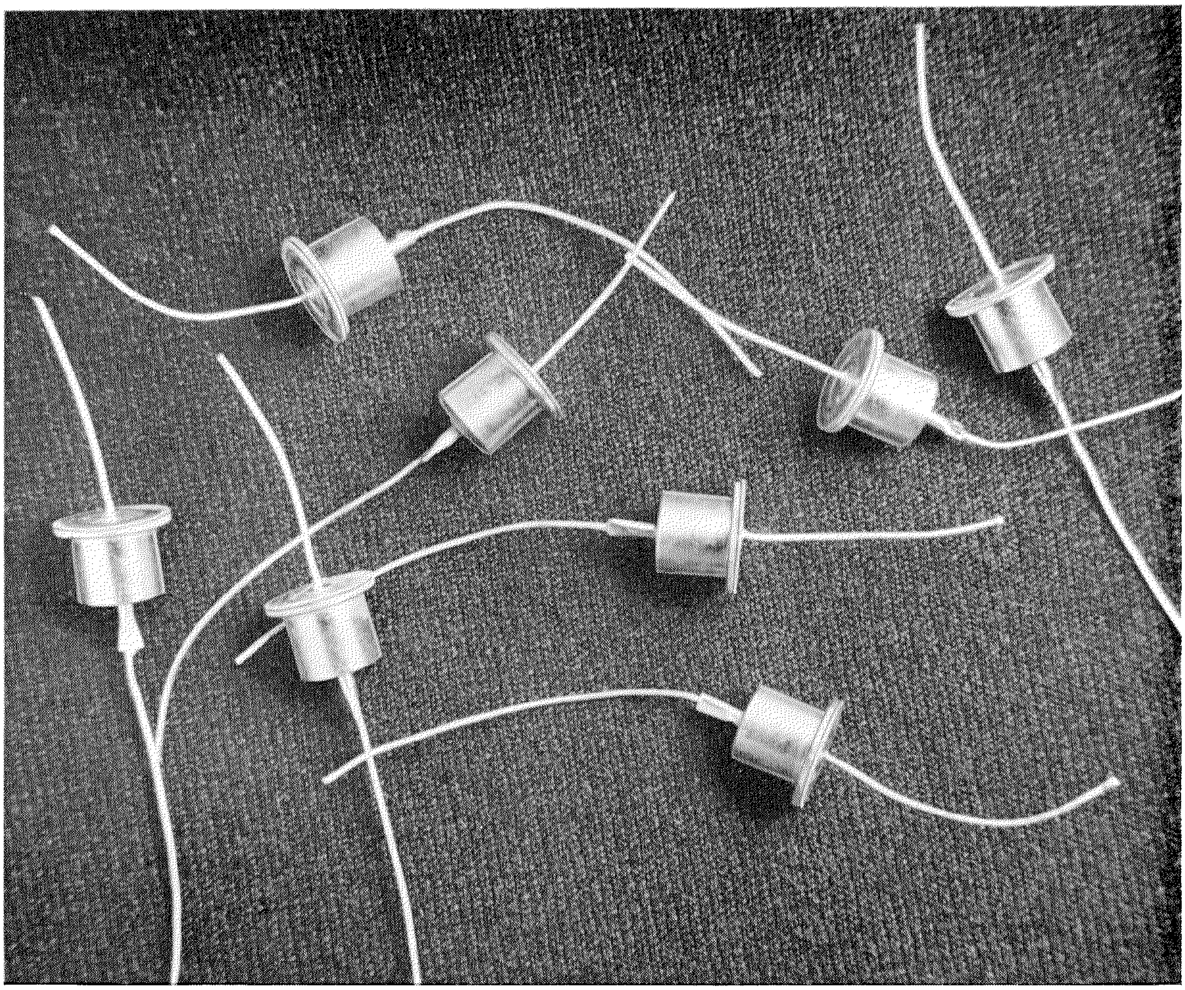
16. After testing, the active parts of the diodes are covered with a protective compound to prevent the adverse effects of moisture and other contaminants and the caps are placed over them. The caps are welded to the bases in the above machine to complete the hermetic seal. At the right is shown, left to right, the cap being slid down the upper lead wire, top view of a cap showing the insulating ceramic seal, and a welded diode. The tube is next mash-welded to the upper lead wire in a machine similar to the one above.



17. After capping, the diodes are plated for appearance and for protection against corrosion and are given several more tests. Shown here is the position for measuring the peak inverse-voltage capabilities of the diode. Two more of the testing positions are shown on the facing page. In the foreground is the preliminary forward-resistance measurement mentioned in step 13. In the rear, the diodes are measured for reverse leakage current under rated operating conditions. During this test, the diodes are placed in an oven to simulate the ambient heat to which they may be subjected during use.



18. One of the panels on which diodes are aged under operation to improve their characteristics is shown above. Finished diodes are illustrated below in about twice actual size. These will be trademarked and stamped with a type number before packaging and shipping to the customer.



Polythene-Insulated Video-Pair Cables

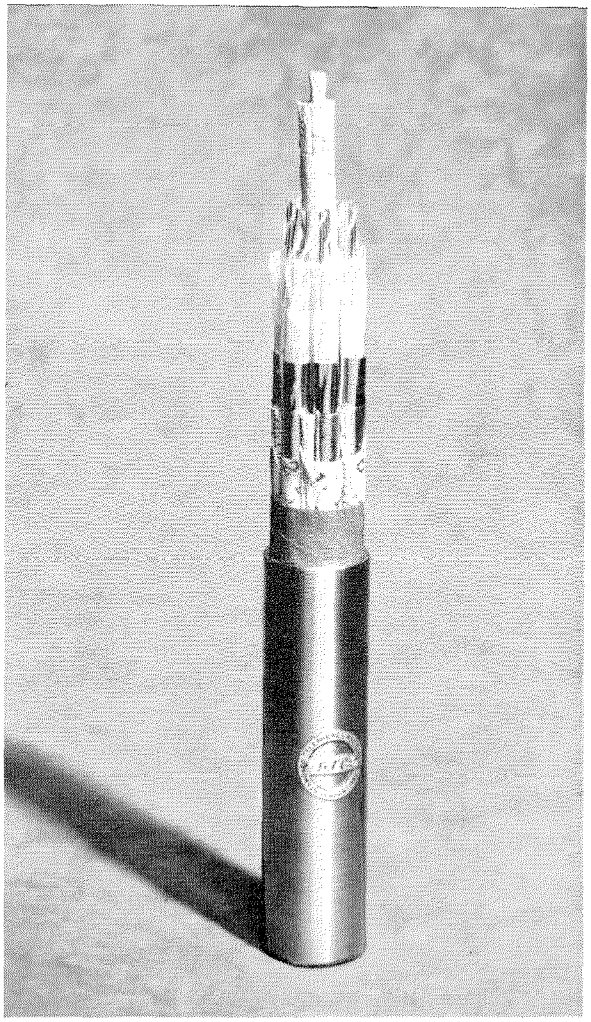
DIRECT transmission of video-frequency signals, consisting of frequency components from a few cycles per second to several megacycles per second, requires circuits that have low attenuation and good crosstalk and impedance-uniformity characteristics, but that can, if required, be connected directly to existing circuits. Conventional paper-insulated pairs are not satisfactory for video-frequency transmission over long circuits because their high shunt attenuation necessitates short spacing of the amplifier equipment and because their inadequate screening results in excessive crosstalk and interference.

Conventional coaxial cables are also unsatisfactory for direct transmission of video signals because of the ineffectiveness of the screening at low frequencies and the need for special apparatus when jointing them to paper-insulated pairs.

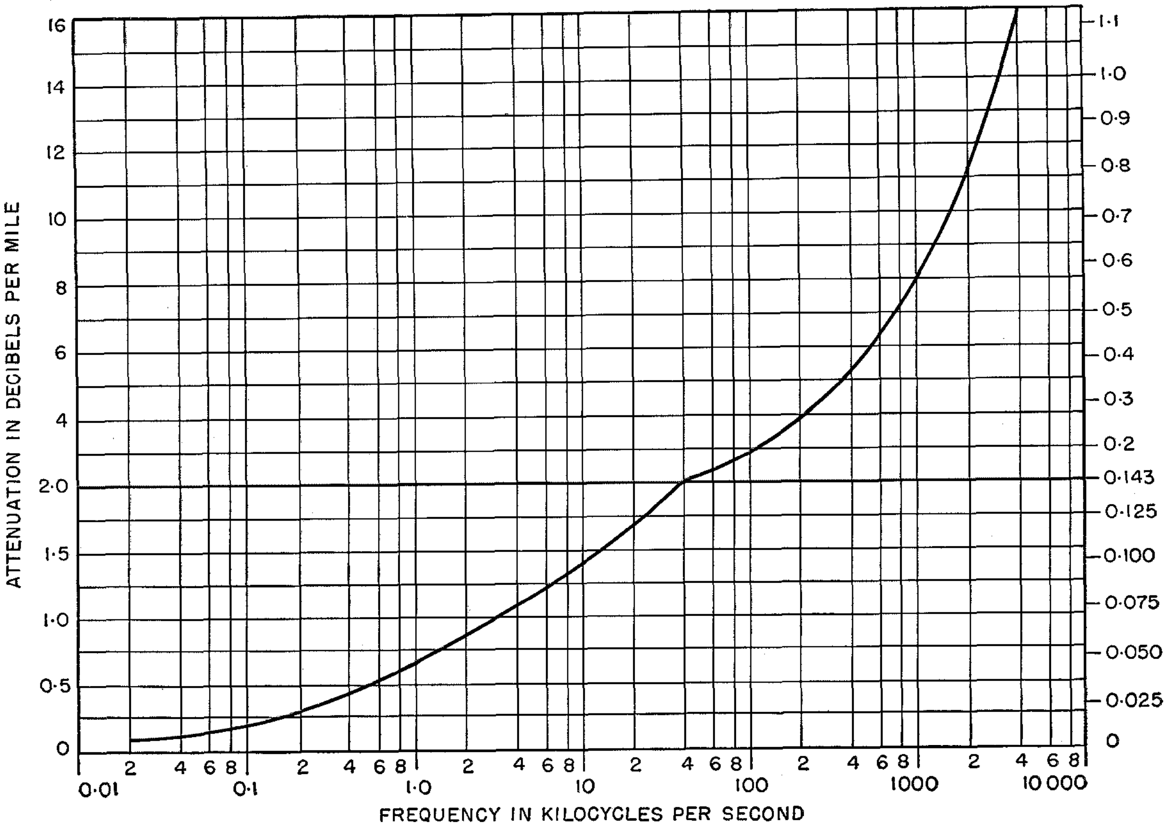
The cable required, therefore, is of a screened-pair construction with a dielectric of low permittivity and power factor, operating from about 20 cycles per second to 5 megacycles per second. A screened video pair with polythene string and tape replacing the paper string and tape of the conventional paper-insulated-conductor type is now manufactured by Standard Telephones and Cables, Limited, London, to satisfy all of these requirements. The design of this pair has the advantage that it can be manufactured on existing paper-insulated-cable plant with only minor modification.

The insulation of the new pair consists of a polythene string applied as an open helix on the conductor, followed by a closed helical wrapping of polythene tape, both applications being made in one operation on a paper-insulating machine. Two such insulated conductors are then twisted together with two fillers consisting of twisted polythene strings in the outer interstices to form a round unit. The pair so formed is lapped with polythene tape, then screened, first with a longitudinally applied copper tape, then with a helically applied copper tape; and finally lapped with two paper tapes, the outer one of which is marked with an identification number. The overall diameter of the pair is approximately 9 millimetres (0·35 inch) for a 1·27-millimetre (0·05-inch) conductor, giving optimum spacing between conductors and screen for minimum attenuation.

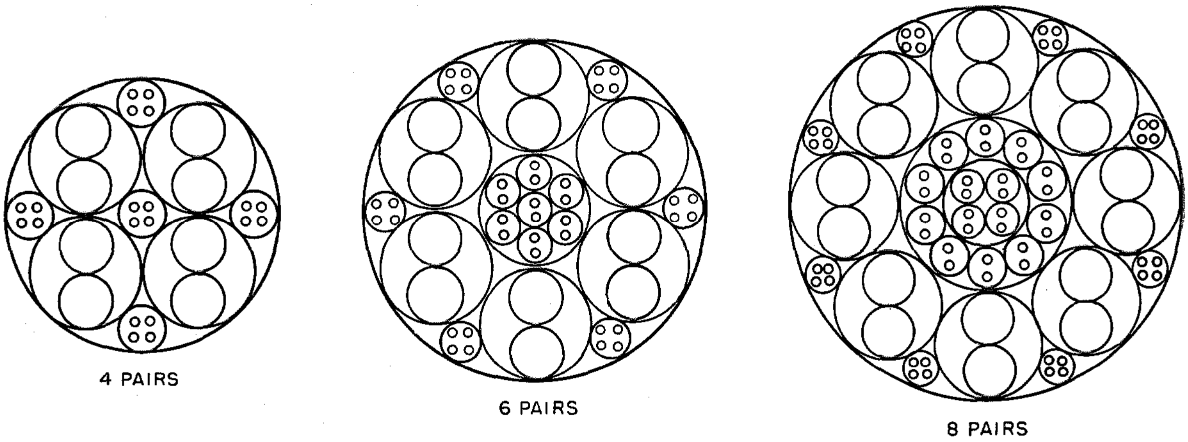
The above design keeps crosstalk to the low values required for the transmission of video-frequency signals. The longitudinal crosstalk due to electric-field coupling between the wires of the pair and their screen depends on the difference in capacitance between each wire and the screen (side-to-earth unbalance) and on the coupling to the pair of a voltage induced on the inner



Cable containing 8 polythene-insulated video-frequency pairs. There are also 8 star quads and 14 pairs in the interstices.



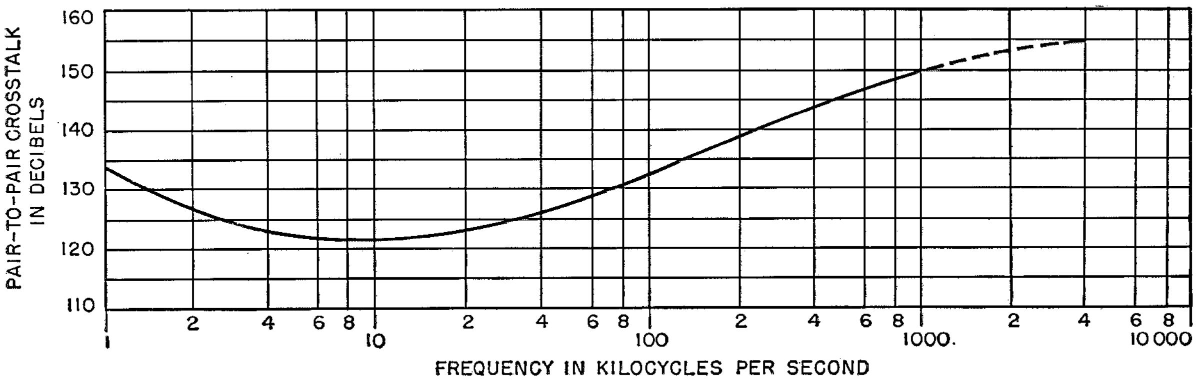
Attenuation of video-pair cable versus frequency; temperature was 60 degrees fahrenheit (15 degrees centigrade). Note expansion of scale below 2 decibels per mile.



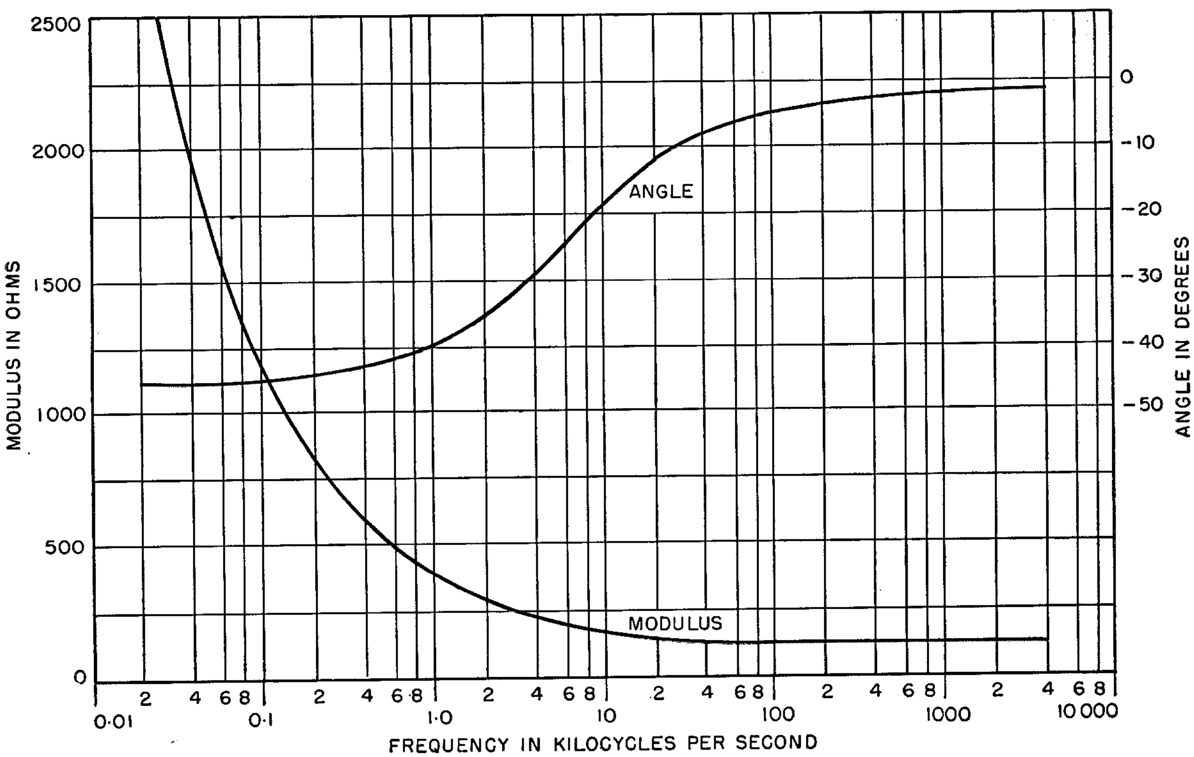
Three typical cable designs incorporating polythene-insulated video pairs. The 4-pair unit has a star quad in the centre and 4 outer-interstice star quads, all with 0·63-millimetre (0 025-inch) conductors; overall diameter is 22 millimetres (0·87 inch) and the weight is 2150 kilograms per kilometre (7630 pounds per mile). The 6-pair unit has 7 pairs in the centre with 0·9-millimetre (0·035-inch) conductors and 6 star quads with 0·63-millimetre (0·025-inch) conductors in the outer interstices. Diameter is 27 millimetres (1·06 inches) and weight, 2940 kilograms per kilometre (10 434 pounds per mile). The 8-pair cable has 14 pairs in the centre with 0·9-millimetre (0 035-inch) conductors and 8 star quads with 0·63-millimetre (0·025-inch) conductors in the outer interstices; diameter is 32·5 millimetres (1·28 inches) and weight is 3850 kilograms per kilometre (13 664 pounds per mile).

surface of the screen from an interference current on the outer surface. This voltage depends on the surface impedance of the screen and for this reason the first copper tape of the video pair is applied longitudinally to give less surface impedance than is obtained from a spiral lapping. With the latter, at the higher frequencies, the current is affected by the contact resistance between turns. The side-to-earth unbalance is kept at a minimum by careful control of the materials comprising each pair and by uniformity of manufacture. Transverse crosstalk due to magnetic coupling between the pairs is reduced by suitable selection of the pair twists and by the adequacy of the shielding.

An alternative insulation for the video pair

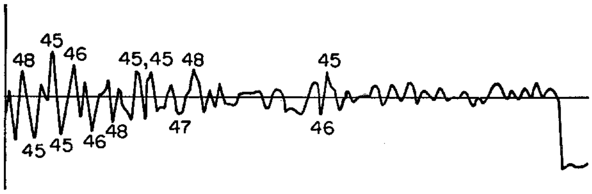


Typical pair-to-pair crosstalk curves plotted against frequency.



Impedance-versus-frequency curve for video-frequency cable.

is expanded polythene. A dielectric of approximately 50-per-cent polythene and 50-per-cent air gives approximately the same attenuation as the polythene string-and-tape insulation.



Pulse-reflection oscillogram illustrating impedance uniformity of the vido-pair cable. The transmitted pulse length was 0·05 microsecond. Cable length was 500 metres (547 yards). The values of the major line reflections (corrected for line attenuation) are indicated in decibels below transmitted-power level.

Any reasonable number of video pairs may be stranded together with their associated control and supervisory circuits to form a composite cable around which paper-insulated pairs or quads may be stranded if required. The cable is then lead-sheathed and armoured in the normal manner.

The accompanying illustrations give the major characteristics of this new type of cable construction. The cable indicated in the curves had 1·27-millimetre (0·05-inch) conductors.

Completed installation showing lifts with intercommunication panels immediately above the lift-operation controls.

Loudspeaker Intercommunication System For Catering Service at London Airport

By J. L. GOODWIN

Standard Telephones and Cables, Limited; London, England

THE South East Face building of the new Central Terminal was designed to handle the bulk of the passenger traffic at that London airport and is completely equipped with full catering facilities. The planning of a modern kitchen invariably divides it into a number of areas devoted individually to such services as food preparation, wine store, and wash-up for tableware. A flexible communication system is required not only among these areas but must include the service lifts or elevators that are part of the food-handling system.

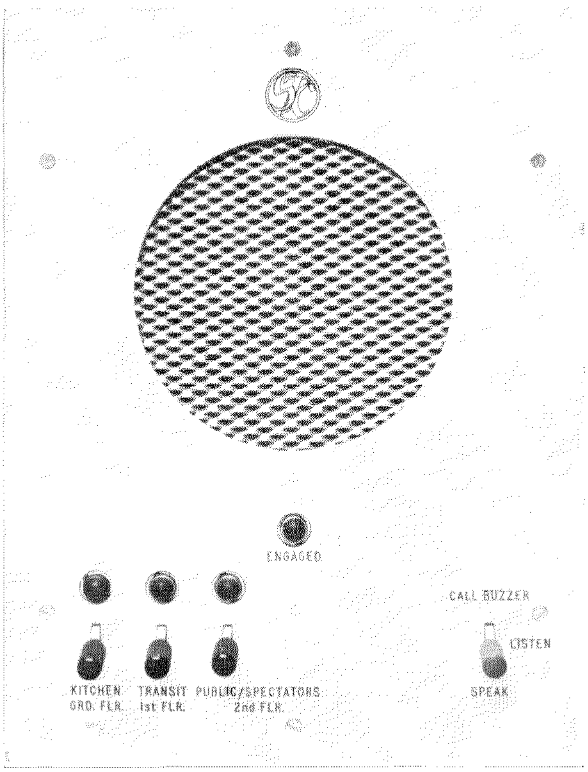
In this particular installation, a loudspeaking telephone having a centralized master station and a number of extensions would not do, for it would not be practicable to nominate one area as the master to which all calls would have to be made. A requirement was that any pair of side stations should be able to talk to one another.

The installation includes a bank of 10 food service lifts and 2 others, all of which are unattended by operators and do not carry personnel. The lifts serve a maximum of 4 floors and in some instances there is access to the car from two opposite sides. An example of this is on the first floor where the transit restaurant is on one side and the wash-up room is on the other.

To cover this installation, 7 loudspeaking telephone systems were required, each system having from 2 to 6 extensions. The design was based on the familiar "talk–listen" principle with the basic circuit and terminals the same for each system. This standardization resulted in manufacturing and installation economies and in simplified maintenance.

In each system, a common amplifier permits transmission between 2 of the extensions at a time. A small loudspeaker at each station also serves as the microphone and relays are actuated to connect the amplifier in the proper direction for speech transmission from the station when the telephone-type key is in the "speak" position. When the key is released, it returns to the "listen" position.

The input and output terminals of the amplifier appear on opposite sides of the same spring set, and whilst this would generally be undesirable when using a high-gain amplifier with a wide frequency response, it is practicable with the amplifiers used for these systems. The gain is of the order of 70 decibels and the high-frequency response is attenuated. In more elaborate systems it has been found necessary to utilize special switches that usually result in the signal passing through a greater number of contacts; if an attenuated high-frequency response can be accepted, that is the better solution to adopt.



Typical control panel for a 4-extension system.

For the convenience of working over long lines and of finding a suitable impedance for transmission and reception, the amplifier is designed to work between impedances of 600 ohms. The transducer units of a nominal 3-ohm impedance are fitted with transformers to convert their impedance to 600 ohms.

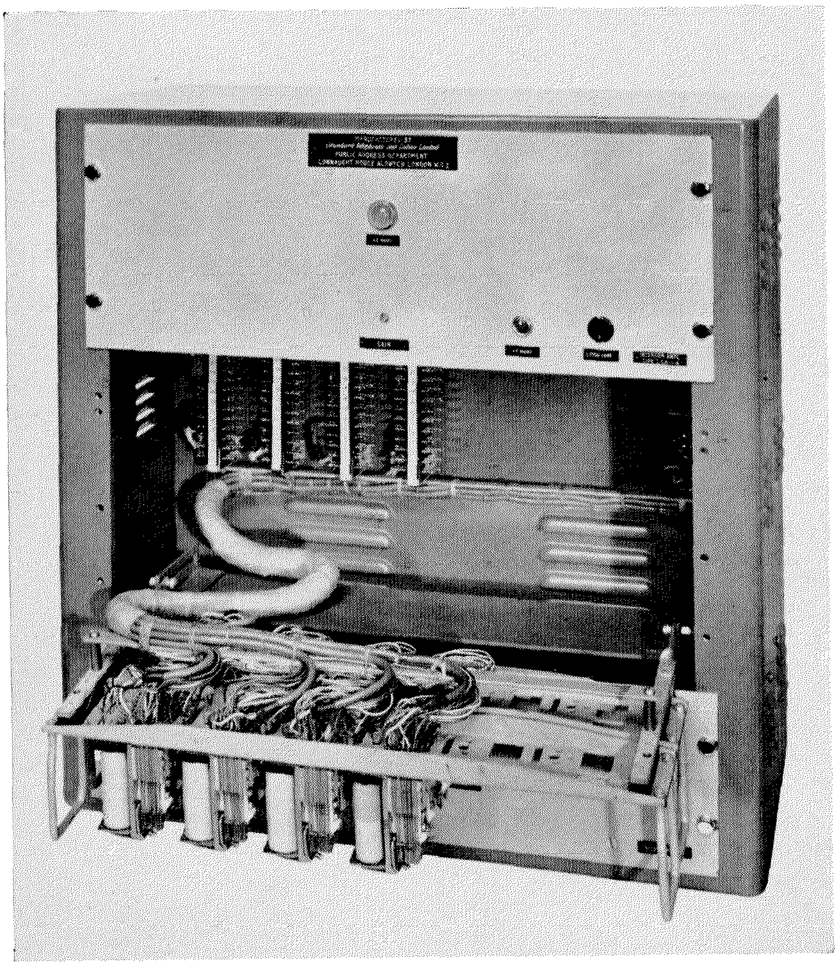
Calling may be effected by using voice or by operating the "Talk/Listen" key in the upward direction to the position "Call Buzzer." Contacts on a relay operate the buzzer and connect its pulsating output through a suitable attenuator to the input of the amplifier. By this means it is possible to employ a central buzzer, using the amplifier to produce the required volume, instead of local buzzers at each extension sounding from within the panel housing.

In common with other intercommunication systems, the extension under call is controlled by the party in the calling position and the called party must therefore pause momentarily before replying to allow time for the throwing of the switch by the caller.

Very simple lamp circuits are used to indicate to the initiator of a call that the system has been seized and further circuits indicate to other parties that the system is engaged.

Bass attenuation has been applied to overcome any low-frequency resonances associated with the loudspeaker unit and its enclosure, and to reduce the effect of the acoustic feedback that can occur through the lift shaft from one extension to another. High-frequency attenuation is included to minimize the effects of response peaks in producing excessive sibilant tones. In addition to this, a decreasing gain with ascending frequency is desirable to ensure that the switched system is stable under conditions of maximum sensitivity.

The power output had to be sufficient to accept without distortion an input from a person speaking loudly or even shouting. A further consideration was that this system be capable of connecting to more than one extension at a time. Under normal operating conditions and for maximum performance, only two extensions are interconnected but it must be possible for the initiator of a call to select more than one extension. In the speaking condition, the amplifier output will be divided among the total number of extensions engaged by the initiator. In the listening condition, all of the called transducers become parallel-connected microphones with a resulting loss in sensitivity and an increased background noise. To meet these requirements a maximum output of 1·5 watts is available and trials have shown this to be a suitable value.



Apparatus cubicle with centre panel removed allowing the relay frames to be hinged forward for servicing and exposure of the terminal strip.

The amplifier, relays, and direct-current supply unit are assembled in a sheet-steel cubicle approximately 22 inches (56 centimetres) square and 10·5 inches (27 centimetres) deep.

The extension units are panel mounted and fitted into flush wall boxes. The panels may be mounted in any suitable enclosure. The flush wall boxes are fixed behind the lift architraves. This is convenient as the controls for the lifts are mounted in a similar manner just beneath the intercommunication panels, presenting a neat and pleasing appearance to these control units.

New Type of Diffusion Cathode *

By A. H. BECK, A. D. BRISBANE, A. B. CUTTING, and G. KING

Standard Telecommunication Laboratories, Limited; London, England

1. *Introduction*

VALVES today are required for telecommunication channels at higher and higher frequencies so the importance of cathodes yielding greater current density with reasonable life under continuous operation increases. Thus in recent years various types of cathodes have been produced which sacrifice emission efficiency for better continuous-service operational characteristics. Notable examples of these are the Lemmens cathode[1,2], the modified *L*-cathode using barium aluminate and tungsten[3] and the moulded cathode mentioned by MacNair, Hannay, and Lynch[4]. The cathode to be described combines some of the features of the barium-aluminate cathodes and of the moulded cathodes in that it may consist of a homogeneous mass of active material as does the barium-aluminate cathode but has as basic material mixtures of alkaline-earth carbonates and iron nickel or cobalt powder with addition of suitable reducing agents. During manufacture the mixed powders are compacted at high pressures, usually into metallic retaining cylinders so that handling is simplified. During decomposition of the alkaline-earth carbonates to oxides and alkaline-earth metal, the nickel sinters into a strong metallic matrix. A diagrammatic cross-section of several possible forms of the cathode are shown in Figure 1, variations of the percentage of nickel powder may be used to vary the porosity of the nickel matrix depending on the requirements for the cathode.

Since the emissive surface of the cathodes studied most was probably barium or barium oxide on nickel, the cathodes have been called somewhat loosely Bariated-Nickel or *BN* cathodes.

2. *Materials*

The iron or nickel powder used is obtained commercially as carbonyl powder by decomposition of the metallic carbonyl. It is very pure and consists of small almost spherical particles of 1 to 5 microns in diameter. The cobalt powder is obtained by reduction from the oxide. The particles are less regular in shape but are about the same size. The packing under pressure of the cobalt powder is rather less than that for the nickel or iron powders for the same pressure.

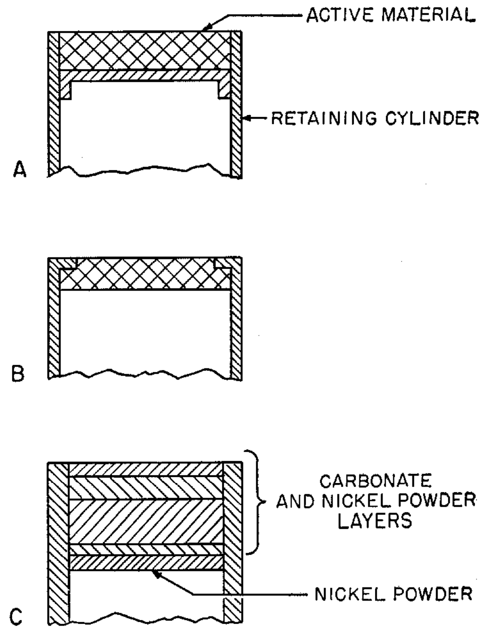


Figure 1—Mechanical construction of *BN* cathodes.

The metallic powders are washed, ball-milled with acetic acid, rewashed with distilled water, and vacuum stoved.

The alkaline earth carbonates (either barium carbonate, co-precipitated barium–strontium car-

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* Reprinted from *Le Vide*, volume 9, pages 302–309; November, 1954.

[1] H. J. Lemmens, M. J. Jansen, and R. Loosjes, "New Thermionic Cathode for Heavy Loads," *Philips Technical Review*, volume 11, pages 341–350; June, 1950.

[2] G. A. Espersen, "L-Cathode Structure," *Proceedings of the IRE*, volume 40, pages 284–289; March, 1952.

[3] R. Levi, "New Dispenser Type Thermionic Cathode," *Convention Record of the IRE 1953 National Convention*, Part 6—Electron Devices—Engineering Management, pages 40–41; 1953.

[4] D. MacNair, R. T. Lynch, and N. B. Hannay, "Molded Thermionic Cathodes," *Journal of Applied Physics*, volume 24, pages 1335–1336; October, 1953.

bonate, or barium–strontium–calcium carbonates) are those used in the preparation of normal oxide-cathode coating. Various reducing agents may be employed: tungsten, tantalum, titanium tend to produce excess barium evaporation, whereas carbon reacts very slowly with prolonged evolution of gas. Zirconium, usually in the form of zirconium hydride (the commercial name for a solution of hydrogen in zirconium), has given best results.

A large number of metals or alloys have been used for the retaining cylinder. With nickel or low-heat-conductive nickel alloys such as kovar, adhesion problems are a minimum. Molybdenum, used because of its low thermal emissivity, requires special treatment (that is, sintering nickel powder onto the part that will be in contact with pressed emission material) in order to ensure good adhesion between the cylinder and the nickel matrix.

The backing plate shown in Figure 1A is only used when heater–cathode insulation is a problem. It may consist of nickel or a nickel alloy, preferably not containing added reducing agents such as magnesium.

The metal cylinder and backing plate if used are hydrogen stoved before use.

3. Preparation of Cathodes

A mixture of 1 part by weight of reducing agent with 30 parts dry carbonate powder is made up. A powder containing 20–40 per cent of this mixture, the remainder being metal powder, is then made up and ball-milled in amyl acetate. The composition of the mixture is varied depending on the final cathode requirements. Large percentages of metal and high pressures give very robust cathodes with very little or no shrinkage on sintering the pressed mixture. Large percentages of carbonates give slightly enhanced emission but are less strong mechanically. Unless otherwise stated, results quoted for these cathodes will refer to the most extensively studied mixture containing roughly 30 per cent by weight co-precipitated double carbonate and 70 per cent nickel powder.

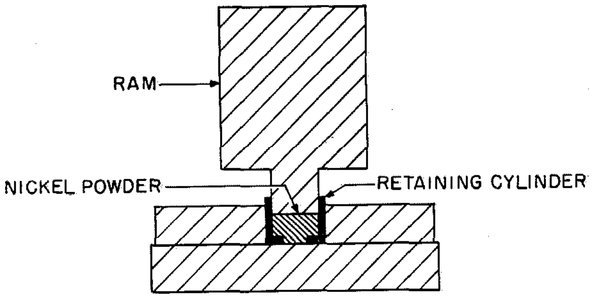


Figure 2—Press for compacting *BN* cathodes.

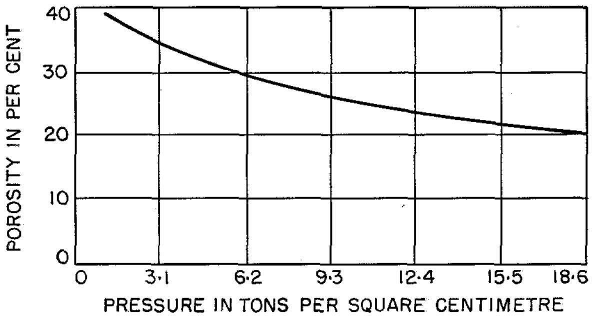


Figure 3—Porosity of pressed powders as a function of pressure.

The mixed, ball-milled powders are dried and compressed in the retaining cylinder as shown in Figure 2. The hardened steel blocks that form part of a hydraulic press have those faces that are in contact with the powder polished, both to prevent adhesion of the pressed powders and for ease of cleaning. The ring is also hardened and is used to prevent lateral expansion of the powder and retaining cylinder. This retaining cylinder may have a lip on its working face as shown in Figure 1B to improve the adhesion of the pressed powder. The backing plate if used is fitted on the nose of the hardened tool and pressed into place during the pressing operation. It is welded to the retaining cylinder after withdrawal from the press.

The pressure used to compact the powders may vary considerably with the composition of the mixed powders. A curve of applied pressure against porosity of the pressed mass of standard 30-per-cent-carbonate mixture is shown in Figure 3. For pressures greater than 50 tons per square inch the change of porosity and therefore change of linear dimensions on sintering is very small and pressures between 50 to 100 tons per square inch are used for the standard mixture and for most other mixtures—the pressures for mixtures containing large percentages of cobalt are, however, usually higher.

The cathode complete with retaining cylinder is ejected from the steel ring and is ready for mounting into a valve. If it is to be stored for a long time it is kept in a vacuum desiccator.

The pressed powders have a bright metallic surface, are easy to handle, and are sufficiently robust to allow a certain amount of surface machining, if required.

Although this description has only considered planar cathodes it is clear that initial shaping of the parts could produce cathodes of a large variety of shapes.

4. Activation

Decomposition of the carbonates and activation of these cathodes is very simple and follows very closely the schedule for standard oxide cathodes. The time required for pumping is very little greater than that required for an oxide cathode. In a typical instance the *BN* cathode can be outgassed in about ten minutes whereas any oxide cathode of similar dimensions would take two minutes. Emissions of 1 ampere per square centimetre are drawn while pumping.

After pumping the emission is normally stabilized by ageing the cathodes for 24 hours drawing 1 ampere per square centimetre. The cathode temperature is reduced from 1100 to 950 degrees centigrade during the first two hours of ageing and maintained at 950 degrees for the rest of the ageing time.

Since, for a plug of active material 1 millimetre thick, the mass of carbonate included is about 20 times the mass of carbonate on an oxide cathode of the same area, the amount of gas evolved during carbonate decomposition is considerable. In some cases it is therefore desirable to decompose the carbonates in vacuum before assembling the cathode in its final valve. Activated cathodes may be opened to air and stored in dry air for several days. On repumping the amount of gas evolved is very greatly reduced and reactivation follows the normal activation schedule.

After carbonate decomposition, the nickel matrix has sintered and the cathodes are mechanically very strong. The active material is easy to machine or polish and connections may be welded to it. Attempts to roll it into thin sheets have, however, not proved successful as the material becomes very brittle by the time it has been reduced to half its original thickness by rolling.

5. Emission Properties of BN Cathodes

In Figure 4 are summarized the pulsed and direct-current emission characteristics of *BN* cathodes as a function of true cathode temperature (that is, pyrometer temperature corrected for emissivity of the active material which, to within experimental error, is the same as nickel for nickel-powder-based cathodes). For comparison pulsed curves representative of good oxide cathodes and the published curves for Lemmens cathodes are shown.

The shaded area in Figure 4 represents the spread found in the pulsed emission results for fully activated *BN* cathodes. Also shown in Figure 4 is a line indicating the generally accepted practical values for the temperature of oxide cathodes for given continuous operating conditions. A similar line has been tentatively

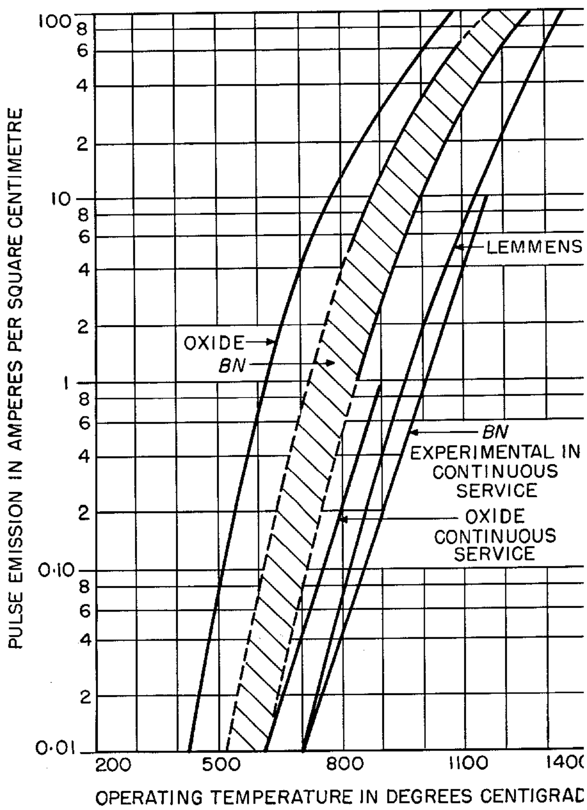


Figure 4—Emission as a function of temperature for oxide cathodes, *BN* cathodes, and Lemmens-type cathodes.

drawn for *BN* cathodes. The exact location of this line clearly depends on extensive life-testing. For the same emission density, *BN* cathodes require a temperature nearly 100 degrees centigrade hotter than the oxide cathodes. In compensation, the lives of *BN* cathodes are greatly prolonged. At 1 ampere per square centimetre and an operating temperature of 1000 degrees centigrade, lives of 5000 hours are obtained and at 1070 degrees centigrade and 3 amperes per square centimetre lives of 3000 hours are obtained.

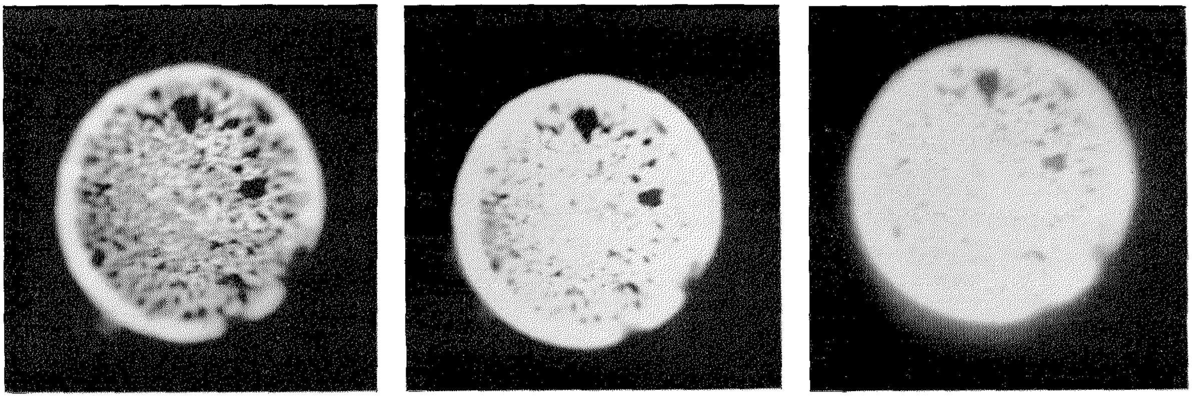


Figure 5—Images of *BN* cathode obtained by electron-microscope technique. Left, before activation. Center, during activation. Right, after activation.

As mentioned earlier, their recovery from oxygen contamination is remarkable and their resistance to poisoning in general is very high. A cathode operated in a pressure of air of about 10^{-4} millimetre of mercury gave an emission of 1 ampere per square centimetre for over 500 hours. *BN* cathodes have been extensively used in demountable vacuum systems with excellent results.

6. *Emission Mechanism*

There are two possible physical pictures of the emission properties of these cathodes: either the cathode is a normal oxide-cathode emitter emitting from small areas between the nickel powder but having a large reserve of free barium to give the required impurity centres for semi-conduction, or it is a genuine diffusion cathode with metallic barium or possibly barium oxide diffusing over the surface of the nickel giving a barium-on-nickel emitting surface. Experimental evidence seems to suggest that the latter is most probable.

Attempts to measure an interface resistance between nickel and oxide material has given no measurable resistance after several thousands of hours even though resistances of less than 5 ohms per square centimetre could be detected. In fact the resistance of the plug is so low that even at continuous emission densities of 3 amperes per square centimetre the cathode cooling due to electron evaporation is greater than the resistive heating and a drop of temperature of up to 20 degrees centigrade is observed on a standard life-test cathode. This in itself is not conclusive evidence as the large area of nickel-oxide surface contact and the volume of free barium would all tend to decrease the apparent interface resistance.

However a measurement of the velocity distribution of electrons at high current densities using the technique of Loosjes, Vink, and Jansen[5] showed no measurable (< 5 volts) distribution of electron energies from these cathodes. Since according to Nergaard[6] this is a property of the emitting surface of an oxide cathode rather than of the oxide–metal interface, the semi-conductor hypothesis seems unlikely.

Furthermore examination of the emitting area of the cathode using a cathode-ray tube as an elementary electron microscope and forming an

[5] R. Loosjes, H. J. Vink, and C. G. J. Jansen, "Thermionic Emitters under Pulsed Operation," *Philips Technical Review*, volume 13, pages 337–345; June, 1952.

[6] L. S. Nergaard, "Studies of the Oxide Cathode," *RCA Review*, volume 13, pages 464–545; December, 1952.

enlarged image of the cathode on the phosphor shows that the emitting areas of the cathode increase greatly during activation. Photographs of a cathode early in activation, partially activated and fully activated are shown in Figure 5 for a magnification of about 250. It is clear that diffusion of emitting material over the surface of the cathode is included in the mechanism of activation.

7. *Work Function and Surface Coverage*

In an attempt to clarify the emission mechanism the work function of a *BN* cathode has been measured in 3 different ways.

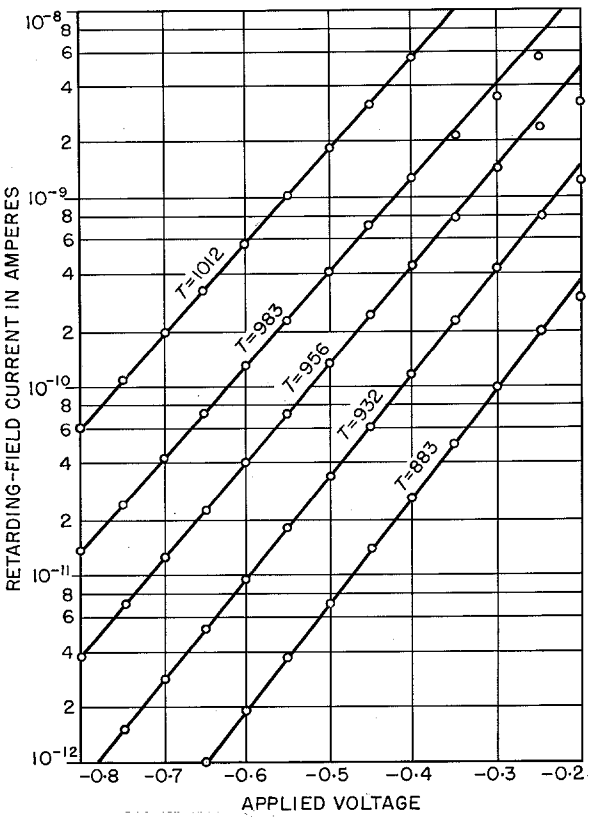


Figure 6—Retarding-field plots obtained from molybdenum cathode. *T* is the temperature in degrees Kelvin.

7.1 Retarding-Field Method

The method described by Benda[7] was used. In this method the *BN* cathode is made up into a diode facing a molybdenum-plate anode that is provided with a heater. In the activation of the *BN* cathode, free barium is deposited on the molybdenum anode, which is subsequently used as a cathode for the retarding-field measurements. It can be shown that

$$i_a = A_a F_a T_a^2 \exp [e (V - \phi_c)/kT_a].$$

Here the suffix a refers to the molybdenum "anode."

A_a = anode emission constant
F_a = anode area
T_a = absolute anode temperature
V = applied anode–cathode voltage
ϕ_c = arithmetic-mean work function of *BN* cathode.

Log i_a is plotted as a function of V for several values of T_a. The values i_{ao} corresponding to $V = 0$ are determined by extrapolation of the

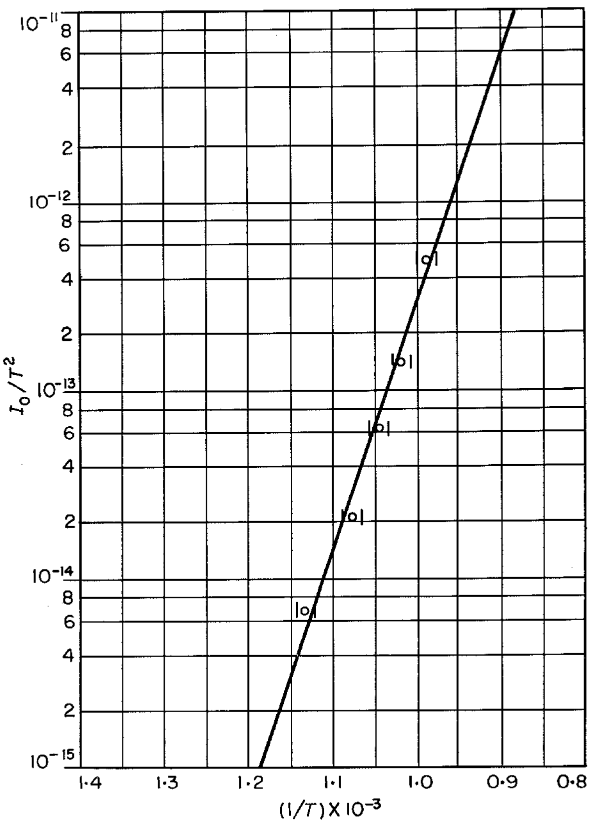


Figure 7—Richardson curves for *BN* cathode obtained by retarding-field measurements. *T* is the temperature in degrees Kelvin.

[7] H. Benda, "Die Emissionskonstanten von Metall-Kapillar-Kathoden," *Frequencz,* volume 7, pages 226–232; August, 1953.

straight-line parts of the curves and ϕ_c is determined from a plot of log (i_{ao}/T_a^2) against $1/T_a$, as usual. The values of T_a are obtained from the slope of the original lines.

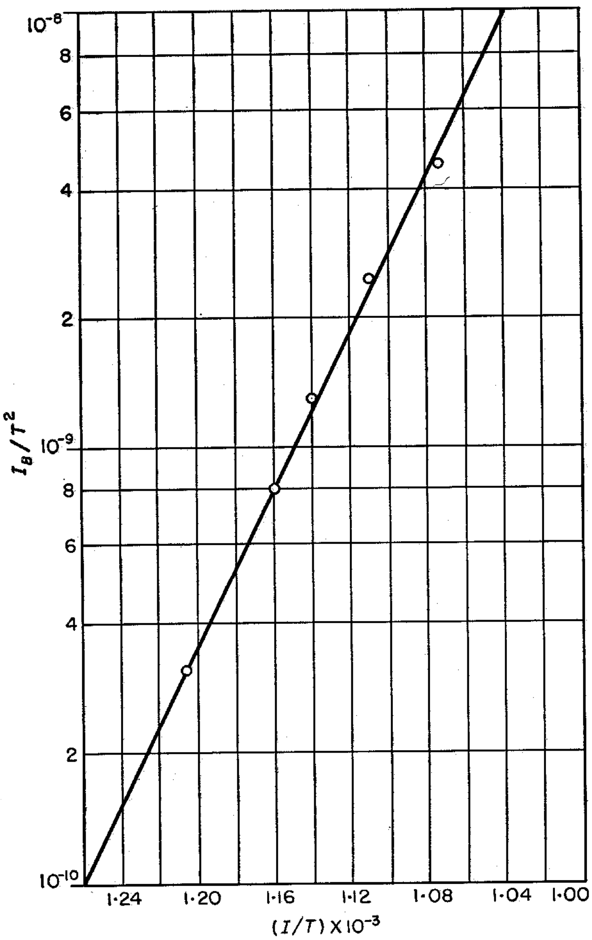


Figure 8—Richardson curve for *BN* cathode.

7.2 Richardson Plot

The data for this plot are taken using both pulse and direct-current operation. The saturation current is assumed to be the current at which the diode characteristic breaks away from a straight line on 2/3-power-law paper. It has been extablished that a Schottky plot gives a zero field current value lower than is obtained from the space-charge break and the temperature deduced from the Schottky line is very low. The zero field current deduced from a Schottky plot is thus unlikely to represent the true saturation emission.

7.3 Comparison with Charts of Loosjes and Jansen[8]

These charts assume that $A = 120$ amperes per square centimetre and that the cathode area is given. If the saturation emission (in our case, the results of the experiment above described) and the temperature are known, the charts yield a value of work function that depends on the temperature. If the work function is plotted as a function of T, the results should show whether there is a systematic variation of ϕ with T and should allow the zero-temperature value of ϕ to be deduced if a straight-line law is obeyed.

8. Results

Typical experimental results for the three methods are shown in Figures 6 through 9.

The mean work function measured by the method of section 7.1 yields a value of $\phi_c = 2{\cdot}8$ electron volts. Method 7.2, which as is well known, yields a value for the minimum work function of a patchy surface gives $\phi_{\min} = 1{\cdot}78$ electron volts. Method 7.3 yields a value about 1·91 electron volts which, when plotted against T, yields a value of $\phi_o = 1{\cdot}79$ electron volts, is in good agreement with the Richardson value. The temperature coefficient of work function is $1{\cdot}36 \times 10^{-4}$ electron volt per degree Kelvin.

Now, let us assume that the cathode can be represented as a patchy surface having two distinct values of work function. One of these has the value 1·78 electron volts and the other is that of nickel, 5·0 electron volts. Let the proportion at the low work function be a. Then,

$$\phi_c = a\,\phi_{\min} + (1 - a)\,\phi_{\text{nickel}} \qquad (2)$$

And, inserting values, $a = 0{\cdot}69$. Thus, the surface coverage is in the region of 70 per cent.

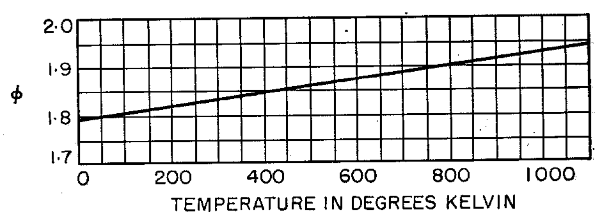


Figure 9—Work function for *BN* cathode plotted against temperature. From the curves of Loosjes and Jansen.

[8] C. G. J. Jansen and R. Loosjes, "Graphs for Rapid Calculation of Work Function of Thermionic Emission," *Philips Research Reports,* volume 8, pages 81–90; April, 1953.

We therefore conclude that the *BN* cathode can satisfactorily be represented as a barium-on-nickel diffusion emitter in which the surface coverage is about 70 per cent, which is in approximate agreement with the electron-microscope photographs, and in which the temperature coefficient of work function is about 10^{-4} electron volt per degree Kelvin.

9. Conclusions

In addition to the use of these cathodes in testing diodes, they have been used in reflex klystrons and television cathode-ray tubes and have given good service in both applications.

It is considered that the *BN* cathode should find a very wide field of application, especially in valves for ultra-high frequencies.

Two New Equations for the Design of Filters

By MILTON DISHAL

Correction for Volume 30, Pages 324–337; December, 1953

IN THE FIRST LINE of Step 8 on page 329, change "left" to "right".

In the numerator of (6), change the plus signs of the second, fourth, sixth, et cetera terms to minus signs.

In the first line of Step 11 on page 330, change "right" to "left" and in the second line replace the first 10 words with "all terms will be positive."

The above changes in no way affect the synthesized network since the actual synthesis is accomplished by using (7) (and (8)). However, it is true that the right-half-plane zeros of Step 7 produce (7) and the left-half-plane zeros of Step 7 produce (8).

There is the following additional point involved: it is sometimes necessary or desirable to know the *terminated* voltage (current) reflection coefficient or the terminated input impedance (admittance) of a particular pair of terminals.

As indicated in Step 10, (7) gives the short-circuit (open-circuit) input immittance of *one pair* of network terminals. The terminated voltage (current) reflection coefficient (complex) of *this same pair of terminals* is given by the ratio of the numerator of (6) (as corrected) to the numerator of (5). The point being stressed is that there must be alternating signs in the numerator of the terminated reflection coefficient (complex) for that pair of terminals that has (7) as its short-circuit (open-circuit) input immittance. Similarly, in agreement with the corrections to Step 11, there must and will be all positive signs in the numerator of the terminated reflection coefficient (complex) for that pair of terminals that has (8) as its short-circuit (open-circuit) input immittance.

The above correction could have been accomplished simply by interchanging (7) and (8). This procedure has not been followed because it is important for practical reasons to use the present (7) for the major part of the synthesis. This is true because (as can be seen) the denominator of (8) is formed by taking *differences*, and significant figures are usually lost in the process.

The author is indebted to Mr. E. Green of Marconi's Wireless Telegraph Company, Limited, for discussions (via mail) that clarified the above points.

Novel Gas-Gap Speech Switching Valve *

By A. H. BECK, T. M. JACKSON, and J. LYTOLLIS

Standard Telecommunication Laboratories, Limited; London, England

A NEW gas-discharge circuit element is described. It has the following properties; low resistance (about 100 ohms) when struck, the resistance being constant for all frequencies up to at least 50 kilocycles per second, the reactive component of the impedance is negligible, and the noise voltages are at least 60 decibels below signal levels. The development of such devices considerably simplifies major problems in electronic telephone exchange design.

. . .

The all-electronic telephone exchange has been the subject of much research in the last decade. One of the simpler ways of making such an exchange seems to be the direct replacement of the crossbar switch by electronic devices, each junction of the mechanical switch being replaced by a suitable electronic device. The more obvious requirements of such a device are:—

A. That it can be switched rapidly from a high (practically infinite) alternating-current impedance to a low (practically zero) alternating-current impedance.

B. That it shall transmit voice frequencies at milliwatt power levels without either frequency or amplitude distortion.

C. That it shall not introduce electrical noise.

D. The device should form part of the line circuit and the switching operation should be performed by pulses in the line circuit.

A cold-cathode gas-filled valve is an obvious device to investigate for the purpose outlined because the glow discharge can be initiated and extinguished directly by pulses and there is a definite break between the conducting and non-conducting state within a time of the order of 1 millisecond. There are other advantages in using a cold-cathode gas-filled valve. When in a non-conducting state the valve consumes no power; rapid fault diagnosis is possible because the glow is easily visible; and such valves have long lives.

The main difficulty to be overcome is that of providing a sufficiently low resistance path for reasonable direct currents in the conducting state. Bell Telephone Laboratories[1] have produced a diode gas-gap using a hollow-cathode discharge, which gives a small power gain in the frequency range of 50 to 3000 cycles per second and operates with direct currents in the range 10 to 20 milliamperes. It is not known whether this valve can be switched and used directly in the line. This article describes an electrode arrangement for gas-filled valves that gives a noise-free, low-resistance path for direct- and alternating-current signals, the upper frequency for the alternating-current signals being at least 50 kilocycles per second. This device satisfies the other requirements stated in this introduction. Several different electrode structures are possible and two practical valves using the arrangement are described.

1. Description and Mode of Operation

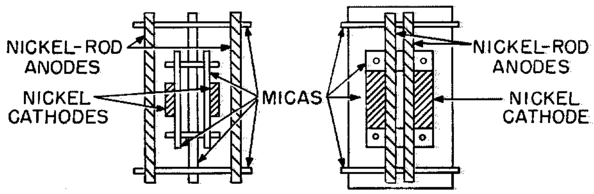
The electrode arrangement consists essentially of a cathode with two electrodes acting as joint anodes equally spaced from the cathode surface. These anodes can be of any geometrical shape provided a normal projection from them in the direction of the cathode will meet the cathode surface. There is a limit to the anode–cathode separation, depending on the nature and pressure of the gas filling. Figure 1 shows various electrode arrangements that have been found to work satisfactorily. Figures 2 and 3 show the variation of anode-gap resistance with cathode current and frequency respectively. The gap

* Reprinted from *Electronic Engineering*, volume 27, pages 7–12; January, 1955.

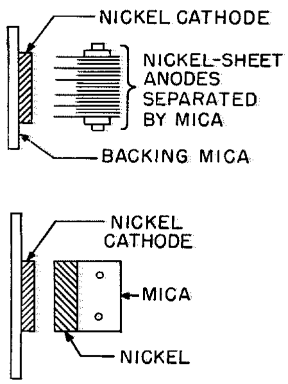
[1] M. A. Townsend and W. A. Depp, "Cold Cathode Tubes for Transmission of Audio Frequency Signals," *Bell System Technical Journal*, volume 32, pages 1371–1391; November, 1953.

resistance can be measured directly using an alternating-current bridge. It can be seen that the gap resistance is substantially independent of frequency but varies inversely with the total current through the valve. The apparent increase of resistance at low frequencies is because the impedance of blocking capacitors used in the measuring circuit becomes appreciable at these frequencies. Below a current of 10 milliamperes there is a rapid increase of gap resistance but above this current, the variation of resistance with current is gradual. The gap resistance is substantially non-reactive, there being no frequency-dependent inductive component present such as is normally found in diode electrode arrangements, It is also noise-free.

Figure 4 shows the direct-current characteristics of a typical valve using the electrode

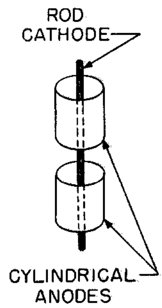


A—Double-assembly valve, rod anodes.

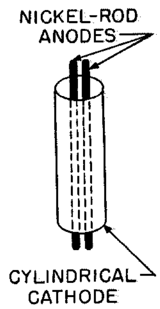


B—Valve with plate anodes perpendicular to cathode surface.

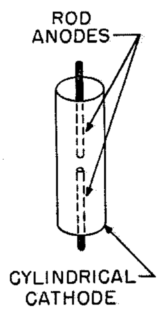
C—(Not illustrated) valve with plate anodes parallel to cathode surface. This valve was similar in construction to *A*, but the nickel rods were replaced by flat anode plates with their flat surfaces parallel to the cathode.



D—Central cathode.



E—Central anodes parallel to each other.



F—Central anodes opposite each other.

Figure 1—Various structures for double-anode valve.

arrangement already described and Figure 5 shows the circuit used to measure the characteristic. The anodes were controlled by separate high-voltage supplies and these supplies were first adjusted so that the currents in the two

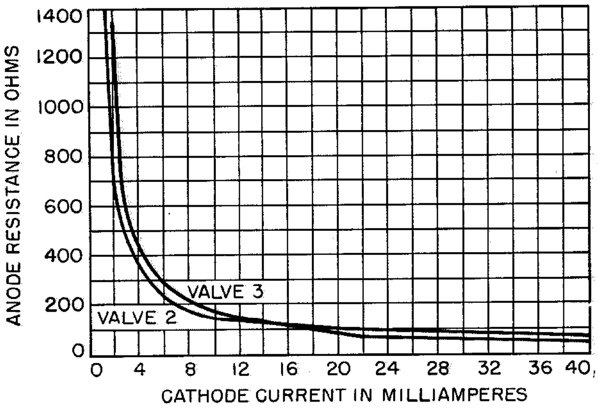


Figure 2—Variation of resistance with cathode current. Valve *2*—0·5-millimetre gap between anodes. 30 millimetres of mercury pressure of neon–argon gas mixture. Valve *3*—0·5-centimetre gap between anodes. 30 millimetres of mercury pressure of neon–argon gas mixture.

anode circuits were equal. The voltage of each supply was then adjusted so that one anode source was raised in voltage by steps of 5 volts, while the other anode source was lowered in steps of 5 volts. The currents in each anode circuit were measured for each 5-volt variation. and the voltage difference between the anodes was also measured. The characteristic shows that about the point where the anode currents are balanced, there is a region where a small change in voltage on one anode produces large current variations in the two anode circuits. It is also obvious that as the current in one anode circuit increases, there is a corresponding decrease in the other anode circuit of approximately

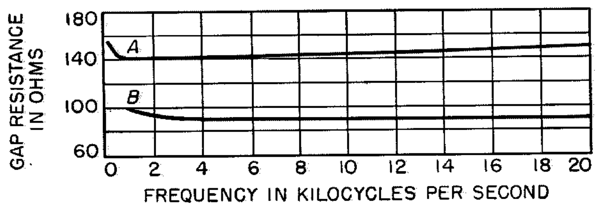


Figure 3—Variation of gap resistance with frequency. Valve *2*—0·5-centimetre anode gap. Neon-argon gas mixture at a pressure of 30 millimetres of mercury. Curve *A* is for 10 milliamperes and curve *B* 20 milliamperes total cathode current.

the same amount, so that the cathode current remains constant. The linearity of the graph in the working region shows that distortion will be negligible. The graph also shows that this electrode arrangement has the important property that a surge of voltage in one anode circuit will not extinguish the discharge so that the valve ceases to operate. Such a surge will merely cause one anode to stop taking current, the whole of the current through the valve being taken by the other anode until the surge has passed. This is a useful characteristic for telephone line operation.

Valves incorporating the double-anode electrode arrangement can be used for speech transmission in a circuit similar to that shown in Figure 6. The valves are acting effectively as switches in the line. In a 10-kilohm-impedance

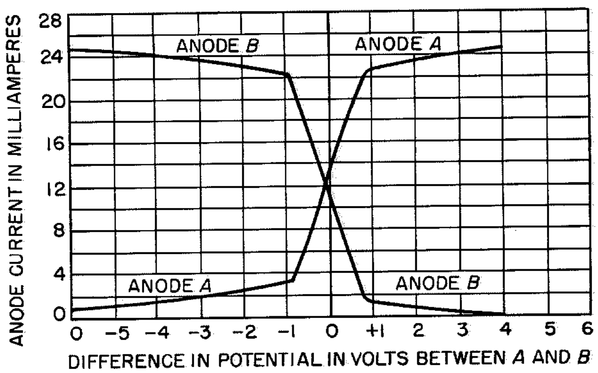


Figure 4—Anode characteristic of double-anode valve. Valve *3*—0·5-centimetre anode gap. Neon–argon at 30 millimetres of mercury pressure.

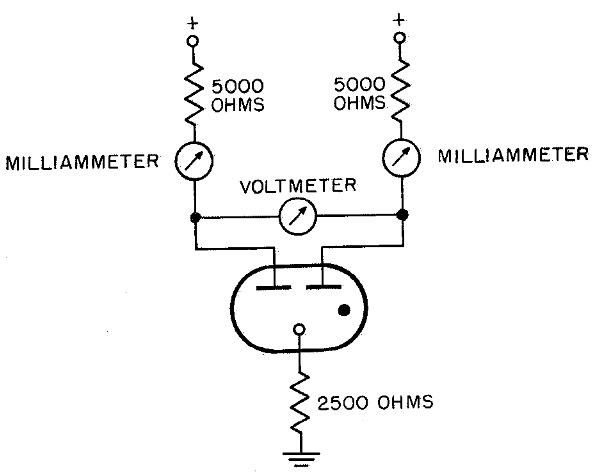


Figure 5—Test circuit for direct-current characteristic.

circuit the total loss due to a pair of valves acting as one stage is less than 0·2 decibel.

To enable valves using the electrode arrangement to be switched directly using pulse voltages in the line, it is necessary to have stable, reproducible, electrical characteristics. For instance, the maintaining voltage V_m of all valves should be within the limits of ± 5 volts if V_m is of the order of 100 volts and the breakdown voltages V_b should be within the limits of ± 10 volts if V_b is of the order of 250 volts. It is also necessary to have a large difference between the maintaining voltage and the breakdown voltage of the glow discharge. V_m is controlled by the nature of the gas filling, current density, cathode material, and state of the cathode surface. V_b is also controlled by these factors, but in addition it is a function of the product of gas pressure and electrode separation.

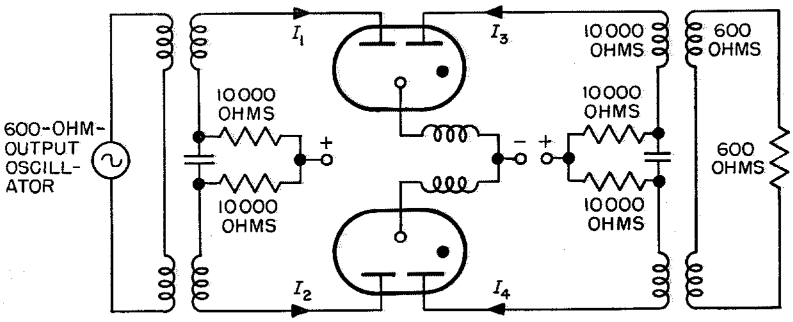


Figure 6—Use of the double-anode valve as a switch directly in the telephone line. I_1, I_2, I_3, I_4 = 10 milliamperes.

During study of the geometrical design factors necessary to ensure a low-resistance gap between the joint anodes, experiments were undertaken to relate gas pressure and electrode separation to gap resistance. This was done by taking a valve with an electrode separation of fixed value and measuring the resistance between the anodes when the valve was still on the pump bench so that the gas pressure could be varied at will. The current in the discharge was kept constant at 20 milliamperes. The results for helium are shown in Figure 7. It shows that there is a sudden sharp increase in resistance when a given pressure is reached for a particular electrode separation. Argon gave similar results. Figure 8 shows a similar graph for neon–argon–hydrogen gas mixture and here, after the sudden increase to a high value there is a rapid decrease to a fairly low resistance once more. It is interesting that there is normally a blue-coloured anode glow on the anode surface when there is a glow discharge in neon–argon–hydrogen gas mixture and yet this glow is not present at low pressures when a low anode-to-anode resistance is observed, but appears suddenly when the sharp increase in anode-to-anode resistance takes place. This argues the sudden formation of a space-charge sheath on the anode, because anode glows are formed when a space charge exists in the anode region. Probe measurements confirmed that space-charge sheaths do form on the anodes at the pressure at which the gap resistance suddenly increases to high values.

The above results show that the anode–cathode gap values cannot be increased indefinitely in order to secure a high V_b because there is a critical distance at a given gas pressure beyond

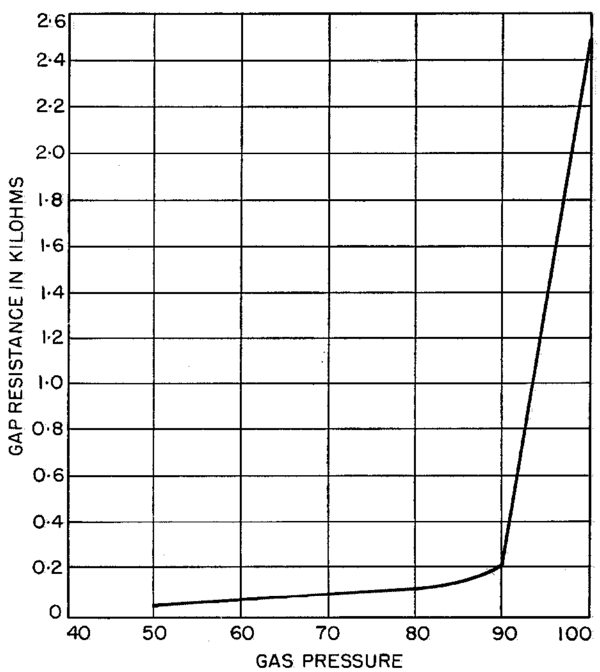


Figure 7—Variation of anode-to-anode resistance with gas pressure (helium). Gas pressure scale is in millimeters of mercury. Anode-to-cathode spacing was 2 millimeters.

which a low gap resistance cannot be obtained. However, it was found that this distance was sufficiently great for a large $V_b - V_m$ difference to be obtained while still retaining a low gap resistance if helium was used as a gas filling and the pressure was suitably chosen.

Another requirement when considering pulse breakdown is that the gap shall always break down when a pulse of given amplitude is applied to one or both of the electrodes.

There are two physical phenomena that introduce a time lag between application of a sufficient voltage on the anode to produce breakdown in a gas-discharge system and the stable establishment of the discharge. These are called the statistical delay time and the formative delay time.

The statistical delay time is dependent upon the presence of an electron or ion in the gap between the electrodes at a particular instant of time. The ion can be accelerated towards either electrode and produces sufficient ionization in the gas to cause breakdown between the electrodes. Ions are normally present in any gas atmosphere because of ionization by cosmic radiation, but the number of ions present per unit volume in a gas-gap can be greatly increased by irradiating the gas with X-rays or ultra-violet light or by photo-electric effects at the electrodes due to either of these agents or visible light. Thus, by suitable priming, the statistical delay time can be eliminated.

The formative delay time depends upon the time taken for the ion or electron that initiates the discharge to produce sufficient ionization to cause breakdown and produce a self-sustaining discharge. This depends upon parameters governed by the pressure and nature of the gas and geometrical dimensions that are normally constant for a given gas-gap, but it is also governed by the voltage applied between the electrodes. Thus, formative delay time cannot be eliminated, but is reduced by increasing the voltage applied between the electrodes to initiate breakdown above the value that will just cause breakdown if applied for a very long time.

Formative delay time necessitates the application of pulse voltages with amplitudes greater than the normal direct-current breakdown voltage of a gas-gap in circuits where pulse techniques are used to initiate the discharge. The difference in voltage between the pulse amplitude required to produce breakdown and the direct-current breakdown voltage is called the "over-voltage". Early measurements upon speech-gap tubes showed that the over-voltage required to produce breakdown depended upon:—

A. History of the cathode (that is, whether it has been conducting or left idle).

B. Pulse width.

C. Light falling on the electrodes; photo-electric effect.

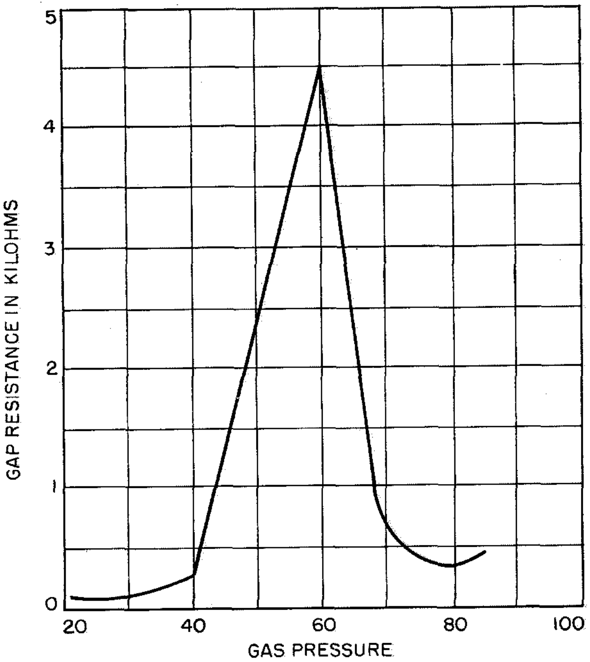


Figure 8—Variation of anode-to-anode resistance with gas pressure (neon-argon-hydrogen in the ratio 92:1:7). Gas pressure scale is in millimeters of mercury. Anode-to-cathode spacing was 2 millimeters.

There are several ways in which the photo-electric priming can be produced. An internal priming source can be incorporated in the valve, or an external agent such as an ultra-violet light source can be arranged to prime a bank of tubes. Another method of priming is to introduce a radioactive substance into the valve, so that ionization is produced by radioactivity. The availability of tritium, a weakly radioactive substance with a half-life of about 12 years and radiation energy of $< 10\,000$ electron volts that can be used with safety within a glass envelope has greatly increased the attraction of this latter form of priming. The advantage of the method is that no electrical power is consumed to produce the priming.

2. *Theory of Low-Resistance Anode-to-Anode Gap*

Probe[2] and other[3] measurements recorded in original published papers have shown that the negative glow is a plasma region, there being high equal concentrations of positive and negative ions here. Concentrations of 10^9 ions per cubic centimetre have been calculated from measurements made, and Emeleus[4] quotes a figure as high as 10^{11} ions per cubic centimetre. The conductivity of the space between the anodes operating under the conditions outlined in this report is therefore high; a low resistance might be expected from this consideration. On the other hand, it is difficult to see why a change of one volt at either anode should cause the anode concerned to take all the current formerly being received by the other anode, which is normally only about 1 millimetre away.

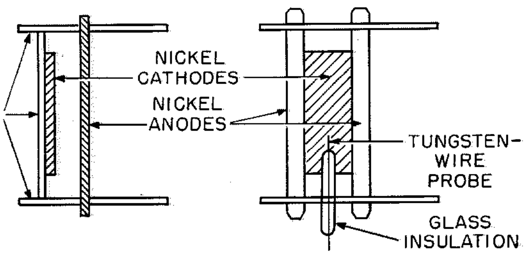


Figure 10—Double-anode valve with probe between anodes.

Figure 9 shows graphs of gap resistance against gap length for an experimental valve containing one fixed and one mobile anode. The electrode assembly was otherwise similar to that of normal valves except that a larger-area cathode had to be used and therefore high total currents were required to cover the whole cathode surface with a glow discharge. At a current of 15 milliamperes per anode an irregular curve was obtained. This was because there was a relatively dark patch on the cathode surface between the anodes; when the moveable anode was over this area the resistance between the anodes was high, but as soon as the anode moved over a part of the cathode where the negative glow was normal, the resistance of the gap decreased. At the maximum spacing the resistance increased suddenly because this was the edge of the negative glow region and the mobile anode was moving out of contact with the plasma. The graphs for higher currents show that there is a steady, small increase of gap resistance with gap width, this increase being only a few ohms per millimetre. The slopes of the graphs at the two higher currents are sensibly the same.

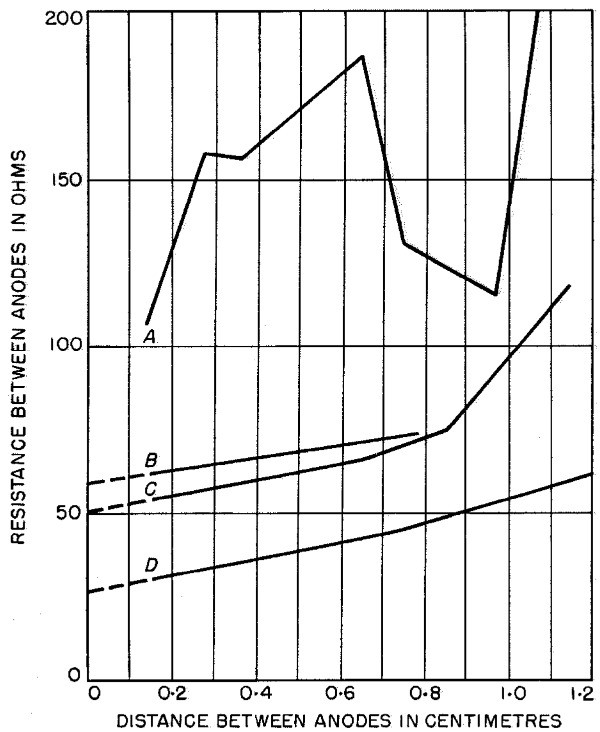


Figure 9—Variation of resistance with gap length. *A*—15 milliamperes per anode. *B*—20 milliamperes per anode; slope = 2·34 ohms per millimetre. *C*—20 milliamperes per anode; slope = 3·0 ohms per millimetre. *D*—30 milliamperes per anode; slope = 2·2 ohms per millimetre.

[2] K. T. Compton, L. A. Turner, and W. H. McCurdy, "Theory and Experiments Relating to the Striated Glow Discharge in Mercury Vapor," *Physical Review*, volume 14, pages 597–615; December, 1924.

[3] B. Van der Pol, "Method of Measuring, Without Electrodes, the Conductivity of Various Points along a Glow Discharge and in Flames," *Philosophical Magazine*, volume 38, pages 352–364; September, 1919.

[4] K. G. Emeleus, "Conduction of Electricity Through Gases," Third Edition, Methuen and Company, Limited, London, England; 1951: see page 61.

The conclusions drawn from the results are that a field theory of the switching action cannot be substantiated, and that a theory on the conductivity of the plasma can only be advanced if some boundary condition at the anodes is postulated to account for the large residual resistance that is obtained for zero length when the graphs are extrapolated.

Probe measurements were next made to elucidate what happens to the plasma potential when one anode is raised and then lowered in potential so that it draws maximum current and then minimum current. A valve similar to that represented in Figure 10 was used and the measuring circuit is shown in Figure 11. The anodes were connected directly to the variable high-voltage sources and the probe potential was varied by means of a potentiometer supplied from a third high-voltage. The currents in the anode circuits were measured by low-resistance milliammeters and the probe current was measured by an accurate multi-range microammeter. The probe potential relative to earth was measured very accurately with a standard potentiometer and the same instrument was used to measure the anode voltages accurately. Since there was practically no resistance in the anode circuits, the shunt current did not affect the anode voltage when the potentiometer was connected between anode and earth. The probe potential measured by the potentiometer must be corrected for the potential drop across the microammeter.

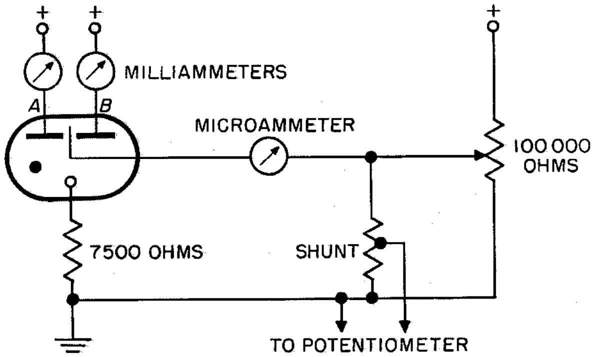


Figure 11—Circuit used for probe measurements.

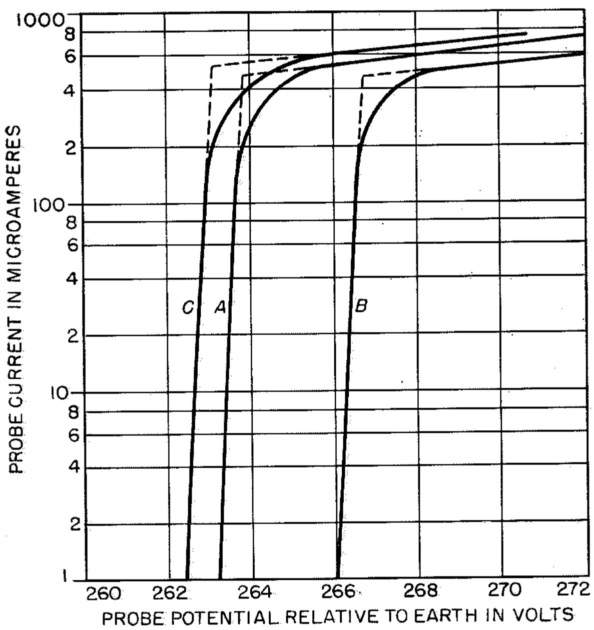


Figure 12—Probe characteristics for balanced and unbalanced states of anode currents in a double-anode valve.

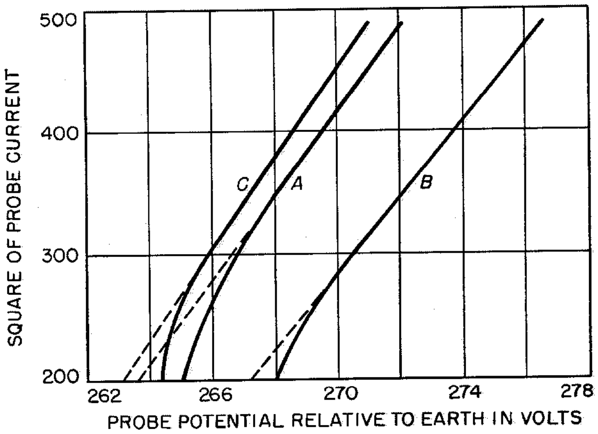


Figure 13—Probe characteristics for balanced and unbalanced states of anode currents in a double-anode valve.

The probe characteristics and anode voltages were measured *A*) with the anode currents balanced at 10 milliamperes, *B*) when anode *A* was raised in potential until anode *A* drew 20 milliamperes current and anode *B* drew zero current, *C*) when *A* was lowered in potential until *A* drew zero current and *B* drew 20 milliamperes current. Figure 12 shows the probe characteristics for these three states and Figure 13 shows the plots of the i^2/V characteristics for the same states for electron currents above the knee part of the probe characteristics. Langmuir probe theory requires that the i^2/V characteristic should be linear except near the space potential, and from the slope of the line the number of electrons per cubic centimetre of the plasma can be calculated.

The results were:—

A.	Potential of anode *A*	= 263·20 volts
	Potential of anode *B*	= 263·12 volts
	Space potential (from graph)	= 263·82 volts
B.	Potential of anode *A*	= 266·90 volts
	Potential of anode *B*	= 263·24 volts
	Space potential (from graph)	= 266·80 volts
C.	Potential of anode *A*	= 258·44 volts
	Potential of anode *B*	= 263·04 volts
	Space potential (from graph)	= 263·15 volts

Electron temperature in degrees Kelvin (from slope of graphs in Figure 12) = 1144

Number of electrons per cubic centimetre (from slope of graphs in Figure 13) = $3{\cdot}7 \times 10^9$

It is seen that the space potential appears to be positive with respect to the anode potential in the balanced state. Contact-potential difference between probe and anodes is concluded to be the cause of this.

The results now make clear the switching action of the double-anode valve. As one anode is raised in potential, the plasma potential follows the potential of the highest electrode with which it makes contact and leaves the other anode at a potential which is now negative with respect to the plasma potential; this latter anode takes a smaller electron current. The other anode takes a greater electron current because the high-voltage supply determines what the cathode current shall be; the current through the valve remains approximately constant. If the plasma potential rises to a voltage such that the lower-potential anode is several volts negative with respect to the plasma potential, then the current to this anode falls to zero; the other anode takes all the current passing through the valve. If one anode is lowered in potential, the plasma potential remains close to the potential of the other anode since it still follows the potential of the highest electrode with which it is in contact. The anode which is being lowered in potential therefore draws less and less electron current as it goes negative with respect to the plasma potential until the cut-off point is reached. The low electron temperature of the space around the anodes shows that the electrons have low energies and therefore a small negative difference of potential between the anode and the plasma will be sufficient to repel the electrons and reduce the electron flow to the anode.

If the anodes are not in intimate contact with the plasma then the above mechanism cannot work. This explains why a space-charge sheath on the anode causes a high anode-gap resistance. It also explains why a patchy glow produces a high gap resistance, because the anode over the part of the cathode where the negative glow is absent or of low intensity will have little influence on plasma potential.

3. Practical Valves

3.1 Multiple-Gap Valve

For convenience in circuit design, it is desirable to have a valve containing a number of gaps; a valve containing ten gaps is particularly suitable for telephone work. To meet this demand the multiple-gap valve shown in Figures 14 and

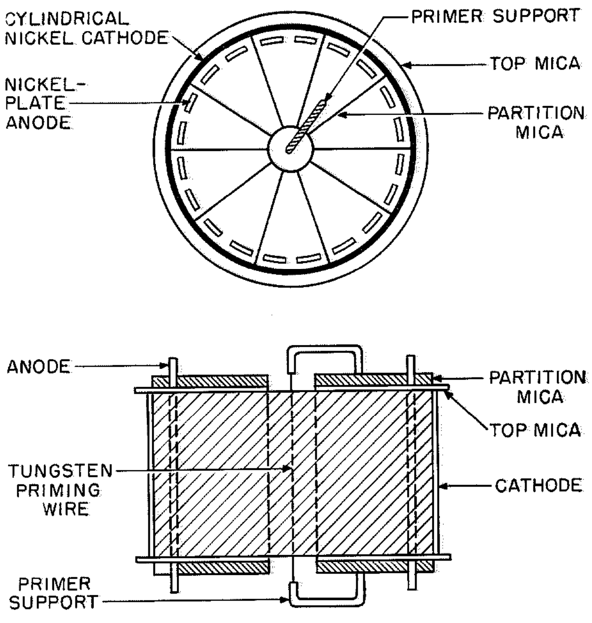


Figure 14—Multiple-gap valve.

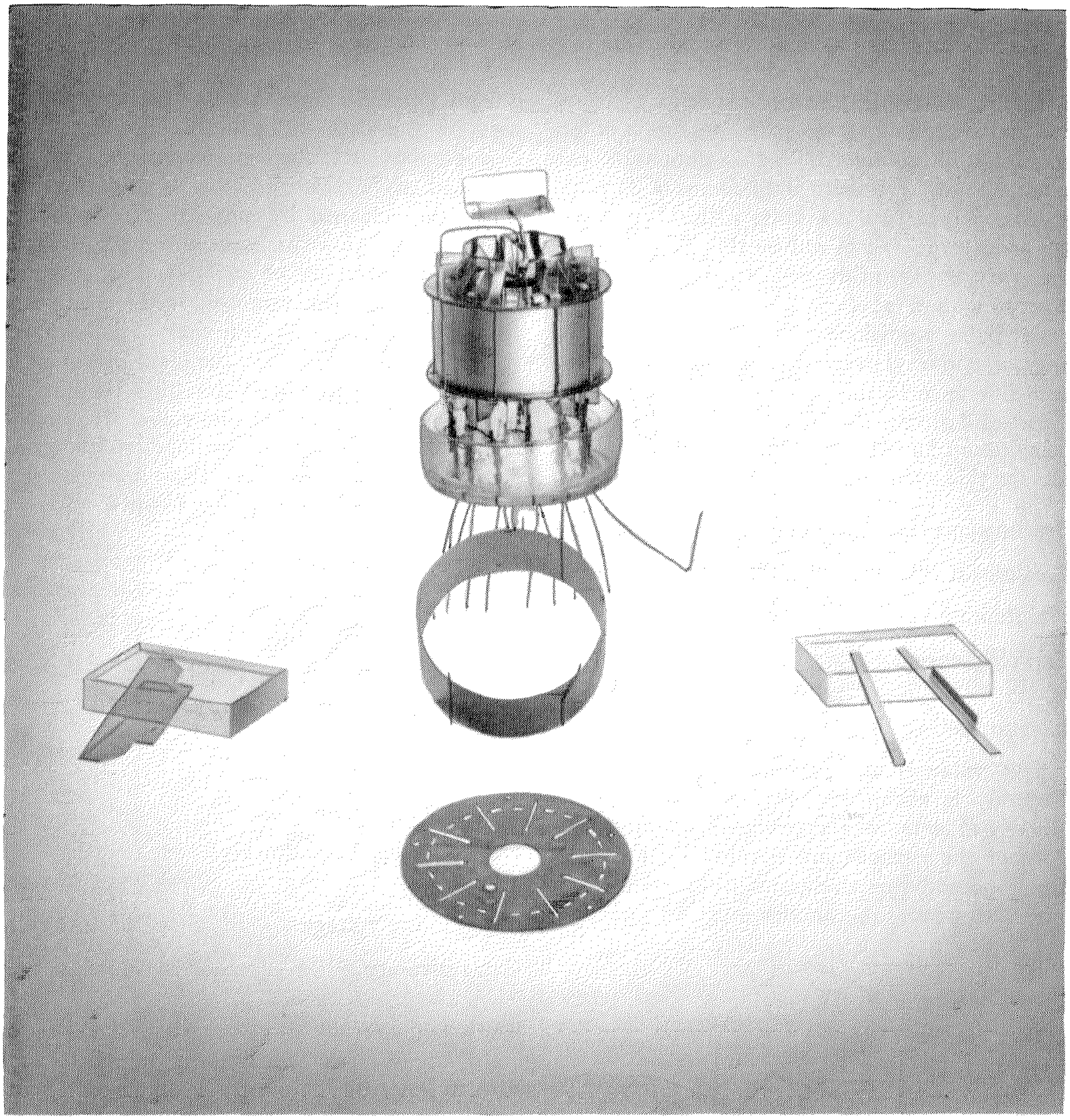


Figure 15—Multi-gap valve and component parts.

15 was devised. The assembly consists of a cylindrical common cathode made of nickel. The surface is divided into ten equal areas by vertical mica partitions that radiate from the axis of the cylinder to make a flush contact with the cathode surface. The partition micas are located in slots cut in the top and bottom micas that close the ends of the cathode cylinder. The anodes, two in each sector, are mounted at a distance of 2 millimetres from the cathode surface and are located in the end micas. At the centre of the cathode cylinder a vertical tungsten wire is mounted along the axis to serve as a photoelectric primer to all gaps. This priming device was for experimental purposes only, because although it is a simple means of priming, its power consumption is too high for practical purposes.

A number of these multiple-gap valves have been made and they have been found to operate satisfactorily in switching circuits. Great care is necessary to keep the cathode spacing constant at 2 millimetres in all ten gaps, because a difference of 0·25 millimetre in this dimension can cause a breakdown voltage difference of 20 volts. V_m is constant to a few volts within one valve, but over a number of valves this characteristic can vary considerably unless care is taken in processing and ageing. The valves work well under pulse conditions with pulse widths of 100 microseconds and overvoltages of 40 volts. There is a difference in the direct-current breakdown before and after a glow discharge, this difference being about 15 volts. Typical values of parameters for this valve are:—

Breakdown voltage before glowing	= 330 volts
Breakdown voltage after glowing	= 315 volts
Maintaining voltage	= 165 volts
Gap resistance	= 100 ohms

3.2 Single-Gap Valve

This valve is illustrated by Figure 16. The assembly is cheap and easy to produce. The pinch seal acts as a convenient mount. Both sides of the cathode are covered by the glow discharge and this eliminates the need for any micas. However, the anodes must be branched

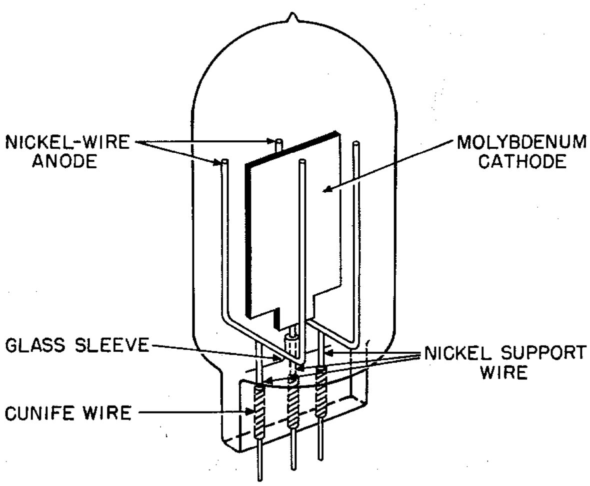


Figure 16—Unit speech-gap valve. The valve envelope is 1·75 inches (4·4 centimetres) tall and 0·354 inch (9 millimetres) in diameter.

as shown so that they are in contact with the glow plasma on both sides of the cathode, otherwise control of the plasma potential is lost and a high gap resistance results. Molybdenum is used as the cathode material because stable, reproducible electrical characteristics can be obtained with this metal if the cathode is thoroughly cleaned by subjecting it to heavy direct-current glow discharge currents. Also, the maintaining voltage of molybdenum cathodes in helium at the pressure used in these valves is 110 volts, which is several tens of volts lower than the maintaining voltage for a nickel cathode. This lower potential simplifies the circuits used in connexion with the valve. Priming is achieved by the use of radioactive material. Typical characteristics for these valves using helium and neon as gas fillings follow.

Helium-filled valves:—

Direct-current breakdown voltage (independent of glow history)	= 240±10 volts
Maintaining voltage	= 113± 2 volts
Gap resistance	= 100 ohms

Neon-filled valves:—

Direct-current breakdown voltage	= 280±20 volts
Maintaining voltage	= 105± 2 volts
Gap resistance	= 200 ohms

These figures relate to valves where the effect of wall charges upon breakdown voltage is not directly controlled by a connexion to the sputtered metal layer in the inner surface of the envelope. Recent work suggests that such control reduces the spread in V_b and can increase the $V_b - V_m$ gap if the potential at which the sputtered layer is maintained is correctly chosen.

4. Conclusion

The operation and characteristics of the speech gap have now been described. Specific embodiments of the ideas presented have been directed mainly towards the utilization of these devices in telephony, but it is clear that there are many other fields in electronic development where their use would be advantageous. It is hoped that this article may stimulate interest in such development.

5. Acknowledgments

The authors wish to acknowledge their debt to their colleagues who have assisted in the design and construction of experimental valves and to the management of Standard Telecommunication Laboratories, Limited, and Standard Telephones and Cables, Limited, for permission to publish this article.

6. Patent References

A. H. W. Beck, T. M. Jackson, and J. Lytollis, British Patent Application 9307/52.

A. H. W. Beck, T. M. Jackson, and J. Lytollis, British Patent Application 24260/53.

Recent Telecommunication Development

"Telecommunications" and "Electronics"

Dr. A. T. Starr, a member of the technical staff of Standard Telecommunication Laboratories is the author of two books recently issued as part of the Pitman Engineering Degree Series. They are companion texts primarily intended for undergraduate students working for bachelor's degrees in engineering at the University of London. They are, of course, suitable for college instruction generally and for private study.

"Telecommunications" is divided into 10 chapters and 9 appendixes under the following titles.

Chapter 1—Communication Systems
Chapter 2—Elements of Telegraph Systems
Chapter 3—Acoustics, Microphones, and Reproducers
Chapter 4—Elements of Telephone Systems
Chapter 5—Lines and Transmission Networks
Chapter 6—Telecommunication Circuits
Chapter 7—Line Telephony
Chapter 8—Basic Principles of Television
Chapter 9—Electromagnetic Waves, Radiation, and Aerials
Chapter 10—Microwave Technique

Appendix 1—Mathematical Formulae
Appendix 2—Steady-State A.C. Theory
Appendix 3—Fourier Analysis
Appendix 4—Power Level: Decibel; Neper
Appendix 5—Operational Calculus: Laplace Transform Method
Appendix 6—Star-Mesh Transformation
Appendix 7—Distortion and Time Delay
Appendix 8—Vectors: Maxwell's Equations
Appendix 9—Waveguide Theory

"Electronics" is divided into 6 chapters and 7 appendixes as follows:

Chapter 1—Physical Fundamentals
Chapter 2—Valves
Chapter 3—Rectification
Chapter 4—Circuit Theory
Chapter 5—Amplifiers, Oscillators, and Detectors
Chapter 6—Electronic Applications

Appendix 1—Mathematical Formulae
Appendix 2—Steady-State A.C. Theory
Appendix 3—Fourier Analysis
Appendix 4—Power Level: Decibel; Neper
Appendix 5—Operational Calculus: Laplace Transform Method
Appendix 6—M.K.S. System: Maxwell's Equations
Appendix 7—Noise

Both books are $5\frac{1}{2}$ by $8\frac{1}{2}$ inches (14 by 22 centimeters) in size and are bound in hard covers.

"Telecommunications" consists of 443 pages of text together with an 8-page index and 9 pages of preliminary material. It sells for 35 shillings in London and $8.75 in New York.

"Electronics" has 388 pages, a 7-page index, and 8 pages of preliminary material. It is priced at 32/6 in London and $7.50 in New York.

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Precision Radar Range Calibrator Incorporating Beacon Function*

By R. D. SINISH

Farnsworth Electronics Company, a division of International Telephone and Telegraph Corporation; Fort Wayne, Indiana

RANGE is perhaps the most accurate information that can be obtained from present radar systems, and extremely precise test equipment for measuring range is therefore necessary. In addition, the test equipment should be compact, portable, and capable of reliable, repetitive operation by technical personnel. The *AN/UPM-11A* range calibrator to be described differs from previous designs in that it includes internal circuits for calibrating the range of beacon navigational radars as well as ordinary radars.

The calibrator operates essentially as a transponder. Radio-frequency pulses from the radar under test reach the calibrator either by coupling to the directional coupler in the radar system or through an external pickup horn included in the test set. Each input pulse to the calibrator develops radio-frequency pulses that act on the radar system as if they were a series of artificial targets having accurately controlled ranges. In normal radar operation, the return pulses are at the same frequency as the radar transmission; while in beacon operation, the return pulses are at a nominal frequency of 9310 megacycles per second.

The calibrator was designed to be operable with pulsed radar systems in the X band, having peak power outputs ranging from 5 to 250 kilowatts, pulse widths of from 0.3 to 3.0 microseconds, and pulse-repetition rates from 300 to 3000 pulses per second. The maximum input radio-frequency power for the required accuracy is 35 microwatts average. At this input level, the return radio-frequency pulses are within 200 kilocycles of the input radio frequency, and the beacon return is within ±2 megacycles of 9310 megacycles.

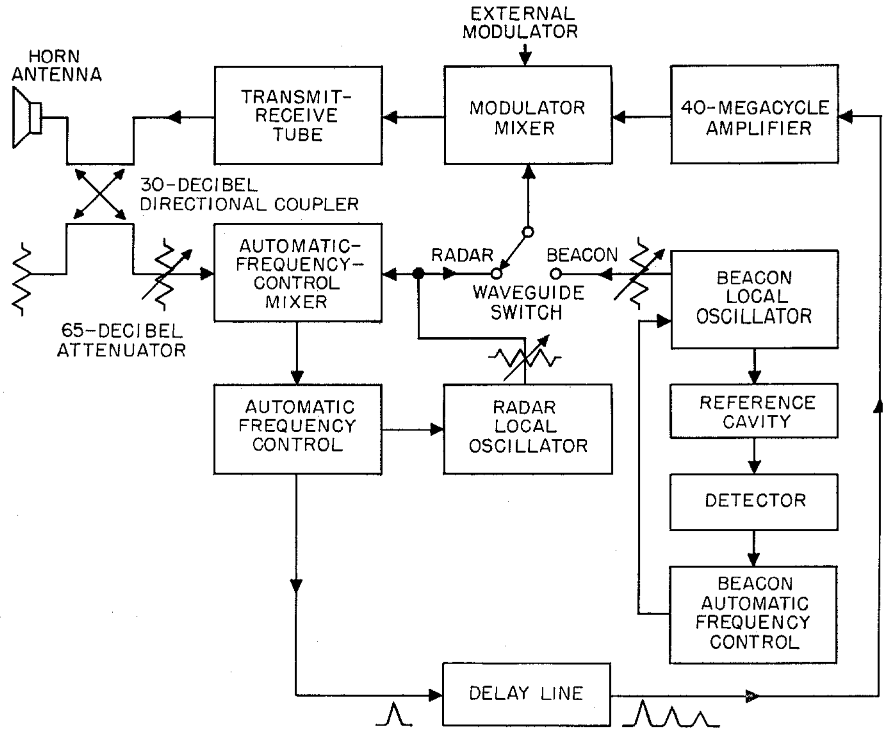


Figure 1—Block diagram of range calibrator.

The range accuracy of the first pulse is held to within ±5 yards of the calibrated value and the spacing between subsequent pulses is held to within ±2 yards. This accuracy, both on frequency and pulse spacing, is held over the en-

* Reprinted in part from "Precision Calibrator Checks Radar Beacons," *Electronics*, volume 28, page 150–153; April, 1955.

vironmental limits of temperature from −40 degrees to +65 degrees centigrade, relative humidity of 95 percent, vibration of 10*g* from 10 to 55 cycles, and shock of 30*g* for 11 milliseconds.

The block diagram of the range calibrator is shown in Figure 1. The signal is introduced into the input waveguide through the horn antenna. A portion of this power passes through a 30-decibel directional coupler to the automatic-frequency-control mixer where it is combined with the output of the radar local oscillator.

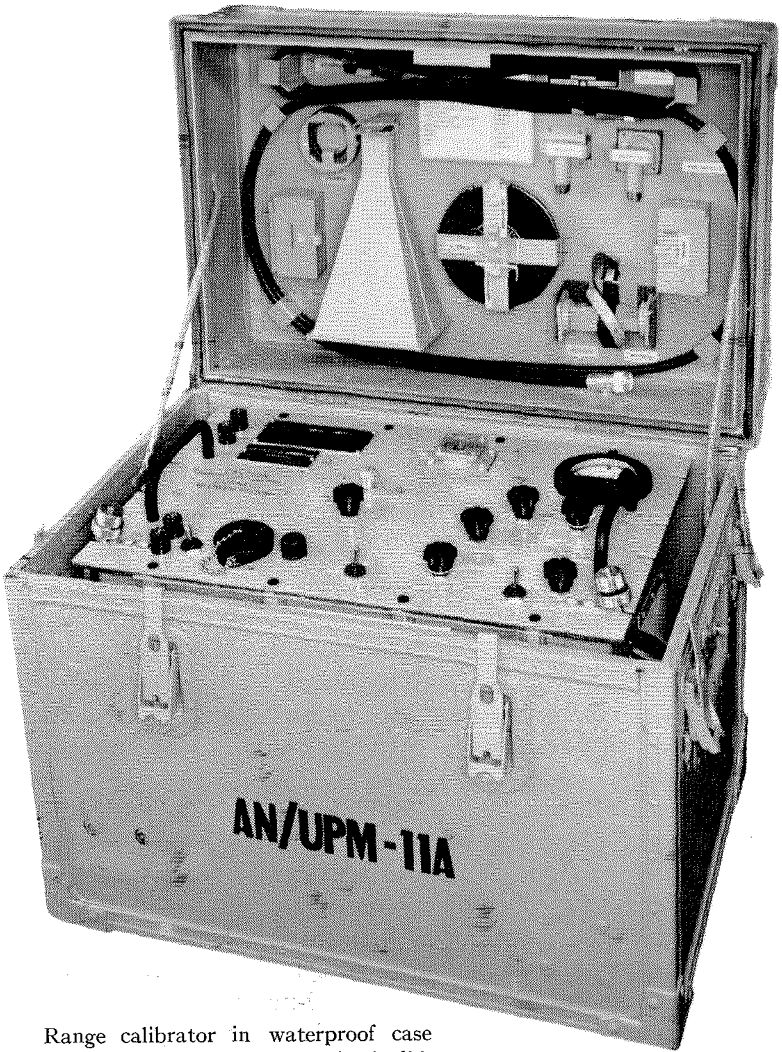
A 65-decibel-maximum attenuator is adjusted to prevent saturation of the crystals in the mixer. Two mixer crystals of type *1N23C* and *1N23CR* connected in reversed polarity permit a push-pull input to the radar automatic-frequency-control chassis.

The automatic frequency control produces an error voltage that maintains the frequency of the radar local oscillator 40 megacycles below the input radio frequency and also furnishes a 40-megacycle pulse to the delay line. The delay line furnishes a series of echo pulses, the first pulse occurring at the approximate one-way delay of the line, with subsequent pulses occurring at intervals of approximately twice the one-way delay of the line. This series of pulses passes through a broad-band amplifier to a *1N23C* crystal in the modulator mixer. The modulator mixer also receives the continuous-wave output of either the radar or beacon local oscillator. The mixing action is such that the pulses go from the mixer through the transmit–receive tube to the radar under calibration. This form of modulation results in two frequencies of output pulses that are 40 megacycles on each side of the local-oscillator frequency. As the local oscillator is held 40 megacycles below the input radar, one pulse output is at the same frequency as the radar while the other pulse output is 80 megacycles below the radar frequency.

On beacon operation, the radar-beacon waveguide switch allows only the beacon local-oscillator output to be injected into the modulator mixer. The beacon local oscillator is held by the beacon automatic frequency control to a nominal frequency of 9350 megacycles, or 40 megacycles higher than the *X*-band beacon frequency. The resultant output pulses from the modulator mixer are thus at 9310 and 9390 megacycles.

1. Radio-Frequency Assembly

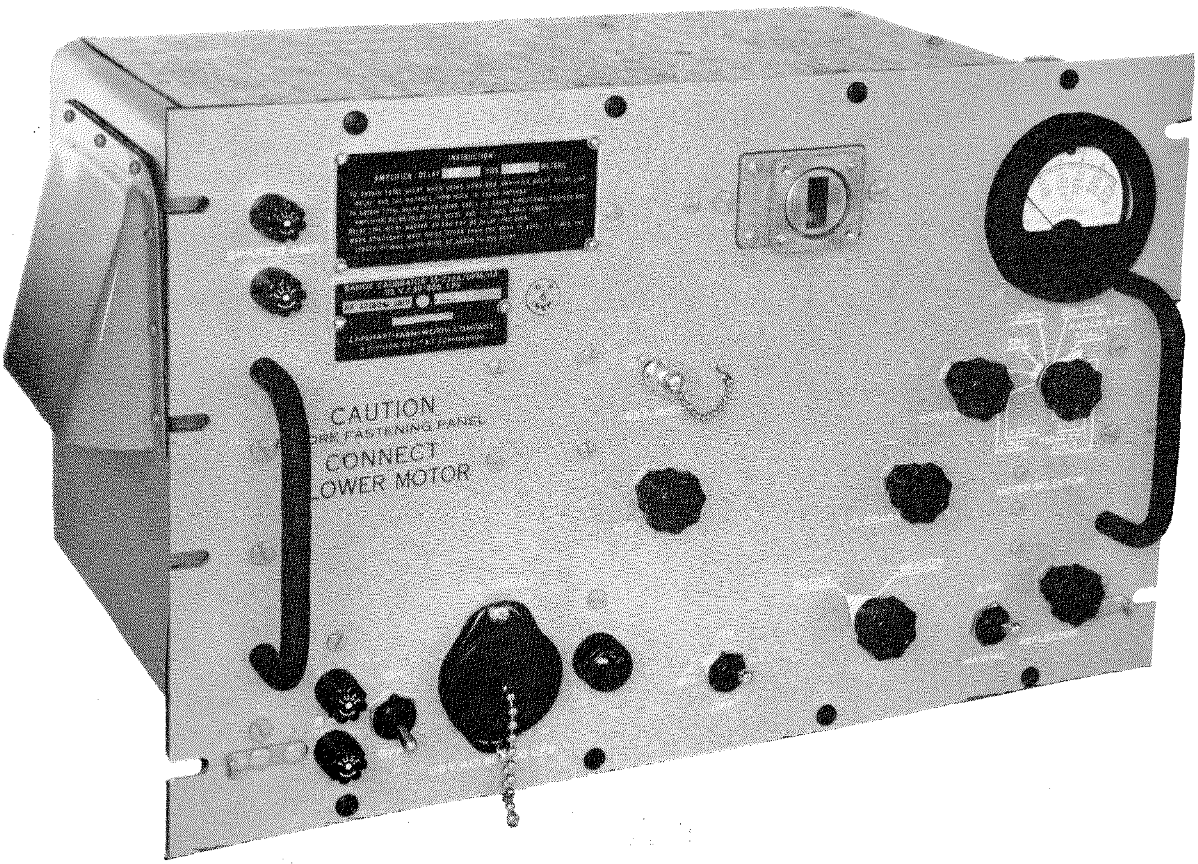
The radio-frequency assembly is fabricated of



Range calibrator in waterproof case with the accessories mounted in the lid.

aluminum. Balanced magic-T mixers are used throughout. A solenoid-operated shutter in the input waveguide assembly provides a 40-decibel insertion loss to protect the mixer crystals when the range calibrator is not in operation. A second attenuator in the input automatic-frequency-control arm, operated by a front-panel control, has a range from 0 to 65 decibels to set the level of the sampled input at a suitable value for the mixer. This control is required with high-power radars to prevent saturation of the crystal mixer and the attendant loss in fidelity of the output pulse from the range calibrator. The radar pulse goes into the E-plane arm of the automatic-frequency-control mixer while the continuous-wave output of the radar local oscillator passes into the H-plane arm of the automatic-frequency-control mixer. Two crystals are mounted in the symmetrical arms of the mixer. The resultant 40-megacycle pulses are carried by a coaxial cable to the radar automatic-frequency-control chassis.

The radar local-oscillator output power goes into the H-plane arm of a modified magic-T mixer. A variable 20-decibel attenuator, adjustable from the front panel, permits setting the output power level. The symmetrical arms of this mixer are connected to the automatic-frequency-control unit and the radar-beacon switch.



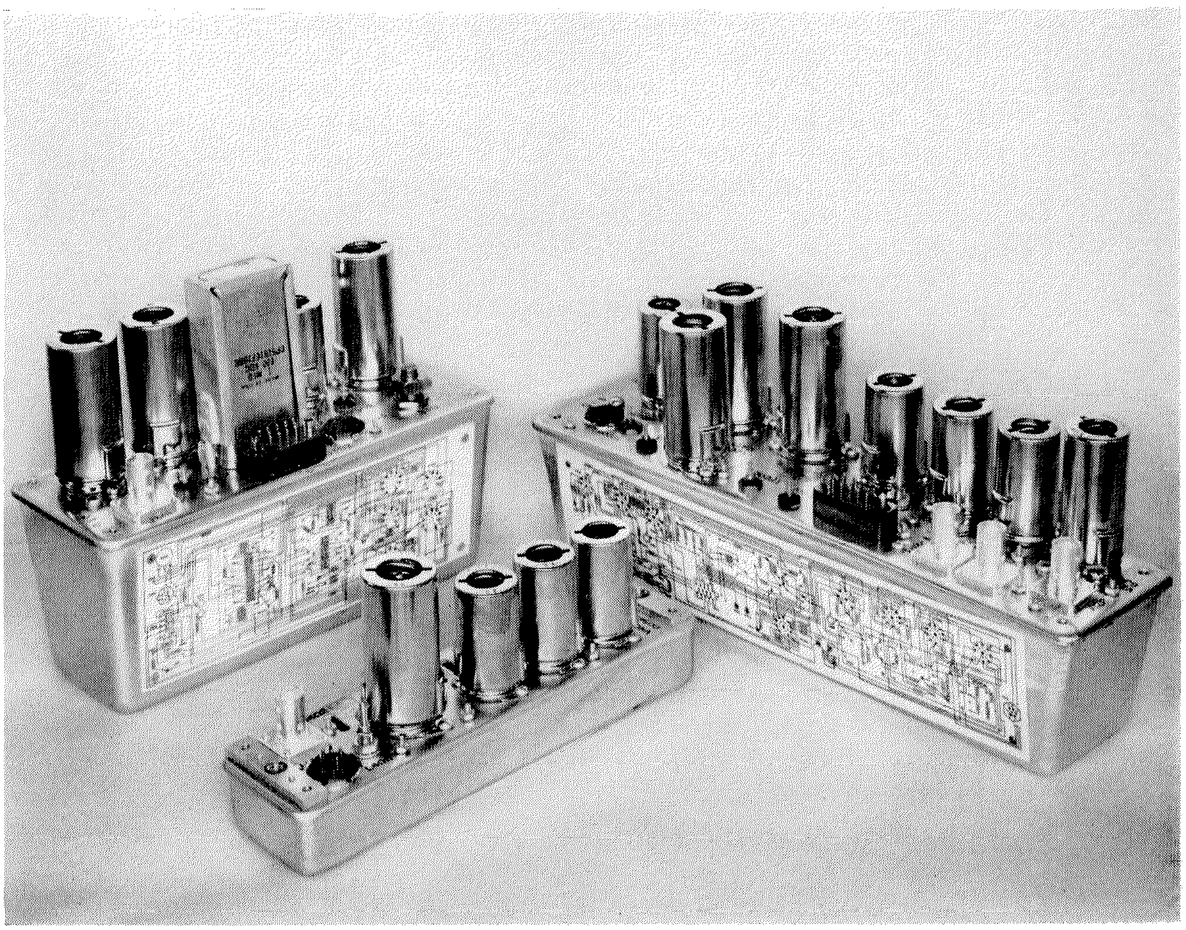
AN/UPM-11A radar range calibrator.

The radar-beacon switch contains a movable vane loaded with polyiron. In the radar position, continuous-wave power from the radar local oscillator is supplied to the modulator mixer. In the beacon position, the vane is switched to present a terminating load to the radar local oscillator and at the same time permits continuous-wave power from the beacon local oscillator to go to the modulator mixer. The 30-decibel isolation provided by this shutter prevents interaction between the radar and beacon local oscillators. As a further precautionary step, when the range calibrator is in radar operation, the repeller of the beacon local-oscillator tube is returned to −300 volts, which prevents the tube from oscillating. Removal of the −300 volts by the action of the radar-beacon switch permits instant operation of the beacon local oscillator with no additional warm-up time. This system of isolating the two local-oscillator outputs was adopted to insure that only one output, either radar or beacon, is transmitted to the radar system.

The modulator mixer is a balanced magic-T with the two *1N23C* crystals mounted in the symmetrical arms. Local-oscillator power passes into the *H*-plane arm while the output signals resulting from the mixing action are transmitted through the *E*-plane arm. The 40-megacycle pulses are applied to one of the crystals. The second crystal is connected to a front-panel jack to allow for external modulation. This permits either sine-wave, square-wave, or noise modulation of the output pulses. It is also useful if the calibrator is to produce a simulated one-way target signal in conjunction with an external pulse generator.

A *1B63A* transmit–receive tube is used to protect the modulator-mixer crystals from burnout by high-level input radar pulses.

The output of the beacon oscillator is split in a modified magic-T mixer, one portion going through the radar–beacon switch to the modulator mixer and the other going to the beacon reference cavity. A special beacon reference cavity operating at 9350 megacycles was developed for use in this equipment. The normal frequency tolerance is ±0.3 megacycle while the maximum deviation from nominal over the temperature range is ±0.4 megacycle. A *1N23C* crystal



The 40-megacycle amplifier is flanked on the left by the beacon and on the right by the radar automatic-frequency-control units.

detector is used with the cavity to provide an input error signal to the beacon automatic-frequency-control chassis.

2. Radar Automatic Frequency Control

The radar automatic frequency control is of the basic phantastron sweep type. The two input pulses from the mixer are combined in a broad-band degenerate π network and applied to the grids of the cathode-follower and the two-stage intermediate-frequency amplifier shown in Figure 2. The cathode-follower drives the ultrasonic delay line. It has a gain of approximately 0.8 and an over-all band pass of 20 megacycles centered at 40 megacycles. The conventional two-stage intermediate-frequency amplifier operates into a Weiss discriminator. The output *S* curve of the discriminator may be examined at a test point provided for monitoring and for initial alignment. The positive-going peak of the *S* curve occurs at 38.4 megacycles while the negative-going peak occurs at 41.6 megacycles. To prevent overloading of the intermediate-frequency-amplifier stages, automatic-gain-control voltage is derived from the negative-going portion of the discriminator curve. This voltage is amplified in one section of a double triode and applied to the second section of the triode operated at a fixed bias. When the amplified voltage is sufficient to overcome the bias, the excess voltage is filtered and applied to the grids of the two intermediate-frequency-amplifier stages. A test point is provided for monitoring the automatic-gain-control voltage.

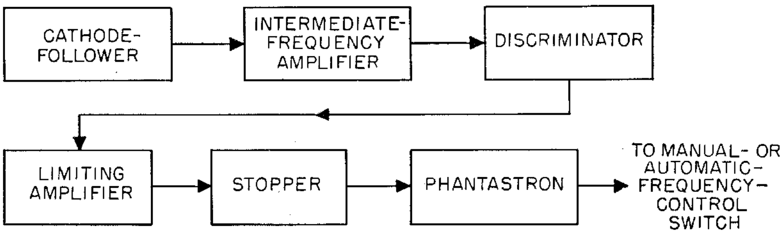


Figure 2—Radar automatic frequency control.

The output of the discriminator goes to a direct-coupled inverse-freeback-pair video amplifier. This amplifier has high gain at low signal levels and low gain at high signal levels to furnish the required dynamic range. The amplified video signals are connected to a peak-limiting amplifier stage and a stopper tube for the phantastron. The negative-going signal at the output of the limiter tube is connected to the cathode of the stopper tube. The stopper-tube output is filtered and then applied to the input grid of the phantastron.

The phantastron operates as a free-running sawtooth oscillator at a frequency of approximately 2 cycles per second. This sawtooth sweep voltage is applied to the reflector control of the radar local-oscillator tube, where the frequency excursion due to the sawtooth voltage is approximately 60 megacycles. A front-panel control sets the direct-voltage level on the reflector.

The bias voltage from the stopper tube effectively lowers the transconductance of the phantastron to the point where it stops sweeping, at which point the phantastron acts as a direct-current amplifier maintaining the correct voltage level on the radar oscillator. For manual frequency control, -300 volts is applied to the screen of the phantastron to prevent sweeping. At the same time, the voltage on the plate of a thyratron is set to its average level by a voltage-divider network, and the same panel control is used for both automatic and manual operation.

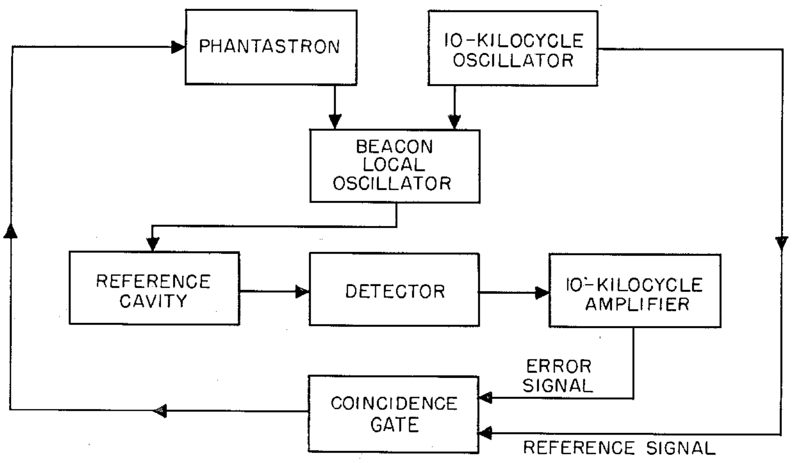


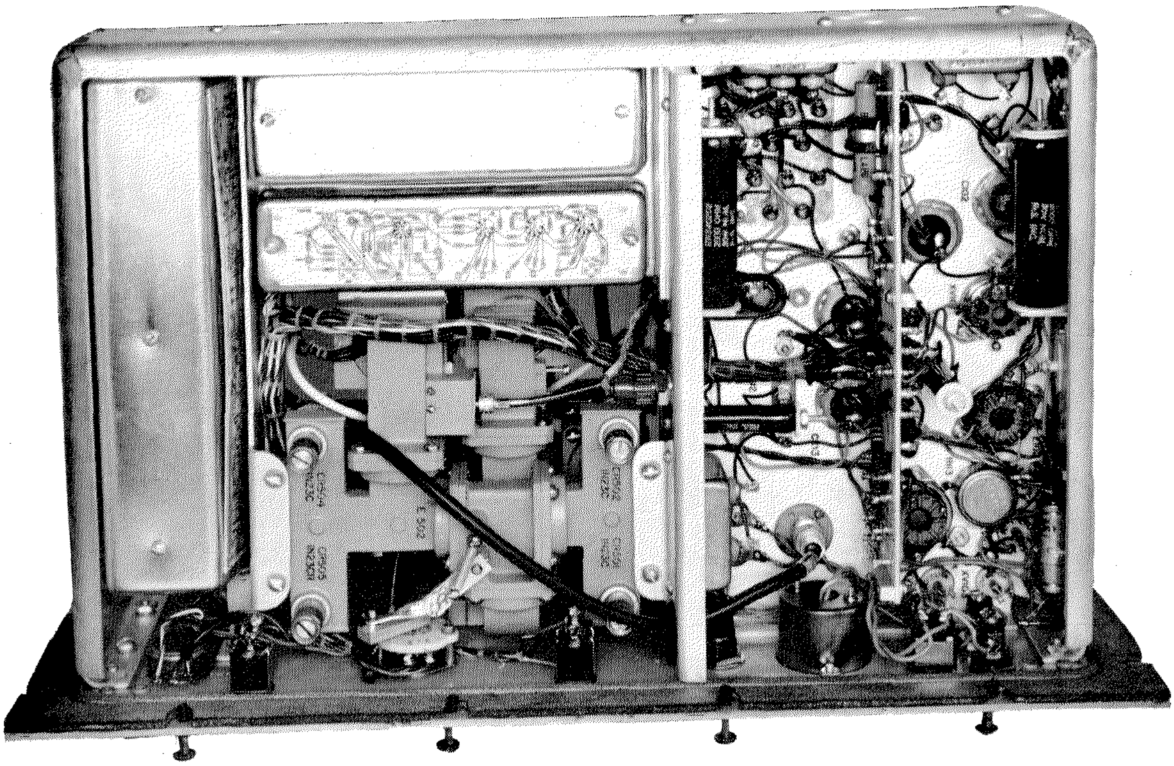
Figure 3—Beacon automatic frequency control.

3. *Beacon Automatic Frequency Control*

An absolute-frequency type of automatic frequency control is used to maintain the frequency of the beacon local oscillator at 9350 megacycles, the frequency of the beacon reference cavity. Again, as may be seen in Figure 3, a phantastron-type sweep circuit is employed. A 10-kilocycle signal is superimposed on the sweep voltage and slope detection in the cavity is utilized to furnish the frequency-control error voltage.

The sawtooth voltage output of the phantastron is modulated with a 10-kilocycle wave and both are applied to the repeller of the beacon local oscillator, thus causing the output frequency of the local oscillator to sweep through the frequency of the beacon reference cavity. The sweep frequency is approximately 13 cycles and the total sweep excursion of the local oscillator is approximately 50 megacycles. As the beacon local oscillator is preset to one frequency, no external controls for this local oscillator are required. Internal adjustments permit setting of the voltage level of the reflector and the injection level.

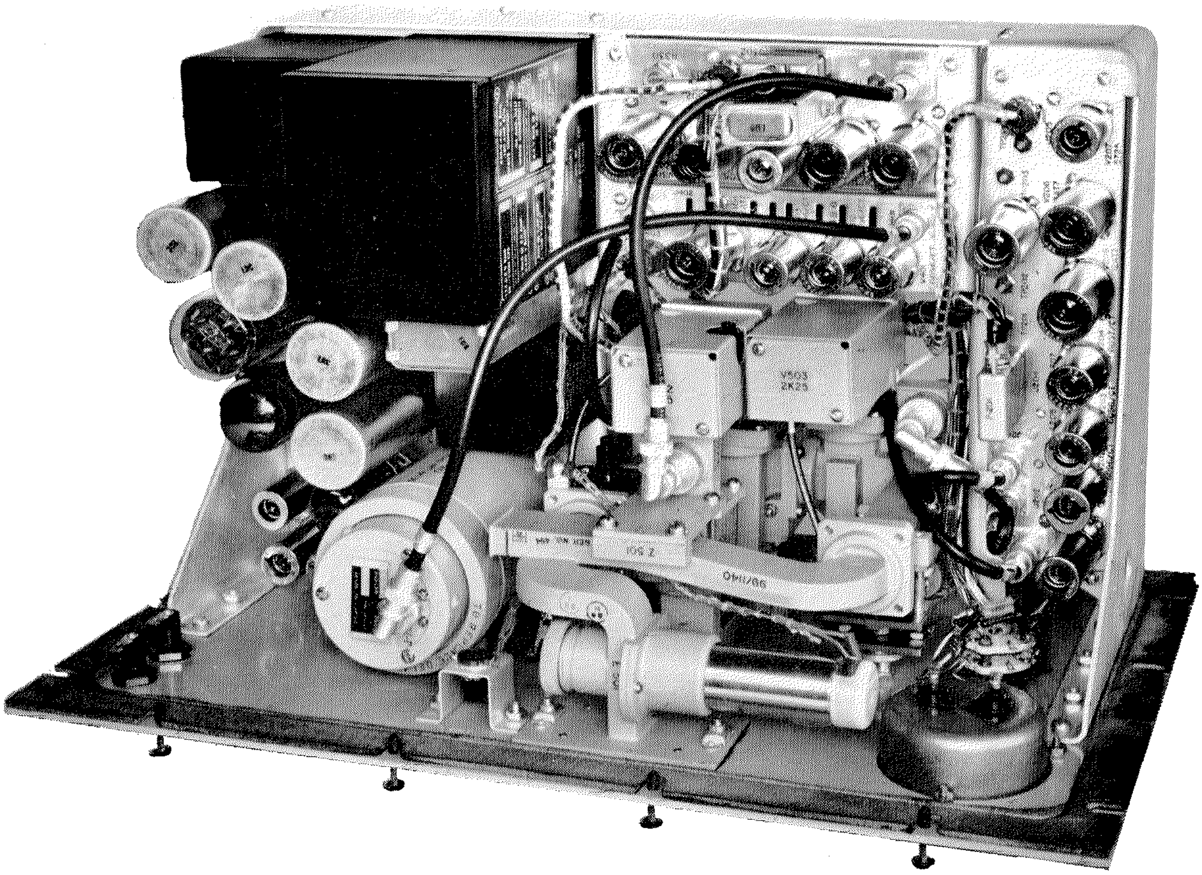
A 10-kilocycle modulating voltage is taken from a Wien-bridge oscillator and is capacitively coupled to the beacon local-oscillator sweep voltage. A similar reference voltage from that same oscillator goes to one of the control grids of a coincidence tube. A phase-shift network is used to correct for any phase difference between the currents in the plate and cathode resistors from which these two voltages are obtained.



Bottom view of the calibrator with the power supply at the right.

The resulting frequency-modulated output of the beacon local oscillator passes through the beacon reference cavity, which has a minimum Q of 1500. By slope detection in the cavity, the frequency-modulated output is converted to amplitude-modulated power, with the phase of the 10-kilocycle amplitude modulation reversing as the local-oscillator frequency sweeps through the resonant frequency of the cavity. When the beacon local-oscillator frequency is higher than the cavity frequency, the 10-kilocycle modulation is in phase with the cavity output.

The 10-kilocycle modulation is detected and applied to the 10-kilocycle amplifier in the beacon automatic-frequency-control chassis. The over-all gain of this three-stage amplifier is approximately 20 000. The amplified error signal is applied to the other control grid of the coincidence tube. When the error signal and the reference signal are in phase, the coincidence tube conducts, applying a filtered voltage bias that stops the sweep action of the phantastron.



The individual chassis are mounted in a U-shaped bracket that is fastened to the front panel. The radar automatic-frequency-control unit is at the extreme right. Next to it, from the panel outward, are the radio-frequency head, 40-megacycle amplifier, and beacon automatic-frequency-control chassis. The delay-line oven is mounted on the panel in front of the regulated power supply.

4. *Delay Line and Oven Assembly*

The delay line is a hermetically sealed unilateral unit approximately 1 inch in diameter by 3 inches long (2.5 by 7.6 centimeters). It employs a *Z*- or *X*-cut quartz bar, the physical length of which determines the delay through the line. A 40-megacycle transducer crystal is bonded to each end of the delay bar to convert from 40-megacycle electric waves to ultrasonic waves and back again to electric pulses. The input pulse from the cathode-follower of the radar automatic frequency control is used to shock-excite the delay line. The first pulse is propagated through the line and picked up by the output transducer crystal. A portion of the pulse energy is reflected back through the line to the input transducer crystal, where it is again reflected back to the output crystal; this process is repeated until the energy is dissipated. It is thus apparent that the second and succeeding pulses are spaced at intervals approximately twice the one-way delay of the line. The total number of useful pulses is in the vicinity of 15 to

20 depending on the length of the delay bar. Each succeeding pulse is of smaller amplitude than the preceding pulse, thus effectively adding to the realism of the pulse simulation of remote radar targets. The pass band of the delay line is a minimum of 12 megacycles centered at 40 megacycles. The insertion loss measured at the tenth pulse is approximately 55 decibels down from the input pulse.

To maintain the required over-all stability of the delay line, a thermostatically controlled oven is used. The temperature of this oven is set to 71 ± 2 degrees centigrade initially and the maximum temperature deviation is held to ± 0.3 degree over the on–off heating cycle. The oven is entirely self-contained with a removable end bell for easy insertion of the delay line. Inductances are included in the oven to tune out the capacitance of the connecting cables and of the delay-line input and output circuits.

5. *The 40-Megacycle Amplifier*

The 40-megacycle amplifier is necessary to overcome the losses in the ultrasonic delay line. This chassis consists of three pentodes and a twin triode. The circuit used is a staggered triple-tuned design followed by a broadband grounded-grid amplifier preceding the cathode-follower output stage. The amplifier has a 10-megacycle pass band centered at 40 megacycles and an over-all gain of 37 decibels.

It was required to maintain the inherent signal delay in the 40-megacycle amplifier fairly constant over a wide variety of input conditions as this delay is the major portion of the calibrated equipment delay. Design tests showed that the impedance of the signal modulator crystal varied considerably with the injection level of the continuous-wave local-oscillator signal. This variation affected the over-all gain–bandpass characteristics, and therefore the signal delay of the 40-megacycle amplifier. The use of the broadband grounded-grid final amplifier stage with cathode-follower output successfully overcame this difficulty.

6. *Mechanical Construction*

To increase the serviceability of the range calibrator, the separate circuits are packaged individually. The range calibrator is composed of six individual mechanical assemblies: radar automatic frequency control, beacon automatic frequency control, 40-megacycle amplifier, oven assembly, radio-frequency head assembly, and the power supply. All units are mechanically bound together by a separate U-shaped wrap-around band that is tied into the front panel. This permits unit substitution of chassis for rapid maintenance. With the exception of the power supply, all units may be removed by loosening the four screws that tie the units to the assembly. In addition, the shields on the beacon and radar automatic frequency controls and the 40-megacycle amplifier are readily removable for circuit checking with the chassis installed in the calibrator. One main cable is used to distribute power to the various chassis.

The *AN/UPM-11A* consists of a transit case, range calibrator, and accessory equipment that is mounted in the cover of the case. The over-all size of the watertight transit case is 22 inches by 19 inches by 15 inches (56 by 48 by 38 centimeters) and the total weight, including the range calibrator and all accessories is 110 pounds (50 kilograms). The range calibrator is designed for standard relay-rack mounting, having front-panel dimensions of $12\frac{1}{4}$ inches by 19 inches and extending 13 inches behind the front panel (31 by 48 by 33 centimeters). The range-calibrator unit weighs 60 pounds (27 kilograms). The accessories include the input power cable, 8-foot (2.4-meter) coaxial cable, 2 waveguide-to-coaxial-cable connectors, measuring tape calibrated in yards and meters, waveguide twist section, 2 quick-disconnect waveguide fittings, and a 20-degree pickup horn.

7. *Acknowledgments*

The development and subsequent production of this range calibrator was done under Air Force Contract AF33(604)-5819. Mr. J. J. Pokorny of the Armament Laboratory, Wright Air Development Center; Dayton, Ohio, served as project engineer for the Air Force. Credit is also due Messrs. E. L. McDirmit, G. A. Richards, D. E. Fisher, R. W. Harpel, R. E. Saxe, and L. D. Stewart, all of Farnsworth Electronics Company, who participated in the development of this equipment.

Cathode-Follower Phase-Shift Oscillator

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CATHODE–FOLLOWER amplifiers may be modified by connecting a phase-shifting network between grid and cathode to produce an oscillator. Analysis indicates that this oscillator is inherently more stable than the standard arrangement in which the phase-shifting network is placed in the plate circuit.

The frequency of oscillation and the necessary gain for sustained oscillations are derived for three typical oscillators. Experimental models are discussed briefly. Preliminary measurements indicated frequency to be very insensitive to supply-voltage variations and reasonably independent of alternating-current loading at the cathode.

• • •

Oscillators derived from the cathode-follower amplifier have been studied by Dunn,[1] Bacon and Salmon,[2] and Schlesinger.[3] The oscillators of Dunn and of Bacon and Salmon utilized networks of resistance and capacitance with overunity gain and are therefore properly called resistance–capacitance cathode-follower oscillators. Preliminary work on these circuits indicated that they are very stable with respect to supply-voltage variations. The oscillator described by Schlesinger employed inductance and capacitance for the elements and was termed a Colpitts-type cathode-follower oscillator. No information was given on frequency stability.

The consequences of modifying the cathode-follower amplifier in such a manner as to produce an oscillator that might properly be termed a cathode-follower phase-shift oscillator will be discussed. It is a counterpart of the standard phase-shift oscillator[4] but with the phase-shifting network in the cathode circuit instead of the plate circuit. The placing of the phase-shifting network in the cathode circuit appears to have a few good points that are not realized when the network is in the plate circuit. Recent synthesis[1,2,5,6,7] of networks of only resistance and capacitance with overunity gain suggests their use with resultant advantages in the phase-shifting network of the cathode-follower oscillator.

1. *Analysis*

1.1 Basic Circuits

The standard phase-shift oscillator[4] is shown in Figure 1. Under the assumptions of the linear-amplifier theory, the frequency of oscillation of the phase-shift oscillator is determined from

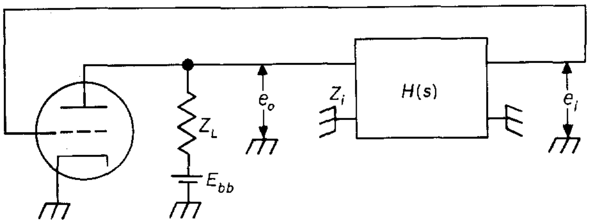


Figure 1—Standard phase-shift oscillator.

$$H(s) + \frac{r_p + Z_0}{\mu Z_0} = 0, \qquad (1)$$

where

$H(s)$ = transfer function of phase-shifting network

s = parameter of the Laplace transformation

r_p = plate resistance

μ = amplification factor

$Z_0 = (Z_L Z_i)/(Z_L + Z_i)$

Z_L = load impedance

Z_i = input impedance of phase-shifting network.

[1] S. C. Dunn, "R-C Cathode-Follower Feedback Circuits," *Wireless Engineer*, volume 30, page 19; January, 1953.

[2] W. Bacon and D. P. Salmon, "Resistance-Capacitance Networks With Over-Unity Gain," *Wireless Engineer*, volume 30, pages 20–22; January, 1953.

[3] Kurt Schlesinger, "Cathode-Follower Circuits," *Proceedings of the IRE*, volume 33, pages 843–855; December, 1945.

[4] E. L. Ginzton and L. M. Hollingsworth, "Phase-Shift Oscillators," *Proceedings of the IRE*, volume 29, pages 43–49; February, 1941.

[5] C. L. Longmire, "An RC Circuit Giving Over-Unity Gain," *Tele-Tech*, volume 6, pages 40–41; April, 1947.

[6] H. Epstein, "Synthesis of Passive RC Networks with Gains Greater than Unity," *Proceedings of the IRE*, volume 39, pages 833–835; July, 1951.

[7] A. D. Fialkow and I. Gerst, "Maximum Gain of an RC Network," *Proceedings of the IRE*, volume 41, pages 392–395; March, 1953.

The cathode-follower phase-shift oscillator is shown in Figure 2.

The condition for sustained oscillations is

$$1 - H(s)\alpha(s) = 0$$

or for linear operation

$$H(s) - \frac{r_p + (\mu + 1)Z_0}{\mu Z_0} = 0, \qquad (2)$$

where

$H(s)$ = transfer function of phase-shifting network

s = parameter of the Laplace transformation

r_p = plate resistance

μ = amplification factor

$Z_0 = (Z_k Z_i)/(Z_k + Z_i)$

Z_k = cathode load impedance

Z_i = input impedance to phase-shift network

$\alpha(s)$ = cathode-follower gain.

An examination of (1) and (2) shows that for the same tube and comparable Z_0, the cathode-follower phase-shift oscillator should be less dependent on tube characteristics than the standard phase-shift oscillator.

1.2 Examples of Cathode–Follower Oscillators

Some typical examples of cathode-follower phase-shift oscillators are given with their frequencies of oscillation as determined from linear theory.

1.2.1 Resistance–Inductance–Capacitance

The frequency of oscillation of the circuit

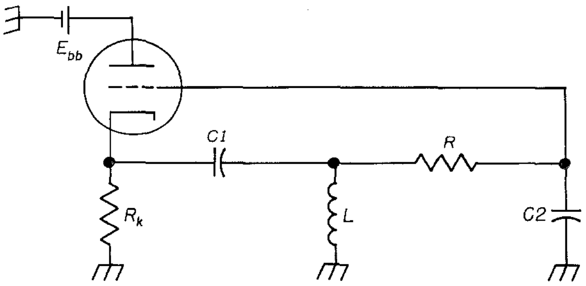


Figure 3—Cathode-follower oscillator using resistance–inductance–capacitance phase-shifting network.

shown in Figure 3 is given by

$$f = \frac{1}{2\pi}\left\{\frac{a - 1}{(a - \bar{\alpha})LC_2 + C_2r_1[C_2r_2 - \beta\bar{\alpha}(C_2R_2)] - (C_2R)^2}\right\}^{1/2} \qquad (3)$$

Since tube characteristics are primarily functions of the supply voltages, the cathode-follower phase-shift oscillator should have better frequency stability with respect to supply-voltage variations than the related standard phase-shift oscillator. It would also appear that the frequency of the cathode-follower oscillator would be less dependent on the effects of loading.

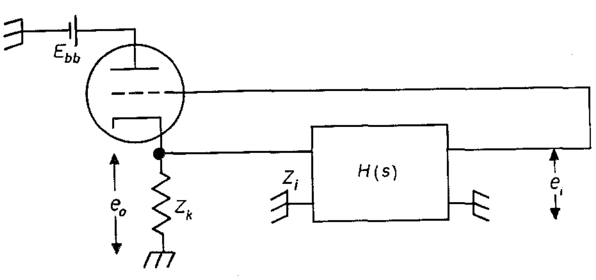


Figure 2—Cathode-follower phase-shift oscillator.

where $\bar{\alpha}$ is a physically realizable solution of

$$\alpha^2 + (\beta_1 + \beta_4 - \beta_3)\alpha + \beta_1\beta_4 - \beta_2\beta_3 = 0. \qquad (4)$$

where

$$\beta_1 = \frac{a(C_2r_1) + C_2r_2 - 2RC_2}{C_2R - C_2r_1 - \beta C_2R_2}$$

$$\beta_2 = -\frac{\mu}{\mu + 1}\frac{r_2}{R_k}$$

$$\beta_3 = \frac{(a-1)(LC_2)(\beta C_2R_k)}{[LC_2 + (\beta C_2R_k)C_2r_1][C_2R - C_2r_1 - \beta C_2R_k]}$$

$$\beta_4 = \frac{aLC_2 + (C_2r_1)(C_2r_2) - (C_2R)^2}{LC_2 + (\beta C_2R_k)(C_2r_1)}$$

$$\beta = \frac{(\mu + 1)}{\mu}$$

$$r_1 = R + R_L$$

$$r_2 = R + R_k$$

$$a = 1 + (C_2/C_1).$$

1.2.2 *Colpitts Type*

The Colpitts oscillator shown in Figure 4 was considered by Schlesinger. The frequency of oscillation is given by

$$f = \frac{1}{2\pi}\left\{\frac{1}{LC_2} + \frac{1}{LC_1}\left[1 + (1-\alpha)\frac{R_L}{R_0}\right]\right\}^{1/2}, \quad (5)$$

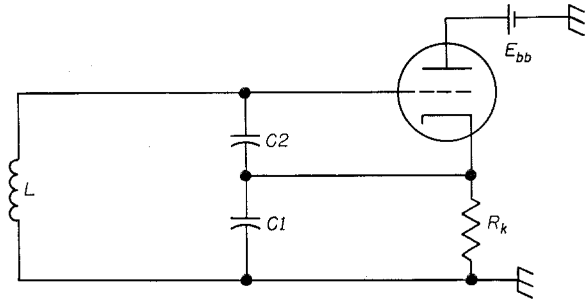


Figure 4—Colpitts-type cathode-follower phase-shift oscillator.

where

$$R_0 = \frac{r_p R_k}{r_p + (\mu + 1)R_k}$$

R_L = coil resistance

and α is a physically realizable solution of

$$\frac{\alpha^2}{C_1^2 R_0^2} - \left(2\tau_1 + \tau_2 + \frac{\tau_3}{\tau_2}\right)\frac{\alpha}{C_1 R_0} + \left(\tau_1 + \frac{\tau_3}{\tau_2}\right)(\tau_1 + \tau_2) - \tau_1 = 0, \quad (6)$$

where

$\tau_1 = 1/(C_1 R_0)$

$\tau_2 = R_L/L$

$\tau_3 = 1/(LC_2) + 1/(LC_2)$

α = required cathode-follower gain.

1.2.3 *Resistance–Capacitance*

In Figure 5 is shown a cathode-follower oscillator using only resistance and capacitance in the phase-shifting network. The corresponding standard phase-shift oscillator is also shown. For equal resistances and equal capacitances, the frequency of oscillation of the cathode-follower circuit of Figure 5 may be calculated from

$$f = \frac{1}{2\pi RC\left[\dfrac{1 + 3R_0/R}{5(1-\alpha) + R_0/R}\right]^{1/2}}, \quad (7)$$

in which

$$R_0 = \frac{r_p R_K}{r_p + (\mu + 1)R_k}$$

and α is the required cathode-follower gain given by

$$\alpha = \frac{1}{60}\left\{59 + 23\left(\frac{R_0}{R}\right) \pm \left[1 - 46\left(\frac{R_0}{R}\right) + 49\left(\frac{R_0}{R}\right)^2\right]^{1/2}\right\}. \quad (8)$$

If $R \gg R_0$, then (8) becomes

$$\left.\begin{aligned}\alpha &= 58/60 \text{ (using minus before radical)}\\ &= 59/60 \text{ (using plus before radical)}\end{aligned}\right\} \quad (8A)$$

and using these values in (7) the frequency of oscillation is

$$\left.\begin{aligned} f &= \frac{1}{2\pi RC}\frac{1}{6^{1/2}} \text{ (} - \text{ sign for radical)}\\ &\text{or}\\ &= \frac{1}{2\pi RC}\frac{1}{2 \times 3^{1/2}} \text{ (} + \text{ sign for radical)}\end{aligned}\right\} \quad (7A)$$

It is of interest to compare these results with those for the corresponding standard phase-shift oscillator. It is found[4] that its frequency of oscillation is

$$f = \frac{1}{2\pi RC[6 + 4(R_0/R)]^{1/2}} \quad (9)$$

and the required gain is

$$A = 29 + 23\left(\frac{R_0}{R}\right) + 4\left(\frac{R_0}{R}\right)^2, \quad (10)$$

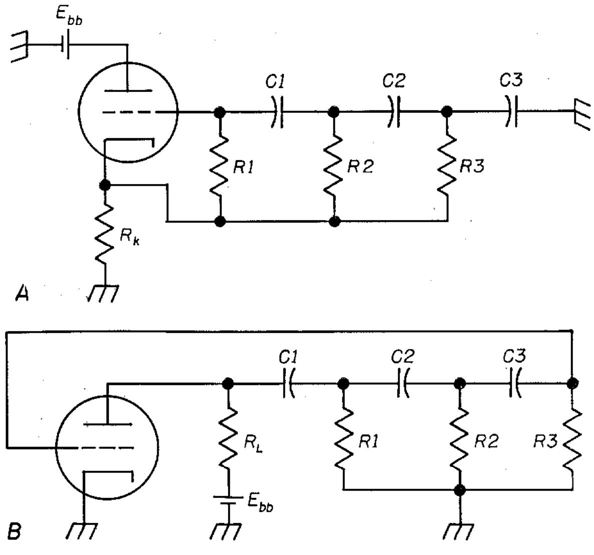


Figure 5—Oscillators using only resistance and capacitance for phase shift. The cathode-follower (grounded-plate) version is at *A* and the standard (grounded-cathode) form is at *B*.

where

$$R_0 = \frac{R_L r_p}{R_L + r_p}.$$

If $R \gg R_0$, then (9) becomes

$$f = \frac{1}{2\pi RC}\frac{1}{6^{1/2}} \qquad (9A)$$

and (10) becomes

$$A = 29. \qquad (10A)$$

The location of the phase-shifting network in the cathode circuit should result in an oscillator more stable than the standard phase-shift oscillator because plate-resistance variations are reduced by the factor $1/(\mu + 1)$.

2. *Experimental*

In accordance with the principles set forth in the preceding portions of this paper, the following oscillators were set up in breadboard form and investigated briefly.

2.1 Experimental Resistance–Inductance–Capacitance

The circuit of Figure 6 had a frequency of oscillation of 4550 cycles per second for $L = 200$ millihenries. It produced a good sine-wave output having a root-mean-square value of about 0.25 volt with 200 volts of plate supply. There was a slight change (less than 0.5 percent) of frequency with a supply-voltage variation from 180 to 270 volts. The long-time frequency stability was good. There was about 0.033-percent change of frequency for a 10-percent change of filament voltage, and about 0.062-percent change for a similar change in plate voltage. A tenfold increase (150 000 to 15 000 ohms) of cathode load produced 1.5-percent change of oscillation frequency.

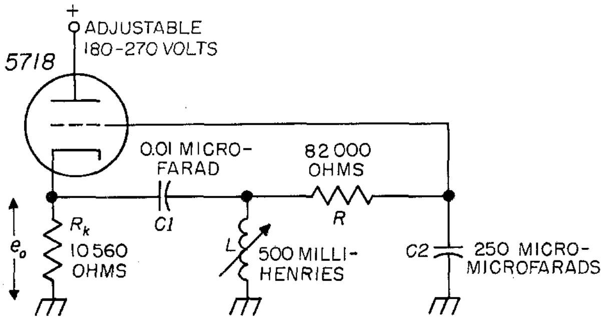


Figure 6—Experimental cathode-follower oscillator using a phase-shifting network consisting of resistance, inductance, and capacitance. The heater was operated with alternating current. The plate supply had negligible alternating-current impedance.

By varying R, L, $C1$, and $C2$, the oscillator was made to cover the frequency range from 41 to 6500 cycles.

2.2 Experimental Colpitts Type

The Colpitts oscillator of Figure 7, which is a modification of the cathode-follower amplifier given in reference 3, was found to give a strong distorted output under the conditions indicated.

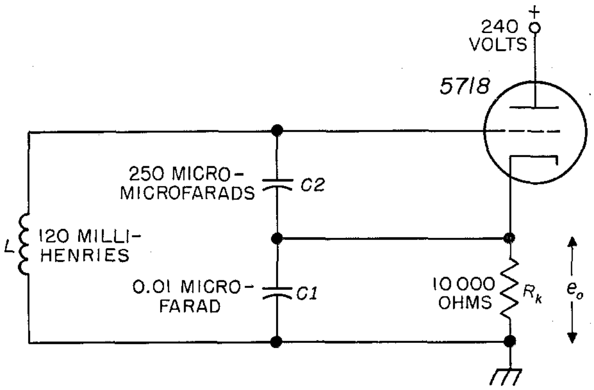


Figure 7—Experimental Colpitts oscillator. The heater was operated with alternating current.

The frequency of oscillation was 24 500 cycles, and the root-mean-square voltage output across the cathode resistor was 6.5 volts. No attempt was made to adjust circuit parameters for best output waveform. The waveform can be considerably improved by adjusting the gain of the cathode-follower so as just to maintain oscillations.

2.3 Experimental Resistance–Capacitance

Figure 8 shows a cathode-follower oscillator using only resistances and capacitances for phase shifting.

Grading of the elements in the phase-shifting network was necessary to reduce the required cathode-follower gain. The use of equal elements in a phase-shifting network of three sections demands a gain of about 58/60 out of the cathode-follower amplifier. This gain is very difficult to obtain with existing vacuum tubes.

The oscillator of Figure 8 had a frequency of oscillation of 137 cycles with a root-mean-square output at the cathode of 20.5 volts. The waveform of the output was approximately sinusoidal.

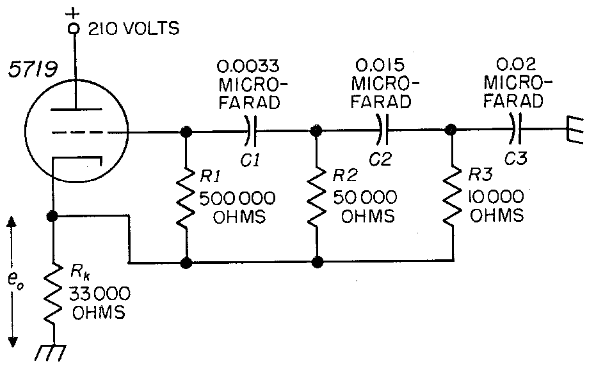


Figure 8—Experimental cathode-follower oscillator using a phase-shifting network of resistances and capacitances The heater was operated with alternating current.

No accurate measurements were made of frequency stability with respect to plate-voltage variations, but there are indications that less than 0.1 percent of frequency change results for a 10-percent increase in plate voltage at some plate voltages. From no load to a cathode load of 43 000 ohms, the oscillation frequency changed from 137 to 161 cycles while the root-mean-square output voltage dropped from 20.5 volts to 7.5 volts.

3. *Concluding Remarks*

The cathode-follower oscillator of the phase-shift type is very stable with respect to supply-voltage variations and is reasonably stable with changes in cathode loading. The location of the phase-shifting network in the cathode circuit should result in an oscillator inherently more stable than the standard phase-shift oscillator. Further development work is required to establish this point more definitely.

Telephone Statistics of the World*

FOR THE SIXTH consecutive year, 5 million telephones were added throughout the world, making a total of more than 89 million on January 1, 1954. More than half of of the world's telephones were in the United States, where some 5100 privately owned and operated systems gave service to almost one out of every three individuals. In Europe, which had some 28 percent of the world's telephones, mostly under public operation, there was a telephone for roughly one out of every 25 individuals.

For the purpose of this compilation, only those telephones that can be connected to a commercial public system are counted. Eleven countries reported more than 1 million telephones in service on January 1, 1953; United States, United Kingdom, Canada, Western Germany, France, Japan, Sweden, Italy, Australia, Switzerland, and Argentina. Of of the world's principal countries, 8 had more than 15 telephones per 100 of the population: United States (31.3), Sweden (27.7), Canada (24.0), Switzerland (21.9), New Zealand (21.9), Denmark (18.7), Australia (16.1), and Norway (15.8).

New York, with more telephones than any other city, had twice as many as Greater London, which ranked second. On a per-capita basis, Washington, District of Columbia, led among the world's large cities with 62.8 telephones per 100 population and Stockholm, Sweden, was first outside the United States with 51.9.

A subdivision in certain of the tables shows the number of telephones operated under private and government ownership. The latter category has reference to municipal and state, as well as national, ownership. Although the American Telephone and Telegraph Company and its subsidiaries operated about 80 percent of the United States' 50 372 972 telephones on January 1, 1954, there were over 5000 other privately owned companies that furnished service in the United States.

The statistics in this compilation are based on questionnaires sent to the telephone administrations of the various countries throughout the world.

TELEPHONES IN CONTINENTAL AREAS

Partly estimated; statistics reported as of other dates have been adjusted to January 1, 1954

Continental Area	Total Telephones			Privately Owned		Automatic (Dial)	
	Number	Percent of Total World	Per 100 Population	Number	Percent of Total	Number	Percent of Total
North America	54 000 600	60.5	30.6	53 460 400	99.0	41 603 500	77.0
Middle America	670,000	0.8	1.2	597 700	89.2	502 000	74.9
South America	2 245 500	2.5	1.9	1 054 500	47.0	1 808 200	80.5
Europe	25 400 700	28.5	4.3	3 982 700	15.7	18 703 800	73.6
Africa	1 181 200	1.3	0.5	22 200	1.9	819 300	69.4
Asia	3 662 200	4.1	0.3	2 729 800	74.5	1 696 700	46.3
Oceania	2 039 800	2.3	14.6	143 500	7.0	1 348 100	66.1
World	89 200 000	100.0	3.6	61 990 800	69.5	66 481 600	74.5
United States	*50 372 972*	*56.5*	*31.3*	*50 372 972*	*100.0*	*39 100 000*	*77.6*

* Abridgement from a booklet issued by the American Telephone and Telegraph Company; New York, New York.

TELEPHONES IN COUNTRIES OF THE WORLD AS OF JANUARY 1, 1954

Country		Total Telephone	Per 100 Population	Percent Automatic (Dial)	Ownership	
					Private	Government
NORTH AMERICA						
Alaska		23 533	14.71	74.3	6 608	16 925
Canada		3 603 900	24.01	69	3 080 780	523 120
Greenland		0	—	—	—	—
St. Pierre and Miquelon		208	4.52	0	0	208
United States		50 372 972	31.27	77.6	50 372 972	0
MIDDLE AMERICA						
Bahamas		5 866	6.82	99.3	0	5 866
Barbados		5 148	2.31	97.3	5 148	0
Bermuda		7 600	18.54	100	7 600	0
British Honduras		854	1.17	0	45	809
Canal Zone	(1) (2)	6 965	24.02	100	0	6 965
Costa Rica		10 600	1.18	0	10 300	300
Cuba		141 055	2.38	89.8	141 055	0
Dominican Republic		8 359	0.36	81.6	8 218	141
El Salvador		15 000	0.72	75	0	15 000
Guadeloupe		1 362	0.50	0	0	1 362
Guatemala		6 428	0.21	84.7	0	6 428
Haiti		4 182	0.12	91	0	4 182
Honduras		7 000	0.44	60	0	7 000
Jamaica		18 020	1.20	97.4	18 020	0
Leeward Islands		902	0.75	0	0	902
Martinique		3 047	1.10	66.2	0	3 047
Mexico		330 221	1.16	72.1	328 675	1 546
Netherlands Antilles		6 905	3.79	96.7	0	6 905
Nicaragua	(3)	3 500	0.30	0	0	3 500
Panama		16 182	1.85	76.9	15 711	471
Puerto Rico		47 367	2.11	52.7	43 909	3 458
Trinidad and Tobago		18 997	2.80	86.3	18 997	0
Virgin Islands (United States)		1 700	6.30	0	0	1 700
Windward Islands						
Dominica		307	0.54	0	0	307
Grenada		875	1.07	0	0	875
Saint Lucia		395	0.47	0	0	395
Saint Vincent		422	0.60	0	0	422
Total		1 999	0.68	0	0	1 999
SOUTH AMERICA						
Argentina		1 001 158	5.39	79.8	75 430	925 728
Bolivia		11 110	0.35	94.3	11 110	0
Brazil		679 540	1.22	82.1	679 540	0
British Guiana		3 908	0.87	7.5	0	3 908
Chile		145 139	2.36	69	144 649	490
Colombia		128 970	1.05	87.6	0	128 970
Ecuador		11 500	0.33	60.9	1 500	10 000
Falkland Islands		319	13.87	0	0	319
French Guiana		336	1.16	0	0	336
Paraguay	(3)	5 800	0.38	86.2	0	5 800
Peru		58 017	0.63	83.4	58 017	0
Surinam		2 741	1.19	91.1	0	2 741
Uruguay		104 510	4.05	75	0	104 510
Venezuela		92 420	1.70	93.7	84 240	8 180
EUROPE						
Albania	(3)	1 700	0.14	0	0	1 700
Andorra	(3)	100	2.00	0	0	100
Austria		458 006	6.58	80	0	458 006
Belgium		777 340	8.84	77.2	0	777 340
Bulgaria	(3)	61 000	0.82	43	0	61 000
Channel Islands						
Guernsey and Dependencies		9 568	20.80	0	0	9 568
Jersey		12 174	21.36	0	0	12 174
Total		21 742	21.11	0	0	21 742
Czechoslovakia	(4)	350 708	2.88	59.4	0	350 708
Denmark		825 879	18.73	39.6	726 144	99 735
Finland		408 531	9.81	62.2	350 700	57 831
France		2 768 951	6.45	64.6	0	2 768 951
Germany, Democratic Republic	(3)	250 000	1.42	60	0	250 000
Germany, Federal Republic	(5)	3 255 971	6.61	89.8	0	3 255 971
Gibraltar		1 399	6.08	89.5	0	1 399
Greece		104 237	1.32	93.7	0	104 237
Hungary	(3)	122 000	1.27	74	0	122 000

(1) Excluding telephone systems of the military forces.
(2) June 30, 1953.
(3) Data partly estimated.
(4) January 1, 1948 (latest official statistics).
(5) March 31, 1954.
(6) January 1, 1936 (latest official statistics). On January 1, 1954, nine cities connected.
(7) Includes the Isle of Man, but not the Channel Islands.

Country	Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership	
				Private	Government
EUROPE—*Continued*					
Iceland	23 774	15.85	64.4	0	23 774
Ireland	103 798	3.52	68.1	0	103 798
Italy	1 774 462	3.74	93.7	1 774 462	0
Liechtenstein	2 334	16.67	100	0	2 334
Luxemburg	28 150	9.26	72.5	0	28 150
Malta and Gozo (5)	7 869	2.46	0	0	7 869
Monaco	6 020	28.67	100	0	6 020
Netherlands	919 572	8.70	91.6	0	919 572
Norway (2)	530 827	15.79	61.5	70 400	460 427
Poland (3)	240 000	0.93	67	0	240 000
Portugal	208 143	2.40	57.4	133 842	74 301
Rumania (3)	141 000	0.85	60	0	141 000
Saar	44 938	4.66	99.6	0	44 938
San Marino (3)	300	2.31	0	0	300
Spain	903 097	3.15	80.3	888 334	14 763
Sweden	1 994 378	27.73	71	838	1 993 540
Switzerland	1 074 216	21.90	98.1	0	1 074 216
Trieste	36 648	9.99	97	35 898	750
Union Soviet Socialist Republics (6)	861 181	0.52	19.9	0	861 181
United Kingdom (5) (7)	6 139 229	12.15	75.3	0	6 139 229
Yugoslavia (3)	149 000	0.87	70	0	149 000
AFRICA					
Algeria	116 889	1.23	75.7	0	116 889
Anglo-Egyptian Sudan	11 648	0.13	76.5	0	11 648
Basutoland	509	0.09	4.9	0	509
Bechuanaland Protectorate	131	0.04	0	0	131
Belgian Congo and Ruanda-Urundi	11 451	0.07	62.6	0	11 451
British East Africa					
Kenya	20 407	0.35	72.3	0	20 407
Tanganyika	7 949	0.10	55.3	0	7 949
Uganda	6 328	0.12	77.6	0	6 328
Total	34 684	0.18	69.4	0	34 684
British West Africa					
Gambia	386	0.14	99.2	0	386
Gold Coast	9 581	0.21	37.1	0	9 581
Nigeria	15 063	0.05	38.5	0	15 063
Sierra Leone	1 573	0.08	84	0	1 573
Total	26 603	0.07	41.6	0	26 603
Comoro Islands	0	—	—	—	—
Egypt	135 388	0.62	79.5	0	135 388
Ethiopia	4 776	0.03	74.6	0	4 776
French Cameroon	2 258	0.07	0	0	2 258
French Equatorial Africal	1 851	0.04	43.5	0	1 851
French Somaliland	570	0.90	0	0	570
French Togo	611	0.06	0	0	611
French West Africa	17 860	0.10	39.7	0	17 860
Liberia	487	0.02	100	0	487
Libya	6 055	0.49	80.6	0	6 055
Madagascar	7 650	0.17	46.8	0	7 650
Mauritius	5 881	1.12	6	0	5 881
Morocco					
French Zone	80 651	1.01	77.6	0	80 651
Spanish Zone	8 076	0.78	56.4	8 076	0
Tangier Zone	9 546	5.19	97.4	9 297	249
Total	98 273	1.07	77.8	17 373	80 900
Nyasaland Protectorate	2 407	0.10	86.2	0	2 407
Portuguese Africa					
Angola	2 268	0.05	85.5	0	2 268
Cape Verde Islands	126	0.08	0	0	126
Mozambique	6 236	0.11	76.6	0	6 236
Portuguese Guinea	257	0.05	91.1	0	257
South Tome and Principe	277	0.46	0	0	277
Total	9 164	0.08	75.9	0	9 164
Rhodesia, Northern	6 077	0.30	94.5	0	6 077
Rhodesia, Southern	37 063	1.62	78.7	0	37 063
Réunion	3 557	1.35	0	0	3 557
Saint Helena	89	1.78	0	0	89
Seychelles and Dependencies	50	0.14	0	0	50
Somalia	856	0.06	0	0	856
Somaliland Protectorate	225	0.03	0	0	225
South West Africa	6 733	1.57	49.2	0	6 733
Spanish Guinea	525	0.26	67.4	525	0
Spanish North Africa	4 320	3.02	58.1	4 320	0

Telephones in Countries of the World as of January 1, 1954—*Continued*

Country		Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership	
					Private	Government
AFRICA—*Continued*						
Spanish West Africa						
Ifni		114	0.32	0	0	114
Spanish Sahara		52	0.10	0	0	52
Total		166	0.19	0	0	166
Swaziland		566	0.28	1.9	0	566
Tunisia		30 666	0.84	59.4	0	30 666
Union of South Africa	(5)	606 152	4.55	68.4	0	606 152
Zanzibar and Pemba	(3)	1 100	0.40	0	0	1 100
ASIA						
Aden Colony		1 550	1.55	100	0	1 550
Aden Protectorate		0	—	—	—	—
Afghanistan		5 822	0.05	30	0	5 822
Bahrain		1 045	0.93	100	1 045	0
British Borneo						
Brunei		125	0.25	0	0	125
North Borneo		971	0.28	66.9	0	971
Sarawak		720	0.12	0	0	720
Total		1 816	0.18	35.8	0	1 816
Burma		7 136	0.04	0	0	7 136
Cambodia		1 884	0.05	0	0	1 884
Ceylon		22 855	0.28	94	0	22 855
China (not including Formosa)	(4)	244 028	0.05	72.9	94 945	149 083
Cyprus		8 145	1.60	79.4	8 145	0
Formosa		34 586	0.41	47	0	34 586
French India		78	0.02	100	0	78
Hong Kong	(2)	40 434	1.84	100	40 434	0
India	(5)	210 868	0.06	50.7	1 966	208 902
Indonesia		63 977	0.08	3.8	0	63 977
Iran		39 300	0.19	57.3	0	39 300
Iraq	(5)	28 010	0.50	67.5	0	28 010
Israel		47 430	2.84	89.3	0	47 430
Japan	(5)	2 594 506	2.96	40.2	2 594 506	0
Jordan		6 536	0.47	75.6	0	6 536
Korea, South		28 461	0.13	31.7	0	28 461
Kuwait		988	0.66	84.8	988	0
Laos		286	0.02	0	0	286
Lebanon		24 368	1.81	79.7	0	24 368
Malaya		40 259	0.69	57.1	1 250	39 009
Maldive Islands		0	—	—	—	—
Muscat and Oman		137	0.02	100	137	0
Netherlands New Guinea		772	0.11	0	0	772
Pakistan		27 886	0.04	56	0	27 886
Philippine Republic		41 807	0.19	64.4	34 753	7 054
Portuguese Asia						
Macao		1 789	0.89	99.5	0	1 789
Portuguese India		224	0.03	0	0	224
Portuguese Timor		312	0.07	0	0	312
Total		2 325	0.18	76.6	0	2 325
Qatar		105	0.58	100	105	0
Ryukyu Islands						
Okinawa		1 500	0.23	0	0	1 500
Others		300	0.10	0	0	300
Total		1 800	0.19	0	0	1 800
Saudi Arabia		7 743	0.13	0	0	7 743
Singapore		28 895	2.52	100	28 895	0
Syria		27 155	0.75	86.5	0	27 155
Thailand		8 020	0.04	100	0	8 020
Turkey		113 609	0.50	79.4	0	113 609
Viet-Nam		13 980	0.06	42.8	0	13 980
Yemen		0	—	—	—	—
Other Places	(3)	32 000	0.16	25	0	32 000
OCEANIA						
American Samoa		323	1.79	100	0	323
Australia		1 432 776	16.07	65.5	0	1 432 776
French Oceania						
French Settlements		710	1.13	0	0	710
New Caledonia and Dependencies		1 830	2.82	0	0	1 830
Total		2 540	1.98	0	0	2 540
Guam		4 412	5.81	90.7	0	4 412
Hawaii		143 461	27.48	96.1	143 461	0
Nauru		0	—	—	—	—
New Hebrides Condominium		108	0.22	0	0	108
New Zealand	(5)	456 289	21.86	59.6	0	456 289

Telephones in Countries of the World as of January 1, 1954—*Continued*

Country	Total Telephones	Per 100 Population	Percent Automatic (Dial)	Ownership	
				Private	Government
OCEANIA—*Continued*					
Pacific Islands (British)					
Fiji Islands	3 210	1.00	0	0	3 210
Gilbert and Ellice Islands	113	0.30	74.3	84	29
Pitcairn Island	0	—	—	—	—
Solomon Islands Protectorate	0	—	—	—	—
Tonga (Friendly) Islands	245	0.46	0	0	245
Total	3 568	0.69	2.4	84	3 484
Pacific Islands (United States Administration)					
Caroline Islands	177	0.42	0	0	177
Mariana Islands (less Guam)	0	—	—	—	—
Marshall Islands	457	4.15	0	0	457
Total	634	1.07	0	0	634
Papua and New Guinea	2 943	0.20	0	0	2 943
Western Samoa	551	0.61	0	0	551

Telephone Conversations for the Year 1953

Conversation data were not available for all countries

Country		Number of Conversations in Thousands			Conversations Per Capita
		Local	Toll	Total	
Algeria		59 000	20 400	79 400	8.5
Argentina		3 106 400	39 100	3 145 500	171.1
Australia		997 700	78 400	1 076 100	121.9
Belgium		444 000	69 700	513 700	58.5
Brazil		2 182 900	38 200	2 221 100	39.8
Canada		5 948 000	131 900	6 079 900	411.3
Ceylon		54 800	3 700	58 500	7.2
Chile		383 500	22 400	405 900	66.8
Colombia		444 200	4 800	449 000	37.3
Cuba		811 200	5 200	816 400	140.0
Denmark		978,200	166 000	1 144 200	261.7
Egypt		271 900	10 900	282 800	12.9
Finland		509 200	77 300	586 500	141.7
France		1 444 900	467 800	1 912 700	44.7
Germany, Federal Republic	(1)	2 077 900	447 900	2 525 800	51.6
Greece		271 900	4 600	276 500	35.2
Iceland		54 900	1 400	56 300	375.3
Ireland		80 300	12 700	93 000	31.6
Israel		74 100	3 500	77 600	47.2
Italy		2 787 000	177 200*	2 964 200	62.7
Jamaica		44 000	700	44 700	30.3
Japan	(1)	8 761 000	568 600	9 329 600	107.4
Luxemburg		10 300	8 600*	18 900	62.6
Malaya		117 900	11 800	129 700	22.7
Mexico		749 300	9 800	759 100	27.1
Morocco		71 400	15 300	86 700	8.7
Norway	(2)	465 200	53 200	518 400	154.9
Peru		211 600	2 900	214 500	23.7
Philippine Republic		280 000	400	280 400	13.3
Portugal		189 800	34 700	224 500	26.0
Puerto Rico		94 300	3 200	97 500	43.5
Saar		55 600	1 800	57 400	59.5
Spain		2 249 000	73 300	2 322 300	81.4
Sweden	(3)	2 096 700	123 900	2 220 600	308.8
Switzerland		437 400	358 400*	795 800	162.9
Tunisia		16 600	7 300	23 900	6.7
Turkey		159 500	5 500	165 000	7.4
Union of South Africa	(1)	731 100	47 000	778 100	59.0
United Kingdom	(1) (4)	3 404 200	279 600	3 683 800	72.9
United States		59 355 000	2 185 000	61 540 000	385.3
Uruguay		306 600	4 600	311 200	123.2
Venezuela		388 600	13 600	402 200	73.9

(1) Year ended March 31, 1954.
(2) Year ended June 30, 1953.
(3) Year ended June 30, 1954.
(4) Includes the Isle of Man, but not the Channel Islands.
* Three-minute units.

United States Patents Issued to International Telephone and Telegraph System February - April, 1955

United States patents numbering 28 were issued between February 1 and April 30, 1955 to companies in the International System. The inventors, titles, and numbers of these patents are given below; summaries of several that are of more-than-usual interest are included.

R. P. Arthur and E. L. Earle, Electrical Terminal Pin Block, 2 703 394.

R. M. Barnard and S. B. Buckley, Metallizing the Surface of Ceramic Bodies, 2 706 682.

A. Brown and T. A. Benner, Panel Control Key, 2 702 321.

M. den Hertog, Two-Stage Group-Selector Circuit, 2 706 748.

L. A. de Rosa, Radio Detection System, 2 703 401.

S. H. M. Dodington, Pulse Repeaters, 2 706 773.

H. Grayson, T. S. McLeod, and V. Vernon-Smith, Frequency-Modulation Carrier-Current System, 2 703 865.

P. W. Hemminger and J. I. Bellamy, Switching System using Capacitor Storage of Digits, 2 700 071.

K. Klinkhammer, Group Selector of Relay Systems, 2 705 743.

H. R. Kough, Electrical Terminal and Terminal Assembly, 2 707 274.

J. S. Macmullan and A. Mortlock, Variable Limiter and Noise Suppressor for *RK9* System, 2 706 776.

F. P. Mason and A. T. Watts, Facsimile Telegraph System, 2 705 739.

W. Master, Dry-Contact Rectifier, 2 707 251.

B. McAdams, Adaptor Circuit, 2 707 211.

K. Muhmer, Contact Spring Assembly, 2 705 602.

A. J. Mullarkey, Axle Counting Arrangements, 2 701 301.

H. L. Overman, Circuit-Component Selecting Unit, 2 706 252.

H. Pletscher, Twinplex Telegraph Signal Receiver, 2 701 276.

W. H. P. Pouliart and F. M. Michiels, Receiving Circuit Arrangement, 2 704 361.

S. J. Powers and W. F. Bonner, Germanium Diode, 2 703 856.

N. Prietzel, Central Switch for Pneumatic Dispatch Systems, 2 703 687.

N. H. Saunders, Conference Provision for Telephone Switchboards, 2 700 072.

K. Seiler, Method of Making Surface-Type and Point-Type Rectifiers, 2 701 216.

K. Seiler and L. Fedotowski, Semiconductor Rectifier or Amplifier of any Desired Surface Profile, 2 702 361.

A. T. Starr and G. King, Switches for High-Frequency Waves, 2 705 776.

V. J. Terry, Electronic-Telegraph Relay Circuits, 2 701 277.

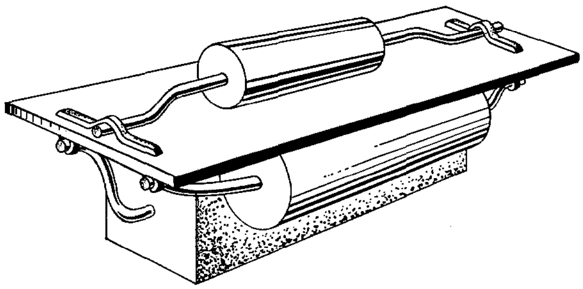
B. V. Thompson and E. W. Deiss, Electrical Coupling Apparatus, 2 707 271.

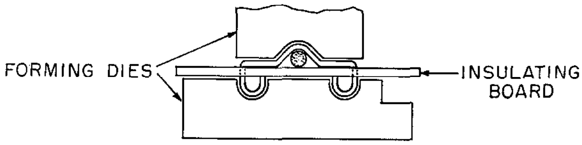
C. deB. White and K. A. Matthews, Amplifier Employing Semiconductor, 2 701 281.

Electrical Terminal and Terminal Assembly

H. R. Kough
2 707 274—April 26, 1955

An electrical terminal in the form of a staple-like member with a bridge portion having a single connecting loop on one side of an insulating sheet, the ends of the staple being curled toward the sheet to provide two connection loops of approximately semicircular outline.





Adaptor Circuit

B. McAdams
2 707 211—April 26, 1955

An adaptor circuit for interconnecting voice-frequency telephone lines with a carrier-frequency system. This development automatically converts the direct-current pulses of the normal dialing and supervisory signals into carrier supervisory signals by means of simple switch connections applying different potentials from separate sources to dial trunk leads under control of dial signals.

Receiving Circuit Arrangement

W. H. P. Pouliart and F. M. Michiels
2 704 361—March 15, 1955

A receiving circuit for use in magnetic-drum storage systems, in which the waveform output alternates to give positive and negative variations. Separate circuits respectively accept the positive and negative parts of the output, each circuit having a short-time-contant storage circuit. Connections between the charging path of each storage circuit and the discharging path of the other storage circuit are included to reduce spurious signals that may occur.

Method of Making Surface-Type and Point-Type Rectifiers and Crystal-Amplifier Layers from Elements

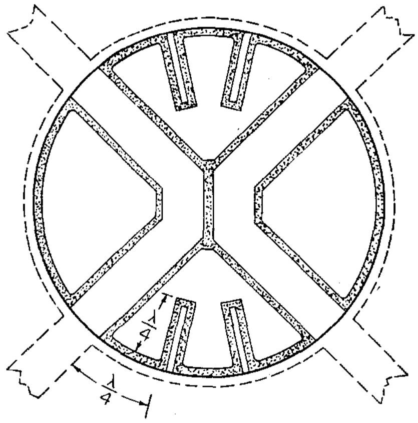
K. Seiler
2 701 216—February 1, 1955

A method of forming semiconductors for point-contact rectifiers and amplifiers to avoid excess impurities that may cause short-circuits in the region adjacent to the point contacts but to preserve adequate conductivity in adjacent conducting zones. The method includes making a vapor deposition of the semiconductor on a base-plate, the conductor substance having a mixture of acceptor or donor impurities and reducing the proportion of these impurities during deposition.

Switches for High-Frequency Waves

A. T. Starr and G. King
2 705 776—April 5, 1955

A switch for waveguides having a rotor carrying passages to connect to different fixed waveguides and having an arrangement to eliminate



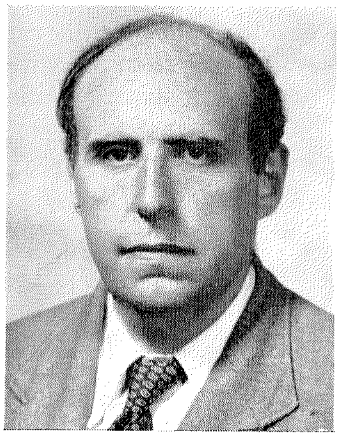
the need of spring contacts between the rotor and fixed waveguides. The switch includes slots cut radially into the rotor drum from the outer circumference, the slots being positioned equal distances from the openings of the passageways and a peripheral distance from these openings of approximately a quarter wavelength.

Contact-Spring Assembly for Automatic Switching of Message Carriers in Pneumatic-Tube Systems

K. Muhmer
2 705 602—April 5, 1955

A pneumatic-tube system in which the message carriers are provided with marking rings set to convey a particular message as to destination or route to be followed, with a detecting arrangement for reading the message consisting of contact springs insulated from one another and held out of contact with the rings. A common actuating device for the contact includes a leaf spring and an actuating device for the leaf spring depresses the insulated springs into contact with the rings when the carrier is properly positioned for contact.

Contributors to This Issue



A. H. Beck

A. H. Beck was born in Norfolk, England, in 1916. He received a B.Sc. degree in engineering from University College, London.

In 1937, he joined the research staff of Henry Hughes and Sons. From 1941 to the end of the war, he was seconded to the Admiralty Signals Establishment and part of his theoretical work there resulted in the publication of his book on "Velocity Modulated Thermionic Tubes."

In 1947, he left Hughes to establish the Enfield valve laboratory of Standard Telephones and Cables, and is now head of the valve division of Standard Telecommunication Laboratories. His interests include microwave valves, special-purpose gas tubes, and vacuum physics applied to valves. His most-recent book is on "Thermionic Valves."

Mr. Beck is an Associate Member of the Institution of Electrical Engineers and a Member of the Institute of Radio Engineers.

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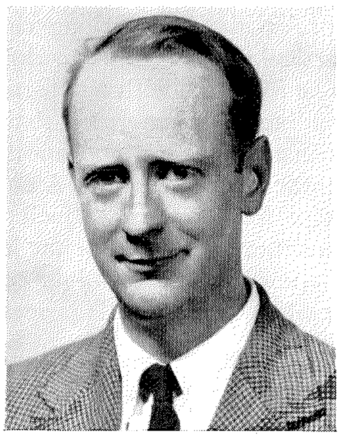

A. D. Brisbane

A. D. Brisbane was born in 1920 and received his early education at Bournemouth. In 1955, he obtained the City and Guilds Final Technological Certificate at Enfield College.

In 1938, he joined the Royal Air Force and after technical training was commissioned in 1942.

After demobilization, he entered the valve laboratory of Standard Telephones and Cables at Enfield, being transferred later to Standard Telecommunication Laboratories. His activities have chiefly concerned thermionic emission.

Mr. Brisbane received the Distinguished Flying Medal in 1941.

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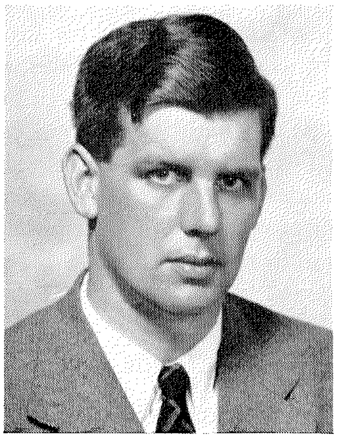

A. B. Cutting

A. B. Cutting was born in Stamford, England, in 1919. He obtained B.Sc. degrees in general science and mathematics from Birkbeck College of the University of London.

After nine years in the research laboratories of the General Electric Company, he joined the research staff of the Enfield valve laboratory of Standard Telephones and Cables in 1947. He was transferred to Standard Telecommunication Laboratories in 1954. His research work has been primarily on microwave valves.

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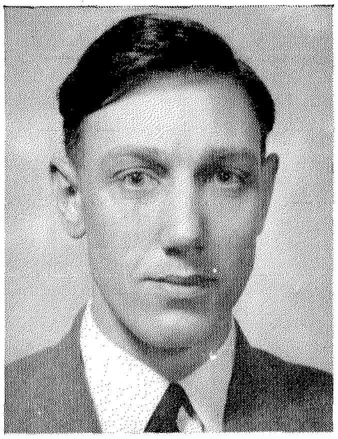
J. L. Goodwin

J. L. Goodwin was born at Chigwell, England, in 1920. He studied electrical engineering at the Borough Polytechnic and Twickenham Technical College, receiving the Higher National Certificate in 1947.

During the war, he did research work on underwater acoustics with the Admiralty and in 1948 he joined the staff



T. M. Jackson

of the transmission division of Standard Telephones and Cables. Mr. Goodwin reports in this issue on a loudspeaker intercommunicating system.

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

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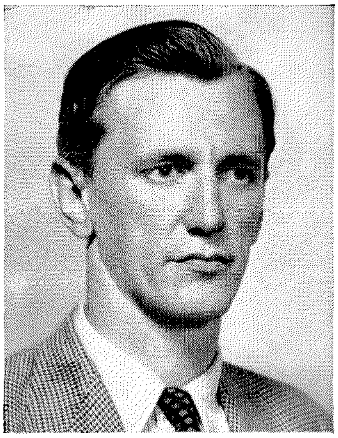
T. M. Jackson was born in Merioneth, Wales, in 1921.

He joined the Royal Air Force in 1940 and spent four years in experimental work on radar systems at the Government Telecommunications Research Establishment.

In 1946, he came to the Enfield valve laboratories of Standard Telephones and Cables and was transferred to Standard Telecommunication Laboratories in 1954. He has been mainly concerned with devices based on electrical discharges in gases.



G. King

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G. King was born in 1925 in London. After technical training at the Royal Technical College in Glasgow, he served the Royal Air Force as a radar fitter.

Later he obtained the Final Technological Certificate in telecommunication engineering from the City and Guilds of London.

On demobilization in 1947, he went to the Enfield valve laboratory of Standard Telephones and Cables and in 1954 was transferred to the engineering staff of Standard Telecommunication Laboratories. He is coauthor of the article in this issue on cathodes.

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J. Lytollis was born in 1924 in Skegness, Lincolnshire. He was educated at Worcester College, Oxford, and received a B.A. in physics from the Honour School of Natural Science in 1945.

After two years with the Philips group of companies and three years as an Instructor Officer in the Royal Navy, in 1951 he was employed by Standard Telephones and Cables in the Enfield valve laboratory. He was transferred to the valve division of Standard Telecommunication Laboratories in 1951. His work has been principally on gaseous discharge devices.

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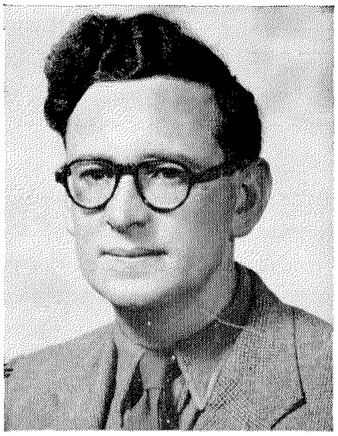
J. C. Samuels received the B.S. degree in electrical engineering from Poly-

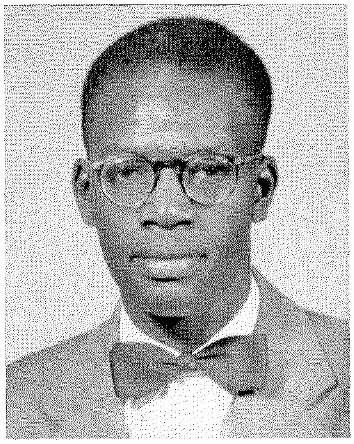


J. Lytollis

technic Institute of Brooklyn in 1948 and the M.S. degree in mathematics and physics from New York University in 1950. During the next two years, he did graduate work in mathematics at Case Institute of Technology. While at Case Institute and for two years thereafter, he worked on jet-propulsion systems for the National Advisory Committee for Aeronautics.

In 1952, he joined the engineering staff of Capehart–Farnsworth Company and is now with Farnsworth Electronics Company. He reports here on cathode-follower oscillators.

Mr. Samuels is a registered professional engineer in Indiana. He is a



J. C. Samuels

member of the Institute of Radio Engineers and a member of Tau Beta Pi and Eta Kappa Nu, both honorary societies.

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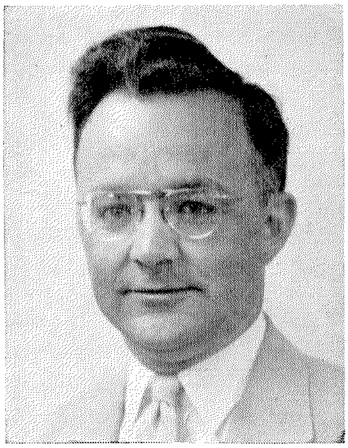
R. DONALD SINISH was born in Leavenworth, Kansas, on May 2, 1919. He received a B.S. degree in electrical engineering from Purdue University in 1942, where he also did graduate work in high-voltage generation and transmission.

From 1942 to 1944, he was employed by the General Electric Company on the design of high-voltage and special-purpose test equipment. In 1944, he joined the Farnsworth Television and Radio Corporation as field engineer in charge of installation and calibration of precision bombing equipment. His design experience includes radio and radar transmitters and receivers as well as precision test equipment for radar systems.



R. DONALD SINISH

Mr. Sinish is a Senior Member of the Institute of Radio Engineers, a Member of the American Institute of Electrical Engineers, and a registered professional engineer.

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Federal Telephone and Radio Company; Clifton, New Jersey

Kellogg Switchboard and Supply Company; Chicago, Illinois

Federal Electric Corporation; Clifton, New Jersey

International Standard Electric Corporation; New York, New York

International Standard Trading Corporation; New York, New York

Kellogg Credit Corporation; Chicago, Illinois

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British Commonwealth of Nations

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AUSTRALIA —

Standard Telephones and Cables Pty. Limited; Sydney

Silovac Electrical Products Pty. Limited; Sydney

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NEW ZEALAND — New Zealand Electric Totalisators Limited; Wellington

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BRAZIL — Standard Electrica, S.A.; Rio de Janeiro

CHILE — Compañía Standard Electric, S.A.C.; Santiago

CUBA — Standard Products Distributing Company; Havana

MEXICO — Standard Electrica de Mexico, S.A.; Mexico City

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Europe

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DENMARK — Standard Electric Aktieselskab; Copenhagen

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Le Matériel Téléphonique; Paris
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CHILE — Compañía de Teléfonos de Chile; Santiago

CUBA — Cuban American Telephone and Telegraph Company; Havana

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PUERTO RICO — Porto Rico Telephone Company; San Juan

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UNITED STATES OF AMERICA —

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All America Cables and Radio, Inc.; New York, New York
The Commercial Cable Company; New York, New York
Mackay Radio and Telegraph Company; New York, New York

ARGENTINA —

Compañía Internacional de Radio; Buenos Aires
Sociedad Anónima Radio Argentina; Buenos Aires (*Subsidiary of American Cable & Radio Corporation*)

BOLIVIA — Compañía Internacional de Radio Boliviana; La Paz

BRAZIL — Companhia Radio Internacional do Brasil; Rio de Janeiro

CHILE — Compañía Internacional de Radio, S.A.; Santiago

CUBA — Radio Corporation of Cuba; Havana

PUERTO RICO — Radio Corporation of Porto Rico; San Juan

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