ELECTRICAL COMMUNICATION

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

•

RAILWAYS MODERNIZE WITH CENTRALIZED SUPERVISORY CONTROL

STYRENE JOINTS AND TERMINATIONS

DESIGN OF A LORAN TRANSMITTER

MAGNETOSTRICTIVE DELAY LINE

MEASUREMENT OF IMPEDANCE WITH THE IMPEDOMETER

PROPERTIES AND TESTS OF MAGNETIC POWDERS AND CORES

ANECHOIC CHAMBER FOR ACOUSTIC MEASUREMENTS

INTERNATIONAL TELEPHONE AND TELEGRAPH

Volume 28 **MARCH, 1951** **Number 1**

ELECTRICAL COMMUNICATION

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

H. P. WESTMAN, Editor

F. J. MANN, Managing Editor J. E. SCHLAIKJER, Assistant Editor

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

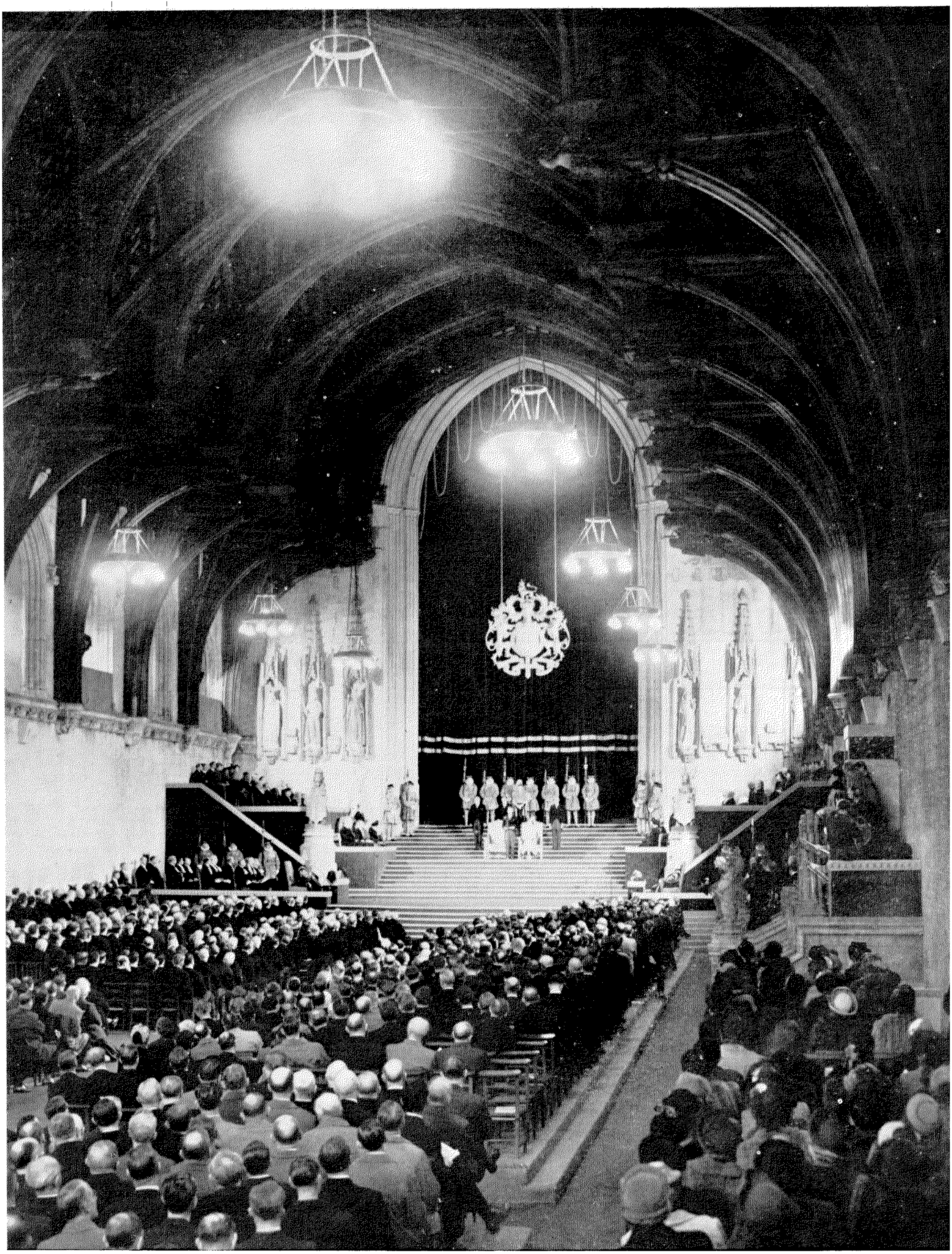
Sosthenes Behn, Chairman William H. Harrison, President
Charles D. Hilles, Jr., Vice President and Secretary

Subscription, $2.00 per year; single copies, 50 cents

Electrical Communication is indexed in Industrial Arts Index

Volume 28 MARCH, 1951 Number 1

CONTENTS



His Majesty, King George VI, addressed both Houses of Parliament on the occasion of the opening on October 26, 1950, of the new chamber of the House of Commons. The original hall was built in the eleventh century and was badly damaged in the last major air assault on London of the second world war. Standard Telephones and Cables was privileged to supply the public-address system for the dedication ceremonies.

Railways Modernize with Centralized Supervisory Control*

By C. GORDON WHITE

Standard Telephones and Cables, Limited; London, England

RECENTLY a number of railway electrification schemes conceived in the late 1930's but delayed by the war have been, or are being, completed, whilst others that have been operating for many years past have been modernized. These schemes, designed to increase the traffic-carrying capacity of the various routes, are the most up-to-date in the world, and it is significant that, without exception, their power-supply networks are supervised and controlled by centralized control systems. This, besides promoting an operational efficiency obtainable by no other means, also contributes to the clamant need for economy both in man-power and in current expenditure, since sub-stations may be wholly unattended.

The latest example of railway electrification is that of the Liverpool Street–Shenfield section of British Railways.[1] This route of approximately 20 miles (32 kilometres) serves the densely populated eastern suburbs of London and is the most heavily loaded suburban service in the world. Besides the suburban traffic, this line also carries nearly all the main-line freight and passenger traffic to the eastern districts of England, and, because of this, it has been planned for main-line electrification. The route served by the initial stage of electrification, its relation to neighbouring routes of the London Underground system,

* Reprinted from *British Engineering*, v. 33, pp. 218–229; August, 1950.

[1] See *British Engineering*, v. 32, p. 900; December, 1949.

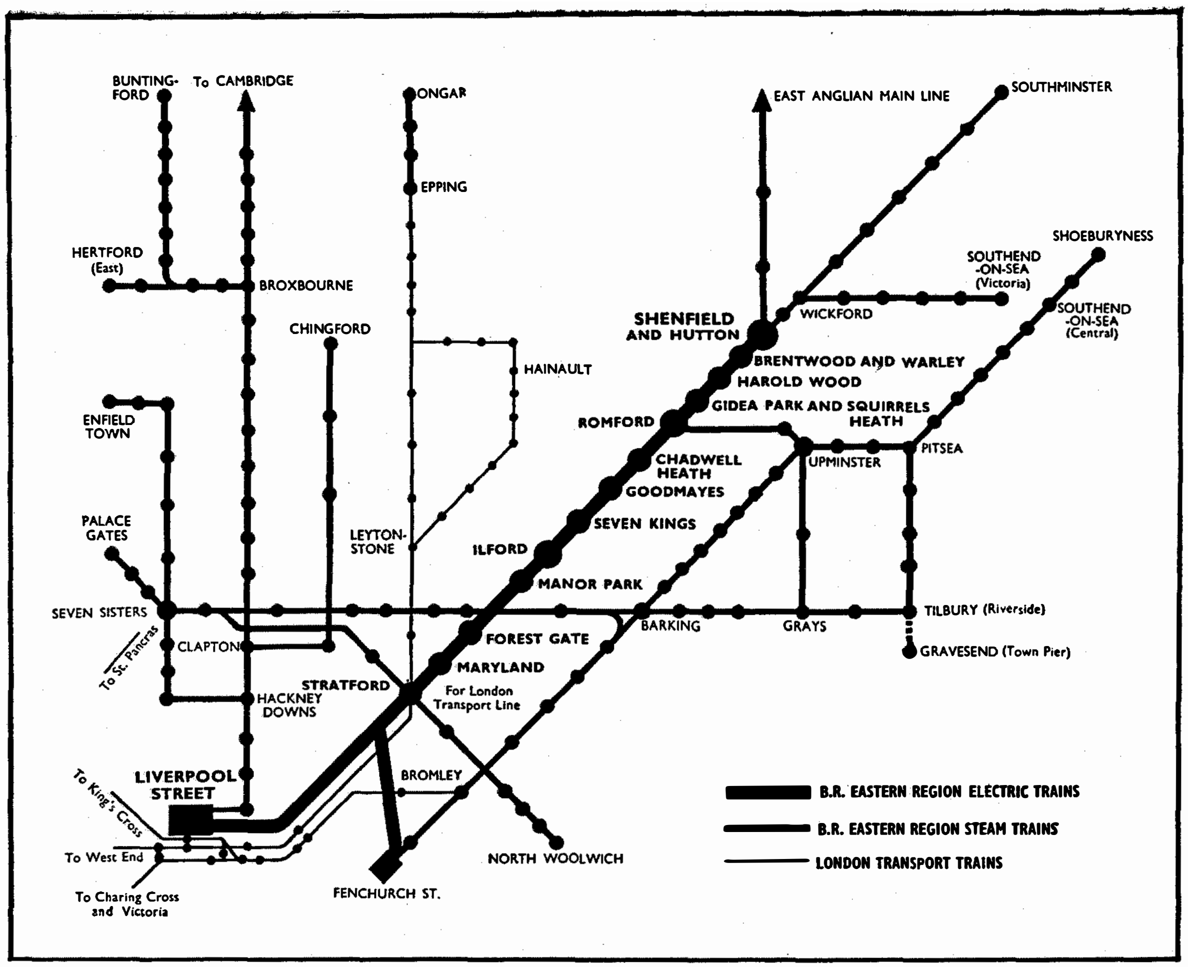
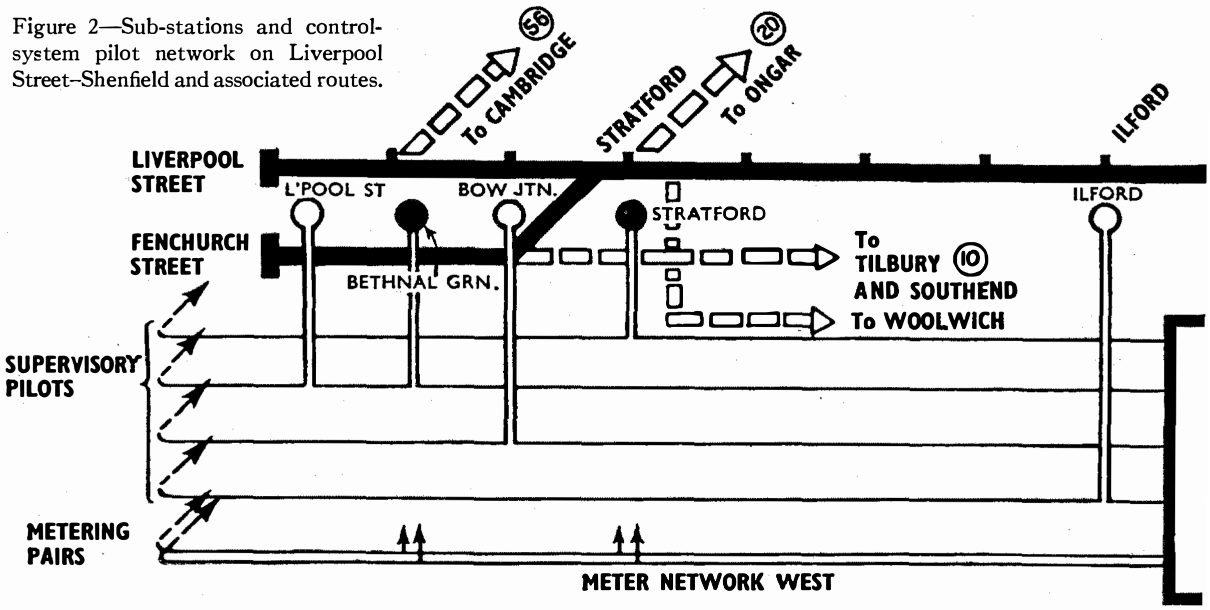


Figure 1—British Railways Eastern Region in relation to the London Transport Railway System. (*Courtesy of British Railways.*)

Figure 2—Sub-stations and control-system pilot network on Liverpool Street–Shenfield and associated routes.



and the main routes to the Eastern Counties can be seen in Figure 1. The location of rectifier sub-stations and track-sectioning cabins between London and Shenfield is indicated in Figure 2.

In planning the remote-control system for this network, attention had to be paid to the probable ultimate scope of the electrification scheme and to the routes likely to be included. These considerations showed that the initial scheme would need 11 remote stations and that ultimately about 33 stations might be required. Because of this, the choice of control system almost inevitably fell on the *common-equipment system*, a system that was developed some 10 years before for just such a set of conditions, e.g., initially, centralized control of comparatively few sub-stations, but with facilities for bringing, as required, many more under control without interrupting the system that must be continuously in operation.

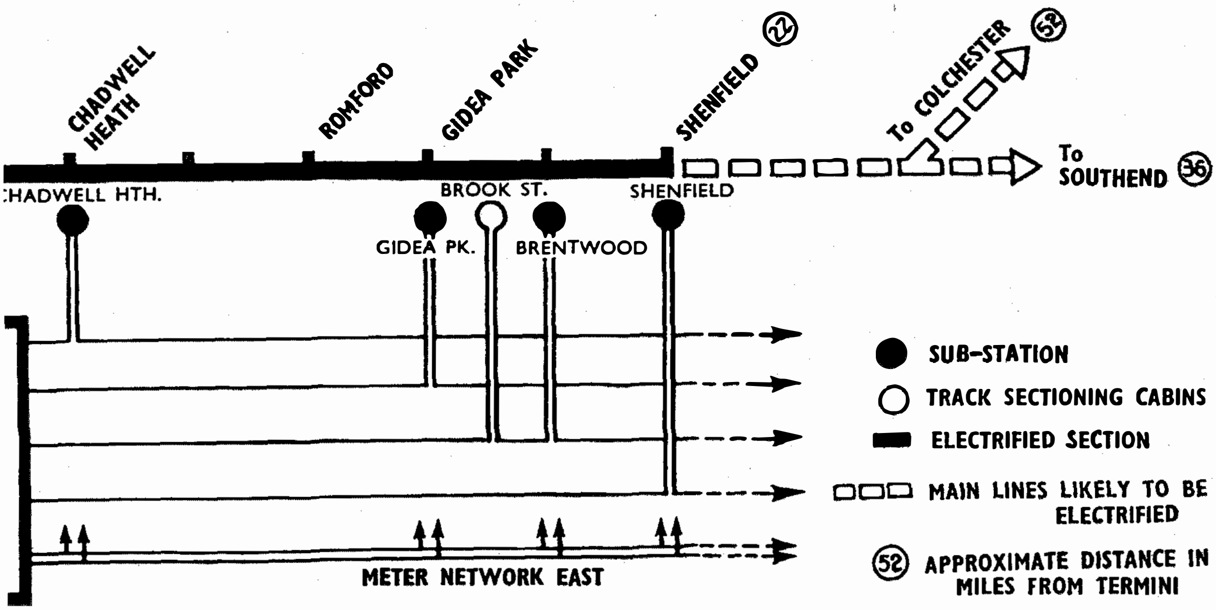
1. Common-Equipment System

The *Standard* common-equipment system (C.E.S.) or common-diagram system (C.D.S.), as it is named, when applied to a city distribution network, was first supplied in 1942 to the Manchester Corporation Electricity Department for the centralized control of their 6·6- and 11-kilovolt distribution networks.[2] Serving initially some 90 sub-stations, the supervisory system was required to cater ultimately for over 600 stations, and since it was manifestly impossible to employ the conventional individual-panel system in which each sub-station has its own mimic panel in the control station, it was decided to design a simple control and indication panel to be common to all stations, but which could be switched as required to any sub-station. This revolutionary approach meant, of course that only one station could be controlled at any given time, but this limitation was accepted on the grounds that rarely, if ever, is it necessary to effect controls simultaneously at more than one station.

Although the needs of a city distribution system differ in many respects from those of an electric traction system, the basic switching problems are similar and involve kindred techniques for their solution so that the common-diagram system in altered guise provides a ready solution to this related but different problem in remote power-gear switching.

The relationship between the main items of equipment that together constitute a typical supervisory system of the individual-panel type and of the common-equipment type, is shown in Figure 3. In the former, stations are shown group served by common code transmitting and receiving apparatus *TR* and provided each with its own control panel, *A*, *B*, *C*, etc., which mounts

[2] See *Journal of The Institution of Electrical Engineers*, Part II, v. 92, n. 28; August, 1945: n. 30; December, 1945: and v. 94, n. 37; February, 1947.



the necessary keys, lamps, and meters for the control and indication of each item of switchgear. Thus, whenever a switchgear unit is added at a station or a new station is connected, a major alteration at the control centre becomes necessary. Also, with a scheme comprising many remote stations, the space required for control-room panels constitutes an important consideration even with the high degree of miniaturization at present attained by the use of telephone-type equipment.

For a similar network, the common-equipment system requires only one set of code transmitting and receiving equipment *TR* and one master control and indication panel. This panel is designed to cater for any station, even the largest, and means are provided to switch it and its set of equipment to any station as required. As the common equipment that is initially installed serves for any number of remote stations, the only changes necessary at the control centre when remote stations are added concern the decoders *D* which, among other things, control the lamps identifying the calling stations.

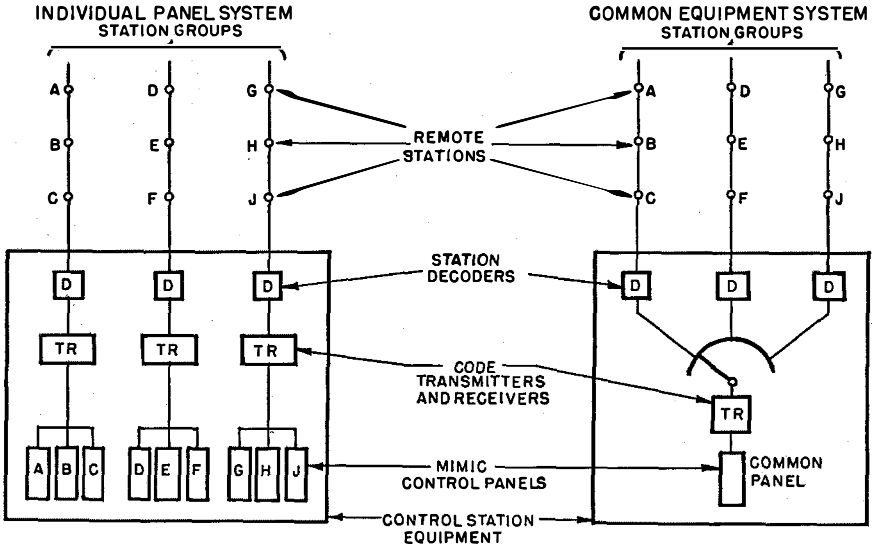


Figure 3—Comparison of the individual-panel and common-equipment systems. Each station group is on a party pilot circuit.

Thus changes at a later date do not materially affect the normal working of the system.

For railway work, where speed in dealing with emergencies is essential, it is considered desirable to use duplicate master panels and associated sets of equipment in order that two operators, acting simultaneously, may deal with two remote stations. A schematic diagram of the common-equipment system as designed for the Liverpool Street–Shenfield scheme is shown in Figure 4.

Furthermore, railway engineers are agreed that an illuminated mimic diagram giving the up-to-date information of the system switching is most desirable. In this arrangement the position information received on the master control panels is manually transferred to the mimic diagram by the centrally located diagram-dressing keys on the designation panel. In fact the diagram may be regarded as an illuminated log-book and the diagram-dressing keys special tools for making entries as and when information is received.

2. *General Arrangement of System*

The capacity limit set to the control equipment of a common-equipment system is determined mainly by the number of remote stations that can be conveniently handled from a point served by two operators, and although this number must be to some extent a matter of experience and conjecture, consideration of previous installations suggests that the limit lies somewhere between 30 and 40. Since, for the network being considered, the probable ultimate need for the various routes was reckoned to be 33 remote stations, it was agreed that the capacity of the common equipment be 36 sub-stations, thus allowing three spare positions. It was also agreed that although the equipment was needed to control initially only 11 remote stations—six rectifier sub-stations, four track-sectioning cabins, and one power-supply distribution sub-station—the control-station apparatus should be fully equipped to its ultimate capacity so that the minimum amount of work would be needed to bring other sub-stations under control at a later date.

The control room is adjacent to Chadwell Heath Station—roughly midway between Lon-

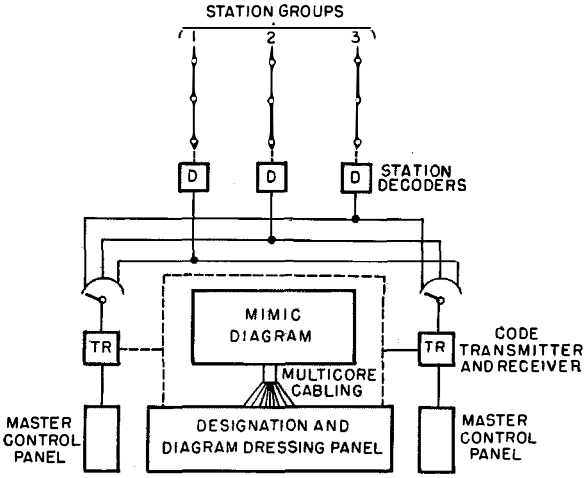


Figure 4—Schematic diagram of common-equipment system.

don and Shenfield. It is a new building and is specially designed as a control-station building with the control room as the central feature and not, as so often happens, a building designed primarily for some other purpose, and converted for use as a control station.

3. *Pilot Network*

Pilot cables, run on posts alongside the track, connect into the various sub-stations and track-sectioning cabins to provide omnibus or party-line circuits for direct-current signalling. The practical upper limit to the number of stations served by each of these circuits is set at six, and a total of eight such circuits, four to the east of the control station and four to the west, amply satisfies present and known future needs. The disposition of these circuits and the stations initially connected thereto is indicated in Figure 2. In general no two adjacent sub-stations are served by the same pilot circuit, thus enabling the traction services to be maintained in the event of one pilot circuit becoming unserviceable. Each pilot circuit consists of two pairs of conductors (one working and one standby) for the transmission of control and indication signals and telephony. Four pilot pairs are provided for telemetering, two pairs being common bus wires to all stations west of the control station and two being common to all the stations east.

A constant line-proving current is applied to each pilot circuit and, in the event of the line characteristics changing beyond predetermined limits, an audible and visual alarm is given on the control desk. The pilot circuit can be switched from the working pair to the standby pair throughout its length by a simple operation from the desk. Stations are readily identified with their pilot circuit by colour code carried on each station connect key escutcheon, and each pilot selection key.

Full protection is afforded to the operators and equipment against induced voltages due to possible surge conditions on the 33-kilovolt

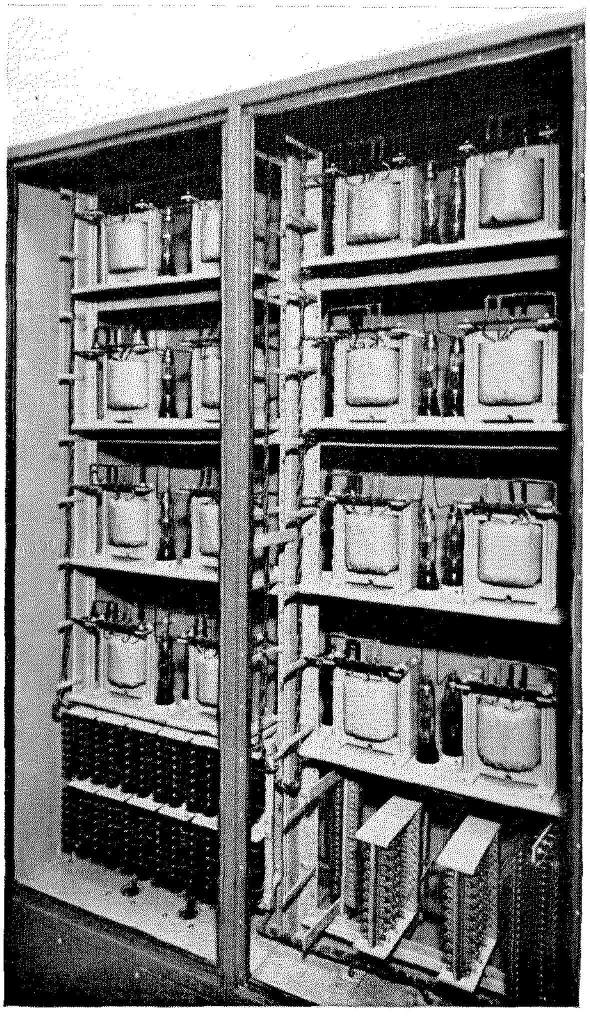


Figure 5—Control station pilot-line protection cubicle.

supply and/or any other power fault that may cause pilot-line potential to rise to a dangerous level. Gas-discharge tubes, fuses, and carbon-block protectors are fitted at all station pilot-line terminations. The protection equipments include line transformers that ensure that the discharge to earth of induced voltages occurs simultaneously on both legs of a line pair. Figure 5 shows the line protection equipment in its cubicle at the control station.

The control-circuit cables consist of high-conductivity copper conductors weighing 40 pounds per mile (11·2 kilograms per kilometer), paper-insulated, the whole lead covered and steel-wire armoured. Because of the likely extension of circuits in these cables to more distant stations in the future, the precaution has been taken at this stage of providing suitable loading coils at approximately 2000-yard (1829-metre) intervals. A loading-coil unit jointed into the cable is illustrated in Figure 6. This also shows the relation of the supervisory control pilot cable to the adjacent feeder protection cable and to the 33-kilovolt station supply cable.

4. *Operational Requirements*

For the effective and comprehensive control of this network the following requirements have been met:—

A. Control and indication of alternating-current feeder breakers; alternating-current and direct-current rectifier breakers; track-feeder breakers; and transformer on-load tap-change equipment.

B. Control—Breakers in selected groups.

C. Telemeter—Selection of rectifier load (0–6000 direct-current amperes per unit); sub-station direct-current busbar volts (1500); and extra-high-tension supply voltage at Crosswall.

D. Miscellaneous—Alarm annunciation; batteries-supply failure alarm; pilot-line switching; continuous pilot-line proving; and interlock on control operations.

E. Selective telephony.

F. Access for routine testing at all locations.

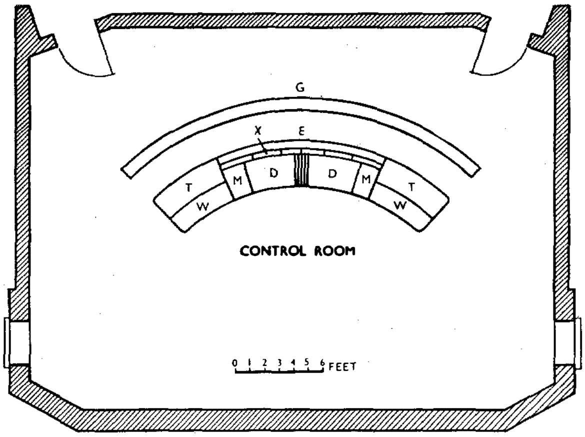


Figure 7—Plan of the control room at Chadwell Heath. *D*—designation and diagram-dressing panel, *E*—central meter indication panels, *G*—mimic diagram, *M*—master control panels, *T*—telephone switchboards, *W*—operators' writing positions, and *X*—miscellaneous services panel.

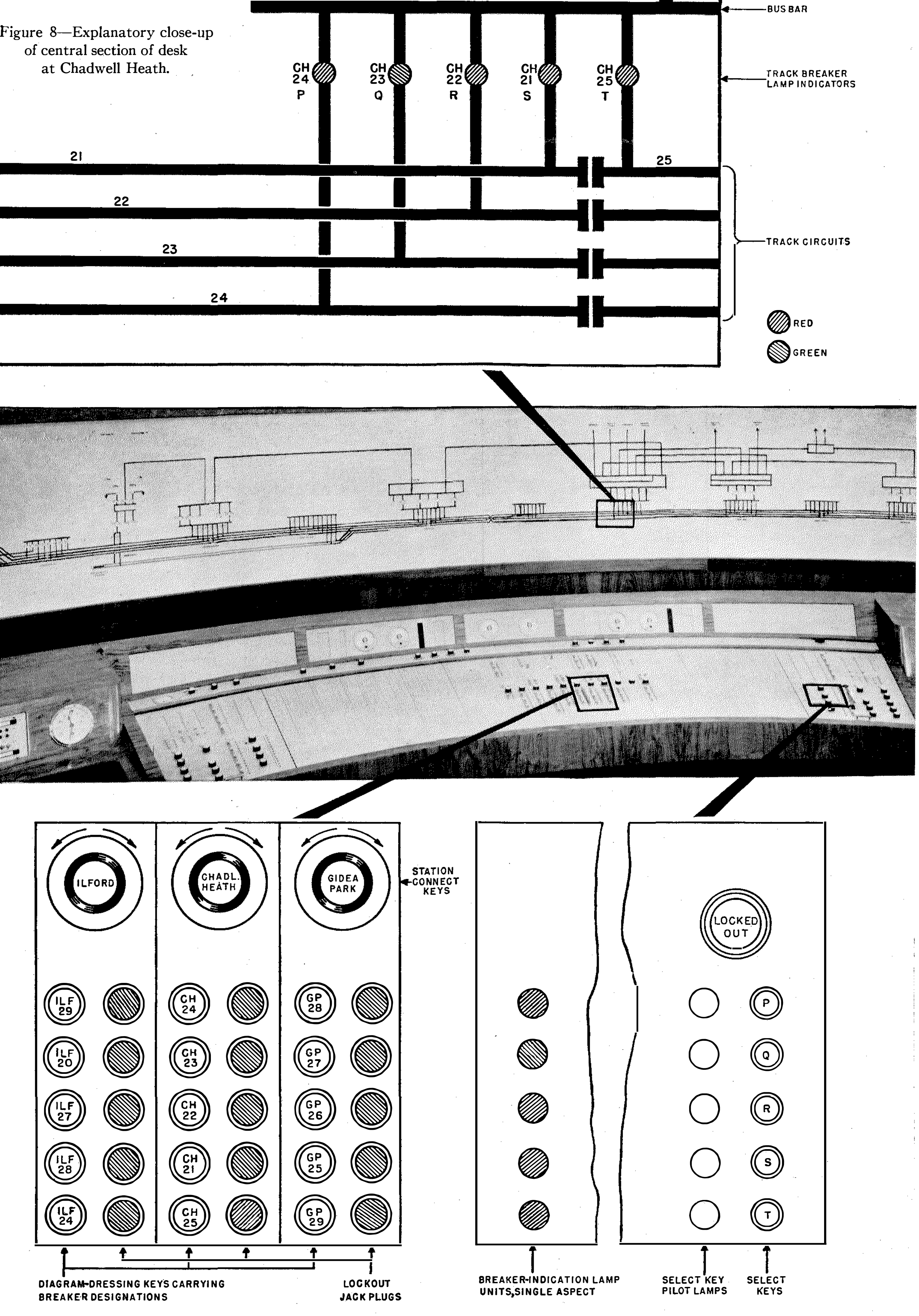
5. *Control-Room Equipment*

On page 3 is a general view of Chadwell Heath control room and Figure 7 shows its plan. These show the 22-foot-long, curved, walnut-veneered, desk- and wall-type diagram housed centrally in the spacious well-lit room. It will be seen that considerable attention has been paid to the external appearance of the equipment and that, in order to promote operational efficiency, the control keys and indicators are arranged within easy reach of the operators and permit of comfortable movements on their part.

Figure 8 shows the central part of the control desk in greater detail. The diagram-dressing or designation panel is in the centre and is flanked by the duplicate master panels from which all operations are conducted. It will be seen that the

Figure 6—Pilot cables showing a loading-coil joint.

Figure 8—Explanatory close-up of central section of desk at Chadwell Heath.



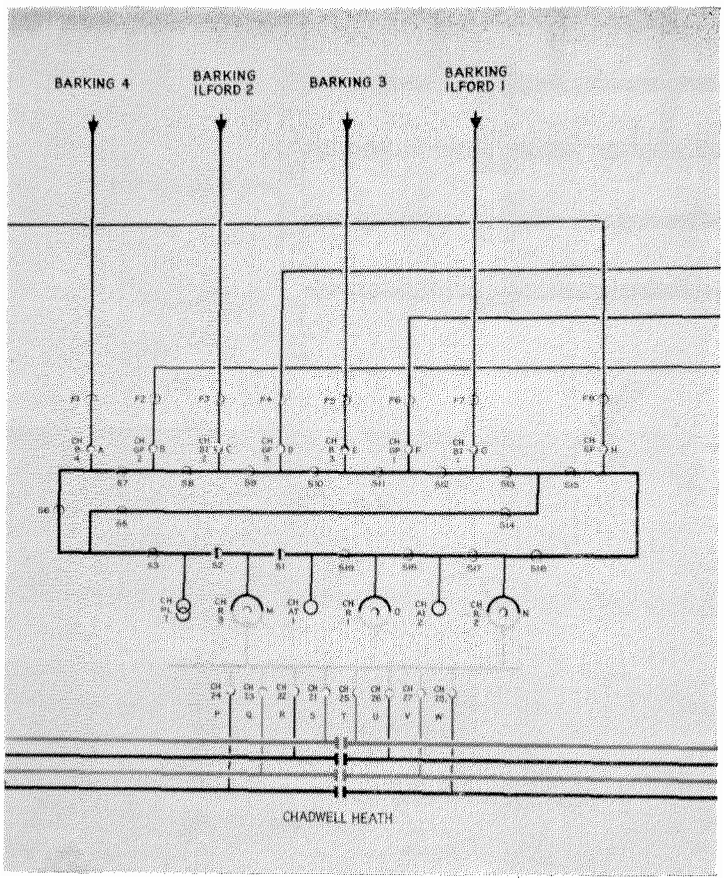


Figure 9—Part of the mimic diagram.

designation panel is composed of strips, one per station, and that only those for the 11 remote stations at present in service are equipped; the remainder will be equipped with the appropriate apparatus as and when the stations concerned are connected to the system. Each strip has a row of plunger-type dressing keys, a row of jacks, and a single rotary-type three-position station-connect key. The dressing keys, one for each controlled breaker, are for dressing the diagram and control the diagram lamp indicators. The station-connect keys connect either the left- or right-hand master panel to the station concerned, according as the key is thrown to the left or right. The jacks are part of the special safety circuit provided to prevent the remote operation of a breaker while a "permit to work" is in force. Normally each jack holds a short jack plug, but if this is withdrawn and replaced by a longer one, then the appropriate breaker-select key on the master panel is automatically made inoperative for the station concerned until the long jack plug is removed. For ease of identification the long jack plugs have red caps and the short ones green, and the operator can therefore tell, by a glance at the designation panel, if a permit to work is in force for any part of the system. Behind, on a horizontal strip, are miscellaneous details, viz., pilot-line selection keys, miscellaneous alarms facia, and keys for local-battery control.

The sloping panels to the back of the desk, and separated therefrom by a trough to minimize the desk height whilst permitting an optimum view of the panels, hold the meters and indication facias for the telemetering system (to be described later). In front of each operator is a telephone switchboard with full facilities for rapid communication with all parts of the network and with connections to outside lines. Behind the desk stands the self-supporting wall-type diagram which shows in mimic outline the extra-high-tension distribution and track circuits of the present network, each circuit-breaker being indicated by a single-aspect two- or three-colour light indicator and interconnections by painted lines. The diagram has not been previously mentioned as a part of the common-equipment system since it is not essential to the operation of the system. The function of the diagram is to display a current record of the condition of all circuit-breakers in the network, a record that cannot be maintained by the master panels since these are common to all stations and are switched as required from station to station. The breaker position indications, i.e., open, closed, locked out, as received at the master panels are manually transferred, by means of the dressing keys on the designation panel, to the mimic diagram. A section of the mimic diagram is shown in Figure 9.

As briefly stated earlier, full facilities are available at each master panel for the complete operation of the system and all indications of breaker positions are received thereon. Figure 10 is a detailed picture of the right-hand master panel; the left hand is identical. Although this panel is common to all remote stations in the network, when it is connected to any one it assumes for the moment that station's identity, a relatively simple achievement, because of the similarity of switchgear arrangement existing between the various sub-stations of a traction network.

Figure 10 shows the vertical row of SELECT keys (with adjacent PILOT lamps) in two groups, *A–O* and *P–CC*, the former for alternating-current breakers and the latter for direct-current track-feeder breakers. The letters *A* to *Z* to *CC* engraved on the key heads serve to identify each key with its breaker indicator on the mimic diagram (the appropriate letter is signwritten against each indicator), and hence with its corresponding actual breaker at the remote station. Breaker positions are indicated on the row of lamp signals on the left of the panel, each of which shows RED for closed or GREEN for tripped, when position-indication signals are returned from the distant station under control. For direct-current track-feeder breakers, in addition to the individual SELECT keys, there are four GROUP-CONTROL keys which, when operated, initiate a special code to select and operate a group of breakers as signified by the sign-written bracket.

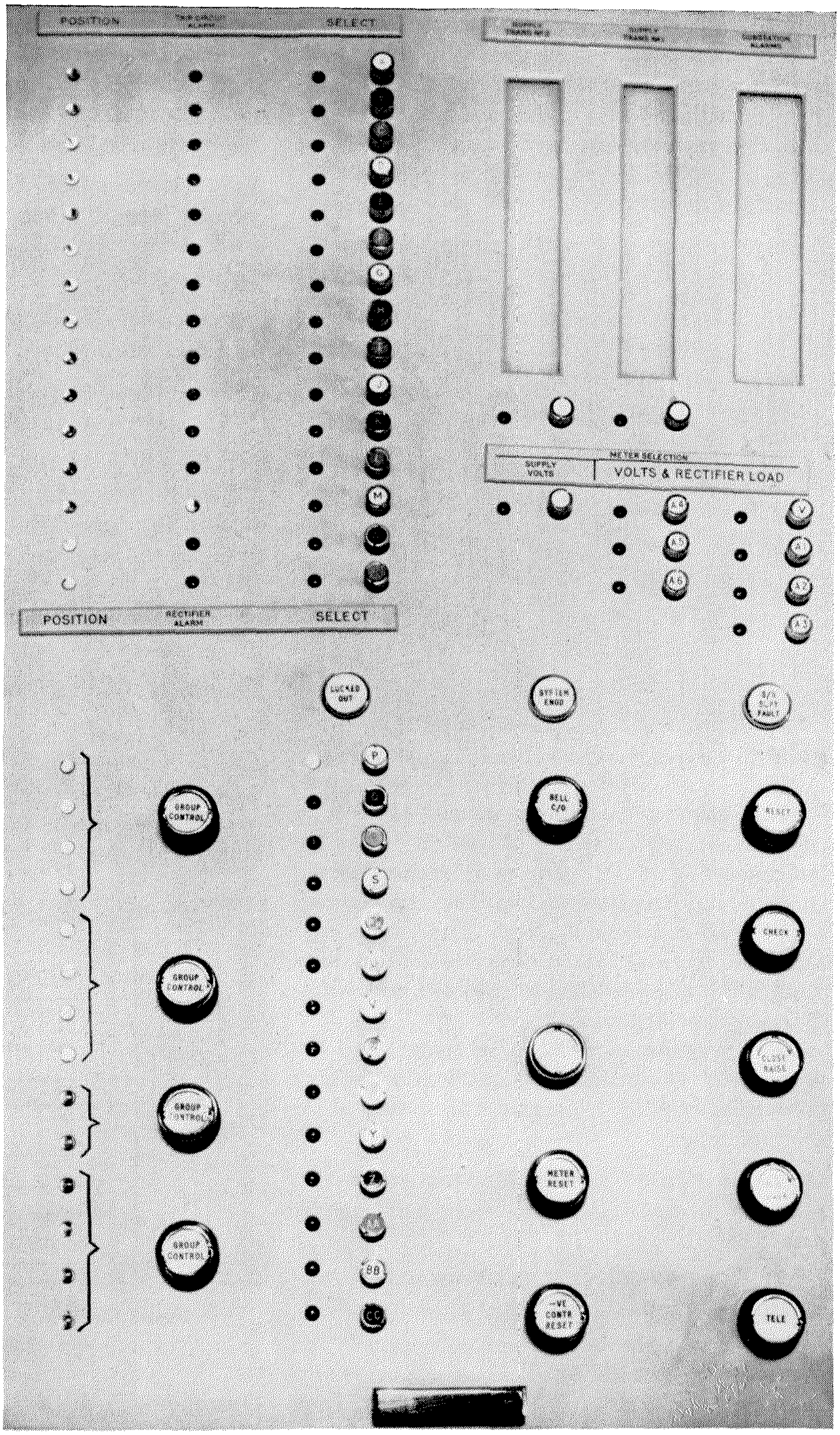


Figure 10—Close-up of master panel on desk showing the vertical row of select keys with the adjacent pilot lamps.

BELL CUT-OFF, RE-SET, CHECK, and COMMON OPERATE keys, each incorporating its own PILOT lamp, are grouped on the lower right of the panel together with the TELEPHONE and METER RE-SET keys. Transformer-tap position and sub-station alarm-indication facias are above the METER-SELECTION keys.

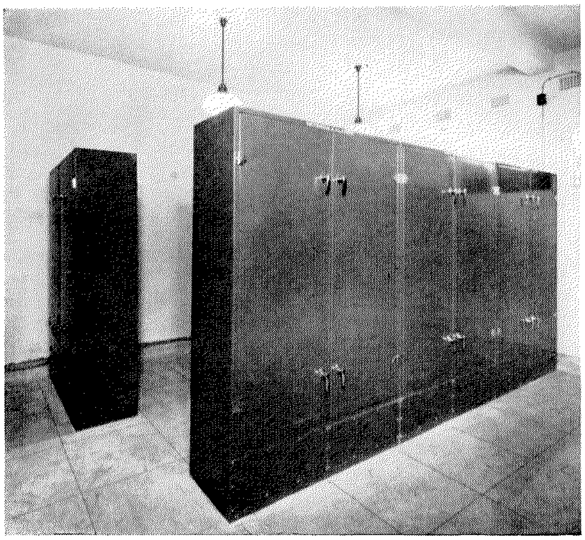


Figure 11—General view of control-station apparatus room.

The foregoing description should be read in conjunction with Figures 8 and 10 when an understanding of the operational features should be apparent, but perhaps this may be clarified by a brief description of a typical operation, e.g., to close Chadwell Heath direct-current breaker number 23, which is shown tripped in the mimic diagram inset of Figure 8.

The operation sequence is as under, and the notes refer to means of identification, which become automatic to operators in a very short space of time.

Operation—Turn Chadwell Heath STATION-CONNECT key to right (to engage right-hand master panel).

Effect—STATION-CONNECT KEY lamp and SYSTEM-ENGAGED lamp on master panel glow steadily.

Notes—Identified by engraved legend on STATION-CONNECT key head. Master panel is now connected preparatory to signalling to Chadwell Heath sub-station equipment.

Operation—Press CHECK key until PILOT lamp flickers.

Effect—Check code transmitted; position indication signals returned; breaker indication lamps glow (RED—closed, GREEN—tripped) SYSTEM-ENGAGED, CHECK and RE-SET lamps are extinguished.

Notes—Received indications are now visually checked against those on mimic diagram by operator and any discrepancy corrected by use of dressing keys.

Operation—Press SELECT key Q until associated pilot lamp flickers.

Effect—Lamps in the common-control keys, viz., CLOSE, TRIP, and RE-SET keys glow indicative of next operational step.

Notes—Breaker now selected for operation; SELECT key is positively identified by (*A*) letter Q associated with CH23 on diagram (Figure 8), (*B*) SELECT key Q—black head on same horizontal row as dressing key CH23. Breaker is now ready for closing.

Operation—Twist CLOSE key until insert lamp is extinguished.

Effect—SYSTEM-ENGAGED lamp flashes; close code transmitted; indication lamps extinguished; position indication signals returned; indication lamps re-lit showing new (closed) position.

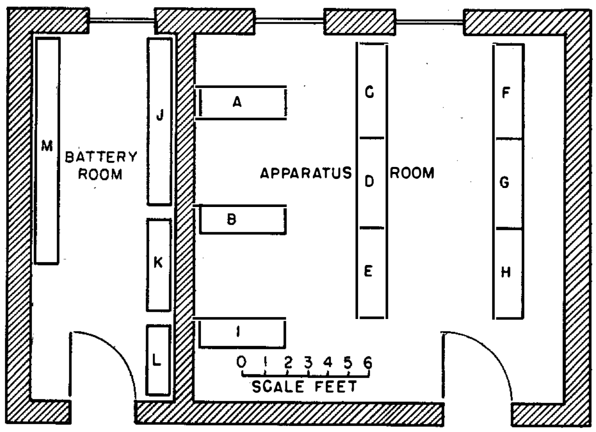


Figure 12—Plan of control-station battery and apparatus rooms.

Operation—Push in DRESSING key CH23 on designation panel.

Effect—Green lamp in breaker indicator 23 on mimic diagram extinguished and red lamp lit.

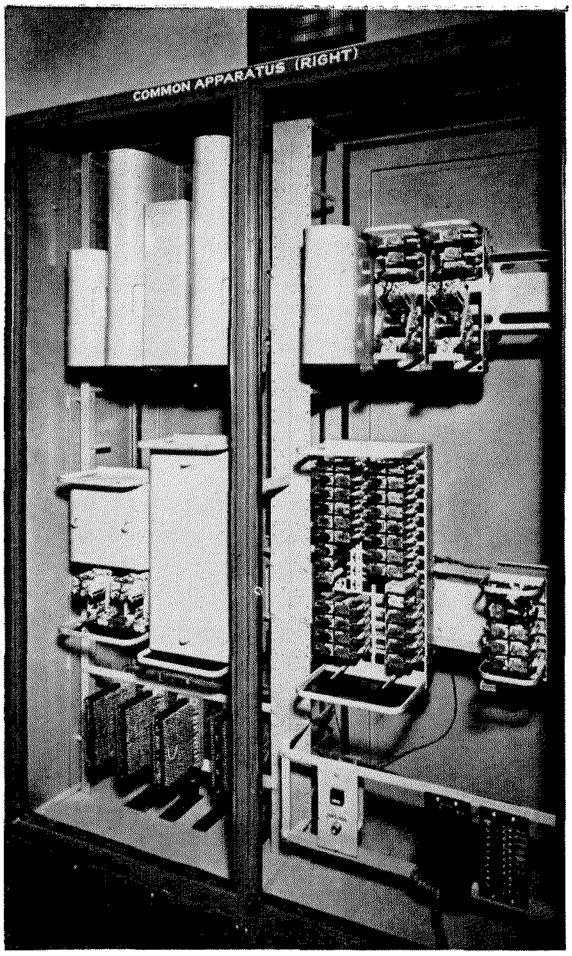


Figure 13—Cubicle housing apparatus for right-hand master panel.

Notes—Mimic diagram indications now agree with actual breaker positions.

Operation—Restore Chadwell Heath STATION-CONNECT key to its central position.

Effect—CONNECT lamp and all indication lamps on master panel are now extinguished.

Notes—System is now restored to normal and free for any other operation.

6. Control-Station Apparatus

All the apparatus for the complete functioning of the system is contained in the apparatus room whence multi-core telephone-type cables connect to the desk in the control room. A general view of this room is seen in Figure 11 and a plan in Figure 12. With the exception of that for battery-charging and line protection, all equipment is mounted on standard telephone-type jack-in panels as used in automatic-switching telephone exchanges. These are contained in seam-welded double-door dust-proof cubicles arranged for bottom entry of cables.

Equipment associated with the right-hand master panel is shown in Figure 13. Individual dust-covers have been removed to show details of some of the jack-in panels. A similar cubicle of equipment is coupled with the left-hand panel and the two sets then feed into the common apparatus cubicle in Figure 14.

A unique feature of this system is the test cubicle that contains a complete set of remote-station jack-in panels, available for rapid replacement when necessary. This cubicle also

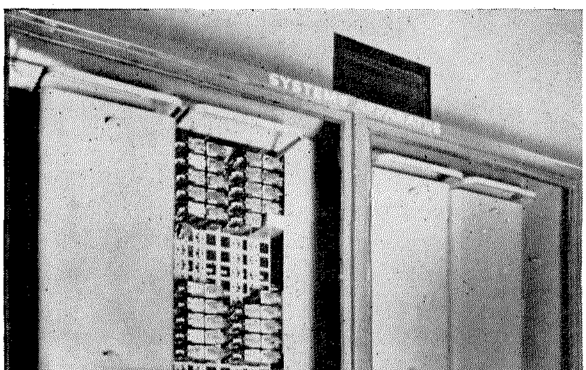


Figure 14—Control-station common-apparatus cubicle.

mounts test panels and associated apparatus on which all control operations can be simulated without interfering with the main equipment and by means of which a repaired sub-station panel can be thoroughly checked functionally before return to its former location. The cubicles are finished outside in glossy grey stoved enamel. Cubicle interiors are finished in a lighter colour. The equipment, which operates at 50 direct-current volts, is fed by a 24-cell 75-ampere-hour lead–acid battery located, together with the telephone speech and ringing batteries, in a separate room next to the apparatus room. The batteries are maintained in condition by the *SenTerCel* selenium-rectifier automatic battery chargers housed in a common cubicle in the apparatus room and shown in Figure 15.

7. Sub-Station Equipment

The Shenfield sub-station supervisory apparatus cubicle, finished glossy black to match the suite of switchgear equipment, is shown in Figure 16. Features of note are the special isolating terminal links with studs for terminating the 3/036 vulcanised indiarubber cables connecting the various switchgear units; and the right-hand portion of the cubicle housing the telemetering jack-in panels (temporarily removed) and above, the power pack supplying the telemetering apparatus.

8. Signalling

Signalling to and from the control station is by means of trains of impulses representing numbers and is analogous to that used in automatic telephone switching. All codes are generated at speeds of 10 impulses per second by code senders that transmit at a closely controlled speed and make/break ratio and, with the relative fixed line conditions obtaining between stations, this gives a degree of stability and reliability not normally encountered on telephone switching systems where many variables such as dial speed and ratio, line resistance, and leakage may be encountered.

The signals from the control station to the party-line system consist of a station code and a function code and only the particular station identified by the former continues in operation until it responds to the latter.

The signals to the control station consist of a station identification signal followed by a train of impulses giving position indication of all the constituent units, viz., breakers, alarms, transformer-tap positions, etc.

Throughout, the coding scheme incorporates the constant-total checking system by which the equipment is safe-guarded and performs only the function required. Where a number of stations has access to a continuous party line for signalling in both directions, due precautions are taken against any possibility of double connection. Changes occurring simultaneously at stations on a party line are stored and then transmitted one after the other in turn.

On the common-equipment system the complete information concerning a position change at a distant sub-station cannot be transmitted

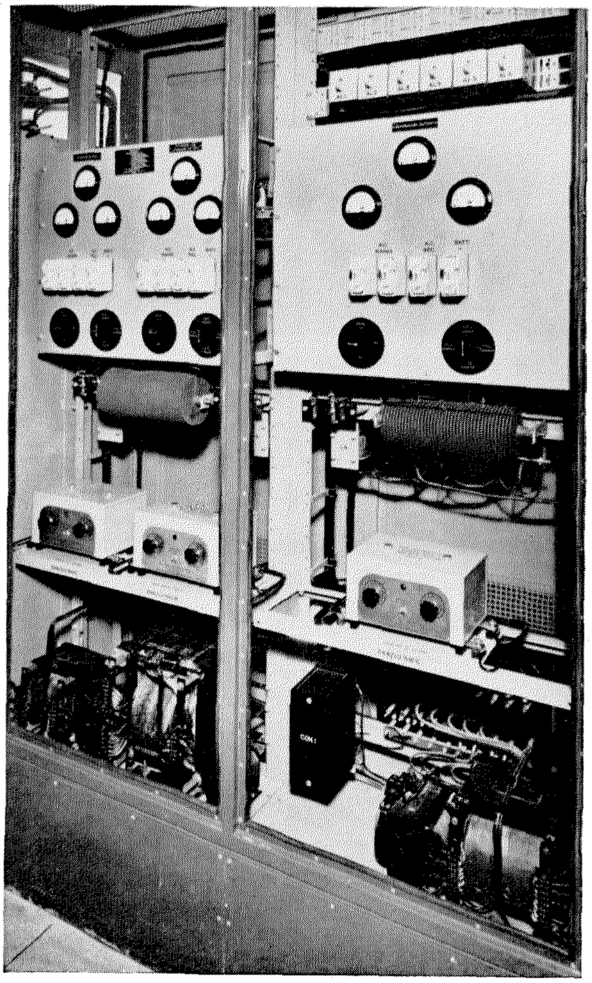


Figure 15—Cubicle for automatic battery chargers.

until the master panel is prepared for its reception. The fact that a change has occurred is notified at the control station (visually and audibly) as a station alarm from the originating station and this is stored until the operator attends to it and prepares the master control panel accordingly. A check signal then to the distant station results in the complete latest information from it being displayed on the master panel.

As thus described, the time required for the display of the new information is longer than with the individual-panel system where indication devices are always available for every remote unit on the system. Experience shows, however, that on metropolitan suburban lines with dense traffic and where speed is essential, the system provides information as fast as can be dealt with by competent operators and with the advantage that the operator may select his information and is not confronted, and possibly confused, by a number of almost simultaneous breaker changes and alarms, as might well happen in times of network disturbance.

9. *Metering System*

On the Liverpool Street–Shenfield line, provision is made for comprehensive remote metering covering the loading of individual rectifiers (2000-kilowatt capacity) and readings of direct-current busbar voltage at each station.

For this purpose the *Standard* impulse-modulated electronic telemetering system (I.M.T.), representing what is claimed to be the most up-to-date technique in this field, has been employed as satisfying fully the essential demands of readings continuously responsive to the load and voltage changes peculiar to a traction system. The main problems in providing a suitable system are encountered in deriving current and voltage readings from high-voltage direct-current circuits and converting them into quantities suitable for transmission over light-current pilot circuits without their being affected by normal variations in line characteristics, and insulating the equipment from the direct-current source. An ideal medium for this purpose is a pulse-rate system, such as the impulse-modulated telemetering system, which readily and accurately measures quantities, provided they can be converted into a simple impulse rate proportional to the quantity being measured.

This system was originally produced to indicate at a distance readings derived from prime-source meters fitted with impulsing contacts as typified by the well-known integrating watt-hour meters in widespread use on power

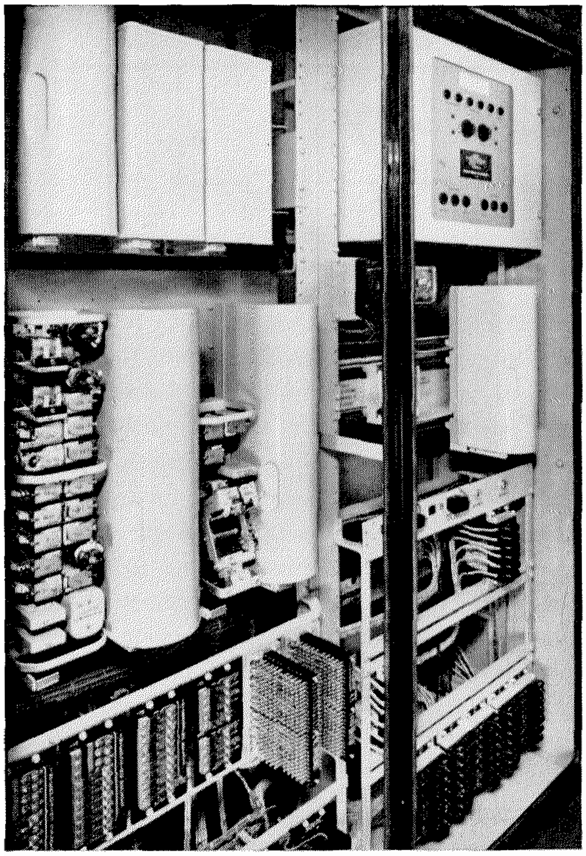


Figure 16—Sub-station supervisory cubicle.

transmission and distribution systems. Where such impulsing meters are not provided at the source or where readings of, say, amperes and voltage not normally requiring such meters are concerned, the quantity to be measured must first of all be converted into impulses by other means ready for transmission to the distant indicating centre.

The complete conversion from source to final reading is therefore in two stages, viz., source to impulse rate at the distant sub-station and impulse rate to meter reading for display at the control room.

It would be as well here to explain the basic principles of the system (see Figure 17).

The telemeter consists essentially of an arrangement in which two opposing currents are balanced, namely (*A*) the average current due to a fixed capacitor charge for each operation of the impulse receiving relay *IR* and (*B*) a steady current obtained from the circuit of the

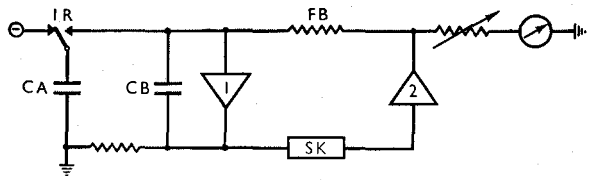


Figure 17—Basic circuit of impulse-modulated electronic telemeter.

indicating meter. Contact *IR*, on a relay sensitive to the incoming impulses from the distant station, charges capacitor *CA* and discharges it into reservoir capacitor *CB* connected across amplifier *1*. The latter has considerable gain and builds up a positive output potential whilst maintaining its input potential at a nearly constant value. This ensures that *CA* delivers always the same quantity of charge so that the average input current depends upon the frequency of operation of contact *IR*. The output voltage of amplifier *1* is applied through a smoothing circuit *SK* to amplifier *2*, a cathode follower, which in turn feeds the average potential to the indicating meter circuit. Part of the output is also fed back to the input via resistance *FB* to neutralize the input due to *IR*. Therefore, for any particular contact rate the

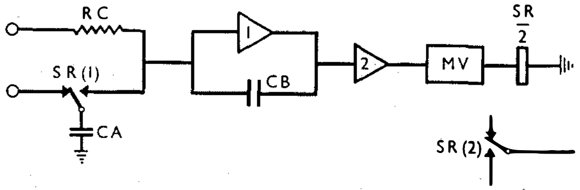


Figure 18—Telemeter sender circuit.

system will establish a balance when the current in the meter circuit is proportional to the contact rate. The saw-tooth effect of the capacitor charge is eliminated in the smoothing circuit, whose time constant is arranged to give a high degree of smoothing combined with a rapid response to changes in input.

The following summarizes the technical features of the telemeter.

9.1 Indicating Instrument

The telemeter gives an output of 5 milliamperes, direct current, at 50 volts, and any instrument of 5-milliampere full-scale deflection may be used as an indicator. If required, a number of instruments may be connected in series.

9.2 Accuracy

The system has an accuracy of ± 1 per cent of full scale. The calibration is linear, the output current being directly proportional to the incoming impulse rate.

9.3 Response

The system operates at a nominal full-scale rate of 20 impulses per second. A sudden change,

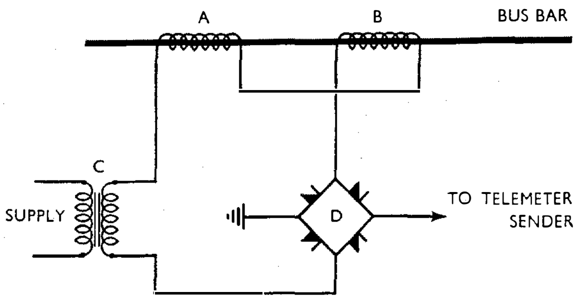


Figure 19—Saturable-reactor coupling for direct-current amperes.

say from zero to full load, will be indicated fully in 2 seconds, the response being sufficiently rapid to follow load fluctuations normally occurring on traction systems. The indicating meter reads zero when disconnected, vibrates at zero when receiving a zero reading correctly and is driven off scale in the event of system failure. This obviates ambiguity between true and false zeros.

9.4 Sub-Station Telemetering Equipment

The relatively fast impulse rate referred to cannot be derived from an integrating meter and so a telemeter sender circuit has been developed to translate the quantities metered

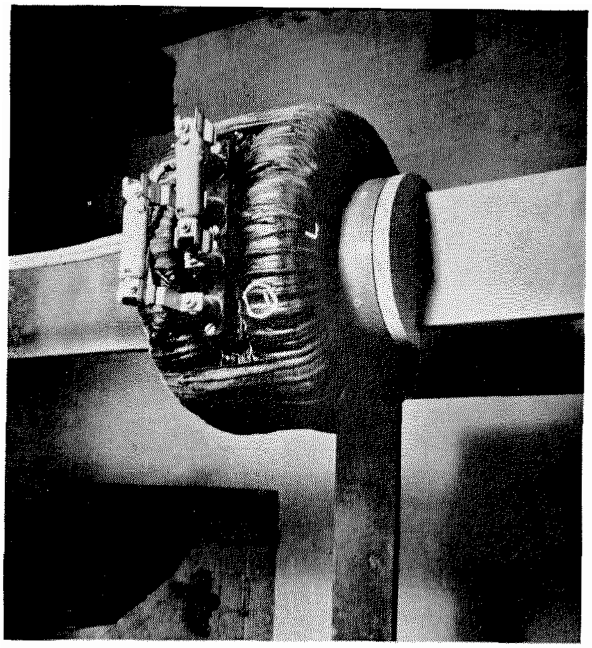


Figure 20—Toroid control winding mounted on 1500-volt direct-current busbar.

into terms of impulses per second. The circuit is shown in Figure 18.

A voltage proportional to the quantity to be metered is applied to resistor *RC* and the resulting current into amplifier *1* is balanced by causing relay *SR* to impulse at a suitable rate. This circuit operates effectively as a reverse telemeter. In this case the applied voltage results in capacitor *CB* charging steadily, the voltage developed across it being applied to amplifier *2* which unblocks the oscillator *MV* at a pre-determined level. The oscillator delivers one impulse that operates relay *SR* and applies an opposing charge to capacitor *CB* from capacitor *CA*. This reduces the potential across capacitor *CB* and causes amplifier *2* to block the oscillator. This process continues at a rate determined by, and proportional to, the input voltage. For each operation of relay *SR* an impulse is fed to the line via contact *SR* (*2*).

9.5 Coupling to Direct-Current Circuit

The telemeter sender must be operated from a voltage source sufficient to swamp variations at the input of amplifier *1*. As these variations are of the order of 0·5 volt, the applied voltage should be no less than, say, 50 volts. This may best be obtained by means of saturable reactors. An arrangement suitable for direct-current amperes is shown in Figure 19. The feeder is taken through two toroid windings, *A* and *B*. thus simulating a single-turn control winding. The coils are fed from an alternating-current supply and are so connected that the direct current flowing in the control windings causes the alternating-current to vary in direct proportion. This current is rectified and provides a direct-current voltage directly proportional to the feeder current except for a small zero error of less than 1 per cent.

A coil designed with a current rating of 6000 amperes is shown in Figure 20 mounted on a rectifier positive busbar. The three essential units required at the source are shown in Figure

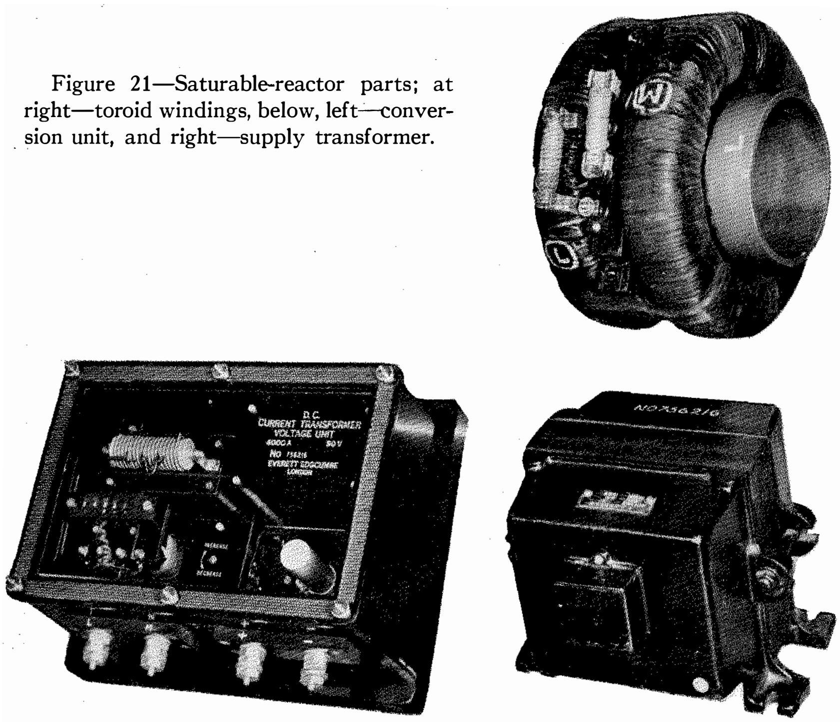


Figure 21—Saturable-reactor parts; at right—toroid windings, below, left—conversion unit, and right—supply transformer.

21 and can readily be identified by reference to the schematic in Figure 19. A similar arrangement shown in Figure 22 is used for measuring busbar voltage, and this use of transformer coupling provides a means of isolating the sender

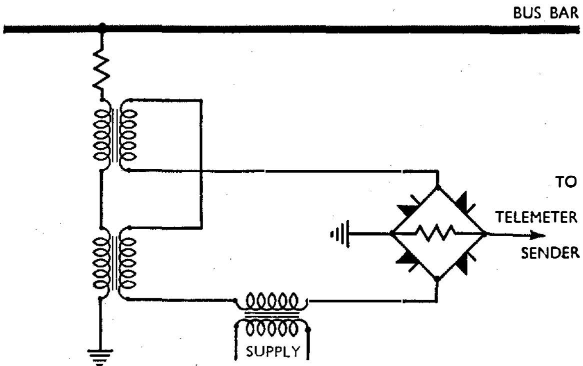


Figure 22—Saturable-reactor coupling for direct-current voltage.

equipment from direct connection to the power circuits; the relatively high voltages of the traction system are therefore not brought into the telemetering cubicles, thus obviating the need for interlocking safety switches. The saturable reactors can provide adequate output for local metering and contrast favourably with shunts and potentiometers heretofore used for this purpose in that they effectively isolate the metering from the power circuits.

9.6 Meter Switching Facilities

The general arrangement of selective meter switching in relation to the control desk, the remote station, and the metering network can be followed by reference to Figure 23. In all, a maximum of four readings selected individually can be displayed on the desk, two each on the east and west networks respectively. Meter display facias which indicate to the operator the source of each reading are mounted alongside each indicating meter.

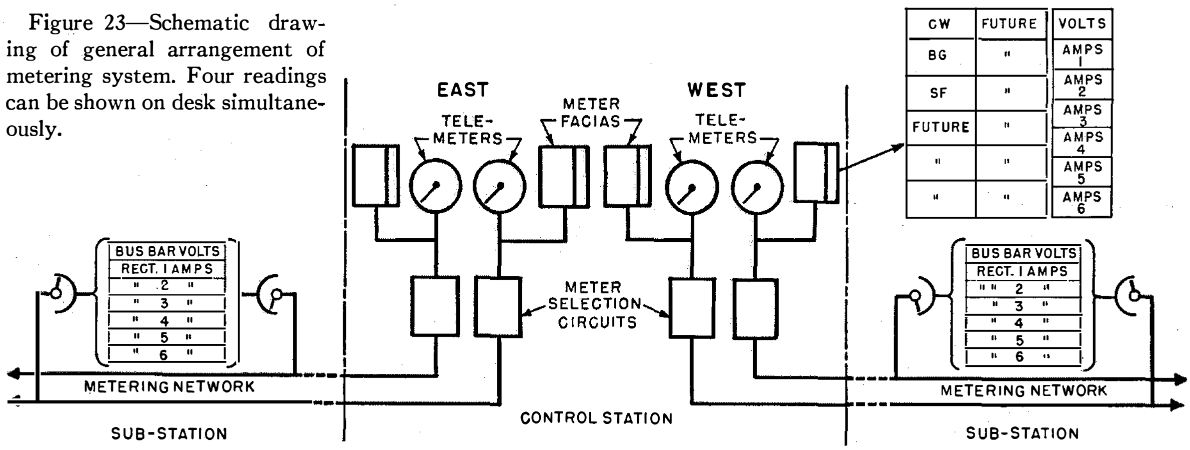
Supply intake from the National Grid at Chadwell Heath (33 kilovolts) is metered via a direct-wire system to a meter on the control desk and that at Crosswall metered via the telemetering system. The tap-change equipments of the two supply intake transformers 22/33 kilovolts at Crosswall are controlled and indicated on the master panels.

10. Other Installations

The common-equipment system has been recently installed on railway networks in Britain and New Zealand and these notes outline special facilities. On the Euston–Watford line, over part of which, incidentally, the London Transport Executive's Bakerloo trains work, centralized control was introduced with the conversion from rotary converters to mercury-arc rectifiers. Three of the 17 unattended sub-stations are much larger than the others and, for convenience, have been divided and each considered as two from the controller's point of view. These divided sub-stations are equipped with two standard remote-station apparatus cubicles and thus all 20 remote-station cubicles have been made to the same standard size and capacity. The control desk and diagram are shown in Figure 24.

The equipment being manufactured for Tyneside Lines (Figure 25) includes special pilot-line switching facilities to cater for the circular route on this network. The substations are connected

Figure 23—Schematic drawing of general arrangement of metering system. Four readings can be shown on desk simultaneously.



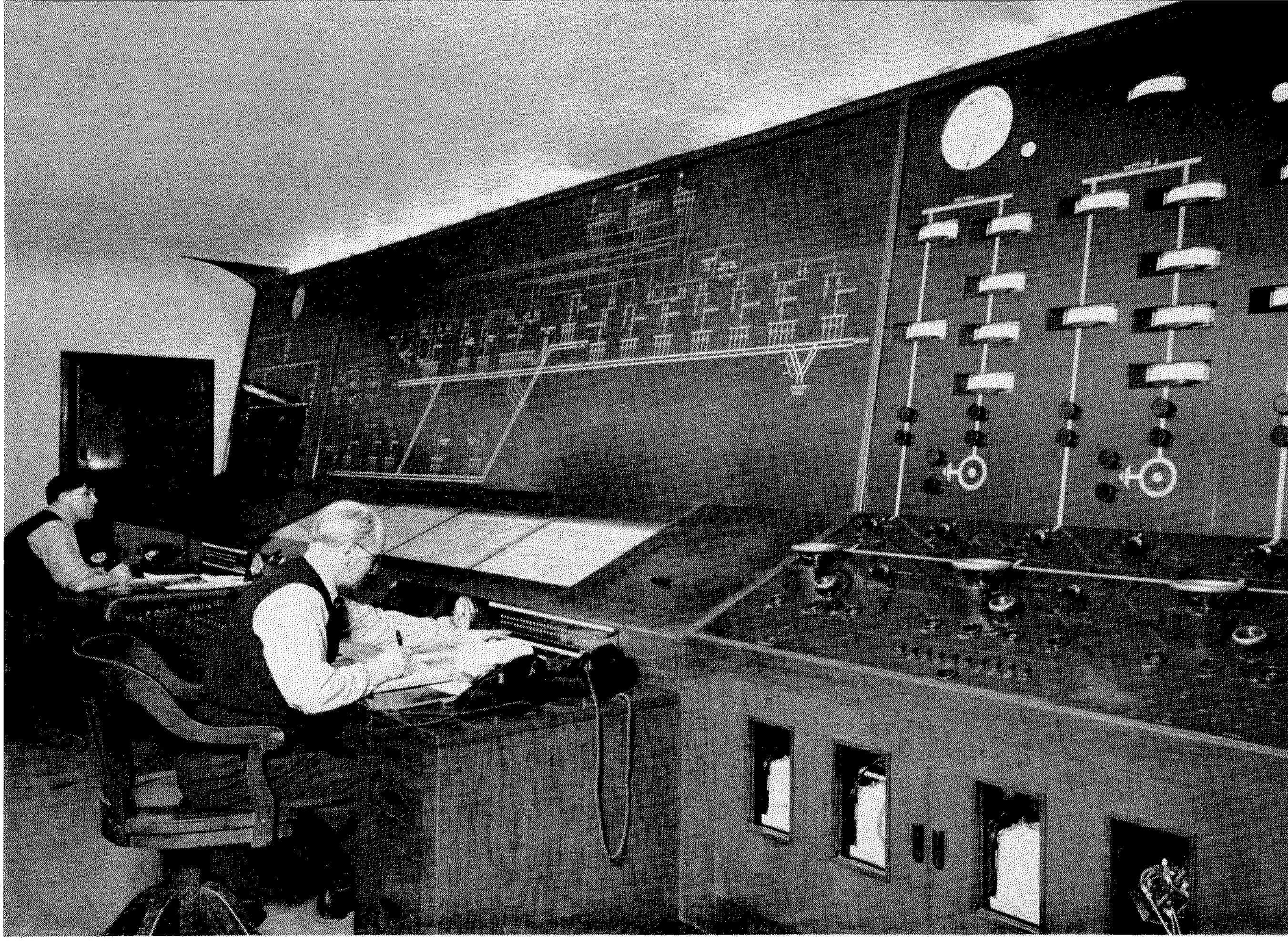


Figure 24—Stonebridge Park control room on the Euston–Watford line.

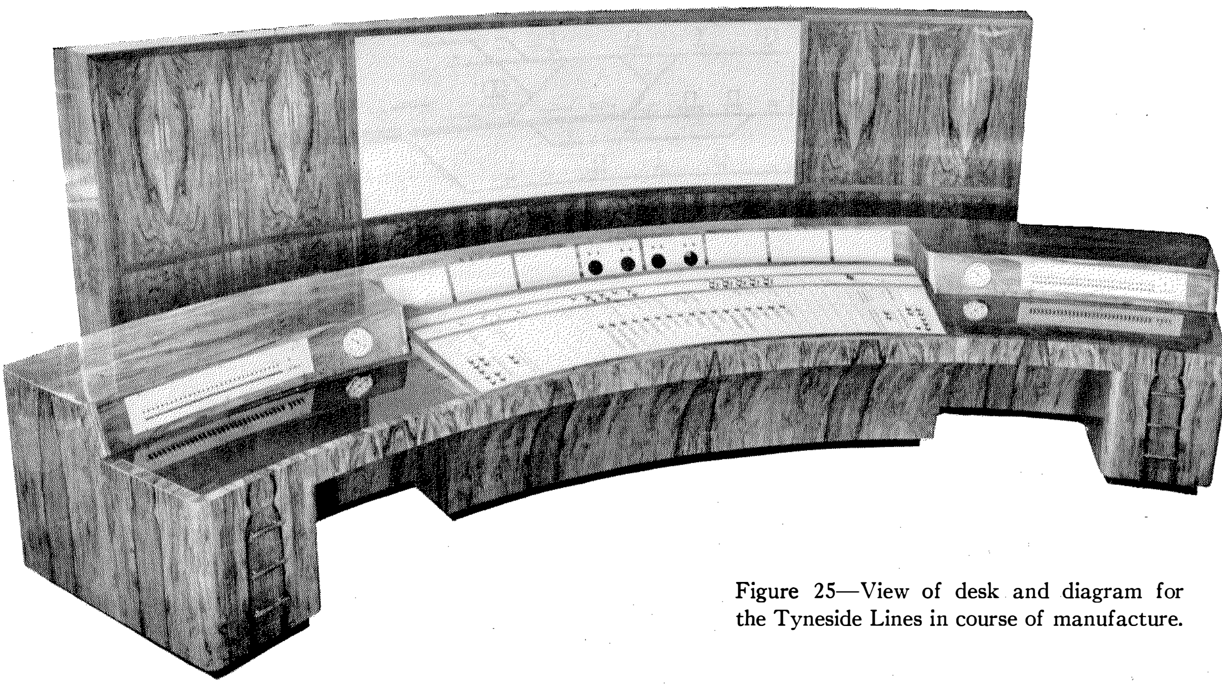


Figure 25—View of desk and diagram for the Tyneside Lines in course of manufacture.

virtually on a ring main, normally open near the most distant sub-station, but with facilities for closing up there and opening the ring at any other part should damage to the pilot cables make this temporary expedient necessary.

The Liverpool–Southport line was also recently converted to operate from mercury-arc rectifiers. On the desk at Hall Road control room (Figure 26) the mimic diagram is in two parts, one showing the alternating-current distribution system and the other the direct-current track system.

For the Hutt Valley scheme in New Zealand the pilot wires used for signalling also serve for the telemetering facilities where the impulse-modulated telemetering system is employed. The metering impulses between sub-station and control modulate a 750-cycle-per-second carrier frequency which is superimposed on the direct-current signalling circuits. The mimic diagram in this installation is so miniaturized that it is easily mounted on the control desk, which is shown in the course of manufacture in Figure 27.

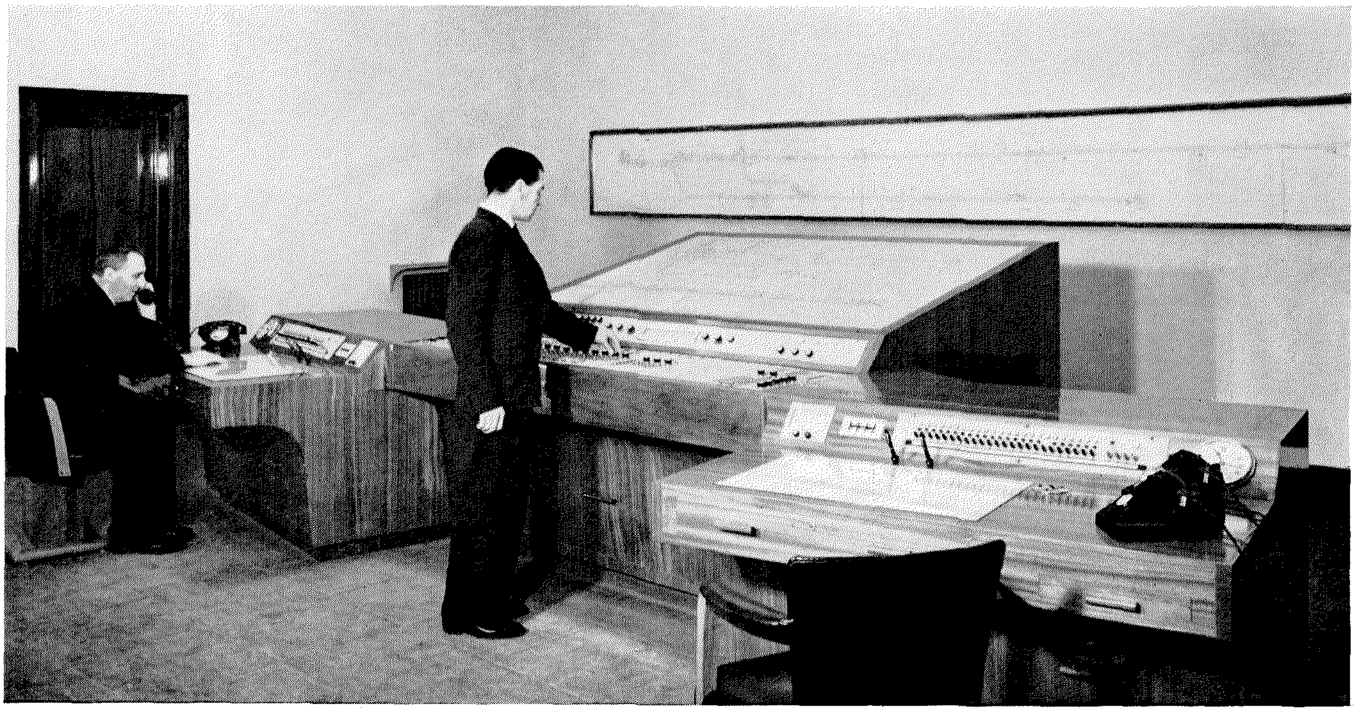


Figure 26—The Hall Road control room on the Liverpool–Southport line.

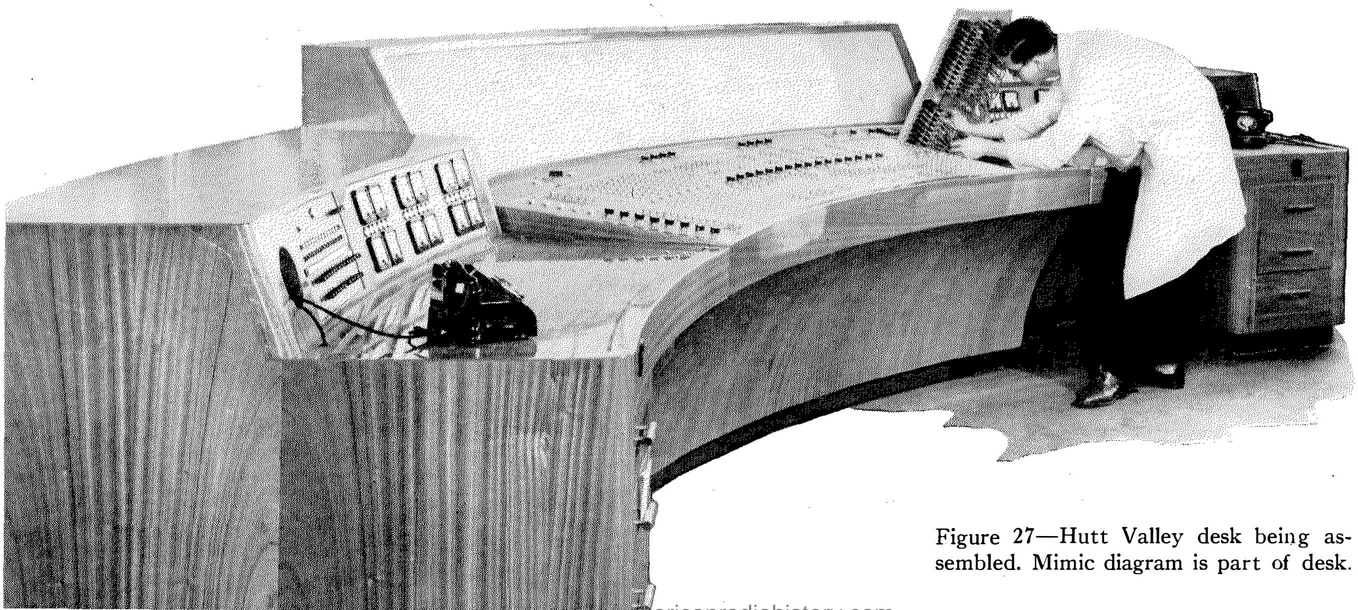


Figure 27—Hutt Valley desk being assembled. Mimic diagram is part of desk.

Styrene Joints and Terminations*

By E. C. LEE and K. KONSTANTINOWSKY

Standard Telephones and Cables, Limited; London, England

STYRENE JOINTS and terminations were introduced many years ago as barriers against the migration of compound in cables. The original processes for making such joints were elaborate and involved lengthy heating on site. Modern methods have eliminated these disadvantages and are easy to carry out. They permit the use of styrene joints and terminations throughout a system, thus offering many advantages over conventional designs with little or no increase of cost. Practical experience with solid-type cables up to 66 kilovolts is good.

• • •

1. *High-Voltage Cable Operating Conditions*

The performance of high-voltage solid-type-cable systems is influenced by several factors including conditions of loading, conditions of laying, e.g., gradients, and by the design of joints. Loading results in permanent distension of the lead sheath owing to the differing coefficients of expansion of the constituent materials, oil being of particular importance in this respect, as will be seen from Table 1.

Some distension is produced by flexing of the cable during manufacture and laying, and the overall effect of distension is a change in the longitudinal fluid resistance of the cable (see Table 2) and the creation of sub-atmospheric pressures during periods of light or no load.

2. *Migration of Compound*

A modern high-voltage solid-type cable can operate successfully under these low-pressure conditions, but if the process becomes cumulative, serious damage to the lead sheath may result. Such an effect is caused by the flow of compound from the joints to relieve the low pressures existing in the cable, or by the migration of compound from a high to a lower level. It is usual to limit the static head to a value of between 40 and 80 feet by means of barrier joints, but it was frequently found that the introduction of a mechanical diaphragm type of barrier joint increased rather than reduced the risk of damage. The reason was that this type of joint, being unscreened, had to be kept filled with compound; and since there was always a tendency for compound to drain into the low-pressure side, frequent inspections and addition of compound were necessary. The cumulative effect of the additional compound in the cable resulted in numerous failures from split lead sheaths at the lower end. The time required to produce these damaging effects was sometimes as short as four years.

Although compound migration is accelerated by severe gradients and by the installation of fluid-compound-filled barrier joints, the effects are by no means negligible on comparatively flat cable routes, when fluid-compound-filled non-barrier joints are installed.[1] Numerous cases have

TABLE 1

COEFFICIENT OF LINEAR EXPANSION
Per Degree Centigrade

Material	$\times 10^{-6}$
Copper	1·7
Impregnated Paper*	0·5
Oil	25
Lead	3

* A. Walraff, "Über die Wärmeausdehnung von Isolierpapier" (Thermal Expansion of Insulating Paper), *Archiv für Elektrotechnik*, v. 37, pp. 458–462; September, 1943.

TABLE 2

FLUID RESISTANCE PER CENTIMETRE
In Centimetre-Gram-Second Units for a 0·2-Square-Inch Single-Core Cable Having 0·57-Inch Radial Insulation

Conductor	12 000
Dielectric Only	1 200
Total	1 100
Total with Loose Sheath	150

* Presented at the Conférence Internationale des Grands Réseaux Electrique à Haute Tension, Paris, France, on July 5, 1950.

[1] C. J. Armstrong and C. T. W. Sutton, "Behaviour of High-Voltage Solid-Type Cable Accessories," *Journal of The Institution of Electrical Engineers*, Part 2, v. 95, pp. 523–533; October, 1948.

been cited where serious drainage from joints has resulted in collapsed and split joint sleeves and burst lead sheaths. The time factor under these less severe conditions is usually longer and is undoubtedly influenced by the shape of the load cycle and the maximum temperature.

The authors are of the opinion that migration along the layers of the dielectric is extremely slow. The conductor path may be of some importance for large cross sections since it is at the highest temperature, and the worming spaces may be of importance because of their low space factor. The path of lowest resistance is that between the insulated core or cores and the lead sheath, and if this path can be kept small by limiting the sheath expansion, then it is believed the time factor associated with migration will be so large as to be of little practical importance. These conditions can be achieved by the use of joints that contain no fluid compound, and which ensure a barrier in both the strand and the dielectric paths. Preferably, these joints should be installed between each cable length. In such an installation the lead sheath will distend to a maximum value when application of full load current corresponds with the maximum ambient temperature. Once this condition has been fulfilled there will be no tendency for further distension since additional compound is not available to flow into the cable, and the paper dielectric and the wormings tend to retain their impregnant.

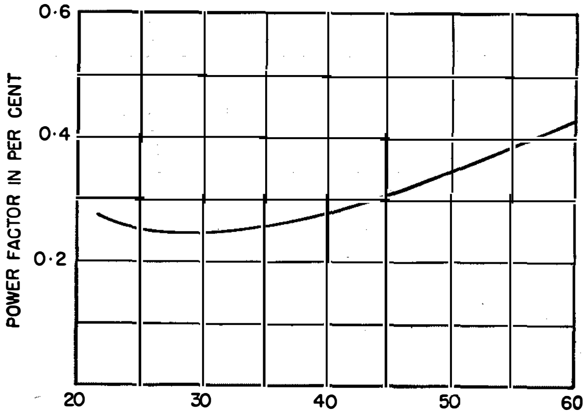


Figure 1—Power factor plotted against temperature for grade-*G* styrenated and lacquered 3·5-mil cable paper at 600 volts and 50 cycles per second.

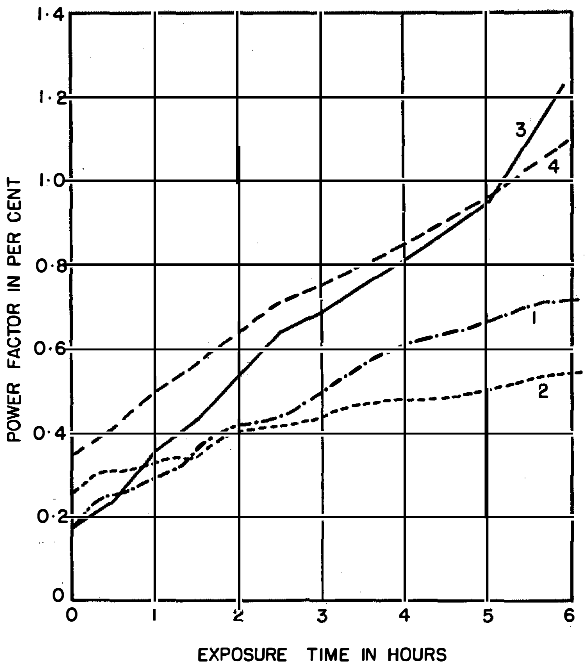


Figure 2—Power-factor changes as a result of exposure of jointing tapes to a relative humidity between 54 and 55 per cent and at 600 volts at 50 cycles per second. 1—Styrenated 3·5-mil low-density grade-*G* paper stressed at 18·2 kilovolts per centimetre. 2—Styrenated 5-mil high-density grade-*B* paper stressed at 33·7 kilovolts per centimetre. 3—Resin-oil-compound-impregnated 3·5-mil low-density grade-*G* paper stressed at 33·7 kilovolts per centimetre. 4—Resin-oil-compound-impregnated 5-mil high-density grade-*B* paper stressed at 21·5 kilovolts per centimetre.

Two layers of tape with their faces adhered to each other were tested. Samples of 1 and 2 separated during the conditioning and thus exposed four surfaces instead of two.

3. Styrene Barrier Joints

An early attempt to design a barrier joint with the required characteristics, using styrene, was described before this Conference in 1935.[2,3] The developments that have been made since that date will be described.

Monomeric styrene, the basic material of these joints and terminations, is a clear mobile liquid boiling at 146 degrees centigrade, and is a good solvent for cable-impregnating compound. It polymerises in time, changing from the liquid to

[2] T. R. Scott and J. K. Webb, Report 209 of the Conférence Internationale des Grands Réseaux Electrique à Haute Tension, 1935.

[3] T. R. Scott and J. K. Webb, "The Application of Styrene to H. T. Cable Systems," *Electrical Communication,* v. 16, pp. 174–179; October, 1937: pp. 276–284; January, 1938: v. 17, pp. 88–95; July, 1938; and v. 19, pp. 108–117; October, 1940.

the solid phase called polystyrene, even at ordinary temperatures. Polystyrene is not dissolved or otherwise attacked by cable compounds but dissolves easily in monostyrene. The formation of polystyrene may be retarded by the addition of certain inhibitors so that the liquid monomer may be stored for long periods; alternatively it can be accelerated by subjection to higher temperatures.

The original styrene barrier joints made use of the solvent properties of monomeric styrene to remove oil from a cable, and the subsequent application of heat converted this monomer to the insoluble polymer, thus forming a barrier to the oil. Although the original joints made by this method were satisfactory—joints made in 1934 are still in operation on 66-kilovolt systems—the long process time and the complications associated with the heating were disadvantages; efforts were therefore directed towards the development of a cold process.

Another objection to polymerisation *in situ* was the contraction (of the order of 15 per cent) that occurs during polymerisation, although in practice the voids so formed were sufficiently well distributed as not to affect the electrical properties of the joints.

Paper impregnated with polystyrene, referred to as styrenated paper, is the basis of the present high-voltage joints and terminations, and its development has led to the present cold process.

It was found that by using suitable plasticisers paper could be impregnated with monomeric styrene, polymerised in bulk, and subsequently unwound. The plasticiser is necessary to give flexibility to the paper in addition to facilitating the final unrolling.

To bond the styrenated jointing tapes without recourse to heating, a film of plasticised polystyrene lacquer is added to one side of the paper. The final product shows excellent electrical characteristics as may be noted from Figure 1.

Styrenated paper will gradually absorb moisture as is shown in Figure 2 and the paper is therefore stored in sealed containers with a silica-gel capsule. Tests of joints made under abnormal conditions show that the effect of moisture absorption is not important for applications at least up to 33 kilovolts.

The method of application to be described applies to a 33-kilovolt joint for a 3-core shielded cable, as shown in Figure 3.

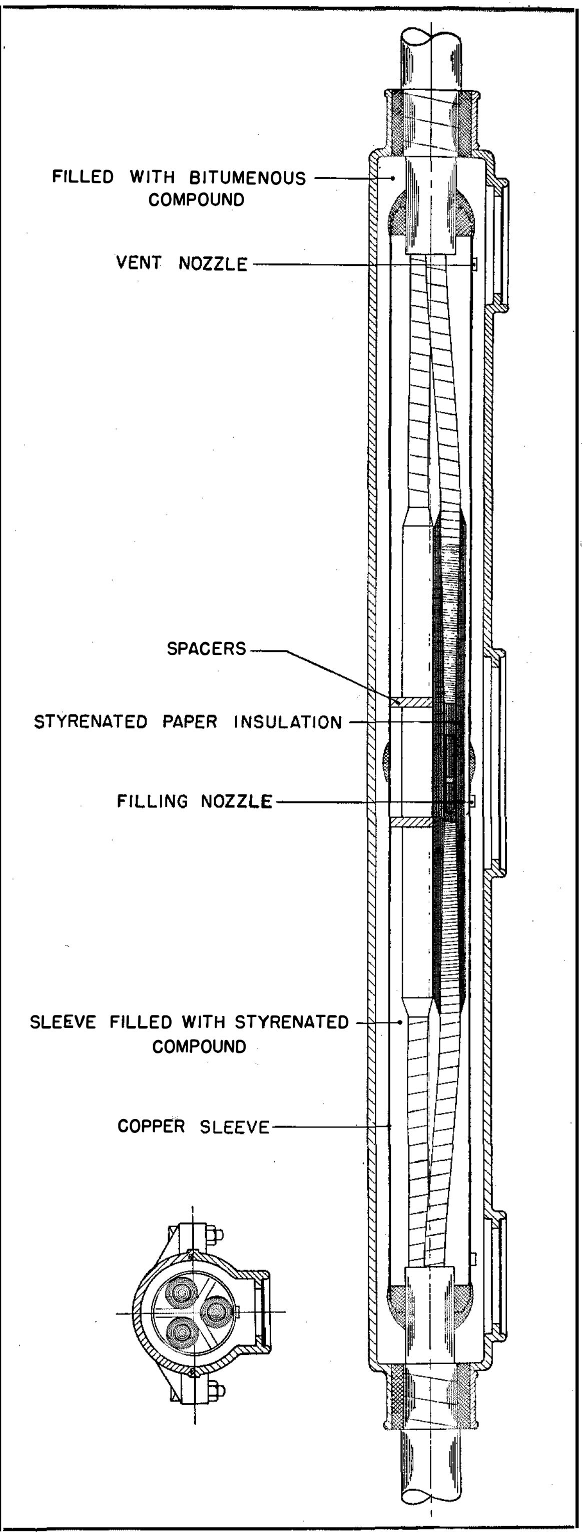


Figure 3—Styrene-barrier joint in a 33-kilovolt 3-core cable. Overall dimension are $69\frac{7}{8}$ inches long by 7 inches high by $9\frac{1}{4}$ inches wide (1775 by 178 by 235 millimetres).

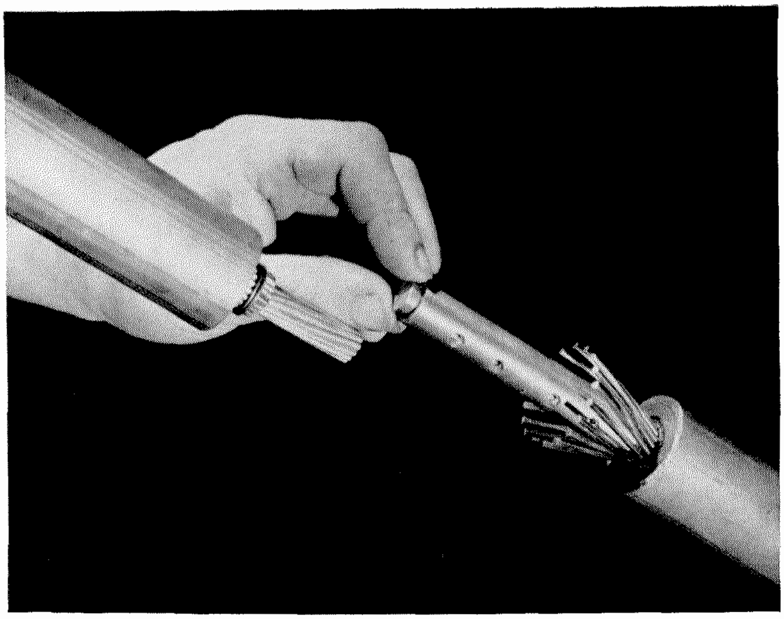


Figure 4—Assembly of conductor barrier ferrule for single-core styrene joint.

3.1 Oil Paths

In a barrier joint there are three main oil paths through which leakage may occur; the stranded conductor, the insulation, and the spaces outside the insulated conductors.

3.1.1 Conductor Path

To block the conductor path a tubular copper connector is used which is of a diameter equal to that of the strand, and which incorporates a loose copper disc to ensure complete sealing at one point, even if the tinning of the conductors is imperfect (see Figure 4). The connector is made flush with the strand to reduce the overall size of the joint and to avoid stress distortion due to change in diameter. Its use also eliminates the awkward recess that would otherwise occur between the normal type of connector and the insulation and which is not easily filled. Sufficient length of strand is exposed to allow the connector to be passed completely over one side, thus avoiding the necessity of bending the core to fit the ferrule.

3.1.2 Insulation Path

The cable insulation is tapered by tearing off each layer against a wire wrapped around the core. This method produces a rough edge with projecting fibres that will interlock with the applied styrenated-paper insulation and improves the bond between the two. The surfaces of the tapers are wiped with monomeric styrene to remove the impregnant from the outer layers of paper. Styrene cement is then rubbed in to replace the impregnant thus removed, and a gradual change from the oil in the cable to the styrene in the applied insulation is achieved with an effective bond at the interface.

Papers are added after brushing the previously applied paper with monomeric styrene and wrapping the next layer onto it with the film surface inside. The process is easier to the jointer than the application of oil-impregnated tapes, since the process is cold and the tapes bond into a hard solid mass onto which the next paper is readily wrapped. It will be obvious that a change will occur due to the polymerisation of the monomer applied during the build-up, and this results in an improvement in electrical characteristics as is evident from Figure 5.

It is usual to apply an insulation thickness 10 per cent greater than that of the cable, and those papers beyond the core diameter are extended to form a gradual taper.

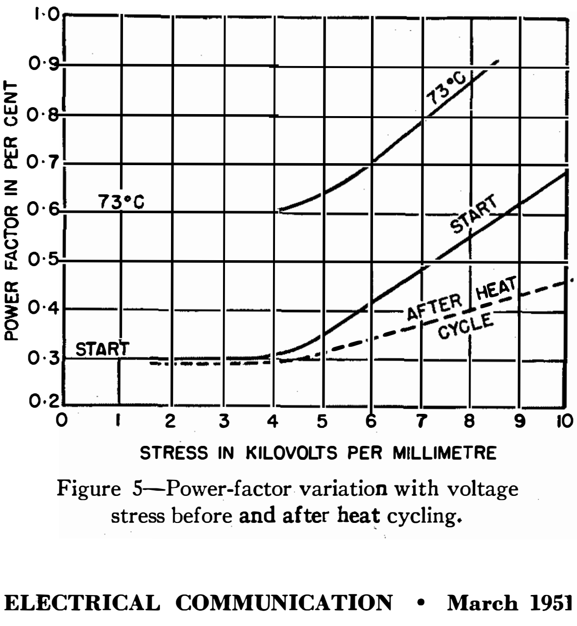


Figure 5—Power-factor variation with voltage stress before and after heat cycling.

A perforated copper-tape screen is carried over the insulated core, and the perforations allow the subsequent filling material to bond firmly to the styrenated paper insulation without interfering with the continuity of the screen. The small increase of diameter of the completed joint ensures a minimum disturbance of the electrostatic field between conductor and screen.

Thus are the conductor and dielectric paths blocked; after completion of the three cores it only remains to seal the third, the external path.

3.1.3 External Path

A copper sleeve is used to envelop the joint, a rigid sleeve being necessary to avoid the distension that would result with a lead sleeve. It can be shown that the forces due to the thermal expansion of the component materials of the joint do not exceed the elastic limit of the sleeve.

The sleeve is filled with a plasticised polystyrene compound too viscous to pour, even at the highest temperature. The method used is first to fill the heated joint sleeve completely with monomeric styrene to displace air; this in turn is displaced by the actual filling compound, which is introduced by a small portable hydraulic extruder fitted at the centre of the sleeve. The monomer is led away through pipe connections attached to orifices at the ends of the sleeves as may be seen in Figure 6.

The compound is shipped in tins, which are removed on site, and the slugs of compound transferred to brass cartridges which are heated in an oil bath to 180 degrees centigrade for at least two hours prior to filling. Several of these cartridges may be required to fill a joint, and their replacement is facilitated by a bayonet connection between the ram element and the base of the extruder. On completion of filling, the drain pipes are removed and compound flushed through until all traces of monomer have ceased. The end orifices are then sealed, and pressure maintained until the joint has cooled.

The process thus described results in a very satisfactory filling, free from air inclusions, and the compound bonds well to the sleeve and to the cores. Any traces of monomer that do remain diffuse into the surrounding polymer and become completely polymerised in course of time.

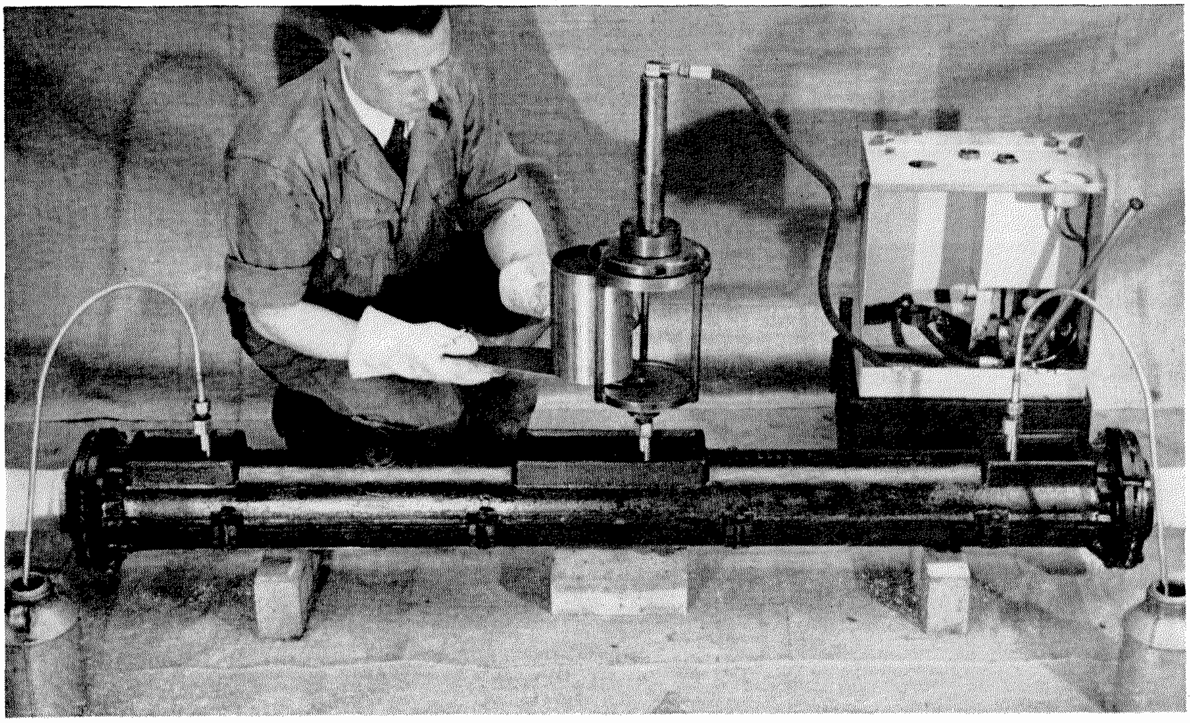


Figure 6—A portable hydraulic extruder is used to fill the sleeve of a 33-kilovolt 3-core styrene joint.

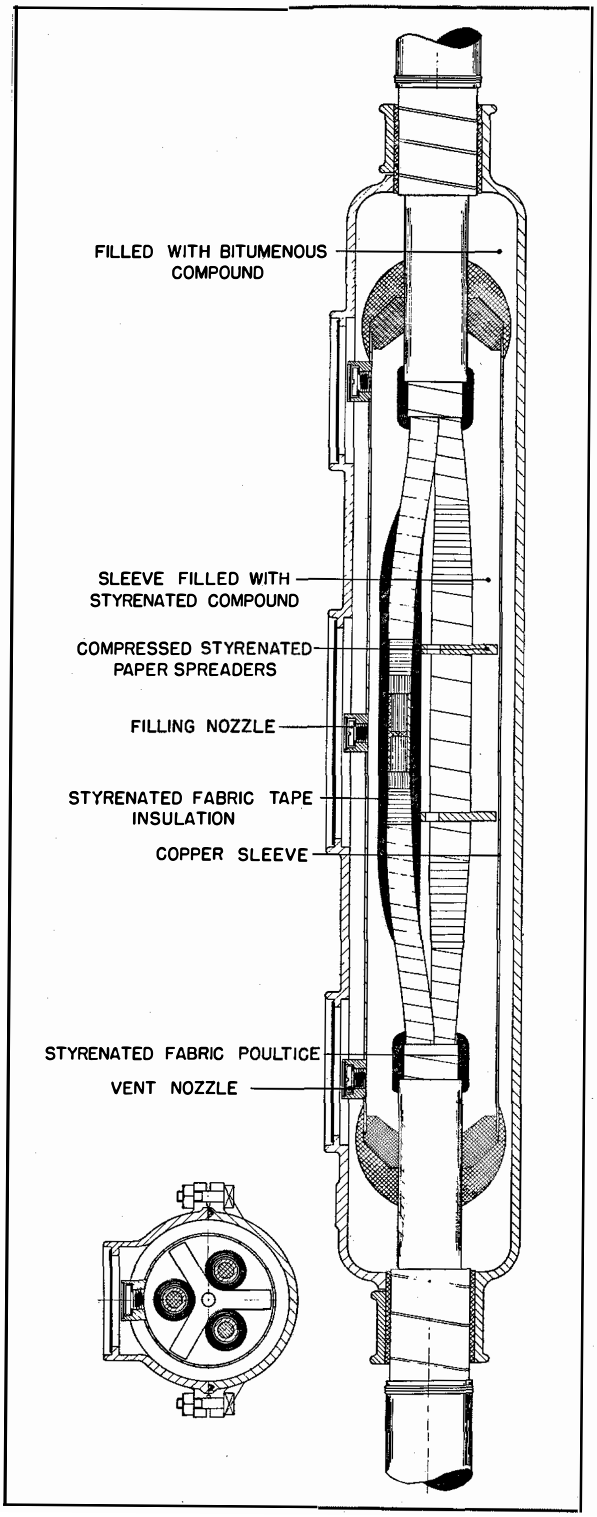


Figure 7—Styrene-barrier joint for 11-kilovolt 3-core cable. The overall dimensions are $32\frac{1}{8}$ inches long by $6\frac{5}{8}$ inches wide by $4\frac{1}{8}$ inches high (816 by 168 by 105 millimetres). Cross-section through centre is at lower left.

4. Joints for Lower Voltages

The design of styrene joints for lower voltages is very similar to that described, the main difference being the use of a styrenated-fabric tape in place of styrenated paper for the replaced dielectric. Unscreened joints have always been used below 22 kilovolts so that the metal spreaders used in the screened joints are replaced by spreaders made of compressed styrenated paper, thus preserving the homogeneity of the insulation. This structure is shown in Figure 7.

5. Miscellany on Joints

Some form of outer protection is required when joints are to be buried; a cast-iron box is usually preferred. This box provides a sound mechanical anchorage for the cable armour and, when finally filled with bitumen, ensures a satisfactory anti-corrosion protection for the joint sleeve. To avoid reheating of the joint sleeve with the hot bitumen the boxes for styrene joints are designed to permit filling of the joint sleeve through openings in the lid, after the box has been partly filled with bitumen. This method also conveniently heats the sleeve prior to filling.

Other types of styrene joints such as those connecting 3-core shielded cable to three single-core cables (trifurcating joints), joints for 3-core single-lead-sheathed cables and between them and 3-core shielded types, or for single-core cables are based on the same principles. Paper-to-varnished-cambric joints have also been installed.

The small size of styrene joints is an advantage and often simplifies installation, particularly in multiple cable runs. Figure 8 shows the difference in size between a styrene joint and a mechanical barrier joint for the same size of cable.

Installations have been made for voltages up to 66 kilovolts and satisfactory experience extends over a period of 16 years at these voltages.

6. Terminations

In addition to the problem of barrier joints there is a considerable demand for barrier terminations. These are required to prevent loss of cable compound from an end where a large difference in level exists near the termination, particularly when the end is subjected to high ambient temperature, or with cables of large conductor cross section.

Barrier terminations in transformer or switchgear cable boxes permit the use of transformer or switchgear oil as a filling medium, and the box may be piped to the main conservator system.

Ratchetting effects have been observed where a bitumenous filling compound has been used in cable end boxes, and in one case on record, fracture of the half-inch-thick cast-iron box resulted.

6.1 Styrenated-Paper Tubes

Early attempts to apply the styrenated-paper-tape technique to terminations were not entirely successful, owing to the difficulty of making a seal to the lead sheath, but it was soon discovered that the styrenated paper in its initial form, i.e., without the lacquer film, could be wound under heat and pressure to form tubes. These tubes, which resemble the well-known phenol-formaldehyde-impregnated-paper product, have excellent mechanical and electrical properties, and, when fitted with metal end caps by a special process, provide a solution to the barrier-termination problem.

For voltages up to 11 kilovolts a plain tube is used, and for higher voltages metal foils are interwoven to provide capacitive grading as in the condenser cone of Figures 9 and 10. The bushes are made with the internal diameter slightly larger than that of the cable core, which is cleaned with monomeric styrene and lapped with styrenated paper as for a joint. The built-up paper is then coated with styrene cement and the bush forced on to the core. A connector previously sweated to the conductor projects through the outer end cap and is held by nuts, a seal being provided by a neoprene *O* ring in a groove. The other end of the bushing is wiped to the cable sheath.

To ensure perfect filling on the high-voltage terminations, a hole is drilled concentrically through the connector, communicating with a lateral hole inside the bushing, so that styrene cement may be forced into the termination from a plug in the sweating gland, thus forcing any air out through the hole in the connector.

An extremely good bond is obtained between the cable core and the applied insulation, and between the latter and the styrenated paper tube, resulting in a solid homogeneous construction of great permanence.

These advantages have been widely recognised, particularly on important applications, such as generator cables, where freedom from maintenance troubles is desirable.

The hydraulic pressures that these terminations will withstand are considerable, and standard terminations are used regularly in the laboratory for pressures of 200 pounds per square inch.

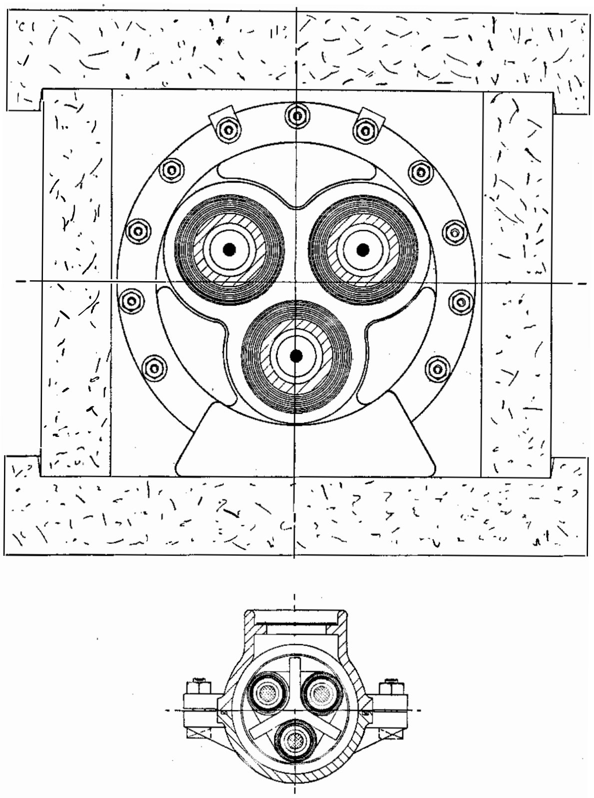


Figure 8—The mechanical-barrier joint at the top has outside dimensions of 24 by 22 inches (610 by 559 millimetres) while the styrene joint is $9\frac{1}{4}$ inches by 7 inches (235 by 178 millimetres). Both are for a 33-kilovolt 3-core shielded cable.

7. *Service Records*

Styrene joints and terminations have been in operational service since 1934, and it may be of interest to review the good record they have established under a wide range of conditions of service.

7.1 Joints

The approximate number of joints in service is 900, made up as shown in Table 3.

All faults so far recorded have been due to causes other than the styrene technique.

Representative 50-cycle-per-second values are shown in Table 4 for 33-kilovolt joints made under normal, good, and abnormally bad conditions. The normal conditions are without delays during jointing, with relative humidity between 55 and 65 per cent. Good conditions are again without delays, but with a relative humidity approximately zero per cent; and under bad conditions the joints were exposed purposely for several days during manufacture, with relative humidity up to 65 per cent.

Typical impulse-voltage results on 33-kilovolt joints subjected to 10 positive and 10 negative impulses of 193·5-kilovolt peak value on a 1/50th-microsecond wave followed by negative impulses successively raised by 20-kilovolt increments until breakdown, are given in Table 5.

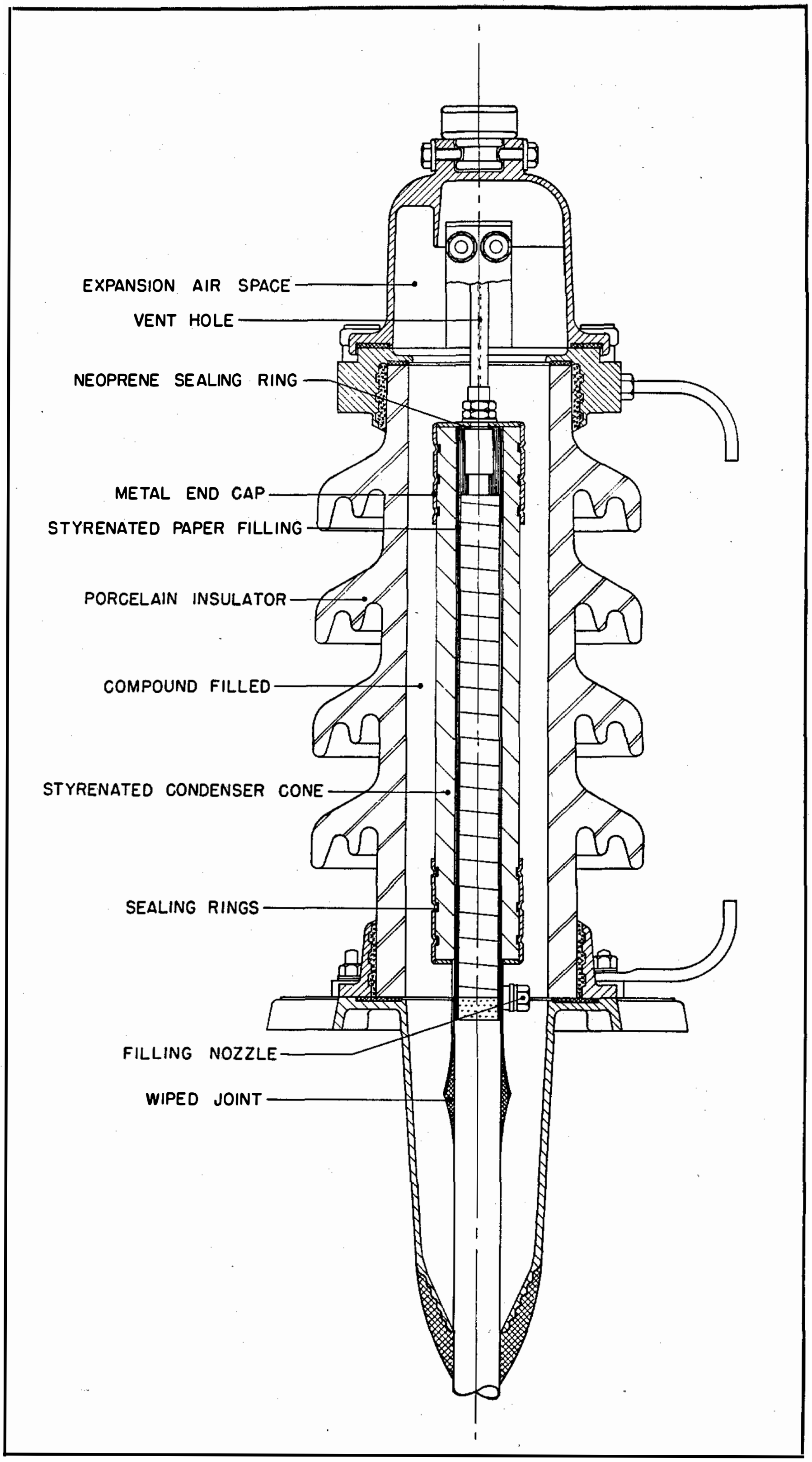


Figure 9—Sealing end of 33-kilovolt bushing incorporating styrenated condenser cone.

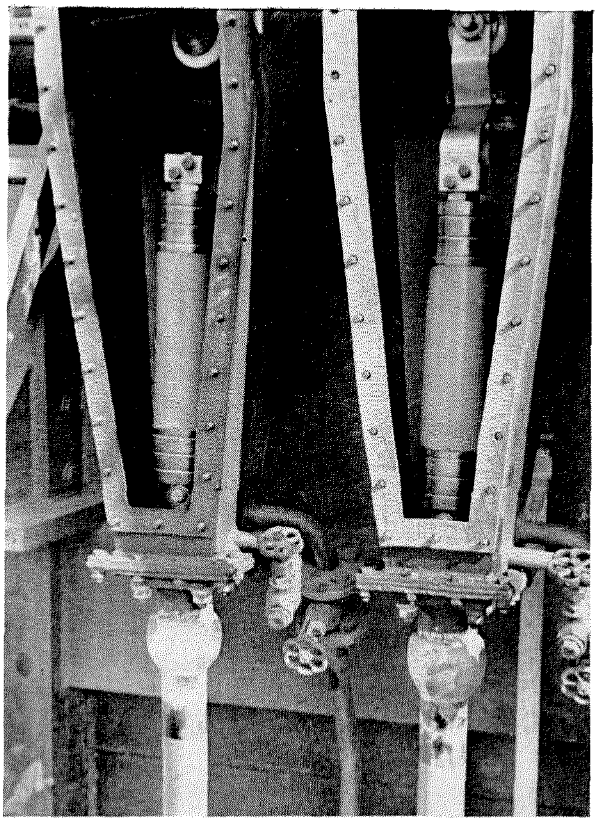


Figure 10—Styrenated condenser terminations in transformer box for 33-kilovolt service. Styrene cement gives a good bond between the cable core and the condenser cone.

The joints are designed to have initial breakdown values approximately equal to those of the cable, but owing to improvement with age, it is usually necessary to use cable rated for a slightly higher voltage if breakdown values of aged joints are required.

Barrier characteristics of joints are measured by applying pressure to the cable on one side of a joint, using a fluid similar to that with which the cable is impregnated. The cable on the other side of the joint is cut short and a lip is formed to enable leakage oil to be collected. The joint is mounted almost vertically to eliminate natural drainage. Provision is made for heating the conductor electrically. Typical results are shown in Table 6.

7.2 Terminations

Details of installed styrenated terminations are given in Table 7. These figures do not include some of the very early styrene cable plugs—some of which are still in service—but refer only to the bonded paper bushes.

No failures have occurred on terminations up to 11 kilovolts and only one on a 33-kilovolt system. The cause has not been determined but it occurred on the only horizontal arrangement so far installed, and incorporated a plastic filling

TABLE 3

Styrene Joints in Service

	3 to 11 Kilovolts	22 Kilovolts	33 Kilovolts	66 Kilovolts
Joints Supplied	385	156	351	16
Joint-Years	710	226	808	168
Breakdowns	—	—	6	—
Failures Per Joint-Year	—	—	0.0074	—

TABLE 5

Impulse Strength of Styrene Joints

Sample	Conditions of Making	Breakdown Value in Peak Kilovolts	Breakdown
1	Normal	485	Joint
2	Normal	520	Cable Failed
3	Normal	530	Cable Failed
4	Normal	430	Joint

TABLE 4

Breakdown Characteristics of Styrene Joints

Conditions of Making	Initial Power Factor at Working Voltage Times		Power Factor After 48 Heat Cycles to 60 Degrees Centigrade at Working Voltage Times		Hours at 3 Times Working Voltage	Failures
	1/2	2	1/2	2		
Bad	0.014	0.03	0.017	0.018	43	Cable*
Normal	0.0045	0.007	0.005	0.005	> 92	None
Good	0.0035	0.0075	0.0029	0.0037	>124	None

* The cable used had been subjected to severe testing previously.

between bushing and cable core, an obsolete method now superseded. The direct-current terminations are installed on electrostatic precipitators, and the high incidence of failures on these is largely due to the rather unusual and strenuous conditions under which these equipments operate.

TABLE 6

Barrier Characteristics of 33-Kilovolt Styrene Joints

Condition of Making	Pressure in Pounds Per Square Inch	Conductor Temperature in Degrees Centigrade	Hours of Test	Leakage* in Cubic Centimetres
Normal	100	Room	680	—
Normal	100	Room	3000	—
Normal	100	Room	3000	—
Normal	100	Room	680	—
Normal	100	Room	2000	—
Normal	100	Room	2000	—
Normal	40	40	100	—
Joints after 3 Years Service under a Static Head of 33 and Transient Pressure Up to 85 Pounds per Square Inch	40	17	351	—
	40	60	48	1
	40	70	48	—
	40	80	28	15
	40	17	24	—
	40	60	62	15
	40	70	48	20
	100	18	120	—

* The leakage figure represents the total oil collected and therefore includes the displacement of oil due to expansion in the adjacent cable.

Improvements have been made in this application and there have been no failures since the solid paper filling was introduced.

Hydraulic characteristics of terminations can be accurately defined. No leakages have been observed during any tests on bushings at pressures up to 200 pounds per square inch over periods exceeding six months. Samples tested to destruction failed by bursting of the copper end cap at between 800 and 1000 pounds per square inch.

Long-time pressure tests, with transformer oil applied both inside and outside, show no leakage, indicating that the sealing rings are not affected. Results of some high-pressure tests are given in Table 8.

Operating experience has shown that a very high factor of safety is obtained, and tests show that the asymptotic breakdown value of a 16-inch barrier condenser cone as used on 33-kilovolt systems (19 kilovolts to earth) is in excess of 50 kilovolts.

8. *Jointing Time*

No mention has so far been made of the time required for jointing. This will depend largely on the individual and his experience, but after the first one or two joints have been made the time should be not more than 10 to 15 per cent longer than a screened oil joint of similar type. In making any comparison it should be noted that load can be applied to a styrene joint about three hours after filling. There is no long cooling period or topping-up process. Barrier terminations are easily fitted in a few hours and are generally suitable for installing in normal cable boxes and porcelain terminations.

TABLE 7

Styrene Terminations in Service

	3 to 11 Kilovolts	33 Kilovolts	30 Kilovolts Direct Current	60 Kilovolts Direct Current
Terminations Supplied	540	97	917	71
Termination-Years	2014	259	1943	108
Breakdowns	—	1	33	—
Failures Per Termination-Year	—	0.004	0.017	—

TABLE 8

Barrier Characteristics of Styrene Terminations

Sample	Final Pressure in Pounds Per Square Inch*	Hours of Final Pressure	Leakage
1	700	5	None
2	500	6	None
3	600	6	None
4	700	6	None
5	500	4	None
6	800	12	None

* Pressure raised in 50-pound increments per 8-hour period from 250 pounds per square inch.

Design of a Loran Transmitter

By R. H. MYERS

Federal Telephone and Radio Corporation; Clifton, New Jersey

VARIOUS PROBLEMS that are encountered in the design of transmitting equipment for loran are discussed. A new type of loran transmitter, using grid modulation in the penultimate stage to shape the transmitted pulse, is described. The pulse shape is controlled to achieve a minimum of sideband radiation with concomitant reduction in interference from the loran transmissions.

• • •

1. *Introduction*

During the second world war, loran[1] became one of the principal long-range navigational systems employed by the allies, and in the postwar period it has remained in widespread use.

After the second world war, the United States attempted to obtain from the International Telecommunications Union the allocation of a frequency band in which the loran service would operate. This was granted with some restrictions; the Acts of the conference specifically stating that such approval is given "provided that . . . all practical measures are taken to minimize harmful interference from loran transmissions to other services operating in the same or adjacent frequency bands and, in particular, to narrow the emitted bandwidth." To assure that it met the commitments that it had made and to improve the equipment performance in other ways, the United States government immediately took action to bring about the following changes in the characteristics of the loran transmitters:

A. Narrowing of the transmitter frequency spectrum to reduce adjacent-channel interference without loss of system accuracy.

B. An increase in transmitted power to about 1-megawatt peak with facilities for operating all stages of the transmitter except the final amplifier to obtain a reduced output of about 160 kilowatts.

C. A transmitting system that would provide a fixed phase relationship between the radio-frequency cycles and the envelope of the output pulse. This would permit a cycle-matching technique to be employed at the loran timer or at a receiver-indicator.

Loran transmitting equipment in service at the start of the present development was of the pulsed self-excited-oscillator type. Power output from these transmitters was of the order of 100 kilowatts, and the approximately rectangular pulses produced a wide radiated spectrum. The pulse rise time from 10- to 90-percent amplitude was about 10 microseconds and the width at 50-percent amplitude, about 40 microseconds.

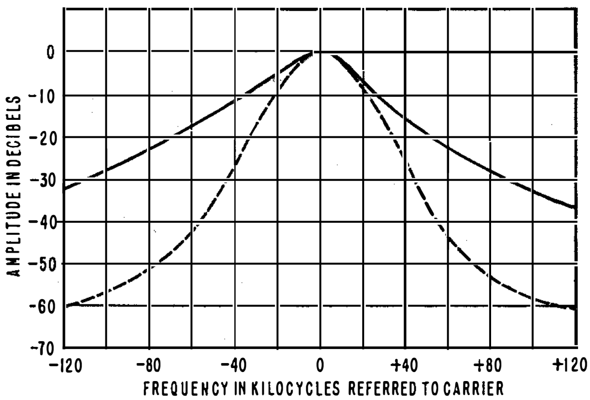


Figure 1—The radiated spectrum of the pulsed-oscillator loran transmitter is given by the solid line. Proposed limits for the new equipment are indicated by the dashed line.

In the new requirements, the transmitted spectrum was to be considerably restricted as compared to the output of the pulsed oscillator. The comparison is illustrated in Figure 1. The rise time of the transmitted pulse was not to exceed 25 microseconds, the half-amplitude width was to be about 40 microseconds, and the width at 10-percent amplitude was not to exceed 65 microseconds. By a process of Fourier analysis applied to relatively simple wave shapes, the pulse spectrum and shape requirements were readily shown to be compatible. In particular, the

[1] John L. Pierce, "Loran Long-Range Navigation," Massachusetts Institute of Technology Radiation Laboratory Series, V. 4, McGraw-Hill Book Company, New York, New York; 1948.

analysis showed that a pulse of the shape of a single cycle of the cosine-squared function was well suited to both the shape and spectrum requirements.

2. *Pulse Shaping and Spectrum Control*

In proceeding with the design of the equipment, the method by which the pulse shape and spectrum width was to be controlled was of basic importance. Two general methods were seriously considered:

A. Obtaining the desired pulse shape and spectrum by generating a pulse in the transmitter whose rise time was considerably faster and whose spectrum was considerably wider than that desired, and by then passing this pulse through a band-restricting filter placed between the transmitter and the antenna.

B. The generation of the desired pulse shape by amplitude modulation of a linear stage in the transmitter and by using linear amplification to raise this pulse to the desired power level.

Preliminary design of a filter for method A showed that to obtain the spectrum characteristics indicated in Figure 1, at least 4 tuned and critically coupled stages of filtering would be required. The entire filter would have to be capable of adjustment to any one of the 5 loran frequencies between 1750 and 1950 kilocycles. The alignment of a filter of this type in the field appears formidable, to say the least. Insertion loss in the filter would approach 3 decibels, and peak voltage would exceed 20 kilovolts. These factors indicated that transmitting equipment utilizing a filter would be relatively large and difficult to handle by field personnel.

Consideration of method B showed that two linear power-amplifier stages would be required; a penultimate stage that would be modulated to obtain the desired pulse shape at a 160-kilowatt level and a final stage that would amplify the pulse to a 1-megawatt peak level with small distortion. Ideally, with zero distortion in the linear stages and no selectivity in the antenna circuit, the envelope of the output pulse would be completely determined by the modulating pulse. Practically, some small deviation could be expected between them, but this could be compensated for by predistortion of the modulating pulse.

For modulating the penultimate stage, grid modulation appeared to be the most practical because of the smaller modulating power required as compared to plate modulation. Since the stage would be operating near peak conditions for a large percentage of the modulation cycle, the efficiency would be comparable to that for plate modulation. Calculations showed that about 50-percent plate efficiency could be expected and that a medium-sized hydrogen

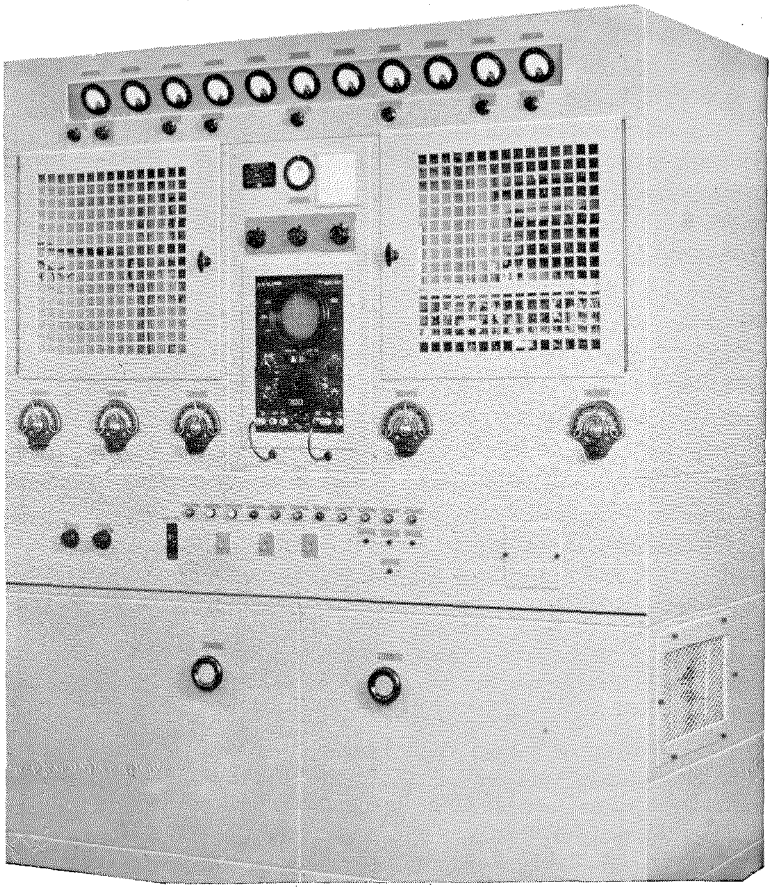


Figure 2—The loran transmitter. The power supplies are placed in the lower part of the cabinet. Monitoring oscilloscope is in the center.

thyratron, the *4C35*, would furnish adequate modulating power. Both of these indications were favorable.

The final 1-megawatt stage would operate essentially as a class-*B* linear amplifier. This stage would give about 50-percent plate efficiency, and since the losses in the plate and output circuits would be only about 0.5 decibel, the linear amplifier appeared quite practical.

On the basis of the foregoing analysis of methods *A* and *B*, method *B* was selected for the equipment design, and the pulse shape was to approximate that of one cycle of the cosine-squared function. The transmitting equipment was divided into two portions. The first of these, called the transmitter proper, contains the frequency-generating stages, two intermediate amplifiers, and the modulated penultimate amplifier. The transmitter may be used separately to give an output of 160 kilowatts. The other portion of the equipment is the linear amplifier, used to amplify the transmitter output to a 1-megawatt level.

A description of the more salient features of the transmitting equipment that evolved from this beginning and the performance characteristics are discussed in the remaining portion of this article.

3. Loran Transmitter

3.1 General

A photograph of the cabinet of the transmitter is given in Figure 2, and a block diagram of the electrical circuits appears in Figure 3. The inputs to the exciters consist of a 100-kilocycle standard frequency and repetition-rate triggers at a nominal 20, 25, or $33\frac{1}{3}$ pulses per second. Eight specific rates have been established in the vicinity of each of these nominal rates.

To provide for double pulsing (simultaneous pulsing at two of the specific rates associated with the same nominal rate), two exciters are included in the transmitter. In the event that double pulsing is not required for the particular station in which the transmitter is installed, the extra exciter is used as a standby.

The output of the transmitter is a radio-frequency pulse recurring at one or two of the above rates, and having a very accurately controlled carrier frequency of 1750, 1800, 1850, 1900, or 1950 kilocycles. Because the carrier frequency is directly related to the 100-kilocycle input, and because the repetition-rate trigger is a direct submultiple, a phase lock exists between the output pulse envelope and the radio-frequency cycles that make up the pulse.

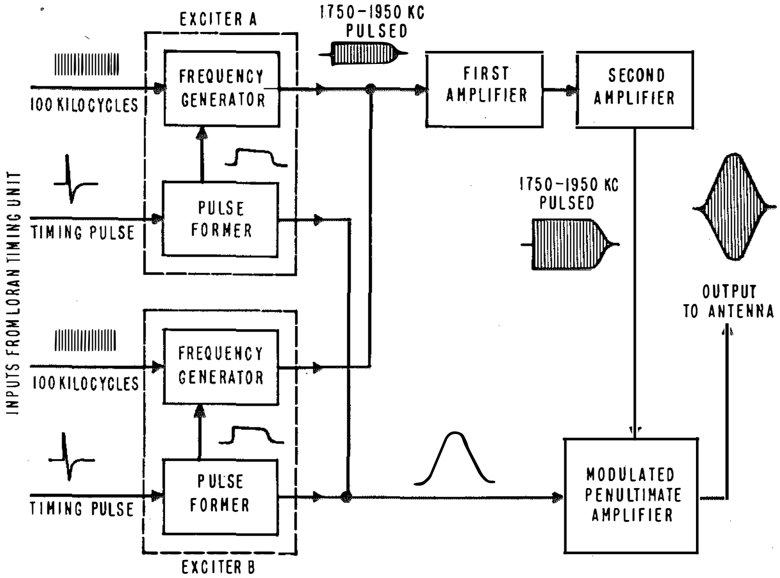


Figure 3—Block diagram of the transmitter. Peak output of 160 kilowatts may be applied either to the antenna or to a power amplifier.

The output pulses have the cosine-squared shape mentioned above, and a peak output of 160 kilowatts is available when the transmitter is single pulsed, or 128 kilowatts when double pulsing is used. The output is applied through a 52-ohm coaxial cable either to the amplifier or directly to the antenna.

3.2 Exciters

Two identical exciters are included in each transmitter. A block diagram of one of these units is given in Figure 4.

3.2.1 Frequency Generator

Those stages that receive the 100-kilocycle standard frequency from the timer and convert it into a carrier frequency for amplification and modulation in the later stages of the transmitter are called the frequency generator.

Provision is made in the frequency generator for choice of any of the 5 loran frequencies by a front-panel switch and two tuning controls. Conventional frequency-multiplier and divider stages are employed, with interstage coupling provided by tuned circuits using iron-dust-cored coils.

This signal is applied to an amplifier–limiter, and thence to a push-push frequency doubler. The doubler stage is normally biased so that plate current does not flow even with the radio-frequency driving voltage present. The stage is then gated "on" with a positive bias pulse applied to the grids. The gating pulse is approximately 100 microseconds long and occurs at the loran repetition rate. The output of the doubler, consisting of rectangular radio-frequency pulses at the station carrier frequency, and repeated at the repetition rate, is applied to the amplifier chain. A considerable gain in the efficiency of the amplifiers, which must furnish radio-frequency drive to the penultimate stage, is obtained by gating them only for a period slightly longer than required by the width of the modulation pulse.

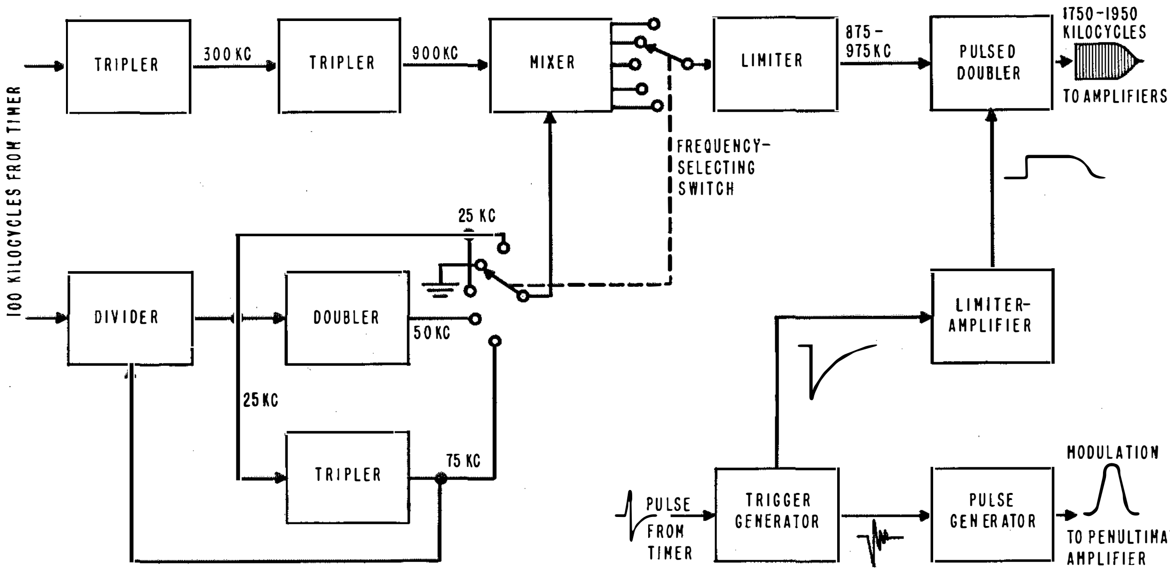


Figure 4—Block diagram of one of the exciters. Those stages that generate the pulse waveforms are called the pulse fomer; these are the three blocks at the lower right. The rest of the circuit generates the carrier frequency.

Two tripler stages multiply the 100-kilocycle input to 900 kilocycles, which is applied to a mixer stage. A combination of a frequency-divider, a doubler, and a tripler can also provide stable 25-, 50-, or 75-kilocycle signals to the mixer. The divider system is of the regenerative type in which the 75-kilocycle signal from the last stage is fed back to the first stage to mix with the 100-kilocycle input and produce 25 kilocycles. Doubling of the 25-kilocycle signal produces 50 kilocycles, and tripling produces 75 kilocycles.

The proper divider output frequency is selected by a 5-position switch and combined with 900 kilocycles in the mixer. Other contacts of the same switch select the proper tuned circuit at the mixer output for rejection of all but the desired frequency from the various sum and difference frequencies present. The output is at any of 5 frequencies at 25-kilocycle intervals between 875 and 975 kilocycles.

3.2.2 *Trigger Generator*

The trigger generator is essentially an isolation stage utilizing a type *2050* thyratron to discharge a resistance–capacitance network in its plate circuit upon arrival of the repetition-rate trigger pulse from the timer. There are two outputs from the trigger generator, one of which is a negative pulse of rather long time duration. This is amplified, clipped, and inverted in the limiter–amplifier stage to form a rectangular pulse about 100 microseconds in duration, and is then applied as the gating pulse to the grids of the pulsed doubler.

The other output of the trigger generator is a very sharp negative pulse that is inverted in a transformer and applied to the grid of the pulse generator, a type *4C35* thyratron.

3.2.3 *Pulse Generator*

The schematic diagram of this stage is given in Figure 5. The *4C35*, with no signal input and zero bias, is normally extinguished, and capacitor *C1* becomes charged to 6000 volts through *R1* and *L1*. The leading edge of the positive pulse at the grid fires the tube, and an oscillatory circuit composed of *L1* and *C1* discharges through the tube, developing a 2250-volt positive pulse across *R2* in the cathode circuit. When the discharge current passes through zero, the tube extinguishes, and the grid regains control until the next pulse.

The cosine-squared shape desired in the modulating pulse is developed by the oscillatory elements in the plate circuit of the tube discharging into the cathode network consisting of *C2*, *C3*, *L2*, and *R2*. The width of the output pulse is adjusted by means of variable inductor *L1*, and the rise time by means of *L2*. These adjustments are only slightly interdependent. In order that the nonlinear loading effect of the penultimate-amplifier grid circuit may not appreciably affect the wave shape, the value of *R2* is made low, being only 100 ohms.

3.3 Intermediate Amplifiers

The two intermediate amplifiers raise the pulsed-doubler output to a level sufficient to drive the penultimate amplifier.

3.3.1 *First Amplifier*

The first amplifier is a conventional single-ended tuned-grid–tuned-plate stage using an *807* tube. The tube is biased to cutoff, so that no plate current flows between pulses of radio-frequency input. Both grid and plate circuits are loaded with resistors to give sufficient bandwidth for passing the rectangular 100-microsecond pulses.

3.3.2 *Second Amplifier*

The second amplifier uses two *715C* tubes in a cross-neutralized push-pull arrangement. The stage is also biased to prevent plate-current flow in the absence of an input. The plate voltage used is 7200 volts, the screen-grid voltage is 1200 volts, and the peak power output is of the order of 14 kilowatts.

3.3.3 *Obtaining Constant Output Pulse Amplitude When Double-Pulsing a Transmitter*

A rather interesting circuit feature has been incorporated in conjunction with the second intermediate amplifier for the purpose of compensating for variations in output pulse amplitude that will occur when a loran transmitter is operated at two repetition rates simultaneously, i.e., double pulsed. A review of the various aspects of the problem may be helpful here.

Double pulsing is usually used at a master station of a chain, with two slave stations each synchronizing with one of the two rates. A complete description of these principles is given in reference 1. The repetition rate of a loran station is nominally 25 pulses per second, and

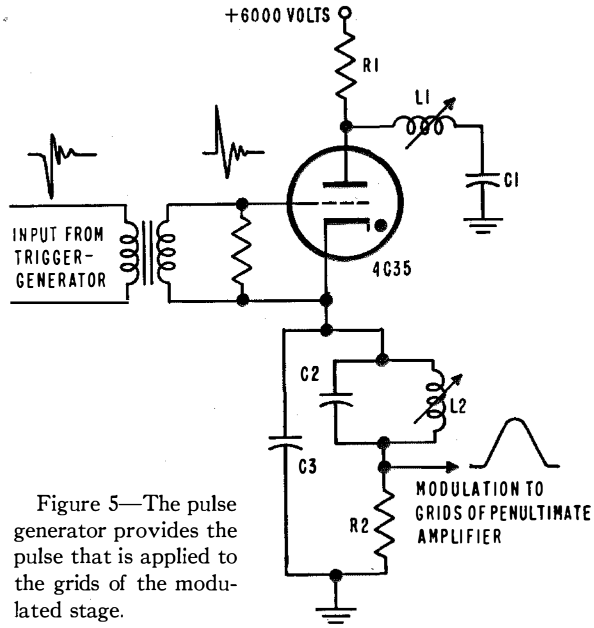


Figure 5—The pulse generator provides the pulse that is applied to the grids of the modulated stage.

in double pulsing the two rates usually differ by about 1/16 pulse per second. Therefore, the pulses corresponding to the two rates are constantly changing position in time with respect to each other and the pulses occur simultaneously about once in 16 seconds.

The direct-current energy to the various stages of the transmitter is supplied from rectifiers with storage or filter capacitors that supply the flywheel action necessary between the relatively low, but continuous, output of the rectifiers and the large, but short-duration, surges of current drawn by the transmitter. Because the various storage capacitors cannot have infinite capacitance, a voltage drop occurs across them with each transmitted pulse, and this voltage recovers toward the initial value in essentially an exponential manner until the next pulse occurs. The voltage levels on the various stages at the instant this second pulse comes along thus depend on the time separating it from the first pulse. When double pulsing, this time separation is constantly changing and causes the transmitted-pulse amplitude to vary at the difference frequency between the two rates. This variation is extremely undesirable because it spoils the accuracy with which master- and slave-station signals can be matched at the indicator.

Basically, the degree of amplitude variation that is present can be controlled by the amount of storage capacitance used in the power supplies. This has practical limitations, however, from the standpoints of cost and space required. With 8 microfarads of storage for the 1-megawatt amplifier and 2 microfarads for the penultimate amplifier, all rated for 20 kilovolts, the pulse amplitude variation was about 5 percent. It was desired that it be 2 percent or less. To obtain the smaller variation by increasing the storage capacitance would require that it be at least

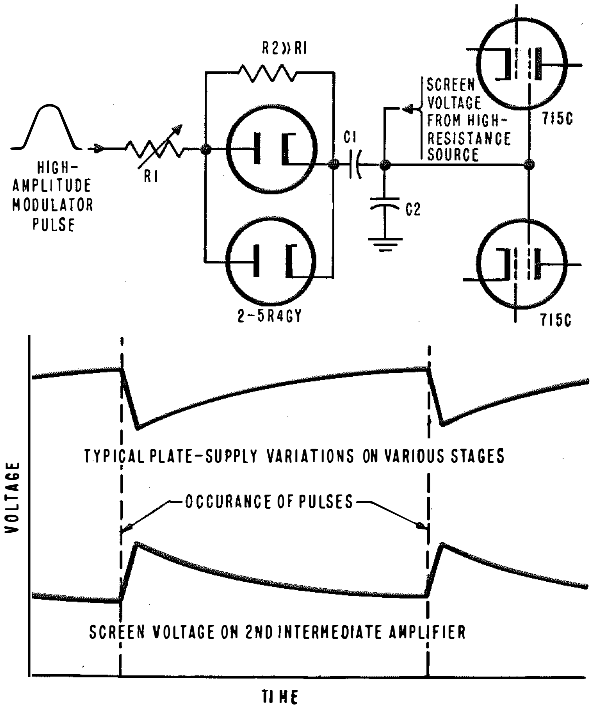


Figure 6—At the top is shown the circuit used to vary the output of the second intermediate amplifier to compensate for changes in plate-supply voltages throughout the transmitter. Below are the voltage waveforms involved.

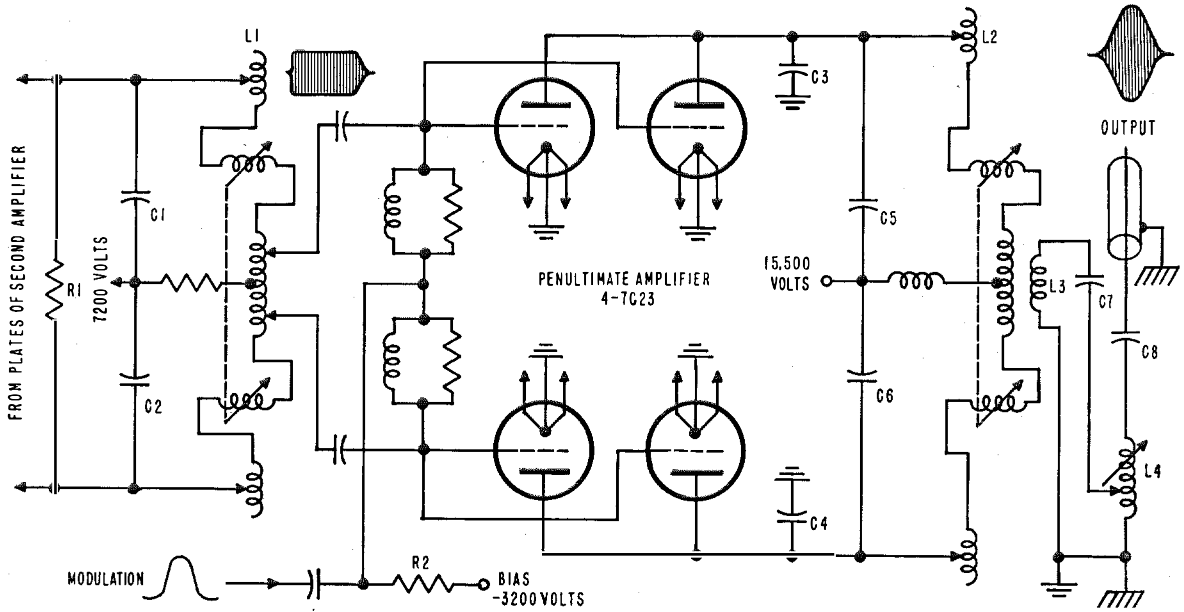


Figure 7—Simplified schematic diagram of the modulated penultimate amplifier.

doubled, which is uneconomical. A method was therefore developed to compensate for amplitude variations rather than to eliminate them by the addition of capacitance.

The principle used was to make the screen voltage of the second intermediate amplifier vary inversely in time to the direct-current plate potentials on this and succeeding stages. Thus, when reduced plate voltages tend to reduce the amplitude of the transmitted pulse, the increased screen voltage increases the radio-frequency drive to the penultimate stage and maintains the output constant. A schematic diagram of the circuit is given in Figure 6.

The positive modulator pulse, whose primary function is to shape the transmitted pulse at the grid of the penultimate stage, also passes through resistor *R1*, the two diodes, and blocking capacitor *C1* to storage capacitor *C2* and the screen grids of the *715C* tubes. Capacitor *C2* integrates the current due to the pulse, causing a voltage rise at the screens. This voltage immediately begins to decay exponentially, as determined by *C2*, *R1*, and the resistance of the direct-current screen-supply source, until the occurrence of the next pulse. The change in the screen voltage, and hence the drive to the penultimate amplifier is, therefore, inverse to the plate-voltage variation as is shown at the bottom of Figure 6. The degree of compensation is controlled by *R1* and the time constant is adjusted by means of *C2* to approximate that of the plate-supply variation.

Actual results using this circuit were that pulse-amplitude variations were reduced to less than 2 percent without any additional plate storage capacitance.

3.4 Penultimate Amplifier

Four type *7C23* tubes are employed in the parallel–push-pull amplifier circuit. A diagram of this circuit is shown in Figure 7. For simplicity, conventional circuit elements such as the bypass capacitors, parasitic-suppressors at the grids, and metering circuits are not shown here.

The input tuned circuit, composed of *L1*, *C1*, and *C2*, is loaded with resistor *R1* to prevent distortion of the driving pulse. The driving pulse is applied through blocking capacitors to the grids of the four *7C23* air-cooled triodes, which

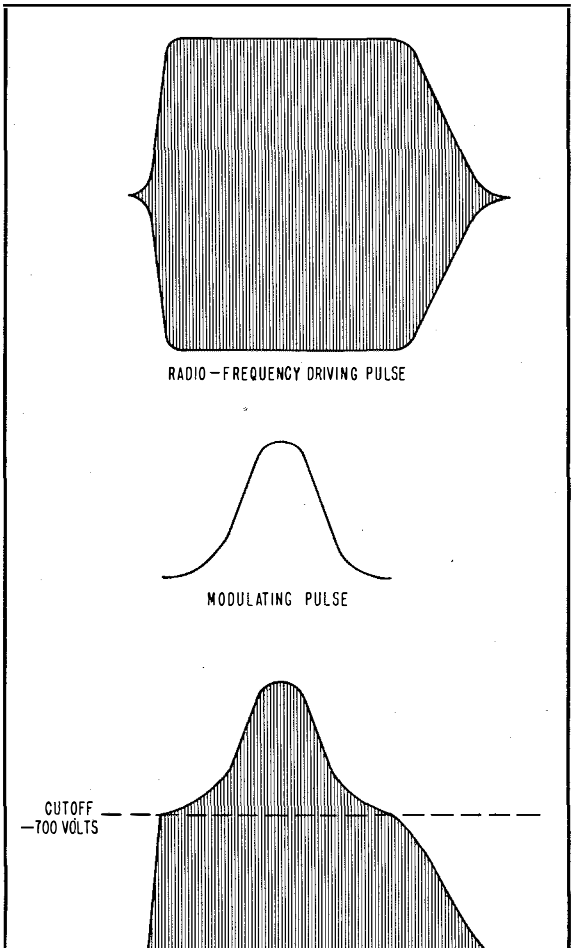


Figure 8—Penultimate-amplifier modulation waveforms.

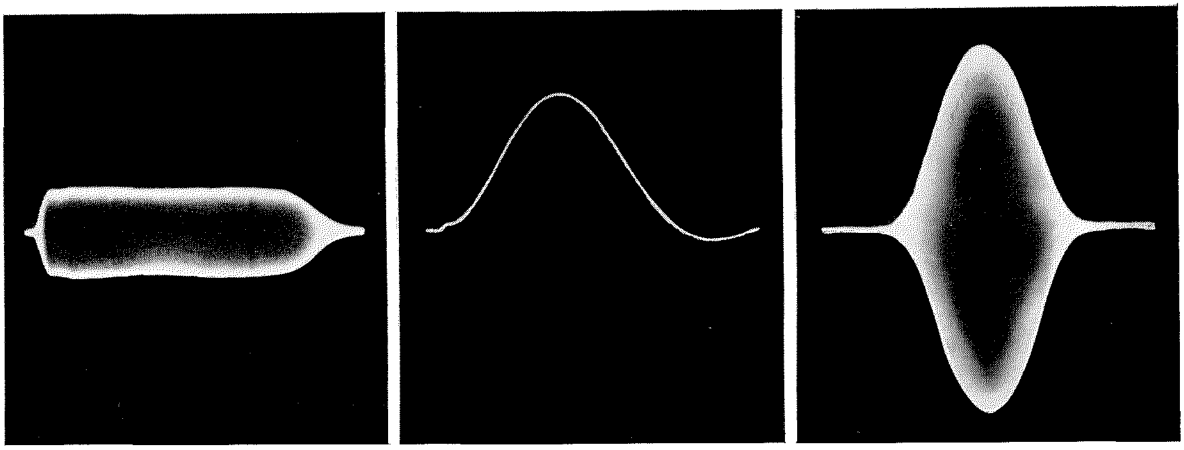


Figure 9—At the left is the carrier-frequency driving pulse at the grids of the amplifier as viewed on the monitoring oscilloscope. At the center is the modulating pulse, and at the right is the transmitter output. The variable-delay oscilloscope sweep is 100 microseconds long in each case.

are biased so that when the driving pulse is present, but with no modulation pulse, the tubes are just on the verge of conduction. This bias, approximately −3200 volts, is applied through isolating resistor *R2*. Application of the modulating pulse has the effect of shifting the bias of the tubes to permit amplification of the input in accordance with the shape of the modulating pulse. This process is illustrated in Figure 8. Actual photographs of these waveforms are given in Figure 9.

Returning to Figure 7, the plate tank consists of capacitors *C3*, *C4*, *C5*, and *C6*, and inductance *L2*. Output is taken from the circuit by inductive coupling between *L2* and *L3*. The reactance of *L3* is series resonated by capacitor *C7* and this circuit is directly coupled to the second series-resonant circuit consisting of *L4*, *C8*, and the load. This double-tuned output circuit alone gives harmonic suppression in excess of 40 decibels; this being in addition to the suppression given by the push-pull class-*B* operation.

A 52-ohm coaxial line connects the transmitter output either to the antenna or to an amplifier. A dummy load is incorporated to dissipate the entire output when testing the transmitter.

3.5 Power Supplies

The total power requirements of the transmitter are 5.5 kilovolt-amperes from a 230-volt, 50/60-cycle line. There are three power supplies for providing the plate voltage, screen-grid voltage, and bias required in the various stages.

3.6 Monitoring Provisions

The high order of pulse-shape stability required of loran transmissions necessitates complete provisions for monitoring. The monitoring oscilloscope, a Dumont type *256-D*, is placed in a compartment in the front of the transmitter cabinet (see Figure 2). The oscilloscope, which incorporates a delayed sweep, permits accurate measurement of the rise time and width of all important waveforms in both the transmitter and amplifier. Front-panel switches in the transmitter permit any of these pulses to be displayed on the oscilloscope.

4. *Power Amplifier*

The function of the power stage is to amplify the 160-kilowatt output of the transmitter to a level of approximately 1000 kilowatts peak power without distortion. An output of 800 kilowatts is available when double pulsing is used. The amplifier may be tuned to any of the 5 standard loran frequencies.

A photograph of the front of the amplifier cabinet is given in Figure 10, and a block diagram in Figure 11. As shown in the block diagram, the transmitter output is applied to the grid circuit of the push-pull amplifier stage. A trigger from the timer unit (or two triggers, where double pulsing is used) is applied to a trigger-generator stage, the output of which is fed to a bias-pedestal generator. The pedestal-generator output is a positive pulse that arrives at the amplifier grid circuit simultaneously with the radio-frequency pulse from the transmitter. The pedestal serves to adjust the bias on the amplifier to the correct value for linear operation, after which the amplifier is again cut off.

The output of this equipment is an amplified reproduction of the input pulse, and this may be applied to a dummy load, or through the antenna-matching unit to the station antenna.

4.1 Bias-Pedestal Generator

4.1.1 Linearity Considerations

During the early development of the power amplifier, it was operated strictly class-*B* push-pull, with no plate current flowing in the absence of radio-frequency driving power. Under these conditions, linearity of the input-to-output characteristics showed considerable curvature in the low-level region. As a result, the pulse shape from the transmitter was distorted in passing through the amplifier to the extent that the rise time (10- to 90-percent amplitude) changed from 21 microseconds to 17 microseconds and the width at 50-percent amplitude from 40 to 44 microseconds.

The effect of the curvature in the low-level region of the characteristic is recognized in any vacuum-tube amplifier operating strictly class *B*. Thus, in class-*B* telephony, the usual practice is to bias the tubes slightly into the conducting region to avoid the nonlinearity. In doing this, the operating static plate current is not increased by a very large percentage and the efficiency is not materially affected. In the loran amplifier, however, where there is a very small duty cycle (nominally 0.005), the average plate current is relatively small and if the tubes were to be biased above the nonlinear region, the static current would far exceed the current drawn as a result of pulsing and the efficiency would become prohibitively low.

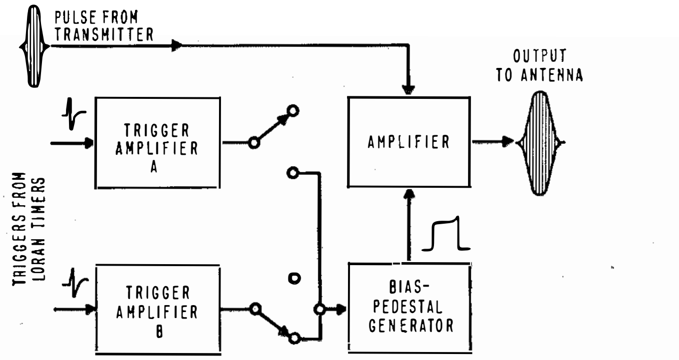


Figure 11—Block diagram of the power amplifier.

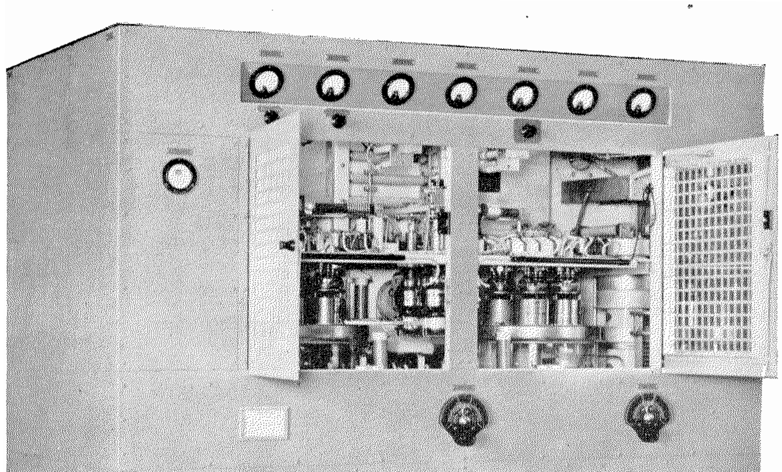


Figure 10—The power amplifier. The parallel–push-pull stage may be seen through the open ventilating doors.

4.1.2 Pedestal-Generator Circuit

The increased linearity obtained by operating the tubes in the conducting region was effected without any loss in efficiency by the use of a bias-pedestal generator. This voltage pedestal decreases the bias level on the amplifier grids for a short period starting just before a pulse is to be amplified and ending just after the pulse itself. This is illustrated graphically in Figure 12. In allowing the desired conditions to exist for only the duration of the pulse, the efficiency is not materially affected because the static current is not increased. With the bias pedestal operating,

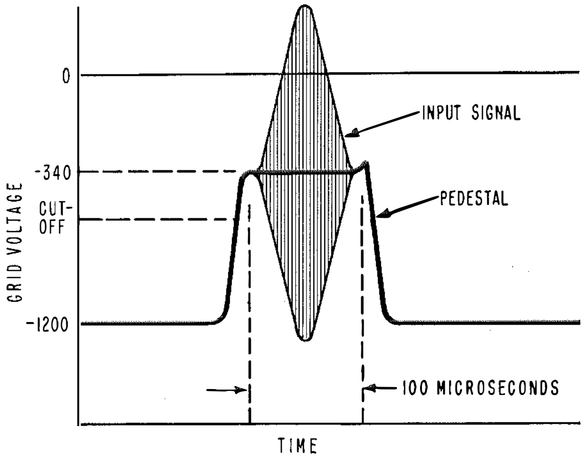


Figure 12—The bias on the amplifier grids is adjusted to the proper value for linear amplification for only the duration of the input signal. At other times the tubes are held below cutoff at −1200 volts.

the rise time and width of the amplifier output pulse are within 1 microsecond of the measurements at the input.

The circuit that generates the bias pedestal is shown in Figure 13. A type *3C45* hydrogen thyratron is used. With no pulse input, the amplifier grids are completely below cutoff, being at the −1200-volt level. Capacitor *C3* is charged through *R2* and *R3* to −1200 volts, and *C1* is charged to the potential difference between the plate and cathode of the thyratron, or 860 volts. When a repetition-rate trigger pulse arrives, the tube ionizes and immediately the potential to the amplifier grids rises to −340 volts as the thyratron becomes virtually a short circuit. Capacitor *C1* and inductor *L1* determine the length of the bias pedestal by putting an oscillatory current through the tube that cuts the tube off after one-half cycle when the current tries to change direction. These constants are chosen to produce a pedestal of about 100-microsecond length. Capacitor *C2* is large in value to carry the large video components of amplifier grid current with little drop in voltage during the pedestal length. A photograph of this pedestal waveform is given in Figure 14.

4.2 Amplifier Stage

4.2.1 Choice of Tubes

In selecting the tube complement for the 1-megawatt amplifier, the prime requisite was that it be capable of generating about 1.2 megawatts peak power in order to furnish a 1-megawatt output with allowance for losses in the output circuit and some safety factor. This peak power had to be realized within operating limits where the input-to-output characteristic of the stage was essentially linear. A push-pull circuit had been decided on to provide attenuation of even-order harmonics, thus setting a minimum of two tubes on the complement.

At the time the development started, there was no tube commercially available that was suitable for the pulse application and could produce the desired output with linear amplification in a complement of less than eighteen tubes. This quantity of the type *7C23* was the only

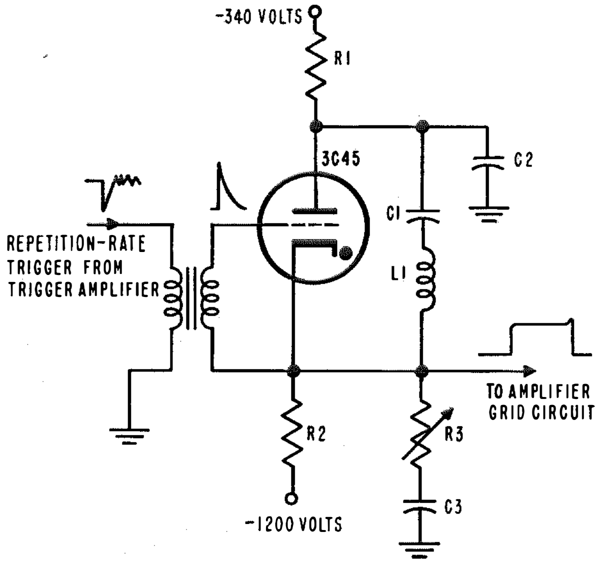


Figure 13—The bias-pedestal generator. Firing of the *3C45* hydrogen thyratron generates the pedestal for adjustment of the push-pull amplifier stage grid bias.

possibility. The total quantity of tubes was dictated by the peak emission necessary to produce the power output. The *7C23*, an external-anode forced-air-cooled triode, had proved itself for loran service, being used in the pulsed-oscillator transmitters that had been in field service for about 5 years. The *7C23* was also being used in the 160-kilowatt penultimate amplifier stage of the new transmitter. However, it was decided that it was inadvisable to use as many as 9 tubes on either side of the push-pull arrangement because of the space required and because of the relatively great electrical distances

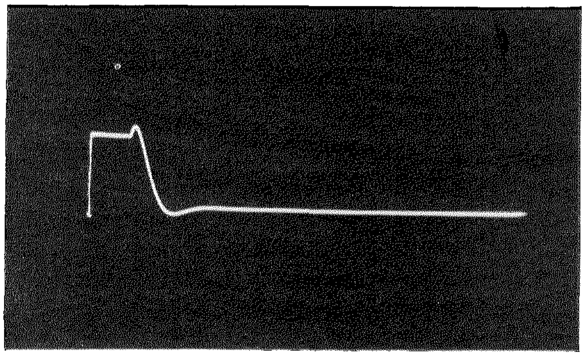


Figure 14—The bias pedestal as viewed on the monitor oscilloscope. The sweep length is 1000 microseconds.

that would separate certain of the tubes in the parallel groups; the latter might be conducive to parasitic-oscillation difficulties. A tube development was therefore undertaken to provide a suitable tube for the amplifier.

It was decided that the maximum number of tubes that would provide a practical arrangement was 12, and the initial question was whether to develop a tube that would provide the desired output with that number, or to develop a large type of which a pair would serve. Design considerations showed that enough additional filament power could be built into the *7C23* to give the required emission in a complement of 12 without any additional changes in the tube structure. This tube would be 6.5 inches high, 3.5 inches in diameter, and would weigh 5 pounds. On the other hand, if a larger tube were developed, of which a pair would furnish the desired output, a tube about 12 inches high and 9 inches in diameter, weighing 30 pounds would be required. Cost-wise, the complement of larger tubes would be about $1\frac{1}{2}$ times more expensive.

From the cost standpoint alone, the use of 12 redesigned *7C23* tubes was the logical choice. However, other factors also weighed heavily in their favor: For shipment to isolated parts of the world, under conditions where rough handling was almost inevitable, the smaller and hence sturdier structure was better suited. For the convenience of the service using the tubes, who were already accustomed to handling the *7C23*, a redesigned version offered no new installation problems as would a larger tube. Finally, preliminary design of the larger tube showed that forced-air cooling of the filament seals in addition to the anode would be required and this would add to the cost of the equipment. Thus, a redesign of the *7C23* was undertaken, and this was carried to a successful conclusion in the *5680* tube shown in Figure 15.

In the 1-megawatt amplifier, an arrangement was devised for mounting six *5680* tubes in parallel by modification of an insulated tube mount normally used with a single 50-kilowatt air-cooled tube. A new top casting was designed

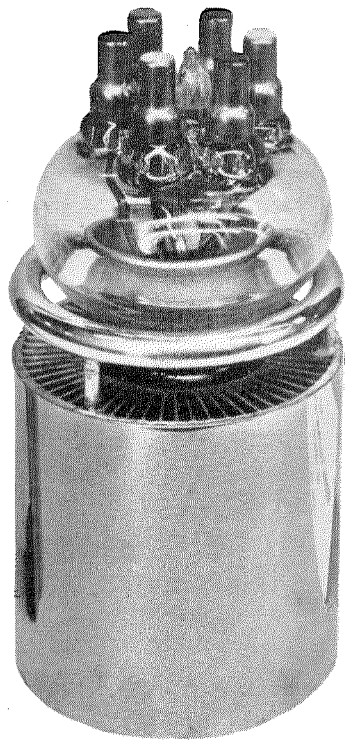


Figure 15—The *5680* tube is a redesigned version of the *7C23* affording larger peak-emission capabilities.

for the mount to hold the six tubes on the circumference of a circle as small as the tube dimensions permit (Figure 16). This connected the anodes in parallel through very-low-impedance paths, and little trouble was experienced

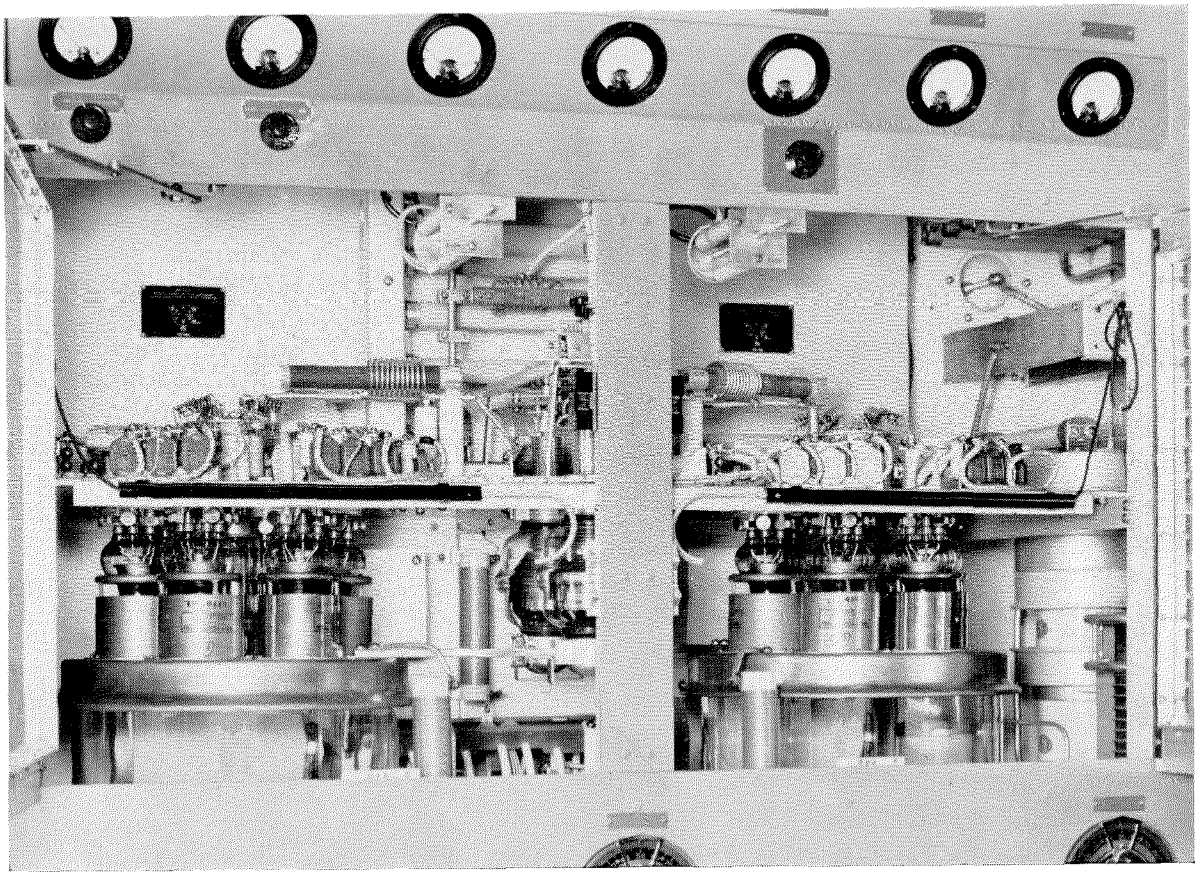


Figure 16—View through the front doors of the amplifier showing the 12 type *5680* tubes in parallel–push-pull.

in freeing the circuit of parasitic-oscillation difficulties.

4.2.2 Amplifier Grid Circuit

The 1-megawatt amplifier and the 160-kilowatt output stage of the transmitter are each designed to feed an antenna through a 52-ohm coaxial cable. Therefore, in order that the transmitter may be used to excite the 1-megawatt amplifier, it was necessary that the push-pull amplifier grid circuit be designed to match the 52-ohm single-ended output of the transmitter.

The normal method of coupling between a single-ended low-impedance circuit and the grids of a push-pull stage at frequencies of 2 megacycles is to employ a balanced tank circuit from grid to grid and to couple inductively the low-impedance source into the electrical center of the tank coil. This method has disadvantages in the loran amplifier. First, an operating Q of about 8 could not be exceeded in the tank circuit without distorting the pulse, and with this Q it was physically difficult to obtain enough inductive coupling between the tank and the input circuit to reflect an impedance to the input circuit as low as 52 ohms. This coupling also had to be variable to give correct operation at each of the 5 loran frequencies. Second, because the grid capacitors were too high in value to be variable, a balanced variable inductance would have to be used, and this, with its associated controls, would be expensive. And third, because the impedance reflected to the source would be a relatively critical function of the grid-tank tuning, it was anticipated that an inexperienced operator would have trouble adjusting it correctly.

The more effective solution to input coupling seems to lie in the technique normally applied in ultra-high-frequency equipment where a quarter-wavelength section of transmission line transforms from a single-ended to a balanced circuit. Such an assembly is normally referred to as a *bazooka* or *balun*, and properly designed and terminated it reflects a particular impedance. The basic circuit shown in Figure 17 is well known.

In the loran amplifier, the bazooka was employed essentially as described in Figure 17. The load at the balanced end consisted of the grids of the tubes with a large percentage of resistive loading to reduce the nonlinearity of the total load. The characteristic impedance of the quarter-wave transmission line was chosen as 72 ohms and the total grid-to-grid load was made 100 ohms, which transforms to the desired 52 ohms at the input. This transformation gives the necessary radio-frequency voltage level at the grids. The characteristics of the 72-ohm line are broad enough so that it could be cut for the geometric-mean frequency between 1750 and 1950 kilocycles and it then gave very satisfactory performance at all of the 5 frequencies employed.

Physically the bazooka is relatively simple, consisting of 90 feet of type *RG-11/U* solid-dielectric coaxial cable wound in pancake fashion on phenolic supports with the assembly mounted in an aluminum enclosure. It is shown in Figure 18. The input is on the left end where the cable shield is grounded. The output is at the lower middle where both the cable shield and inner conductor are brought out to separate insulated terminals.

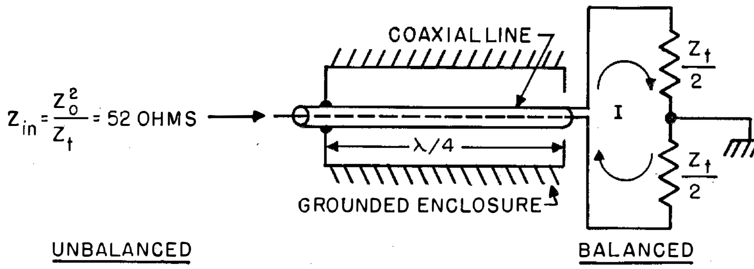


Figure 17—Basic circuit of balancing section. The term Z_0 is the characteristic impedance of the quarter-wavelength line; total load impedance is Z_t.

A simplified schematic of the input circuit of the amplifier including the bazooka is given in Figure 19. Inductances *L1* and *L2* tune out the capacitive reactances of blocking capacitors *C1* and *C2*. A tuned circuit (consisting of *C3*, *C4*, *C5*, and *L3*) is used from grid-to-grid of the amplifier tubes, which, in conjunction with the effective grid-to-grid resistance of 100 ohms, has a Q of about 5. This provides low-impedance paths for the grid-current harmonics. When the tank is at resonance and its impedance is high, the 52-ohm termination is still wholly established by the 100 ohms.

The inductance required in the tank circuit is low enough, due to the low impedance of the circuit, to permit the use of a parallel-rod line section with a short at the far end. The inductance is adjusted by moving the short, and positions have been calibrated for each of the 5 frequencies. The operator has only to position

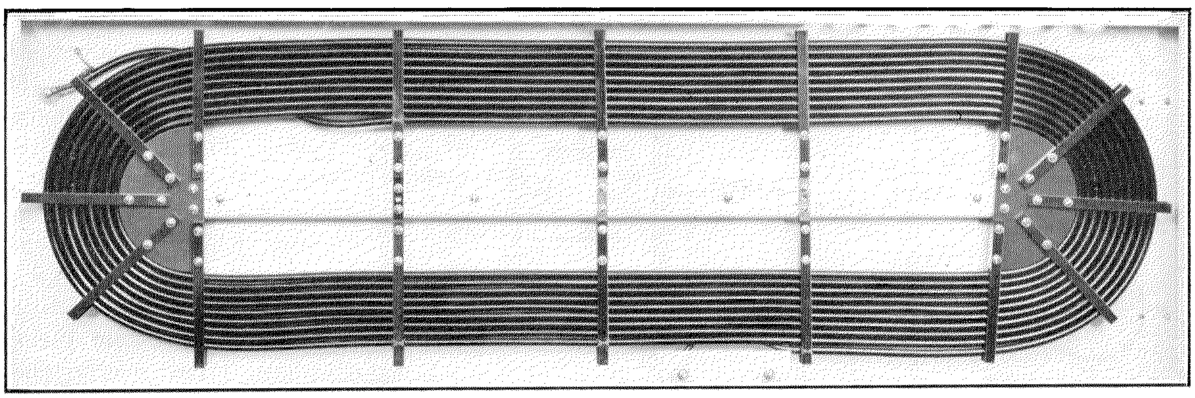


Figure 18—A bazooka is used to match the 52-ohm transmitter-output line to the grids of the amplifier tubes.

the short for a particular frequency and automatically the penultimate stage is presented with the proper load impedance of 52 ohms and the driving voltage relations are correct at the amplifier-tube grids.

4.2.3 *Amplifier Plate Circuit*

A simplified schematic diagram of the plate circuit is given in Figure 20. Plate voltage is fed to the parallel–push-pull *5680* tubes through inductances *L1* and *L2*. Variometer *L3* has a sufficient range of inductance to tune with capacitors *C1*, *C2*, *C3*, and *C4* to any of the 5 loran frequencies. The amplifier output is coupled to the 52-ohm coaxial line by two resonant circuits consisting of *L4–C5* and *L6–C6*. A photograph of the output waveform is given in Figure 21.

4.3 Additional Circuits

Provision is made in the amplifier to feed various waveforms back through a coaxial line for display on the oscilloscope mounted in the transmitter.

The total power requirements of the amplifier are 17 kilovolt-amperes from a 230-volt, 50/60-cycle line. There are two power supplies in the amplifier. One of these, the low-voltage rectifier, supplies bias voltage for the amplifier and bias and plate voltages for the trigger generators and the pedestal generator. The other provides 15,500 volts for the plates of the amplifiers.

5. Conclusion

The 1-megawatt loran equipment has now been in field service since January of 1950, and performance has fully justified the methods by which the required characteristics were met.

With respect to the transmitted pulse shape, the cosine-squared characteristic obtained by modulation in the penultimate amplifier has been illustrated in Figure 9. Comparison of this waveform with the pulse at the output of the amplifier, Figure 21, illustrates the small amount of distortion introduced by the amplifier. In the field, a loran chain is in operation wherein one master station does not employ a 1-megawatt amplifier but a mating slave station does. Pulse overlay between the two signals is completely satisfactory.

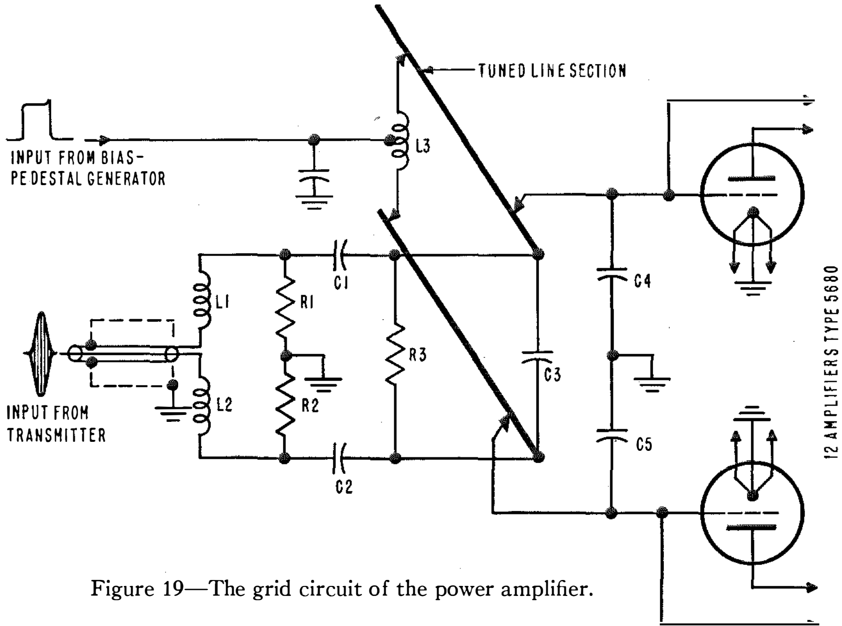


Figure 19—The grid circuit of the power amplifier.

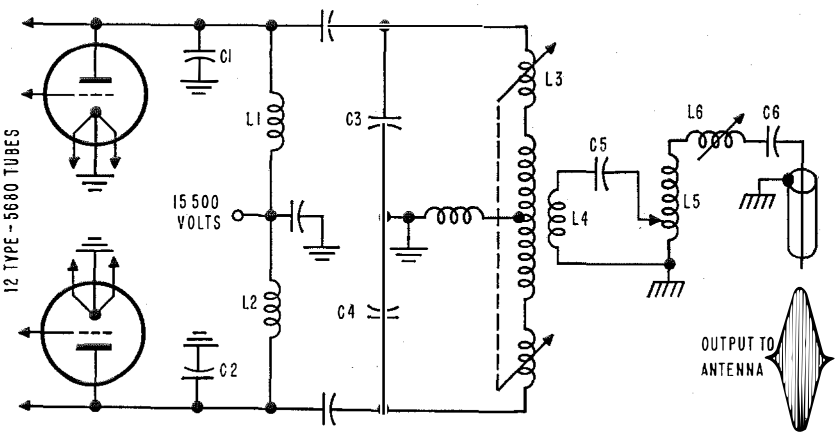


Figure 20—The plate circuit of the amplifier is shown above.

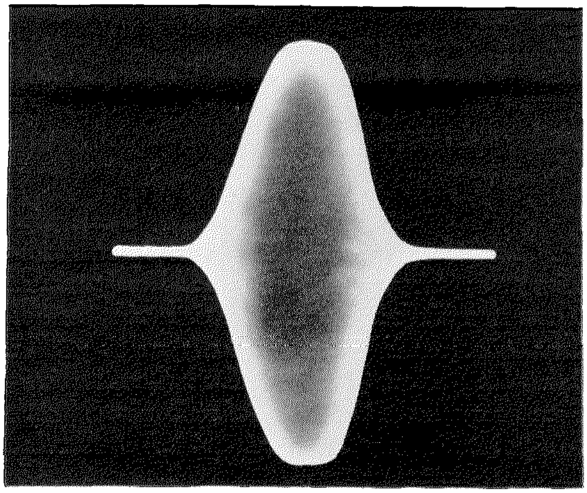


Figure 21—Output waveform of the power amplifier on a 100-microsecond sweep.

The measured frequency spectrum from the equipment was well within the original spectrum calculated from the mathematical cosine-squared function. The measured spectrum is compared with the maximum proposed limits in Figure 22. The older type of pulsed-oscillator transmitters are still used along the northeastern seaboard of North America, and comparison of these signals with those of the new equipment on a communications-type receiver or on a loran indicator readily shows the improvement in spectrum width.

6. *Acknowledgments*

The author wishes to acknowledge the early work done on controlling the spectrum of loran signals by the Naval Research Laboratory, Washington, D. C., as summarized in their report number *R-2781*. Assistance in the development of the subject equipment was given by Messrs. F. Davidoff and L. Everitt; and development of the type *5680* tube was done by the tube division of Federal Telephone and Radio Corporation.

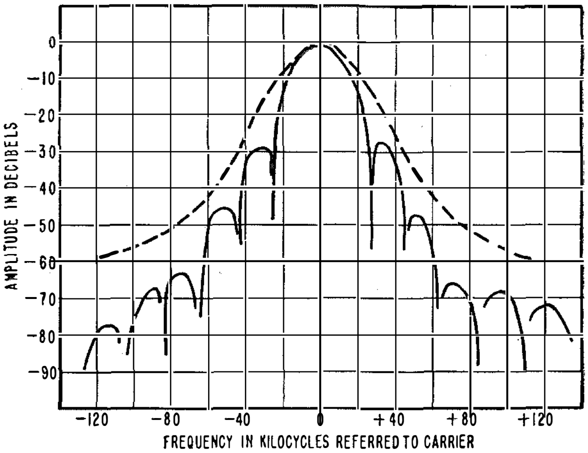


Figure 22—The measured spectrum of the transmitter is well within the limits set by the proposal (dashed line).

Recent Telecommunication Development

Etchings of Oliver Heaviside

OLIVER HEAVISIDE (1850–1925) is the subject of the latest etching in the series issued by the International Telecommunications Union.

Born in 1850 in London, England, the value of his theoretical work in the cable, telegraph, and telephone fields was obscured in part through his use of mathematical processes that were often original and not readily understood and accepted by practical readers. This resulted in some difficulty in getting his papers published.

In the field of radiant energy, he suggested that the earth was surrounded by a conducting layer that would reflect electromagnetic waves and permit signals to be transmitted over long distances despite the curvature of the earth.

This Heaviside portrait is the sixteenth in a series started in 1935. Each etching is on a good grade of paper and with margins measures 9 by $6\frac{5}{8}$ inches (23 by 17 centimeters). All of the etchings are available at 3 Swiss francs each from Secrétariat général de l'Union internationale des télécommunications, Palais Wilson, 52, rue des Pâquis, Genève, Suisse. The etchings are of Ampère, Baudot, Bell, Erlang, Faraday, Ferrié, Gauss and Weber, Heaviside, Hertz, Hughes, Marconi, Maxwell, Morse, Popov, Siemens, and Tesla.

Magnetostrictive Delay Line

By E. M. BRADBURD

Federal Telecommunications Laboratories, Incorporated; Nutley, New Jersey

MAGNETOSTRICTIVE delay lines permit pulse signals to be delayed over a smoothly adjustable range of time. They constitute passive systems that will accommodate a multiplicity of pulses simultaneously. A comparison is presented of various signal-delaying systems. The fundamental design equations for the magnetostrictive delay lines are given together with methods for eliminating the damaging effects that might be caused by echoes from the ends of the line.

• • •

1. *Need for Continuously Variable Delay of a Passive Nature*

In certain types of electronic equipment, there is a need for a device that can produce a continuously variable delay of pulse signals and, in addition, will be composed of passive elements yielding a delay independent of the past history of signals applied to the circuit.

A need for a device meeting requirements of this type is encountered in certain types of pulse decoding equipment used in a system of aerial navigational aids known as "Navar."[1] In this system, the altitude of a cooperating aircraft is transmitted to the ground station by the use of a group of pulses arranged in a code and transmitted by the aircraft in reply to a ground-radar challenging pulse.

The pulse group is decoded by delaying the incoming pulses in a network by definite and predetermined amounts depending on the anticipated information. These delayed signals are then compared to the undelayed signal by the use of coincidence-recognition and gating circuits. The magnitudes of the delays to be encountered vary continuously, depending on the height of the aircraft, and the maximum value of the delay is in the neighborhood of 150 microseconds.

The pulses entering the decoding equipment are in a sense random, since the information being relayed by a given aircraft and the position of the aircraft (and, hence, the time of return of their coded replies) is continuously varying in an indefinite pattern. The problem is further complicated by the fact that sets of pulses from many other aircraft may be present in the system at the same time.

For these reasons it is necessary to develop a method of delaying pulses that will meet the conditions outlined above and will be capable of accepting and accurately delaying one pulse even while other pulses are passing through other portions of the delay circuit. This condition is not met by most vacuum-tube delay circuits, for they cannot accept a second input pulse until the first pulse has finished its delay period. Passive delay lines meet this requirement and, where continuously adjustable delays exceeding 50 microseconds are desired, the magnetostrictive delay line seems especially attractive.

The magnetostrictive delay line is not a new device to electronic engineers. It was first introduced to us by Mr. Gordi of Watson Laboratories, Red Bank, New Jersey. Our additions to the development have been concerned with improving the pulse response and eliminating echoes from the system.

2. *Available Types of Delay Networks*

Most available systems of time delay are not capable of functioning properly with signals of the kind described because they suffer from some combination of the following defects.

A. The delay is not continuously variable.

B. The delay circuit has a long recovery time, i.e., the circuit has no storage capacity for a definite time after a pulse has entered the delay network.

C. The attenuation of the delay network varies widely with the delay.

D. The pulse distortion is severe due to the amplitude and phase characteristics of the system.

[1] H. Busignies, P. R. Adams, and R. I. Colin, "Aerial Navigation and Traffic Control with Navaglobe, Navar, Navaglide, and Navascreen," *Electrical Communication*, v. 23, pp. 113–143; June, 1946.

E. The circuits are quasi-active, i.e., the output pulse is triggered by a waveform produced by the input pulse, this waveform being generated by a resistance-capacitance or an inductance–resistance–capacitance network.

F. The insertion loss is high.

We list these defects in Table 1 to permit the available delay methods to be compared qualitatively.

When pulse reshaping is permitted to overcome the effects of phase and amplitude distortions, it can be seen that the magnetostrictive delay line is excellently adapted to the problem of long, continuously variable, passive delays.

3. Magnetostrictive Delay Line

Basically, the magnetostrictive line is a supersonic device, i.e., the pulse signal is conducted from input to output transducer in the form of a sound wave. The transducers described here make use of the magnetostrictive properties of some metals (such as nickel or monel) to produce this sound wave and to convert it back to an electric signal.

The magnetostrictive effect is described briefly as a change in length of a material when subjected to a magnetic field. The reciprocal effect is the change in flux in the material (if in a magnetic field) when subjected to a mechanical strain.

The transducers used here consist of coils through which the sound-conducting and magnetostrictive material is threaded. When a pulse of voltage is applied to the transmitting transducer, a magnetizing field is produced in this coil and creates a magnetic flux in the piece of magnetostrictive material within it. The material reacts to this flux by contracting or expanding. For nickel, the effect is a contraction. This mechanical action is transmitted as a sound wave in the material, which is presumably in the form of a ribbon. This wave is launched in two directions, as shown in Figure 1, one wave from each end of the coil. One of these waves can reach the receiver only after reflection from the end of the line. The latter is termed the indirect wave and the other, the direct wave.

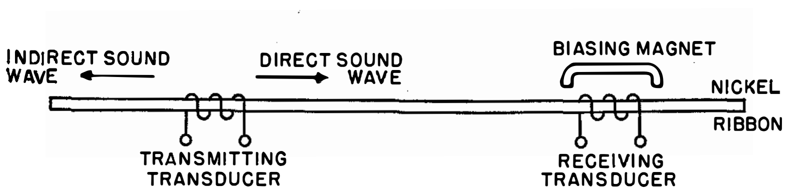


Figure 1—Magnetostrictive delay line showing general arrangement of transmitting and receiving transducers along the ribbon of magnetostrictive material, which may be nickel.

The direct wave propagates along the ribbon to the receiver coil. The nickel in this coil is subjected to a magnetic biasing field. The mechanical strain associated with the sound wave

TABLE 1

QUALITATIVE COMPARISON OF AVAILABLE PULSE-DELAY SYSTEMS

Type of Line	Delay Continuously Variable	Recovery Time	Attenuation as Function of Delay	Pulse Distortion	Amplitude Distortion	"Activity" of Circuit	Insertion Loss	Size of System
Wound Lines	No	Excellent	Poor	Excellent	Excellent	No	No	Large
Liquid Supersonic Lines with Carriers	Average	Excellent	Excellent	Excellent	Excellent	No	Poor	Excellent
Solid Supersonic Lines; Glass, Quartz	Average	Excellent	Excellent	Excellent	Excellent	No	Poor	Excellent
Metal Ribbons as Magnetostrictive Lines	Excellent	Excellent	Excellent	Average	Average	No	Poor	Excellent
Delay Multivibrator and Phantastron	Excellent	Poor	Excellent	—	—	Yes	Excellent	Excellent
Pulse-Excited and Resonant Circuit	Excellent	Poor	Excellent	—	—	No	Excellent	Excellent

reaches the receiver coil and causes a change in magnetic flux in the piece of ribbon within the coil. This change in flux is responsible for the output voltage of the receiver transducer. The velocity of sound propagation for a longitudinal wave along a ribbon can be shown to be

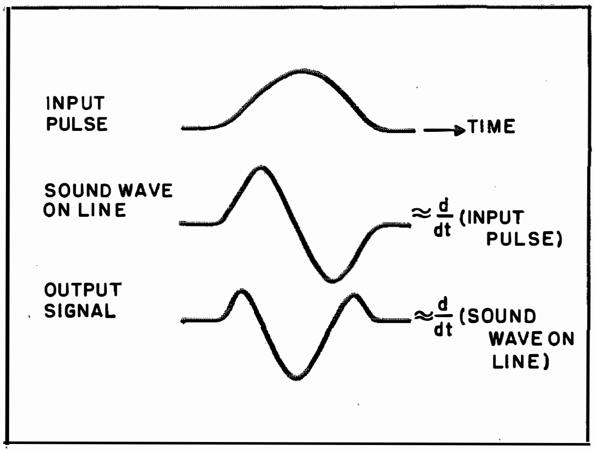


Figure 2—Pulse signals in the system. The output pulse approximates the second derivative of the input pulse.

$$v = (E/\rho)^{\frac{1}{2}}, \tag{1}$$

where

E = Young's modulus for the material of the ribbon

ρ = density of the material.

The time required for a pulse to be propagated along the ribbon is

$$\left.\begin{aligned} t &= L/v \\ &= L/(E/\rho)^{\frac{1}{2}}, \end{aligned}\right\} \tag{2}$$

where

t = delay time

L = distance between input and output transducers.

Thus, for 99.9-percent-pure-nickel ribbon and for a temperature of 60 degree centigrade, v is calculated to be 1.58×10^4 feet per second or a time delay of 5.27 microseconds per inch of path length. From this figure, it is evident that sizable delays can be contained in reasonable lengths of ribbon. For instance, one foot of ribbon yields 63 microseconds delay.

A comparable wound line (such as the commercially available *RG/62U*) would be about 100 feet long and have approximately 100 decibels of attenuation. A supersonic line such as described here will have 35 decibels of attenuation, which is fixed and will not vary appreciably with delay.

4. *Nature of Pulse Response*

The fact that the receiving coil in the magnetostrictive line is sensitive to variations in strain in the coil aperture, means that the output voltage is essentially the derivative of the sound wave in the ribbon, i.e.,

$$e = K\frac{d}{dt}\left[f\frac{\Delta L}{L}\left(t - \frac{L}{v}\right)\right], \tag{3}$$

where the expression in square brackets is a wave strain function.

In addition, it can be shown that the transmitting transducer behaves in the same manner, i.e., produces a sound-wave response that is approximately the derivative of the input signal.

The result is that the output pulse approximates the second derivative of the input pulse. This is shown diagrammatically in Figure 2. This response, while showing considerable phase and amplitude distortion as mentioned above, is

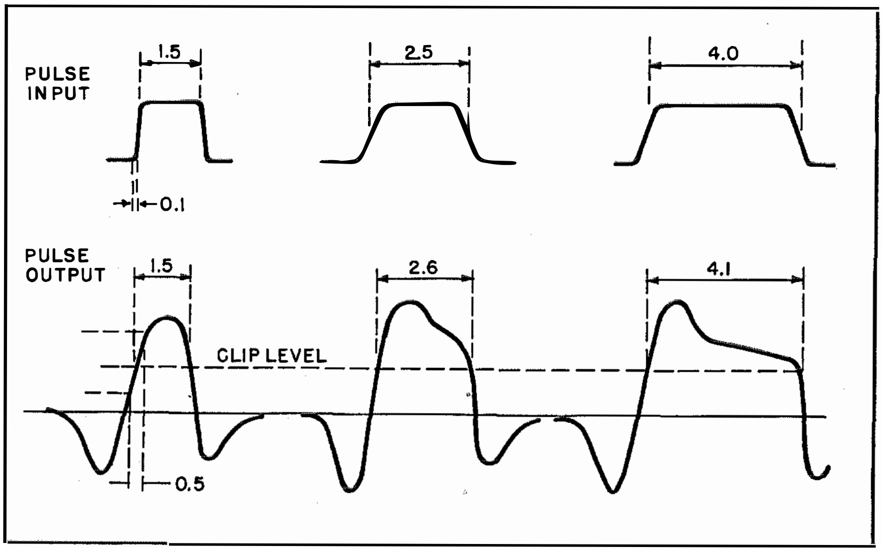


Figure 3—Pulse response. The numerical time values are all in microseconds.

useful in many applications because of the possibility of using reshaping circuits.

Oscillograms of pulse input and output shapes are shown in Figure 3 for input pulses of 1.5-, 2.5- and 4-microseconds duration. It is evident that the poor low-frequency response of the system restricts its use to narrow pulses of relatively constant duration. However, in the Navar application, it was quite feasible to use the line for delaying pulses varying between 1.5 and 4 microseconds in duration.

An analysis[2] of the magnetostrictive transducer gives an equivalent circuit for the transmitting and receiving portions as indicated in Figure 4.

It is interesting to note that the mechanical load on the transducer, namely, the ratio of the velocity of the wave in the ribbon at the end of the transducer to the force it produces on the load, is in this case the mechanical characteristic impedance of the ribbon to sound conduction, i.e., the mechanical Z_0 of the ribbon. But the antiresonant element of the drive circuit as shown in Figure 4 has the same electrical characteristic impedance. The equivalent electrical impedance of this resonant section is given in electrical units by

$$Z = \frac{j}{Z_0} \tan \frac{\omega l}{v} \tag{4}$$

where

Z_0 = mechanical characteristic impedance of line

$\omega = 2\pi f$ of the signal

l = mechanical length of transducer

v = velocity of sound in transducer (equal to that in load).

(We are considering the electromechanical analogy where force is equivalent to current and velocity to voltage.) Therefore, the equivalent electrical Z_0 of the ribbon is also expressed as $1/Z_0$ electrical ohms. This means that the Q of the combined antiresonant section plus the mechanical load due to the ribbon has a value close to one.

However, it is found in practice that the inductance Z_L (as shown in Figure 4) combined with the effects of the parasitic load capacitances, is an important factor in determining the bandwidth of the line.

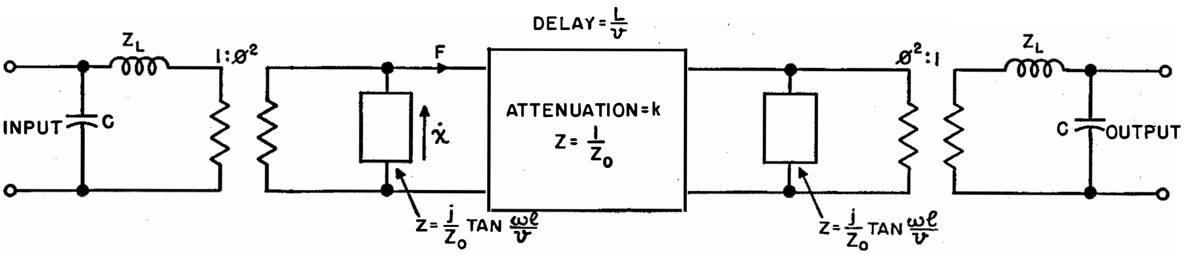


Figure 4—Electromechanical equivalent circuit of a magnetostrictive delay line. In the equations, a=area of the ribbon, C=parasitic input and load capacitance, E=Young's modulus for the material of the ribbon, $Z_0 = a/(\rho E)^{\frac{1}{2}}$, Z_L=clamped inductance of the transducer, ϕ=electromechanical coupling, and ρ=density of the material of the ribbon.

The over-all equivalent circuit of the delay line consists of two such units back-to-back and separated by an attenuating network having a characteristic impedance equal to the reciprocal of the mechanical impedance of the ribbon. This is also shown in Figure 4.

From the equations for the resonant frequency of the shunt element (which is just the mechanical resonant frequency of the transducer element), we note that as the length of the ribbon in the transducer is increased, the low-frequency response will be improved at the expense of the high-frequency response. It is thus possible to stagger the responses of the input and output transducers and get an over-all bandwidth that is noticeably larger (approximately 1.2:1) than that of either network alone. Of course, the amplitude of the response is deteriorated in about the same ratio.

Inasmuch as the Q of the antiresonant element of the network is not dependent on the dimensions of the ribbon (if we assume complete flux penetration of the nickel) the only other way in which the bandwidth can be improved is by affecting the value of the transducer inductance and stray capacitance. The amplitude of the response will also be improved if the magnitude of the inductance Z_L is decreased provided the

[2] W. P. Mason, "Electromechanical Transducers and Wave Filters," D. Van Nostrand Company, New York, New York; 1942: p. 215.

value of ϕ is kept low (large electromechanical coupling). This means that the value of stray magnetic field must be as low as possible and, in addition, we must use an inductance as small as is consistent with good electromechanical coupling.

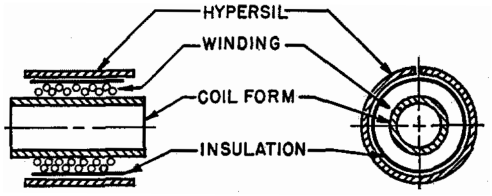


Figure 5—Constructional details of transducer coils.

This is accomplished by using a minimum number of turns and a magnetic shield around the transducers to confine the flux to the nickel path. This is illustrated in Figure 5 showing the construction of the coil.

In addition, the bandwidth is improved when the network is driven by a source of low impedance, since in this way the shunting effects of the parasitic input and output capacitances are minimized.

5. *Echo Suppression*

The transmitting transducer emits a sound wave in two directions, one being called the direct and the other the indirect wave, as mentioned above. The echo of the indirect and direct wave, each from its end of the ribbon, respectively, may or may not be a useful signal depending on the particular application. It will be assumed that these are not useful and that it is desired to suppress all echoes. Some techniques for the elimination of or suppression of the effect of these echoes are described below.

First, it is evident that the self-damping in the ribbon is quite small. To attenuate the echo by 20 decibels, compared to the direct signal, approximately 2000 microseconds of delay is necessary.

To increase the attenuation to echoes only, damping pads are brought to bear against the ends of these ribbons. These pads are made of neoprene and are held in place by a clamp system. The attenuation of such a pad 1.25-inches (3-centimeters) long is approximately 7 decibels. Thus, the echo of a sound wave will be 14 decibels below the direct signal inasmuch as the echo must pass through the damping cell twice if it is located at the end of the line.

However, even echoes of this level can be objectionable. For instance, if a signal contains many pulses, each having a definite and important time relation with respect to the others, the echoes will tend to disturb this time relation and, in addition, cause a level of confusion

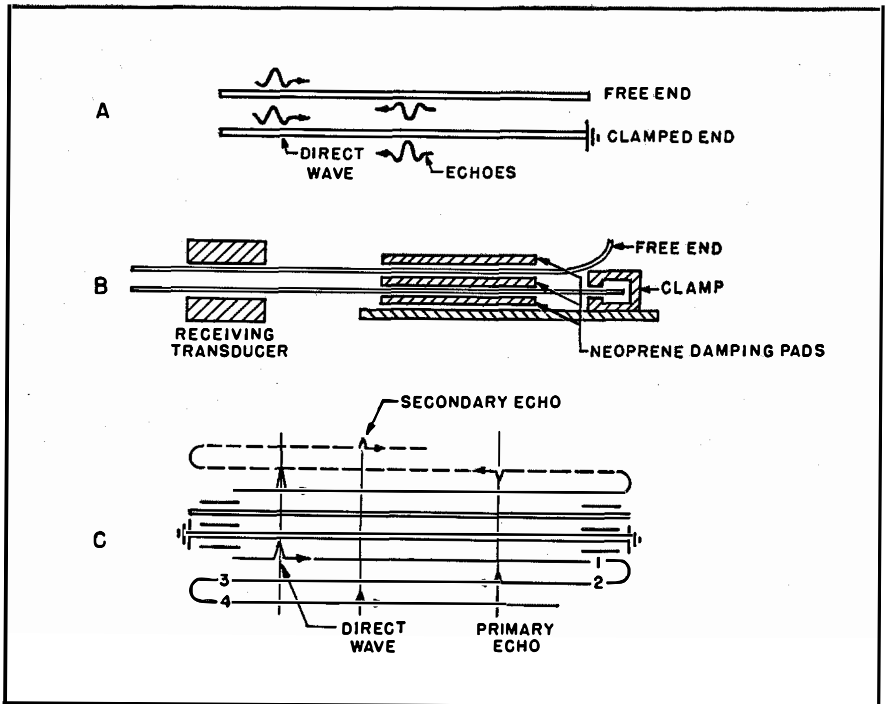


Figure 6—Echo-suppression system. The echo from the free end of the ribbon in *A* is of opposite phase to the direct wave and the echo from the clamped end is in phase with the direct wave. The effects of the echoes thus cancel. *B* shows the mechanical structure of a 2-ribbon line and *C* shows the polarities of the primary and secondary echoes, the latter two being in phase with the direct wave.

similar to noise on the line. To reduce the effect of the echoes even further, the systems described below can be used.

Basically, one method attempts to cancel echoes against one another. We note that if a ribbon conducting a sound wave is clamped at its end in a relatively large elastic mass, the echo returning from this end is in phase with the direct pulse on the line. On the other hand, if the end of the ribbon is left free, the echo will be found to be of opposite phase to that of the direct signal. This is shown in Figure 6A. If, as in 6B, two identical ribbons are used in parallel, the distance between the receiving transducer and the ends of the ribbons is identical, and the end of one ribbon is clamped while the corresponding end of the other ribbon is left free, the two echoes, one on each ribbon, exactly cancel each other insofar as their effect on the receiving transducer is concerned.

This means that the effect of the primary echo returning from the end of the line can be completely suppressed. However, the echoes are still present on the lines; they merely cannot produce an output for this first reflection.

In cases where it is desirable to suppress the indirect wave, it does not matter which end of which ribbon is clamped or free. All that will happen is that the second-order echo (that resulting from a second reflection) will be either positive or negative with respect to the direct signal. This is easily shown to be the case by referring to Figure 6C. Mechanically, it is best to have a given ribbon of the pairs arranged with both ends clamped or free to prevent wandering of the ends of the two ribbons with respect to each other in an echo-suppression unit.

The phases of the echoes from the first reflection are of opposite polarity on corresponding pairs of ribbons. Considering the echo of the direct wave only, and confining ourselves for the moment to the sequence of events at the transmitting end of the line, we find that if we again clamp one of the ribbons and allow the other to be free, the phase of one of the echoes is again reversed. The second-order echoes for this arrangement must, therefore, be in phase and, hence, will produce an output in the receiver coil. If a damping pad is arranged at either end of the line immediately preceding these echo-cancellation terminations, the echoes will have passed through the damping pads a total of 4 times before they can produce an output in the receiver coil. In this way, an echo reduction of 28 decibels on the line is obtained.

If we do not treat both ends of the delay line similarly with respect to this echo-cancellation scheme, then a first-order echo or indirect wave will always be present in this system. The mechanical arrangements of supports and damping pads are shown for a typical line in Figure 7. Of course, this method can be extended to any even number of ribbons in a line system.

A second method of echo reduction is accomplished by the use of noncoherent reflection. Many ribbons are used in this case, each threading through the transmitting and receiving transducers as for the two-ribbon case described above. Here, however, the lengths of the ribbons are randomly selected, and the resultant group

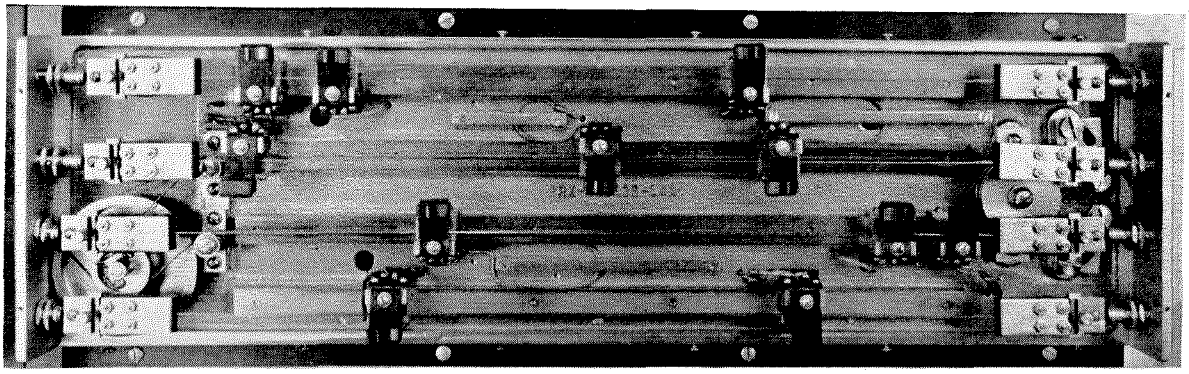


Figure 7—A typical magnetostrictive delay line. The pulleys and cables permit the positions of three receiving coils to be varied with respect to the transmitting coils and thus adjust the delay time.

of echoes from the open ends are not coherent. As a result they produce no appreciable output in the receiver coil. A damping pad is still used to attenuate these echoes as much as possible to prevent the formation of a high noise level on the line.

An additional advantage resulting from the use of many ribbons is the decrease on the insertion loss of the network. The parameter ϕ is decreased thereby giving a larger coefficient of electromechanical coupling. This is also true of the preceding method of echo suppression.

Hence, it can be said that

$$\frac{\partial t}{\partial T} \approx -\frac{L}{2vE}\frac{\partial E}{\partial T}. \tag{6}$$

This is approximately 5 parts per 10^4 per degree centigrade per microsecond delay.

For this reason, it is necessary to operate the lines in an oven that is temperature controlled if precise values of delay are to be obtained. For a 100-microsecond line, however, with a 20-degree centigrade temperature variation, only

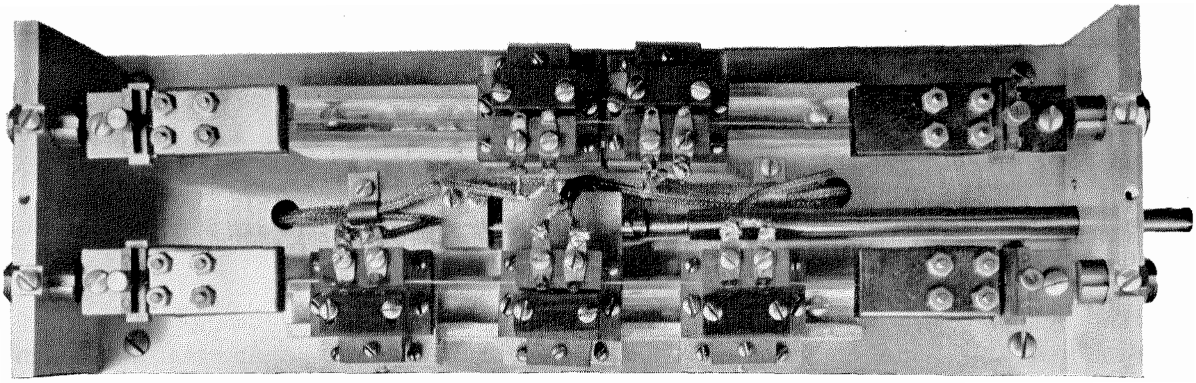


Figure 8—In this line, the sum of the delays of two pulses is a constant, the position of the transmitting transducer being adjustable between the two receiving coils.

6. *Variability of Delay with Temperature*

From (2), the variation of delay with temperature is given by

$$\left.\begin{aligned} \frac{\partial t}{\partial T} &= \frac{1}{v}\frac{\partial L}{\partial T} + \frac{1}{2}\frac{L}{\rho v}\frac{\partial \rho}{\partial T} - \frac{1}{2}\frac{L}{Ev}\frac{\partial E}{\partial T} \\ &= \frac{L}{v}\left[\frac{\partial L_0}{\partial T} + \frac{1}{2\rho}\frac{\partial \rho}{\partial T} - \frac{1}{2E}\frac{\partial E}{\partial T}\right], \end{aligned}\right\} \tag{5}$$

where

T = temperature
L_0 = reference length
ρ_0 = reference density.

But

$$\frac{\partial \rho}{\partial T} = -3\rho_0\frac{\partial L_0}{\partial T}$$

$$\therefore \frac{\partial t}{\partial T} = -\frac{L}{2v}\left[\frac{\partial L_0}{\partial T} + \frac{1}{E}\frac{\partial E}{\partial T}\right].$$

For nickel

$$\frac{\partial L_0}{\partial T} \approx 10^{-5} \text{ per degree centigrade}$$

$$\frac{1}{E}\frac{\partial E}{\partial T} \approx 10^{-3} \text{ per degree centigrade.}$$

about ½-microsecond variation in delay will be observed. For some applications, this is not considered serious.

7. *Description of Practical Delay Line*

Three versions of a practical delay line are shown in Figures 7, 8, and 9. Two of these lines are arranged for a variable delay. In Figure 7, three coils are varied in position with respect to transmitting coils by use of a pulley-and-cable arrangement. A definite ratio of the delays for

TABLE 2
CHARACTERISTICS OF A TYPICAL NICKEL-RIBBON DELAY LINE

Delay	90 microseconds
Insertion Loss	35 decibels
Bandwidth	300 kilocycles
Center Frequency	200 kilocycles
Ribbons	
Material	99-percent nickel
Number	2
Size	0.008 by 0.020 inch (0.2 by 0.5 millimeter)
Input Impedance	500 ohms
Output Impedance	50 ohms

two coils is obtained by arranging a definite ratio between the diameters of two pulleys and driving the pulleys from a common shaft. Each coil is driven by the appropriate cable.

Both transducers were wound with size 40 (American Wire Gage) enamelled copper wire to an inner diameter of 1/16 inch (1.6 millimeters). The transmitting coil had 200 turns in a form 3/16-inch (4.8-millimeters) long and the receiving coil had 100 turns in a length of 1/8 inch (3.2 millimeters).

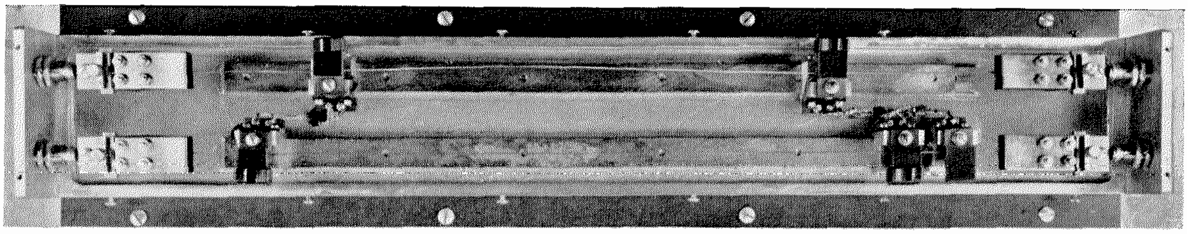


Figure 9—A fixed-delay line.

In another of these lines, shown in Figure 8, the delay between two pulses is varied in such a way that the mean value of the delay for both pulses is a constant. This is done by varying the position of the transmitting transducer, which is located between the two receiving transducers. A fixed line is illustrated in Figure 9.

A typical design using nickel ribbons had the characteristics and performance data given in Table 2.

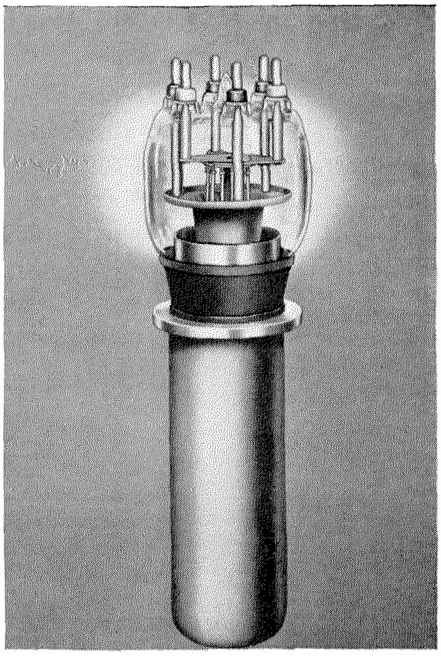
8. Acknowledgments

Without the assistance and encouragement of Messrs N. H. Young, I. Krause, R. Alter, and N. Weintraub, this work would not have been possible.

Recent Telecommunication Development

200-Kilowatt Triode

THE NEED for increased transmitting power, particularly for international high-frequency broadcasting, prompted the development of a new triode designated the *F-5918*. It is suitable for use as a radio-frequency amplifier, oscillator, and class-*B* modulator, operating at maximum ratings at frequencies as high as 22 megacycles per second.



A multistrand filament of thoriated tungsten produces adequate emission with a power of 4275 watts; 150 amperes at 28.5 volts. The several hairpin filaments can expand individually, thus avoiding stresses that are conducive to filament warping.

The anode is of high-conductivity copper and is water cooled. It is capable of dissipating 60 kilowatts in continuous commercial service. Power outputs of the order of 200 kilowatts may be obtained in telegraph operation with driving powers of less than 4 kilowatts.

The tube was designed and is manufactured by Federal Telephone and Radio Corporation. Its first commercial application was to increase to 200 kilowatts each the power outputs of three of the *WLW* "Voice of America" high-frequency transmitters.

Note on the Measurement of Impedance with the Impedometer*

By W. SICHAK

Federal Telecommunication Laboratories, Incorporated; Nutley, New Jersey

OBTAINING an impedance from measurements made with the impedometer[1] is simplified by using the Smith chart. The impedometer gives directly the reflection coefficient and the voltage at a point on the line. Draw a circle with radius equal to the reflection coefficient around the center of the chart. Draw another circle with the radius equal to the probe voltage, using the $R/Z_0 = 0$ and $X/Z_0 = 0$ point as the center. One of the intersections is the impedance at the probe. If the reflection coefficient decreases (or probe voltage increases) when shunt capacitance is added at the probe, the impedance is inductive. If the reflection coefficient increases (or probe voltage decreases), the impedance is capacitive.

The basis for this construction is illustrated in Figure 1, which shows a Smith chart (the R and X coordinates are omitted for clarity). The incident wave is a vector, with length equal to the radius of the chart, lying along the real axis. The reflected wave, of length K, is a vector whose phase depends on the position along the transmission line. The voltage V at any point on the line is the vector sum of the incident and reflected waves.

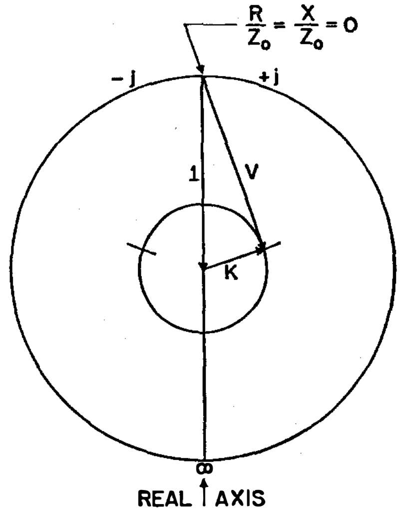


Figure 1.

This method is useful when the Smith chart is used either to present impedance data or as an aid in designing matching networks.

To avoid calculation, a scale can be constructed and used repeatedly. The scale can be marked in voltage, power, or decibels. A distance equal to the radius of the Smith chart corresponds to a reflection coefficient of unity (0 decibel), a distance equal to half the radius corresponds to a reflection coefficient of $\frac{1}{2}$ (-6 decibels), etc. The length of the scale is twice the radius, so that the same scale can be used for the probe voltage. A Smith chart with such a scale printed in a convenient position is available commercially.[2]

* Reprinted from *Proceedings of the I. R. E.*, v. 38, p. 951; August, 1950.

[1] B. Parzen "Impedance Measurements with Directional Couplers and Supplementary Voltage Probe," *Electrical Communication*, v. 26, pp. 338–343; December, 1949: *Proceedings of the I. R. E.*, v. 37, pp. 1208–1211; October, 1949.

[2] The Emeloid Company, Hillside 5, New Jersey.

Some Properties and Tests of Magnetic Powders and Powder Cores*

By C. E. RICHARDS,
Post Office Engineering Research Station; London, England

S. E. BUCKLEY,
Standard Telephones and Cables, Limited; London, England

P. R. BARDELL,
General Electric Company, Limited; Wembley, England

and

A. C. LYNCH
Post Office Engineering Research Station; London, England

POWDERS of iron or iron-nickel alloys can be made either mechanically or chemically. The particles are then insulated and pressed into cores; high-permeability cores must be annealed after pressing. The permeability of the core depends mainly on the effectiveness of packing of the particles.

Various systems of expressing the electrical losses are in use and conversion factors are given. In laboratory work, bridges and a Q-meter technique can be used to measure the permeability and losses; for control of core production, a permeameter is useful as well as the bridges.

Data are given for various materials, both commercial and experimental, including some ferrites; and it is shown that, in typical inductors, the losses in the windings may exceed the losses in the cores.

• • •

1. List of Symbols

h = Legg's coefficient for hysteresis
c = Legg's coefficient for residual loss
e = Legg's coefficient for eddy-current loss
e' = a similar coefficient for residual loss proportional to f^2
e_t = a similar coefficient for total eddy-current loss of coil and core
d = particle diameter
f = frequency
$\hat{B}$ = peak flux density
C = self-capacitance
G = conductance associated with the self-capacitance
L = total inductance
L_m = inductance due to core
ΔL = increase of inductance due to hysteresis
R = total resistance
R_m = component of resistance of wound core, due to losses in core
ΔR = change of R_m due to hysteresis
γ = (volume of particles)/(total volume)
μ = permeability
ρ = resistivity of particles.

(Other symbols are introduced and defined in Table 1.)

2. Introduction

Powder cores† for transformers and inductors are not a new development. They were first mentioned in the literature by Heaviside in 1886, though they did not come into general use in telephone transmission until about 1915, when iron-powder cores were introduced in the United States to replace wire cores. Iron powder gave way to the lower-loss nickel-iron in 1925. Since then changes have been in detail, until the last year or two when the ferrite-type cores were introduced.

Although some of the earliest materials are now seldom used, the others all have their own particular fields, and to-day the selection of iron powder, nickel-iron powder, or ferrite, depends on the exact performance required from the finished coil.

* Reprinted from *Proceedings of The Institution of Electrical Engineers*, Part II, v. 97, pp. 236–245; April, 1950. Presented at a Symposium on Ferromagnetic Materials, London, England, on November 7, 1949.

This is an "integrating" paper. Members are invited to submit papers in this category, giving the full perspective of the developments leading to the present practice in a particular part of one of the branches of electrical science.

† The term "powder core" is used in this paper in place of the more usual "dust core." Some of the powders described are too coarse to be reasonably given the name of "dust."

In telephone transmission systems, powder cores are used to reduce attenuation in cable circuits by adding inductance (loading coils), and to select appropriate frequencies in multi-channel carrier-frequency systems (filter coils). To a limited extent they are used for transformer cores for high frequencies. Loading-coil cores, particularly those for audio-frequency circuits, are generally of relatively coarse powder of the high-permeability type. This results in eddy-current losses which can, however, be tolerated, and small size, so that a large number of coils can be accommodated in a small manhole. Hysteresis is important as it causes harmonic distortion. Filter coils must normally have low losses, and, although hysteresis is important, other losses cannot be neglected and finer powders are more usual. Ferrites have recently been used for this purpose on the Continent.

In radiocommunication powder cores are used in filter circuits and perhaps for transformer cores; also they are becoming popular for permeability tuning.

3. *Manufacture of Materials*

For the purpose of this paper magnetic powders will be considered to consist of all those materials that can readily be produced in the form of a fine powder and which, when in that form, have "soft" magnetic properties. These materials include iron, certain iron alloys, and some non-metallic ferromagnetic substances such as the ferrites, which are salts of the hypothetical ferric acid $HFeO_2$.

There are three commerically important ways of producing metallic magnetic powders:

A. Mechanical disintegration of the massive metal in various kinds of attrition mill.

B. The production of the metal, either directly in the powder form or as a sinter cake, by chemical reduction of oxides or other suitable compounds, followed when necessary by mechanical disintegration.

C. The spontaneous decomposition of certain iron compounds or mixtures of them with nickel compounds, during which the metal is deposited as a fine powder.

Other methods which may be used, or are in the process of development, are:—

D. Electro-deposition.

E. Atomization of the molten metal.

3.1 Mechanical Disintegration

There are many variants of this process, and only those known to have been used for making magnetic powders will be mentioned.

The Hametag (eddy mill) process was evolved and operated in Germany. The magnetic material is fed into the mill in the form of wire which is chopped into short lengths as it enters. In the mill, the particles are whirled against one another until they reduce themselves to a fine powder, and are swept out of the chamber by a current of inert or reducing gas which flows throught it. The powdered metal may be fractionated by sifting, winnowing, or elutriation.

3.2 Ball–Milling

This is operated on a large scale in this country for alloys rather than for iron. It is usual to make the alloys by melting together the constituents in a high-frequency furnace, and to cast ingots. To assist the subsequent pulverizing process, some embrittling agent such as sulphur may be added, which migrates to the grain boundaries and leaves the metal in a brittle state. The ingots are hot-rolled to a convenient thickness, heat-treated to bring the metal into the brittle condition, and pulverized in a series of attrition mills. For example, large pieces may be broken down in a swing-hammer mill until they are small enough to be handled by a large ball-mill. The final powdering may be done in a ball-mill swept by air or inert gas to remove the finer particles and keep the mill cool. The powder is screened, the smallest particles being collected from the air-stream by fabric filters. In certain cases the powder may be annealed and repulverized to break down any consequent slight sintering.

3.3 Reduced Oxides

Both iron and alloy powders may be made by the chemical reduction of oxides. The process for both types is basically the same, though alloys may need several additional steps. The raw material for iron powder is iron oxide in one of its purer forms, e.g., high-grade natural red oxide, clean mill-scale, or chemically prepared oxide. This substance, powdered if necessary, is exposed to the action of a reducing atmosphere (generally hydrogen or cracked ammonia) at a moderately low temperature. The resulting product is a bulky sponge-iron powder which has for some purposes very desirable characteristics. For core-making in general it is, however, usually too bulky to be handled conveniently in the tools, and it absorbs more binder than is desirable. It must, therefore, be densified. This may be achieved to some extent by milling, but the process has limitations and a better way is to sinter in hydrogen. The resulting dense sinter-cake is broken down in ore-crushers and finally reduced to powder in a ball-mill.

When an alloy powder is required the oxides of the constituent metals are mixed before the first reduction. If this reduction takes place at a low temperature a mixture of metal powders will result; the subsequent sintering process both converts this mixture by diffusion into an alloy, and imparts the necessary density.

3.4 Spontaneous Decomposition of Iron or Nickel Compounds

The only commercial example of this is in the carbonyl process which has been used since the end of the 19th century to manufacture nickel, and which can be modified to handle iron. It gives a finely divided powder. The raw materials for this process are coke, steam, and iron ore or scrap-iron. Coke and steam are used to produce water-gas which is then separated into its constituents, carbon monoxide and hydrogen. Iron ore, after a preliminary purifying and crushing operation, is reduced in a current of hydrogen to iron sponge; this sponge, after cooling somewhat, is subjected to the action of carbon monoxide under pressure. Direct combination takes place and the iron is converted into iron pentacarbonyl, which condenses to a liquid at temperatures below about 100 degrees centigrade. The liquid iron carbonyl is atomized into a hot vessel known as a decomposer, where it breaks down into its constituents, iron and carbon monoxide. The gas is pumped away for re-use, while the iron powder falls to the bottom of the decomposer and is periodically removed.

The whole process needs careful control if a uniform useful product is to be made, as the decomposition of iron pentacarbonyl may be associated with an exothermic secondary reaction which, unless controlled, will lead to rapid heating of the decomposer with consequent variation in the quality of the product.

Iron powder made in this way consists of spherical particles having an average diameter of about 5 microns. It is mechanically very hard, the estimated hardness being over 850 V.P.N.* It requires no further treatment beyond a light milling to break up clusters, and sifting to remove large or foreign bodies. The metal contains about 0·8 percent carbon and about 0·25 percent nitrogen, and has a characteristic "onion skin" structure of concentric shells.

A soft, pure-iron powder can be made from the hard one by heat-treating in a current of hydrogen at a temperature just below the sintering point. This treatment also removes the nitrogen, breaks down the onion-skin structure, and produces spheres of homogeneous solid iron.

Using the carbonyl process, it is possible to mix nickel and iron carbonyls and so obtain a mixed metal from the decomposer. It seems certain that the metal as deposited is actually a mixture of iron and nickel which must be heat-treated to give a nickel-iron alloy; the alloying need not change the spherical shape of the particles.[1]

3.5 Degussa Process

This process, developed in Germany, is used in this country mainly to produce non-ferrous metal powders. It has been used experimentally for producing magnetic iron and iron-alloy powders, so far without full success, but it is a method which should be kept in mind. The raw material is melted and heated until very fluid, usually in an induction furnace, and is poured through a vortex of water on to a fast-moving steel rotor. As the molten metal strikes the blades, it is broken into minute globules which are immediately quenched in the water. The process gives roughly spherical particles of metal; but they are covered with oxide, which must be removed by heat-treating in hydrogen.

3.6 Electro-Deposition

Under suitable conditions iron and nickel-iron alloys can be electro-deposited either as powders

* Vickers diamond-pyramid hardness test.

[1] Numbered references are to the Bibliography, Section 10, pp. 68–69.

or as brittle slabs which can readily be pulverized. Electrolytic iron was at one time used for telephone-line loading-coils.[2] An electrolytic method of depositing Permalloy* has been worked out[19] but it is rather complicated and does not seem to be in use commercially.

3.7 Other Methods

Powdered iron and some of its alloys can be produced by other methods, such as metal-spraying and chemical reduction of iron salts; these methods do not seem to have been developed commercially and may, for the present, be neglected.

3.8 Non–Metallic Magnetic Substances

Although most magnetic materials with which we are familiar are iron or alloys of iron, the first magnets ever known were non-metallic. Lodestone, or magnetic oxide of iron, is typical of a family of magnetic substances known as ferrites. They are salts of the hypothetical ferric acid $HFeO_2$ and are made by the interaction of ferric oxide with the oxide of a metal such as nickel, copper, or cobalt. In general, they have the composition $RO \cdot Fe_2O_3$, where R represents a divalent metal. Magnetic iron oxide is just a special case, being in fact ferrous ferrite, $FeO \cdot Fe_2O_3$. Not all the ferrites are markedly magnetic, zinc and cadmium ferrites being notable exceptions.

Ferrites can be made by grinding together appropriate quantities of the necessary oxides and then heating in air or oxygen to about 1100 degrees centigrade. In many cases their magnetic properties are very dependent on their thermal history. It is not possible in this brief review to describe their production in detail, but an indication of their properties will be given later.

An analogous material is gamma ferric oxide, a metastable variety of ferric oxide which, being produced by the gentle oxidation of magnetite, retains its basically cubic structure and is magnetic. If heated to 500 degrees centigrade it recrystallizes to the normal hexagonal form, which is non-magnetic.

4. *Core Manufacture*

This is a branch of powder metallurgy that is usually left severely alone in any description of the art. This omission, at first sight strange, is reasonable since the general aim of the powder metallurgist is to produce either a homogeneous metallic mass or a sponge in which there is a continuous metallic phase, whereas the core maker tries to produce a dense compact of discrete metallic particles, each one insulated from all the others. The art is perhaps more akin to synthetic-resin moulding than to metallurgy.

Powder cores tend to fall naturally into two classes according to the pressure used in forming. The high-pressure type needs pressures up to 100 tons per square inch, whilst the low-pressure type is generally made at pressures below 20 tons per square inch. The high-pressure type nowadays usually requires heat-treatment to anneal the metal after pressing, but low-pressure cores need only gentle baking to cure the resinous binder.

High-pressure powder-cores were developed first and satisfied the need for high-permeability cores for audio- and carrier-frequency uses; the low-pressure cores followed when it was found that they could be used at radio frequencies.

4.1 High–Pressure Cores

The great majority of high-pressure cores made to-day are of the toroidal type and are pressed in sectionalized tools. The first commercial production of high-permeability powder-cores yielded a permeability of 35, the particles being composed of electrolytic iron powder; pressing was in special steel tools and the pressure used was around 100 tons per square inch. This pressure was necessary to obtain close packing of the particles and to reduce the air-gaps between them. Before pressing, the particles were coated with an insulating layer that had to withstand the strains during pressing while maintaining separation between the particles, and yet this layer had also to allow sufficient interlocking of the particles to ensure adequate mechanical strength of the core.

When alloy powder-cores of the Permalloy type were developed, it was found that the pressed core needed to be heat-treated to obtain the properties required. This is not surprising, since the particles are strained beyond their elastic limit during pressing and all high-permeability materials are sensitive to the effects of strain. The requirements for the insulating layer are therefore severe; it must be capable of being

* Nickel-iron alloy.

applied as a very thin layer that will withstand heat treatment of the pressed core, without breaking down or fluxing, and must also be such that the core is mechanically strong at all stages of manufacture. Inorganic silicates and oxides are suitable as insulating layers; they are applied to the particles from solutions or suspensions in water. Often, successive layers are applied. During the heat-treatment the insulating materials produce a ceramic-like substance. High-permeability nickel-iron powder cores are heat-treated at temperatures often as high as 600 degrees centigrade in a controlled atmosphere.

4.2 Low–Pressure Cores

Many magnetic powders can be packed by hand into a container; they then yield an effective permeability of about 10. Since this is a useful working permeability for many cores for radio frequencies, it follows that much lower pressures can be used for such cores than are necessary for the higher-permeability types. The magnetic powder is first insulated with a thin layer of, usually, inorganic material. A binder, normally a phenolic resin, is next added either dry or else as a solution which is then dried off to a controlled degree. If a dry resin is used the method of producing cores follows normal hot-moulding technique, while if the second method of applying the binder has been used it is often possible to mould in the cold state and then to complete the curing of the resin by a subsequent baking at 100 to 140 degrees centigrade for 3 to 18 hours.

5. *Magnetic Properties at Low Flux Densities*

By "low" flux densities are meant those for which the hysteresis loop is well represented by Rayleigh's approximation of two parabolae.[3] In practice this implies flux densities of the order of 10 gauss or less.

5.1 Permeability

If particles of magnetic material are uniformly distributed in insulating material, then the permeability of the resulting body is a function more of the ratio (volume of particles)/(total volume) than of the permeability of the particles.[4–8]

The permeability can, however, vary greatly according to the particle shape and manner of packing. Hence the possibility of increasing the permeability depends mainly on the use of insulation as thin as possible in relation to the particle size (this will usually involve using larger particles), on the mixing of particles of different sizes in proportions that result in close packing, and on the mechanical softness of the particles and the application of high pressure.

Within the range of frequencies in which powder cores can be used without excessive loss, the permeability is almost independent of frequency; for the skin effect in small particles is usually negligible, and even when the effective permeability of the particles begins to fall, the permeability of the mixture is not greatly affected. For example, for 10-micron particles of iron of permeability 1000, the quantity $\pi d(\mu f/\rho)^{\frac{1}{2}}$ is only 0·2 at 10 megacycles per second. The permeability of a particle would then be reduced by about 0·1 per cent and that of a material having $\mu = 20$ by 0·002 per cent.

For a toroidal core, the permeability is a useful quantity to know, because the inductance of a coil can be calculated directly from it; but in other shapes the magnetic leakage may be large, and then the "inductance ratio," i.e., the ratio in which the inductance is increased by the presence of the core, this ratio being valid only for a particular shape and permeability, is a more useful piece of information.

5.2 Hysteresis

The calculation of hysteresis loss in powder cores is of interest chiefly at low flux densities. It can be expressed by an equivalent resistance, the resistance usually being taken as proportional to the maximum flux density. Thus the hysteresis resistance is proportional to current, and the hysteresis loss to the cube of the current. The difference between this and the Steinmetz relation [in which the loss varies as $(\text{current})^{1\cdot 6}$] is due to the difference in the shapes of loop under consideration. If the Rayleigh loop applies, then there is a relation $\Delta R = \frac{8}{3} f \Delta L$ between the increases of resistance and inductance at frequency f. This holds in practice for many materials.

The hysteresis loss appears to be influenced by mechanical strain and therefore by the magnetostriction of the material.[9]

5.3 Eddy–Current Loss

If the particle shape is known approximately, the eddy-current loss is calculable from the resistivity and the particle size; for spherical particles,

$$\mu e = \frac{2\pi^3 d^2 \mu}{5\rho\gamma^{\frac{1}{3}}},$$

where

e = Legg's eddy-loss coefficient
μ = permeability
ρ = resistivity
d = particle diameter
γ = (volume of particles)/(total volume).

This quantity may be very small. There is also the possibility of eddy currents flowing in larger paths around the whole cross-section of the core; this proves to be unimportant in metallic powders with insulating binders, but in ferrites it becomes appreciable at frequencies above about 1 megacycle.

5.4 Residual Loss

This is another source of loss not accounted for by hysteresis and eddy currents. This is known as "residual loss," "viscosity loss," or "*Nachwirkung*." This type of loss is not usually considered in stampings, because they are usually tested at flux densities at which residual loss is masked by hysteresis. In iron powders this loss is assumed to be proportional to frequency; in ferrites it includes terms proportional to the square, and perhaps to higher powers, of the frequency.

Possible explanations are viscous resistance to the movement of magnetic domains, and change of permeability with time. The existence of the latter effect has been established independently.[10]

5.5 Representation of Losses

The losses are independent of each other at all low frequencies; the equivalent resistance that they add to the circuit can therefore be represented approximately by an expression of the form

$$R_m/L_m = \mu h\hat{B}f + \mu cf + \mu e'f^2 + \mu ef^2,$$

where R_m, L_m are the resistance and inductance due to the core, and μh, μc, $\mu e'$, μe are coefficients expressing hysteresis, residual, residual varying with f^2, and eddy-current losses. $\hat{B}$ is the peak flux density. (This expression is adapted from that of Legg by the inclusion of the term in $\mu e'$.) For the purposes of this paper μe and $\mu e'$, which cannot be separately measured, can be considered as a single quantity μe.

For other forms of this expression see Table 1.

6. *Apparatus for Measurements: Research and Development*

6.1 General

Core-loss constants should be measured in the same frequency range as that in which the cores will eventually be used, so that any departures from the assumed law of variation of resistance will not be too serious. If the eddy-current loss and the residual loss are to be separated, the test frequencies should be of the same order as the frequency at which these two losses are equal. Thus high-permeability nickel-iron cores are usually measured at audio frequencies, and carbonyl-iron and the lower-permeability nickel-iron cores at carrier frequencies.

All the methods described for measurements of losses in powder-core materials are in fact methods of measuring the resistance of inductors wound on toroidal cores. From the values of this core-loss resistance at various frequencies and currents, the core-loss constants of the material can be determined.

6.2 Audio–Frequency Bridges

The test coil is preferably of about 100 millihenrys inductance and wound in sections to have a self-capacitance of not more than 100 micromicrofarads. The corrections for self-capacitance and conductance will then be negligible. Tests may be carried out on a direct-comparison bridge. Its inductance standards should be large, of high quality, and of low losses, as they can be if the bridge is permanently set up and the screening is adequate. A filter is required between the oscillator and the bridge to reduce harmonics to a tolerable value. A frequency run at 1 milliampere testing current at frequencies of, say, 500, 1000, 1500, and 2000 cycles, and a current run at 1, 2, 3, and 4 milliamperes at two of the frequencies (for example 1000 and 2000 cycles) are required. Alternatively audio-frequency tests may be made

TABLE 1

Losses in Powder Cores at "Low" Flux Densities and Conversions between Various Sets of Units

				In This System	Use of System	
Given the coefficient	μh	μc	μe			
of which the units are	$\frac{\text{ohms}}{\text{henrys}\times\text{c/s}\times\text{gauss}}$	$\frac{\text{ohms}}{\text{henrys}\times\text{c/s}}$	$\frac{\text{ohms}}{\text{henrys}\times(\text{c/s})^2}$	$\frac{R_m}{L_m}=\mu h\hat{B}f+\mu cf+\mu ef^2$	Originated in U.S.A.	Both of these systems are in use in Britain
or, more briefly,	1/gauss	pure number	sec	(a is often written instead of h)		
To obtain	h	v	e_c			
of which the units are	$\frac{400(2)^{\frac{1}{2}}\pi}{\text{oersteds}}$	10^3	10^6 sec	$\frac{R}{L}=\frac{hINf'}{l}+vf'+e_cf'^2$	Originated in Germany	
multiply by	1775μ	10^{-3}	10^{-6}			
To obtain	ρ_h/H_m	dn	v			
of which the units are	2π/oersteds	2π	$4\pi^2$ sec	Power factor $=R/\omega L$ $=\rho_h+dn+v\omega$	Originated in Sweden	
multiply by	$\mu/2\pi$	$1/2\pi$	$1/4\pi^2$			
To obtain	c_h	c_r	c_e			
of which the units are	$\frac{\text{oersteds}}{\text{gauss}^2}$	$\frac{\text{oersteds}}{\text{gauss}}$	$\frac{\text{sec}\times\text{oersteds}}{\text{gauss}}$	$R/\mu fL=c_h\hat{B}+c_r(f)+c_ef$ (In this system c_r is a function of f, and c_e is to be calculated from the resistivity and length of conducting path)	Intended for ferrites	
use:	$c_h=\frac{1}{\mu}\mu h$	$c_r=\frac{1}{\mu}(\mu c+f\mu e)$				
also:	$C_h=10^6c_h$	$C_r=10^4c_r$				
To obtain	F_h	F_c	F_e			
of which the units are	$\frac{1}{\text{amp}\times(\text{henrys})^{\frac{1}{2}}}$	pure number	sec	$R/L=I(L)^{\frac{1}{2}}fF_h+fF_c+f^2F_e$	For design of inductors on a given core. (The conversion factors hold for toroids only)	
multiply by	$15{,}800\ \mu/Al$	1	1			
To obtain	P_h	P_c	P_e			
of which the units are	watts	watts	watts	Power loss $=P_h+P_c+P_e$	For comparison with properties of laminated cores	
multiply by	$15{,}800(\mu L/v)^{\frac{1}{2}}fLI^3$	fLI^2	f^2LI^2			

Note on units: the quantity $\frac{\text{ohms}}{\text{henrys}\times\text{cycles per second}}$ is dimensionless.

f = frequency in cycles per second (c/s)
f' = frequency in kilocycles per second
I = root-mean-square current in amperes
$\hat{B}$ = peak flux density

N = number of turns
l = length of magnetic path in centimetres
v = volume of core in cubic centimetres
A = cross-section of core in square centimetres.

by the series-resonance method. But the elimination of harmonics is then more difficult; also, a mica capacitor will be necessary to resonate with the specimen, and great care is needed to ensure that its conductance is accurately measured and allowed for.

6.3 Series–Resonance Bridge

A series-resonance bridge for powder-core specimens has been described by Welsby.[11] The essentials are shown in Figure 1. There must be two equal fixed resistors, a variable resistor, and a good variable air capacitor connected in series with the specimen. An oscillator of good short-period stability is essential; it may be thought worth while to use a set of crystal oscillators. The amplifier used with the bridge must discriminate strongly (e.g., 60 decibels) against harmonics of the working frequency. If the specimen arm is double-screened (or triple-screened for use with an auxiliary "ballast coil"), the resistance of the specimen is measured directly on the variable resistor. The shunt conductance of the variable capacitor, and of any strays in parallel with it, must be kept low and corrected for in the result. Unless the range of the capacitor is extended by the use of, for example, mica capacitors, the correction for self-capacitance of the specimen is large. If the range is extended in this way, the correction for shunt conductance of the variable capacitors becomes uncertain; if a "ballast coil" is used to extend the frequency range, the correction becomes very large.

The inductance of the specimen may conveniently be between 0·2 and 10 millihenrys, according to the frequency range required. Its self-capacitance must be kept small—say below 10 micromicrofarads. The wire losses may be either measured on a similar coil wound without a core (in which case stranded wire should be used), or calculated (in which case solid wire is better).

This bridge can be made to measure resistance to within about 2 per cent.

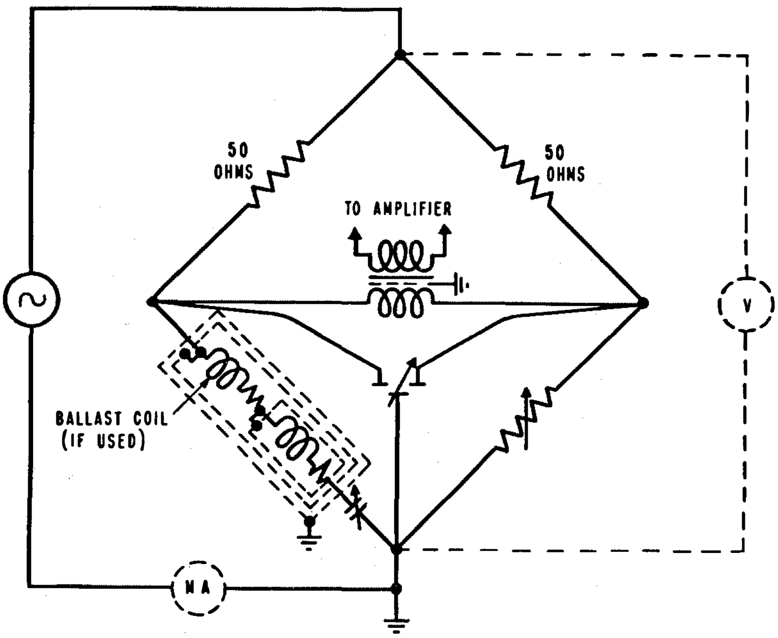


Figure 1—Series-resonance bridge for measurements on powder cores.

6.4 *Q*–Meter with External Galvanometer[12]

In some *Q*-meters the injected current is measured by a thermocouple. If such a *Q*-meter is provided with an additional galvanometer connected to the thermocouple output, and this galvanometer can be shunted to reduce its sensitivity in known ratios (as shown in Figure 2) then the *Q*-meter can be used for measurements by the reactance-variation method without any use being made of the calibrations of either the built-in meters or the external galvanometer. The difference from the usual reactance-variation method is that, instead of the voltage in the resonant circuit being measured with a constant injected voltage, it is kept constant by detuning the circuit while the injected voltage is increased in known ratios.

An auxiliary coil is used, and the difference when the specimen is connected is observed. The two coils may be either in series or in parallel. This method measures the resistance within about 3 per cent, and the result requires a large correction for the self-capacitance of the specimen. However, the high frequencies available with this method make it very suitable for measurements of low eddy-current losses. It is not suitable for measurements of hysteresis.

6.5 Wide–Range Maxwell Bridge

A Maxwell bridge for frequencies of 1 kilocycle to 1 megacycle can be made by using radio-type resistors in both fixed and variable resistive arms. The product of the resistances of the two fixed resistors must be so chosen that the bridge can be balanced with an air capacitor in the variable arm. Capacitance or inductance must be added so that this product contains no reactive term. Two arms—the specimen arm and one of the fixed resistors—must be of low impedance, and their junction is earthed (see Figure 3). If a transformer were used between the non-earthed ends of these two arms, it would introduce large capacitance and conductance to earth. This can be avoided by using, as the first stage of the amplifier, a valve whose grid and cathode are connected to these two points.[13] The resistance of the specimen is read directly on a conductance box in the variable arm. A correction is required for the series resistance of the variable capacitor; this correction becomes large at frequencies above 1 megacycle. The correction for the self-capacitance of the specimen may be made very small. A Wagner earth has intentionally not been used, as it has been found to introduce errors, of unknown origin, in the measurement of the component in quadrature with the predominant component of the impedance of the specimen. The errors introduced by direct earthing are of known origin and can be made tolerably small.

The accuracy obtained in resistance measurement is to within about $\frac{1}{2}$ per cent. The wide frequency range available with any one specimen assists in the measurement of residual loss. As the use of radio-type resistors limits the attainable accuracy to within about $\frac{1}{4}$ per cent, a separate bridge, operated at a frequency such as 300 kilocycles, is needed for hysteresis measurements.

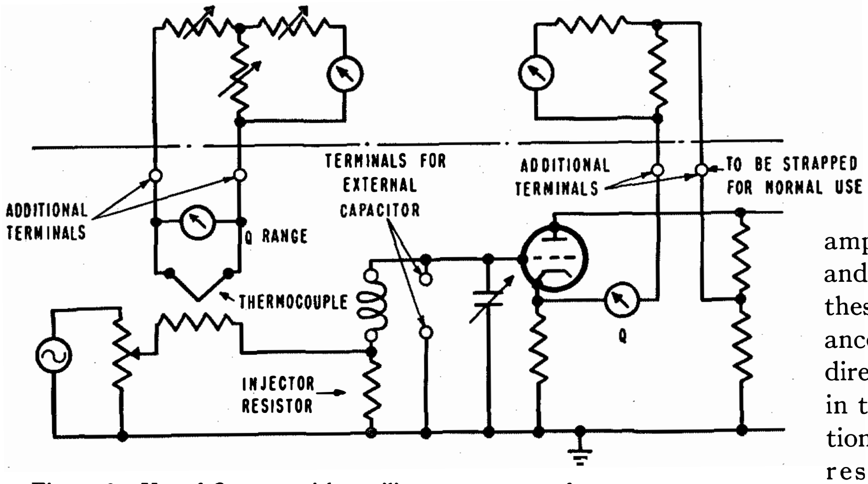


Figure 2—Use of Q-meter with auxiliary components for measurements on powder cores.

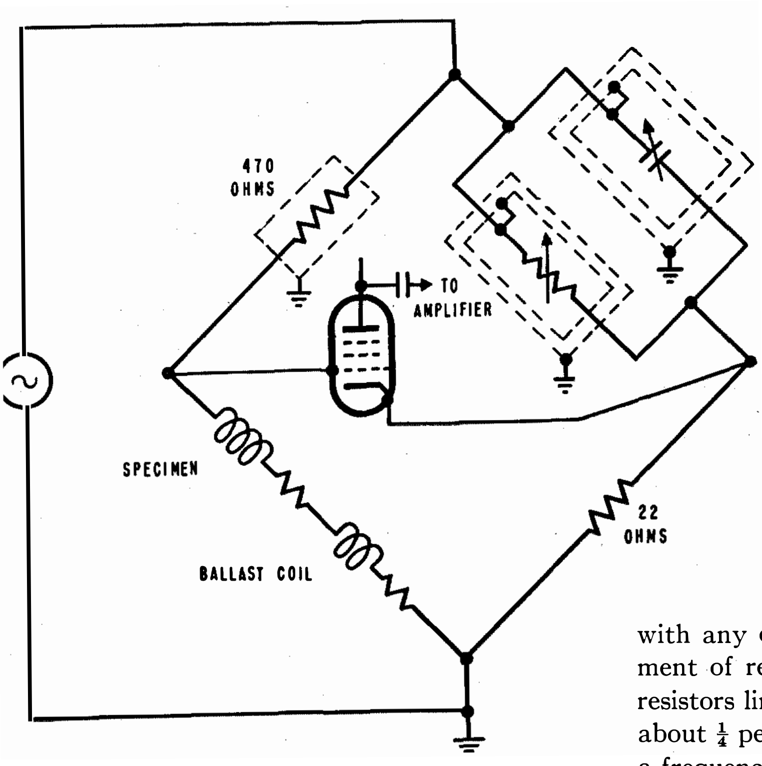


Figure 3—Wide-range Maxwell bridge for measurements of losses in powder cores.

Useful confirmation of the residual-loss measurements can be obtained from a further bridge, measuring the change of resistance between two audio frequencies (such as 1 and 4 kilocycles).

6.6 Temperature Coefficient of Permeability

The temperature coefficient of the inductance of a coil is not necessarily the same as that of the permeability of the core on which it is wound. Measurements on the core alone can be made if the temperature of the coil is kept constant while that of the core is varied. However, the shape of the test core should resemble those used in practice, as it is not certain that the material is isotropic.

Measurements on a complete inductor involve the mechanical behaviour of the winding and also the temperature coefficient of its self-capacitance. To eliminate the latter effect, we may measure the temperature coefficients of the apparent inductance at two frequencies; we then have two equations from which the temperature coefficients of the permeability and the self-capacitance can both be calculated.

For either type of measurement, the change in inductance must be measured to within about 1 part in 10 000. This may be done with a bridge circuit, or alternatively by using the inductor in the tuning circuit of an oscillator running at about 100 kilocycles. In the latter case it is essential either to use an oscillator of which the frequency is independent of the magnitude of the losses in the tuning circuit or to keep the losses constant throughout a measurement.

7. *Apparatus for Measurements: Control of Production*

The following tests are carried out:—

A. Measurement of permeability

B. Determination of core losses, including the resolution of core-loss coefficients

C. Measurement of magnetic stability

and, in special cases only,

D. Determination of temperature coefficient of permeability, as described in Section 6.

The tests are made on sample cores from a sample mix of each batch of powder manufactured, and check tests are made on each production lot of cores from each accepted batch of powder. For loading coils, the final acceptance tests include measurement of total effective resistance and measurement of increase in effective resistance brought about by a given increase in coil current.

Tests in all cases are made on ring specimens.

7.1 Measurement of Permeability

The result of the measurement of permeability on the trial core is used to make any necessary modification to the processing of the insulated powder for subsequent cores to be made from the same batch of metal powder. Two types of permeameter will be described, since the first method, although it has now been abandoned by the authors, is of considerable historic interest.

7.1.1 Kelsall Permeameter[14]

In this permeameter a primary coil of very-high inductance is enclosed in a copper shell which acts as a very-low-impedance secondary circuit. The shell is so arranged that a lid can be removed and a core inserted to be interlinked between the primary and secondary. From measurements of the primary inductance before and after insertion of the test core the permeability of the core can be assessed. Unfortunately the results vary according to the resistance of the secondary (i.e., with the fit of the lid), and this tends to change, particularly with wear. For this reason, the use of this type of permeameter has been abandoned by the authors.

7.1.2 "Twenty-Turn" Permeameter

The core is placed in a hinged jig, which when closed consists essentially of a number of turns embracing the toroid. The turns are series-connected and the inductance is measured (at 800 cycles) on an inductometer giving a direct reading. The apparatus is calibrated by inserting cores of known permeabilities which have been assessed by applying single-layer windings and measuring their inductances on a Campbell inductometer.

7.2 Determination of Core Losses

7.2.1 Separation of Loss Coefficients

Tests to determine the contribution of the three sources of loss—hysteresis, residual, and eddy current—to the total effective resistance of a coil wound on the test core are made on a small number of cores selected at random from each production batch.

In one example of routine measurements, the test core is wrapped with 4 thicknesses (2 lapped layers) of paper tape and is then wound in four sections (balanced in two halves) to an inductance of 10 millihenrys ($\pm 0{\cdot}2$ millihenry). The wire is single-strand 28 Standard Wire Gauge copper. The effective resistance of the winding is measured at 4 and 8 milliamperes at each of the frequencies 4, 8, and 16 kilocycles on a series-resonance bridge.

Care must be taken with the contact resistance in the resistive arm, which may introduce errors of the order of 0·01 ohm which are important when low-loss cores are being measured.

The 4–8–16-kilocycle oscillator feeding the bridge is crystal-controlled.

A heterodyne detector is used, giving a 1-kilocycle signal which is measured by visual (meter or "magic eye") or aural means.

The readings of effective resistances are used to evaluate the three coefficients μh, μc, and μe, corresponding, respectively, to hysteresis, residual, and total eddy-current loss of coil and core.

7.2.2 Measurement of Core Eddy-Current Loss Factor

For coils wound on cores of low permeability (below 50) the eddy-current loss in the winding swamps that in the core and a separate measurement of eddy-current loss, using a low-loss coil, is made on such cores. For this purpose a winding of 30 turns of 243-strand litz wire is used and is tested at a frequency depending on the permeability of the core, namely

Nominal permeability 50: test frequency = 700 kilocycles

Nominal permeability 25: test frequency = 1000 kilocycles

Norminal permeability 14: test frequency = 1500 kilocycles

Under these conditions the loss resistance is mainly due to eddy-current losses in the core. A correction is made for the loss in the winding.

7.2.3 Degree of Accuracy

The method of measurement of loss coefficients outlined above is open to errors that are difficult to assess accurately as they depend on stray resistances in capacitors and contact resistances in the decade resistances and slide-wire. As a result of general experience and the analysis of results obtained on a reference core measured from day to day over a long period, the degree to which the various results should be expressed is about 10 per cent for μh, about 1×10^{-3} for μc, about 5 per cent for μe_t, and about 1×10^{-9} for low values of μe or 10 per cent for high ones.

7.3 Magnetic Stability

For many requirements, e.g., for loading-coil cores as described in the specification[15] of the Comité Consultatif International Téléphonique, it is important that the permeability of a core should not be greatly changed by the passage of a large surge of direct current through the coil winding. The usual pulse given is 1000 ampere-turns for 5 seconds. The coil is then allowed to recover for 5 minutes and the inductance is remeasured. The change is expressed as a percentage of the inductance before polarizing and is usually less than 1 per cent.

7.4 Determination of Particle Size

Various methods are available and have been used from time to time, the most popular being sedimentation, air permeability, and microscopic examination.

7.4.1 Sedimentation

The rate at which a small particle falls through a viscous liquid under the influence of gravity depends on the densities of the particle and the liquid, the diameter of the particle, and the viscosity of the liquid. For given materials the rate of fall is proportional to the square of the diameter of the particle. Sedimentation rate can be measured, and from it the distribution of the size of particles can be calculated.

7.4.2 Air Permeability

A current of air flows more readily through a bed of coarse powder than through an otherwise similar bed of fine powder. Gooden and Smith[16] have derived the mathematical formulae and developed a convenient instrument in which air at constant pressure is led to the packed powder sample in a tube. The flow of air through this packed bed of powder is measured by a flow meter, calibrated directly in particle diameters. The particle diameter so measured is a statistical average.

7.4.3 *Microscopic Examination*

The powder is spread very thinly on a microscope slide, and the diameters of (say) 100 randomly selected particles are measured using a micrometer eyepiece. The method is tedious, and it is difficult to be certain that a random selection has been made, but it is probably the simplest way of determining particle-size distribution. This method is more difficult with the very fine powders, which contain a high proportion of particles of less than 1 micron diameter.

8. *Summary of Properties*

It may be easier to determine the magnetic properties of a given material in sheet than in powder form, but the measurements on one are not readily converted to be useful for the other. Probably the most noticeable differences between measurements on sheet and powder are in the permeability and the eddy-current losses. Even with the thinnest laminated core the permeability in the required direction is substantially that of the massive material, but, no matter what the permeability of the solid, it is not possible to make a metal powder core with a permeability much exceeding 125. This effect is clearly shown in Table 2, which gives values of permeability for

TABLE 2

Material	Permeability of Solid[17]	Permeability of Experimental Powder Core
Iron	250	75
Nickel	150	30
50/50 Nickel-Iron	2000	80
80/20 Nickel-Iron (Annealed)	2000	80
80/20 Nickel-Iron (Flatter Particles)	2000	110

material in solid (sheet) form and in the form of powder cores. Each of the powder cores was insulated by the same process, using the same percentage of filler.

The permeability of 80/20 nickel-iron, as generally used in laminated cores, is increased to at least 8000 by quenching from about 600 degrees centigrade, a process which cannot be applied to powder cores.

It is also difficult to correlate data on loose powder materials with test data on finished cores, but in spite of this there are certain guiding lines that can be followed. As will be seen from the data, the same arguments cannot always be applied to alloys and to iron powders.

Particle size has approximately the predicted effect on the eddy-current loss. With nickel-iron powders, low hysteresis loss seems to be associated with large particles, but with carbonyl iron grade E the reverse is true (Table 3).

TABLE 3

Material	Mean Particle Diameter, Microns	Hysteresis-Loss Coefficient, $\mu h \times 10^3$
Nickel-Iron	>15	0·070–0·079
	<15	0·093–0·118
Carbonyl E (Experimental)	5–9	0·28
	3–6	0·07
	2–5	0·025
	1–3	0·035

There are also one or two simple empirical relationships, e.g., for nickel-iron, hysteresis loss increases directly with oxygen content, and with carbonyl iron (partly decarburized) hysteresis loss and peremability vary inversely as the carbon content, for a given forming pressure.

To the designer of coils and equipment magnetic data of the core material are important, and Table 4 gives a review of the properties of many materials that can be used. Other points to be watched are the strength of the finished cores and the ease of manufacturing them. As a very broad generalization, the mechanically disintegrated and the soft powders are easier to form into strong cores than the hard spherical ones like Carbonyl E. As this is to some extent due to the ease with which the particles can be deformed, it is liable to be associated with difficulty in insulating.

Another point to be borne in mind is that nickel-iron alloys that are of the 78/22 class of alloy generally need heat treatment after pressing if their best properties are to be achieved. Thus it may be difficult to decide whether a core of permeability 50 should be made of soft iron, not needing heat treatment, or 78-per-cent nickel-iron, which must be heat treated after pressing.

The magnetic data given include some of the ferrite type of magnetic powders. These do not all represent commercial grades, and are only included as typical of the various materials. Ferrites are so susceptible to heat treatment and admixture that they cannot be adequately dealt with in a table of this kind.

When coils are to be designed it is important to remember that the total effective resistance is not composed entirely of the direct-current resistance and the core loss, but may have components due to three other factors, namely eddy-current loss in the winding and the effects of self-capacitance and its associated conductance. Thus,

$$\text{effective resistance} = \frac{R+\omega^2L^2G}{(1-\omega^2LC)^2},$$

where

R = direct-current resistance + copper eddy-current loss + core loss

L = inductance of the coil

C = self-capacitance of the coil

G = conductance of the self-capacitance = $\omega C \times$ power factor. (In Figures 4 and 5 the power factor is taken as 1/120.)

In Figures 4 and 5 the coil performance is shown as the variation of power factor with frequency. The lower (unshaded) portion gives the direct core losses, while the upper (shaded) portion represents the effects arising from the winding.

The direct-current resistance depends on the core size and permeability and on the gauge of wire used. The copper eddy-current loss is determined by the degree of stranding. The self-capacitance and its conductance are dependent on the arrangement of the winding and the nature of the insulation between turns.

TABLE 4

	Binder, in Per Cent by Weight	Permeability μ		Hysteresis Coefficient $\mu h \times 10^3$		Residual Coefficient $\mu c \times 10^3$		Eddy-Current Loss Coefficient $\mu e \times 10^6$	
Pressure in Tons per Square Inch		20	50	20	50	20	50	20	50
80/20 Nickel-Iron, 100 Tons per Square Inch Pressure, Heat-Treated		120		0·24		6		—	
		90		0·19		5		—	
		50		0·17		3		0·25	
		25		0·14		2		0·08	
		12		0·16		1		0·03	
2/81/17 Molybdenum-Nickel-Iron, 90 Tons per Square Inch, Heat-Treated	1	125		0·15		4		2·7	
	3	26		0·18		2		0·2	
	3	26		0·16		6		0·13	
	12	14		0·14		1·5		0·10	
	12	14		0·12		3·3		0·05	
Electrolytic Iron, 80 Mesh, 90 Tons per Square Inch	1	35		1·7		3·8		3·1	
Electrolytic Iron, 200 Mesh, 90 Tons per Square Inch	3	26		2·1		3·6		0·8	
Electrolytic Iron, 200 Mesh	4	17	24	4	4·6	30	30	0·5	0·6
Electrolytic Iron, 325 Mesh	4	16	23	3·8	3·5	20	30	0·3	0·4
Carbonyl Iron MC	4	14·5	22	0·53	0·55	14	13	0·01	0·02
Carbonyl Iron ME	4	10·8	12·7	0·05	0·06	2·3	2·5	0·002	0·002
Carbonyl Iron MF	4	9	10·5	<0·02	<0·02	<1	<1	<0·002	<0·002
Carbonyl Iron, Foreign C	4	16	22	1·6	1·4	12	10	0·04	0·04
Carbonyl Iron, Foreign E	4	10·5	12	0·1	0·1	4	4	0·005	0·006
Carbonyl Iron, Foreign H	4	10	11·5	0·07	0·04	2	2	0·002	0·002
Carbonyl Iron, Foreign SF	4	9	10	0·02	0·02	1	2	<0·002	<0·002
Carbonyl Iron, Foreign L	4	25	30	3	3	12	11	0·15	0·13
Reduced Iron		19	—	1	—	14	—	0·06	—
Sendust		60		0·34		11		0·2	
Magnetite	8	4·2	—	0·04	—	7	—	<0·002	—
γ Ferric Oxide	8	4·9	—	0·8	—	24	—	0·006	—
Copper Ferrite	Sintered	8		20		50		0·35	
Nickel Ferrite	Sintered	12		5·4		22		0·04	
Magnesium Ferrite	Sintered	6		4·4		35		0·005	
Manganese Ferrite	Sintered	100		4·5		40		0·2	
Grey Cast-Iron, 140 Mesh	4	12·5	14·5	3·5	4·0	20	30	0·12	0·16
White Cast-Iron, 200 Mesh	4	7	8	2	2	20	30	0·01	0·04
Cast Iron, Degussa Process, Annealed	4	7	—	0·6	—	30	—	0·006	—
Ferroxcube 1*	None	1000		—		400		3	
Ferroxcube 3*	None	1500		2·5		15		1·5	
		600		—		20		0·5	
Ferroxcube 4*	None	80		—		25		0·01	

* Deduced from data given by Snoek.[16]

The hysteresis losses correspond to a current of 1 milliampere, so that $\hat{B}$ is 12 gauss for the core of Figure 4 and 1·5 gauss for that of Figure 5.

As shown by Figure 4, at audio frequencies it is easy to arrange for the power factor to be determined almost entirely by the direct-current resistance and the core losses, provided that the self-capacitance is not more than 100–200 micromicrofarads for a 100-millihenry coil. However, Figure 5 illustrates the importance of keeping the self-capacitance and its conductance as low as possible when the term $\omega^2 LC$ becomes appreciable.

9. Acknowledgments

The authors wish to thank the Engineer-in-Chief of the Post Office, the Directors of Standard Telephones and Cables, Limited, and the Manager of the Research Laboratories of The General Electric Company, Limited, for permission to publish the paper.

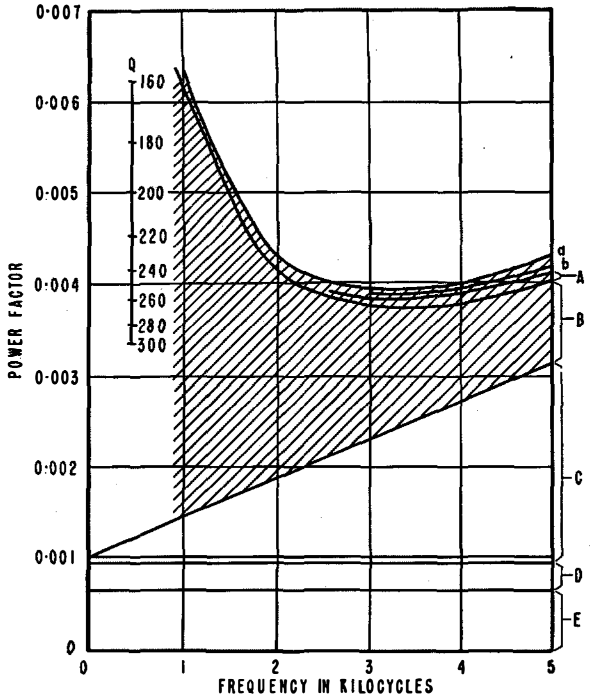


Figure 4—Variation of power factor of a 100-millihenry inductance on a Permalloy powder core for frequencies between 1 and 5 kilocycles due to: *A*—copper eddy-current loss, *B*—direct-current resistance, *C*—eddy-current loss in core, *D*—hysteresis, and *E*—residual loss in core. Values of *Q* are plotted for self-capacitances of *a*—200 and *b*—50 micromicrofarads.

10. Bibliography

1. L. B. Pfeil, "Symposium on Powder Metallurgy," *Iron and Steel Institute Special Report No. 38*; 1947: pp. 47–51.

2. B. Speed and G. W. Elmen, *Transactions of the American Institute of Electrical Engineers*, v. 40, p. 1321; 1921.

3. E. Peterson, "Harmonic Production in Ferromagnetic Materials at Low Frequencies and Low Flux Densities," *Bell System Technical Journal*, v. 7, p. 762; 1928.

4. V. E. Legg, and F. J. Given, "Compressed Powdered Molybdenum Permalloy for High-Quality Induct-

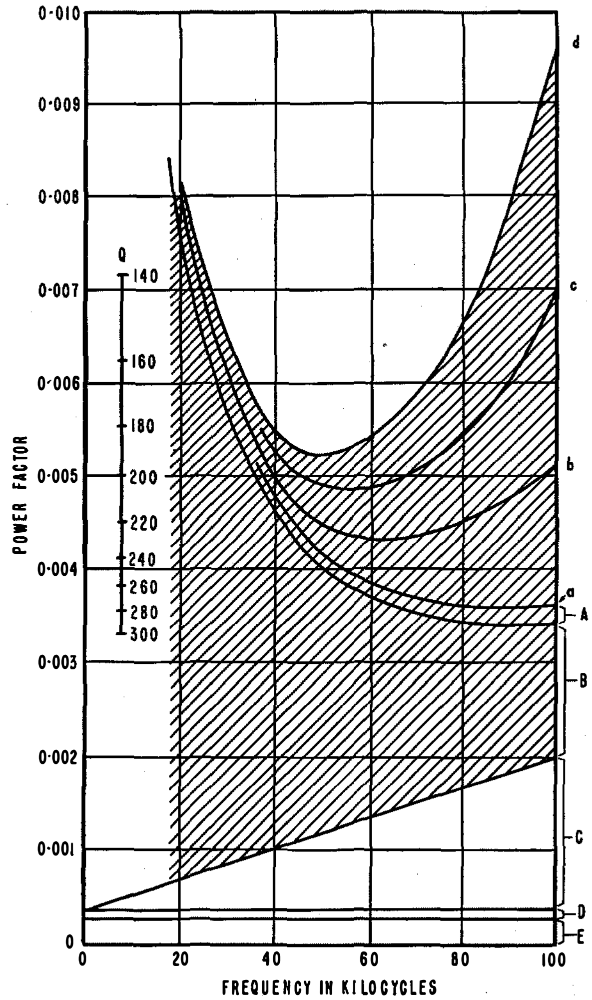


Figure 5—Variation of power factor of a 20-millihenry inductance on a Permalloy powder core for frequencies between 20 and 100 kilocycles due to: *A*—copper eddy-current loss, *B*—direct-current resistance, *C*—eddy-current loss in core, *D*—hysteresis, and *E*—residual loss in core. Values of *Q* are plotted for self-capacitances of *a*—0, *b*—10, *c*—20, and *d*—30 micromicrofarads.

ance Coils," *Bell System Technical Journal*, v. 19, p. 385; 1940.
5. G. W. O. Howe, "Iron Powder Compound Cores for Coils," *Wireless Engineer*, v. 10, p. 1; 1933.
6. P. R. Bardell, "Permeability of Dust-Cores," *Wireless Engineer*, v. 24, p. 63; 1947.
7. W. Doebke, "Ferromagnetische Mischkorper," *Zeitschrift für technische Physik*, v. 11, p. 12; 1930.
8. F. Ollendorff, "Magnetostatik der Masskerne," *Archiv für Elektrotechnik*, v. 25, p. 436; 1931.
9. J. L. Snoek, "New Developments in Ferromagnetic Materials," Elsevier Publishing Company, New York and Amsterdam; 1947: p. 7.
10. J. L. Snoek, "New Developments in Ferromagnetic Materials," Elsevier Publishing Company, New York and Amsterdam; 1947: p. 54.
11. V. G. Welsby, "Dust-Cored Coils," *Electronic Engineering*, v. 16, p. 281; 1943.
12. A. C. Lynch, "Improved Accuracy with a 'Q'-meter by the use of Auxiliary Components," *Electronic Engineering*, v. 21, p. 91; 1949.
13. A. C. Lynch and E. J. Frost, British Provisional Patent 32319/47.
14. G. A. Kelsall, "Permeameter for Alternating-Current Measurements at Small Magnetizing Forces," *Journal of the Optical Society of America and Review of Scientific Instruments*, v. 8, p. 329; 1924.
15. Comité Consultatif International Téléphonique, Réunion d'Oslo (1938), tome 1 ter, p. 126.
16. E. L. Gooden and C. M. Smith, "Measuring Average Diameter of Powders: Air Permeation Apparatus," *Industrial and Engineering Chemistry* (Analytical Edition), v. 12, p. 479; 1940.
17. G. W. Elmen, "Magnetic Properties of Iron, Nickel and Cobalt," *Journal of the Franklin Institute*, v. 207, p. 602; 1929.
18. J. L. Snoek, "New Developments in Ferromagnetic Materials," Elsevier Publishing Company, New York and Amsterdam; 1947: pp. 77–98.
19. C. L. Faust, "Electrodeposition of Alloys, 1930–1940," *Transactions of the Electrochemical Society*, v. 78, p. 383; 1940.

Recent Telecommunication Development

Acoustic Mill Feed Controller

Solid fuel for modern draught-fed furnaces passes through a mill in which steel balls cascading in a rotating tube or drum reduce the fuel to the consistency of fine powder. The efficiency of the mill is almost entirely dependent on the amount of raw coal within the tube.

A set rate of feed is not practicable as the furnace demands are variable and an excess of coal in the mill must be cleared manually. Mechanical gauging proved to be unsatisfactory and best results were obtained with a skilled operator who relied almost entirely on his sense of hearing to judge when to admit more coal.

An acoustic feed controller has been designed to make a direct assessment of the mill contents based on the noise level in the vicinity of the drum. A microphone, adequately protected against dust and dirt, is placed directly beneath the drum. A high-pass filter segregates the significant high-frequency noises emanating mostly from inside the drum from the low-frequency noises associated chiefly with the external gearing and drive of the mill.

The most significant noise frequencies are in the vicinity of 4 kilocycles and their intensity varies over a range of some 14 decibels between the full and empty condition of the tube. If the entire noise spectrum is sampled, the comparable volume range is only about 3 decibels. Small wonder, therefore, that mill feed controlling was so highly skilled an occupation and, incidentally, that the skill diminished when the operator had a cold in his head.

The output from the amplifier and filter actuates a relay that alters the speed of the feed motor and keeps the drum loaded to an optimum value. Alternatively, a suitable gauge may be actuated to afford visual indication of the contents of the mill for manual control of the feed mechanism.

This equipment was designed and installed by Standard Telephones and Cables, Limited, and has proved so completely effective that it has already been scheduled for inclusion in a number of the most modern fuel processing plants.

Anechoic Chamber for Acoustic Measurements

By D. W. ROBINSON

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

ACCURATE acoustic measurements may be made out of doors or in rooms having surfaces that absorb substantially all the sound energy impinging on them. A room of this type has been constructed in accordance with fundamental designs outlined by Beranek and Sleeper. Its structure is described and some results of tests made to prove its performance are given.

• • •

1. *Introduction*

In recent years a number of acoustic rooms have been described in the technical literature. An important contribution to the art was made by Beranek and Sleeper[1] who published data demonstrating the advantageous properties of "fibreglass" wall treatments for obtaining high sound absorption at low audio frequencies. Acoustic rooms at the Acoustics Laboratory of Harvard University and at the Parmly Sound Laboratory in Chicago were among the first to be built to the specifications of Beranek and Sleeper.

The acoustic room described here was erected in the acoustics laboratory of Standard Telephones and Cables at New Southgate, London, during 1947 and 1948; it was designed and the construction supervised by the company's engineering staff. It was desired to provide a chamber which, for its size, was comparable in performance with those mentioned above and to those being erected at the same epoch by other authorities in England. The basic design was accordingly modelled on the same lines as these rooms.

The published data related particularly to fibreglass of one density ($2\frac{1}{2}$ pounds per cubic foot), and it was recognised that the validity of the design information possibly rested on this datum. At the time of construction, fibreglass of that density was not available in England. Accordingly, it was found necessary to use a slightly denser material, namely resin-bonded fibreglass with a density of 3 pounds per cubic foot. Some factors other than the density may also considerably modify the absorption characteristic, for instance the mean diameter of fibres and the bonding medium control the specific resistance, and in these there may have been appreciable differences from the American material. However, the dimensional design criteria of Beranek and Sleeper were utilised, and despite the gaps in our theoretical armoury a very effective acoustic room has successfully resulted.

2. *Functions of the Acoustic Room*

The size of an acoustic room should be as large as cost will allow. A lower limit is set by the functions it is required to perform, in this case problems arising in the following categories:—

A. Development of telephone transmitters and receivers.

B. Maintenance and calibration of standards of free-field sound pressure measurement.

C. Calibration of various types of microphones, loudspeakers, and miscellaneous transducers, but excluding equipment of really large dimensions.

D. Psycho-acoustic investigations mainly relating to speech transmission, under silent or controlled conditions.

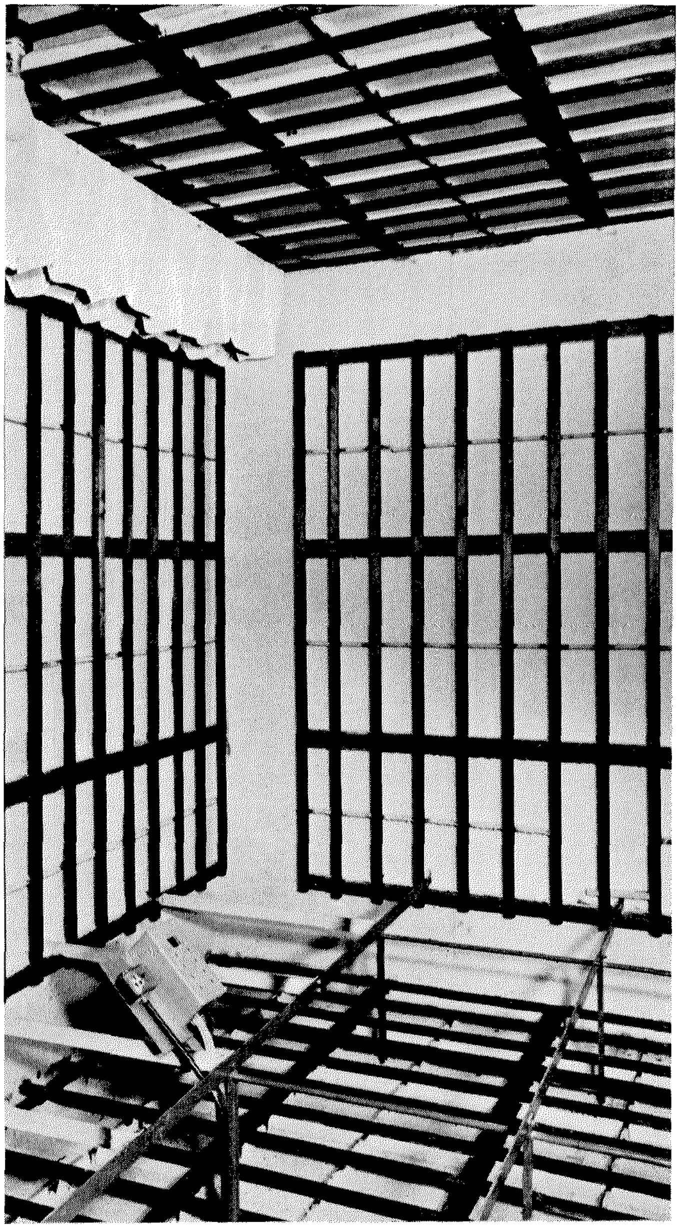
These requirements can be met with a working space of about a cubic metre, provided that it is not too close to any wall. The gross floor area available was about 260 square feet, and the free interior dimensions of the finished room are 12 feet 3 inches by 8 feet 9 inches by 8 feet high (3·7 by 2·7 by 2·4 metres). The laboratory in which it is situated is in a comparatively quiet location free from earthborne vibration. Tests on the transmission of airborne sound using a level noise spectrum showed a reduction of 65 to 70 decibels when the double door was closed. The ambient noise level in the interior is normally below threshold.

* Now with Physics Division, National Physical Laboratory; Teddington, England.

[1] L. L. Beranek and H. P. Sleeper, "The Design and Construction of Anechoic Sound Chambers," *Journal of the Acoustic Society of America*, v. 18, pp. 140–150; July, 1946.

3. Structural Details

In order that the interior floor level of the chamber should coincide with the main laboratory floor, a reinforced concrete pit, 2 feet 6 inches (76 centimetres) deep, tanked with asphalt and slightly larger in plan than the external dimensions of the chamber, was sunk in the main floor. The bottom and sides of this pit were lined with a 2-inch (5-centimetre) thickness of cork insulation and a 4-inch (10-centimetre) concrete loading slab was laid on top of the cork bottom. This loading slab carried the constructional walls of the chamber. It was originally intended to construct these outer walls in cavity brickwork but owing to a shortage of bricks at the time 9-inch-thick (23-centimetre) hollow precast concrete block construction had to be substituted. The roof of the chamber is an 8-inch-thick (20-centimetre) reinforced concrete slab cast *in situ*. The slab bearing on the walls was insulated with 2 inches (5 centimetres) of cork.



As may be seen in Figure 1, the acoustic treatment begins with a layer of Celotex board over the whole of the interior surface. A lattice of wooden battens supports the fibreglass wedges which are arranged with adjacent dihedral angles at right angles. The wedges are 8 inches (20 centimetres) square at the base, and each is built out on a $\frac{3}{4}$-inch (2-centimetre) timber frame. The cross-section of the fibreglass is square for the first $2\frac{3}{4}$ inches (7 centimetres), the remainder is a triangular prism 13 inches (33 centimetres) in length. The direction of the fibres is approximately normal to the walls. With the wedges mounted on the battens an air space of $2\frac{3}{4}$ inches (7 centimetres) is left between the base of the fibreglass and the Celotex lining. This dimension has an important effect on the low-frequency absorption characteristic. The air space is honeycombed in 16-inch (41-centimetre) squares by strips of Celotex to break up possible resonant modes parallel to the long dimensions of the air space.

The wedges are sufficiently short to be self-supporting, and each is enclosed in a muslin bag. The treatment is applied to flat wall, floor, and ceiling surfaces. No attempt has been made to turn the corners by setting wedges at oblique angles. This results in a small area of wall surface that has no acoustic treatment. The proportion so affected is, however, small owing to the comparative shortness of the wedges; no part of the uncovered wall is directly visible from the centre of the room. The shortness of the wedges possibly confers a net advantage, despite the inferior absorption characteristic for normal incidence.

Figure 1—The walls have been covered with Celotex board and a lattice of wooden battens has been erected to support the fibreglass wedges that provide most of the sound absorption.

Long wedges must result either in a much greater proportion of uncovered area, or, if the corners are rounded, in a bunching of wedge tips.

Access to the interior of the chamber is by double door. The outer one has a heavy timber construction in order to have sound reduction comparable with the walls. The inner one is a basket of light tubular construction hinging inwards and continuing the acoustic wall treatment without interruption. The walking floor consists of removable slatted boards resting on a minimum angle-iron framework level with the tips of the floor wedges. A number of hooks in the ceiling support removable iron rails running parallel and diagonally. The other permanent fittings are a lighting socket and cord connection frame taking eight screened and three unscreened pairs to the operating bench outside.

Ventilation has not been specially provided; it has been found adequate to run a fan in the interior when the room is not in use.

4. *Acoustic Properties of Anechoic Chambers*

Before describing the performance of this anechoic room, it is pertinent to consider what is its *prima facie* prospect of achieving a good result, which is to represent faithfully the acoustic conditions in free air. This criterion is the axiom of the art. To satisfy it exactly, the refractive index of the medium beyond the boundary, that is to say, the fibreglass wedges and interstices, would have to be the same as that of free air. For any medium that depends for its absorption properties on porosity and viscous power loss, the phase velocity, and hence the refractive index, necessarily differs from that of air.

It is the practice to study the reflection coefficient of a sample of the wall treatment for normal incidence of a plane wave, and very low values are known to apply to the type of treatment we are considering. The behaviour of the same material for oblique and random incidence does not necessarily follow the same laws. It has been noted by several observers that at low frequencies an anomalous sound pressure distribution sometimes exists in the space between a small source and a wall. It consists of a decline in the root-mean-square value of sound pressure on a line away from the source more rapid than in free air. This is contrary to the intuition that a greater pressure should exist in the presence of a forward obstacle than in a free field, and it serves to illustrate that considerations of plane waves are no necessary guide to the properties of diverging waves. A theoretical investigation of this phenomenon has been made by Rudnick,[2] who confirms it for certain conditions. It has not been observed at New Southgate.

The behaviour at higher frequencies may be explained by the orthodox theory of statistical mechanics. Olson[3] quotes a result for the mean ratio of reverberant-energy density to direct intensity in terms of the reflection coefficient of the boundary, for normal incidence, namely

$$\frac{E_r}{E_d} = 16\pi r^2 . \frac{1-a}{Sa}, \tag{1}$$

in which

E_r = energy density at a point due to reverberant energy
E_d = sound intensity at the same point due to a point source
r = distance of the point from the source
S = surface area of the enclosure
a = energy absorption coefficient of S.

This equation applies to a rectangular room with uniform absorption at the walls. If this is written in the form

$$\frac{E_r}{E_d} = 16\pi r^2 . \frac{\Sigma\Delta S - \Sigma\alpha.\Delta S}{\Sigma\Delta S . \Sigma\alpha.\Delta S}, \tag{2}$$

in which

ΔS = an element of surface area
α = energy absorption coefficient associated with ΔS,

some estimate of the order of magnitude of the ratio may be made; (2) is not rigorous because (1) depends on symmetry of the eigenfunctions. We shall err on the side of over-estimating the reverberant effect in our chamber if the projected area of the walls is taken as 550 square feet (51 square metres) with an energy absorption coefficient of 0·9975 (this corresponds to a

[2] I. Rudnick, "The Propagation of an Acoustic Wave Along an Absorbing Boundary" (Summary) *Journal of the Acoustic Society of America*, v. 18, p. 245; July, 1946.

[3] H. F. Olson, Elements of Acoustical Engineering, Van Nostrand, New York; 1940; p. 305.

pressure reflection of 5 per cent, which is less optimistic than Beranek's best figure for fibreglass wedges at mid-frequencies), and include a term of 38 square feet (3·5 square metres) at zero absorption coefficient to represent the total exposed surface area of all iron work in the room. The floorboards are supposed absent. With r equal to one metre, these data give a reverberant-energy density 12 decibels below the direct-sound intensity. The reverberation is in random phase, and so we may expect irregularities of the order $\pm 0{\cdot}7$ decibel about the mean line in the curves relating pressure with distance from a point source. This must certainly be an upper limit to the error at frequencies for which the statistical equation is valid, since it is evident that the angle iron of the floor framework, which is the dominating factor in the calculation, cannot all be equally effective in reflecting sound; it is half buried in the floor treatment and in any case exposes only about one sixteenth of the allotted area to normally incident waves. In practice, on the other hand, one does not have a point source or a point probe, and contributions to observed error from the finite size of these and their mountings are at least as likely as from wall reflection. We conclude that errors exceeding the order of magnitude $\pm 0{\cdot}7$ decibel will need other explanations.

5. *Criteria of Comparison of Anechoic Rooms*

The performance of anechoic chambers is most satisfactorily described in relation to certain rather artificial conditions. It is not practicable to compare the radiation fields of practical sources in the chamber and in the open air because of the complexity and the difficulty of the outdoor measurement at a sufficient height above ground.

Various methods suggest themselves for assessing the extent of the departure from free air conditions. A property characterising an unrestricted medium is the propagation of spherical waves from a point source. The loosely termed inverse-square-law test consists of measuring the root-mean-square sound pressure as a function of distance along a radial line from the source, which may be placed anywhere in the enclosure. For this test, one needs an effective point source, a probe microphone, and a means of varying and accurately determining the position of the latter. The position of the source can usefully be restricted to the position of practical sources in the normal use of the room, and the radial line should cover the prospective working space, that is, not much more than a metre in appropriate directions. The independent variable in this method of test is most conveniently taken to be the distance; measurements are made at a series of fixed frequencies. This method of investigation has been found to be by far the most sensitive and satisfactory.

Among other possible methods of approach is the procedure in which an arbitrary receiver–microphone system is moved about in the enclosure and the overall response between electrical terminals is observed. If the acoustic room is effective, the transducer system is indifferent to the position of the boundary relative to itself and indicates a constant overall response for all positions in the chamber. The *modus operandi* is to put the transducer system at a number of arbitrary positions and measure the overall response–frequency characteristic. In this way, it is claimed that the performance of the chamber is presented as a continuous function of frequency. In practice, it has been found inferior to the inverse-law test for two main reasons. To set up a receiver and microphone in constant relative position by independent mountings for direct mensuration may be ruled out owing to the fact that the sound field is sensitive to small changes in the position of clamps, hooks, etc., thus obscuring the effect under observation. A rigid connecting framework is necessary and this suffers from a similar disadvantage; the receiver and microphone must be sufficiently far apart to bring the reverberant response into relative prominence. This usually means having a robust framework, which introduces irregularities in the total response due to local reflections. These local reflections obscure the effect of wall reflections, the more so the more effective the chamber. These objections may be overcome to a large extent if the transducers are directional and face away from each other, so increasing the relative importance of the reflected term.

No suitably directional sound sources or microphones were available at the time the room was tested, but some tests were made by the method just described using an omnidirectional receiver and microphone. The results added little to the data obtained from the inverse-law tests.

The second principal objection is that the observed effects are less amenable to interpretation. The difficulty created by response irregularities due to reflection from the rigid mounting can of course be circumvented by reverting to the use of fixed frequencies, with distance as the independent variable. There is then no advantage in the method. Using directional transducers, a direct indication of 1-percent pressure reflection at the boundary requires a back-to-front discrimination ratio of -50 decibels at all frequencies, a stringent requirement.

6. *Performance of Room*

The tests to be described were based on the inverse law of pressure against distance as described above, for which a good point source and probe are necessary to ensure sharply localised measurement and spherical symmetry. It has been found that the closest attention must be devoted to these items and to their mountings to avoid experimental errors exceeding the effect under observation. This was the point of Section 4, in which some estimate of the effect attributable to the room, as opposed to the measuring technique, was made in advance.

A high degree of resolution was aimed at, to show the fine structure of standing-wave patterns when they occurred. The frequencies of interest ranged up to 10 kilocycles per second. Increments of distance of 0·5 centimetre were required at the highest frequencies, increasing to 10 centimetres at the lowest (125 cycles per second) frequency. The positional accuracy of the microphone, which was suspended on a trolley of steel strips from the diagonal ceiling rail and manipulated from outside by a steel wire and balance weight, was about 0·1 centimetre. The probe microphone was of the type with a tube about 20 centimetres long by 1·5 centimetres bore. It was variously arranged, with or without a small-bore probe tip 3 inches (8 centimetres) in length, and parallel or transverse to the line of the sound source. An alternative type with a probe tube of smaller diameter (0·35 centimetre, length 40 centimetres) was occasionally used but was somewhat restricted in application owing to its inherent insensitivity combined with the lower power of the diminutive sound source, which was essential to the success of the experiments. Small corrections were made at the highest frequencies when the 1.5-centimetre probe was used transversely, to allow for reduction of the observed amplitude of stationary-wave peaks due to the phase difference across it.

The source of sound presented the major problem. An electrodynamic loudspeaker with undamped 2-centimetre-diameter throat was rejected, except for some low-frequency measurements, owing to strong reflections from the face and other anomalous properties. An open-ended tube connected to it was satisfactory at low frequencies. The speaker unit was concealed between two wedges. At higher frequencies, the radiated power was only sufficient when the tube was resonant, but facilities for the adjustment of its length were not available.

Satisfactory measurements at 125 and 250 cycles per second were obtained by the aid of the hornless loudspeaker unit and the aperiodic tube. From 375 cycles per second upwards a point source was approximated by a miniature electromagnetic receiver, the overall dimensions of which are 2·75 centimetres in diameter by 1·5 centimetres thick. At frequencies below about 6 kilocycles per second this unit could certainly be considered as a point, and no account need be taken of its orientation, or of whether sound emanated from part or whole of the surface. At the higher frequencies there was evidence of a virtual source at a distance up to 0·5 centimetre in front of the emitting face, and care was needed in the axial alignment of the receiver and microphone. The existence of a focus was demonstrated by plotting reciprocal sound pressure at the probe tip against distance from the apparent or physical source; the relation was linear with positive intercept on the axis of distance. The electromagnetic receiver was suspended on twine using its shielded input cord as a guy rope. The greatest sound pressure at 1 metre distance delivered by this receiver was -15 decibels relative to 1 dyne per square centimetre at 500 cycles per second, 0 decibels at 9000 cycles per second, with a maximum of $+15$ decibels at 2500 cycles per second.

All the measurements were taken over the range from 10 to 100 centimetres between source

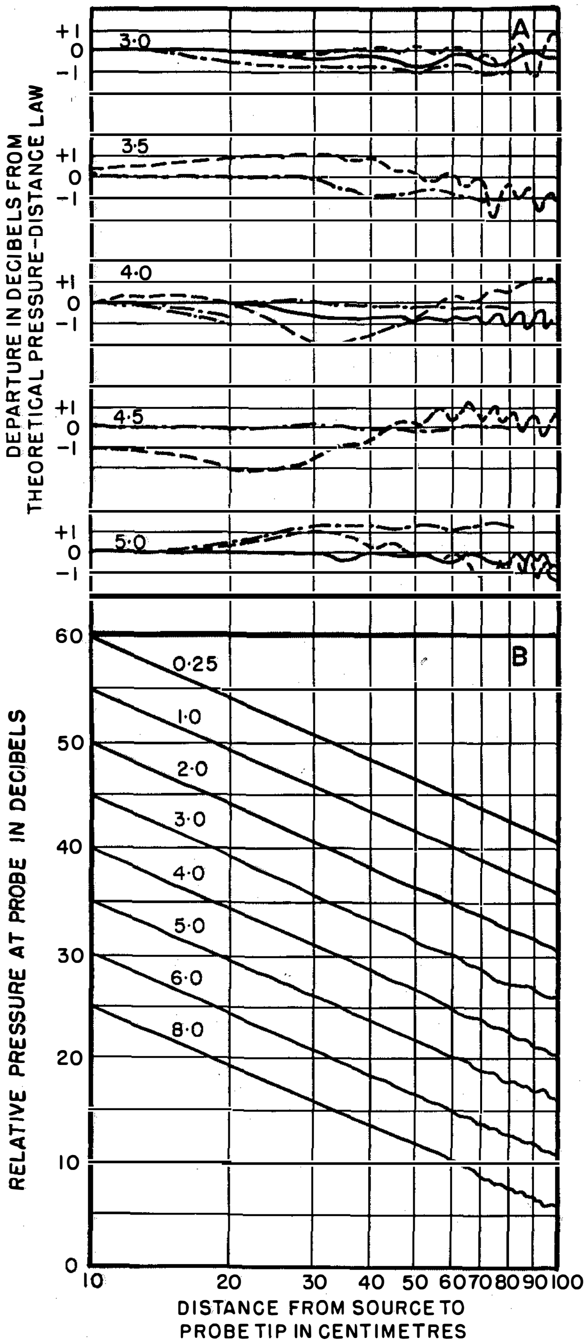


Figure 2—Departures from the theoretical relation of pressure at the probe tip being proportional to the inverse distance to the sound source are plotted against distance between the probe and source for the indicated frequencies in kilocycles per second in *A*. The dotted lines are with the chamber empty of superfluous fittings. The final results with all reflecting surfaces covered are given in solid lines and some intermediate conditions are indicated by the chain lines. The lower set of curves (*B*) correspond to the solid-line graphs of *A* at the frequencies indicated.

and probe tip, most of them along a horizontal diagonal of the room. Owing to the necessity of frequently overdriving the small source, the circuit of the probe microphone included a variable band-pass filter, and linearity checks were made over the range between the receiver overload point and the microphone-circuit noise level. This was not less than 20 decibels. The stability and calibration of the measuring circuit was better than 0·1 decibel.

An initial series of measurements at 25 frequencies from 125 to 9000 cycles per second gave some anomalous results in the region of 3 to 5 kilocycles per second, some extremely close approaches to the inverse law below 2 kilocycles per second, and evidence of stationary waves of small amplitude at higher frequencies amounting to less than ± 1 decibel about a mean line, which was itself in very close agreement with the inverse law. The anomalous result was in the nature of a total departure of the experimental results from point-source theory over a certain distance range (about 20 to 60 centimetres), amounting to perhaps 2 decibels at most. The departure was above or below the theoretical curve according to frequency, with which it varied critically. The cause was found to be associated with the mounting of the microphone (the moving trolley), and to some extent the permanent fixtures in the room, although these had already been reduced to apparent insignificance. The floorboards were naturally removed throughout these proceedings.

It was discovered that by modifying the mechanical set-up very slightly the anomalous results could be shifted in frequency and in position along the scale of distance as shown in Figure 2A. A second series of measurements was made in the same way as the first after every possible reduction of exposed reflecting area had been made. Figure 2B shows a few typical results from this series. That is, such permanent fittings as were not already covered were wrapped in cotton-wool, the moving trolley was simplified and similarly treated; the probe microphone was also wrapped in absorbent material. The effect of these minor adjustments, which were made progressively in a number of steps, was both to shift the frequency at which the abnormal effect was most pronounced and to reduce the magnitude of the departure from the theoretical to about 0·7 decibel. There was no appreciable change in the pattern of the curves previously ascribed to stationary waves; in fact, the fine structure of the pressure distribution was very closely repeated.

In Figure 2A the dotted lines show the departure from the inverse law in the initial experiments at the frequencies where the effect was important. The full lines show the departures in the improved set-up and the chain lines represent various intermediate conditions. It will be seen that the best agreement is not obtained with the same set-up at each of the frequencies illustrated, but that the final set-up is generally an improvement on the initial.

At this stage it was possible to segregate effects proper to the anechoic chamber from those due to adventitious causes, and the conclusions are summed up as follows. The central space of the room for a distance of one metre from a given point is equivalent to free air within these limits:—

A. ±0·2 decibel from 125 to 1500 cycles per second. There was no reliable data for lower frequencies.

B. ±0·5 decibel from 2 to 10 kilocycles per second. The observed departures from free-air conditions are consistent with a regime of stationary waves in the direction joining the point of observation to the source.

It appears surprising that any regular system of stationary waves should reveal itself, in view of the multiplicity of possible reflection paths contributing to the total pressure at the probe microphone. There is little scope for misinterpretation: each experimental curve of sound pressure against distance exhibits periodic fluctuations about the theoretical mean curve, the amplitude of which increases with distance from the source, and the period of which corresponds exactly to the wavelength of sound in air. If the oscillations were due to reflections from the microphone and source, the amplitude should not vary with distance in this manner.

7. *Axial Radiation of Some Practical Sound Sources*

In the course of the experimental work described, a variety of sources had been used in addition to the types already mentioned. The observed axial pressure distributions were in certain cases very irregular and illustrate the unsuitability of these conventional radiators as sound sources in acoustic experiments in which microphones of high resolution are involved. The irregularities greatly exceeded the effect of room reflections.

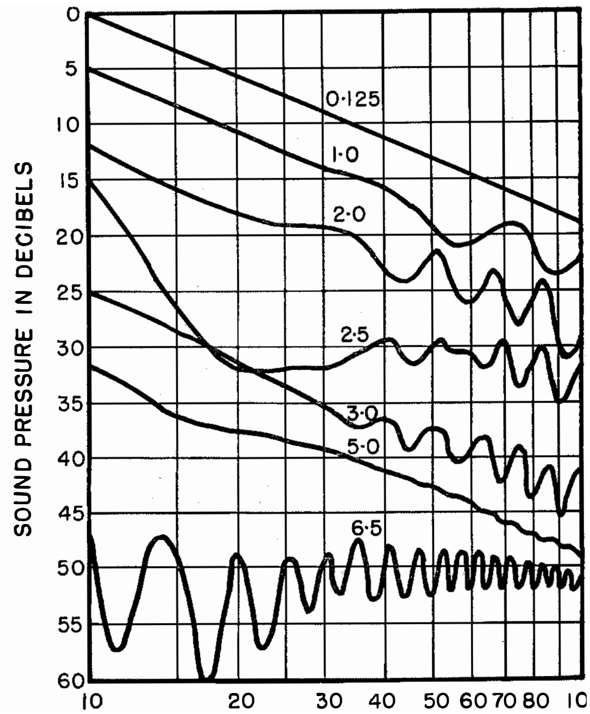


Figure 3—Axial pressure distribution of *RR-101-C* loudspeaker unit with undamped throat for the frequencies indicated in kilocycles per second.

An 8-inch (20-centimetre) cone-type loudspeaker in a 36-inch (91-centimetre) square baffle gave a normal axial distribution in various frequency ranges from 1 to 10 kilocycles, from as short a distance as 30 centimetres, frequently accompanied by fluctuations about the theoretical curve of 2 or 3 decibels, correlated with wavelength. The curves for 62·5 and 125 cycles per second fell too steeply; that for 250 fell too slowly. Curves for 500, 1500, 6000, and 8500 are entirely anomalous, exhibiting neither periodic fluctuations nor even general adherence to the theoretical curves. The theoretical curve for this loudspeaker was taken to be the axial pressure of a flat-piston radiator in a baffle; this approximation, though still somewhat crude, improved the agreement. At certain frequencies the agreement was improved by shifting the origin of distance, indicating focussing of the source not indicated by simple flat-piston theory. The shift, of up to 5 centimetres, was sometimes in one direction and sometimes in the other.

The permanent-magnet speaker without horn, already mentioned as having been used in the crucial experiments at low frequencies, had some remarkable properties at higher frequencies as may be seen in Figure 3. At 1000 and 2000 cycles per second the oscillations of the curve occur at wavelength intervals. The extraordinary departure of 12 decibels at 2·5 kilocycles per second and 20 centimetres was confirmed in a further series of measurements. The pattern of oscillations about the theoretical curve for 6·5 kilocycles per second decreases with the distance. The interpretation of this is that reflections are set up from the face of the source, which is a metal plate having a diameter of 6 inches (15 centimetres). Curves for frequencies in the intervals between those illustrated were comparatively normal.

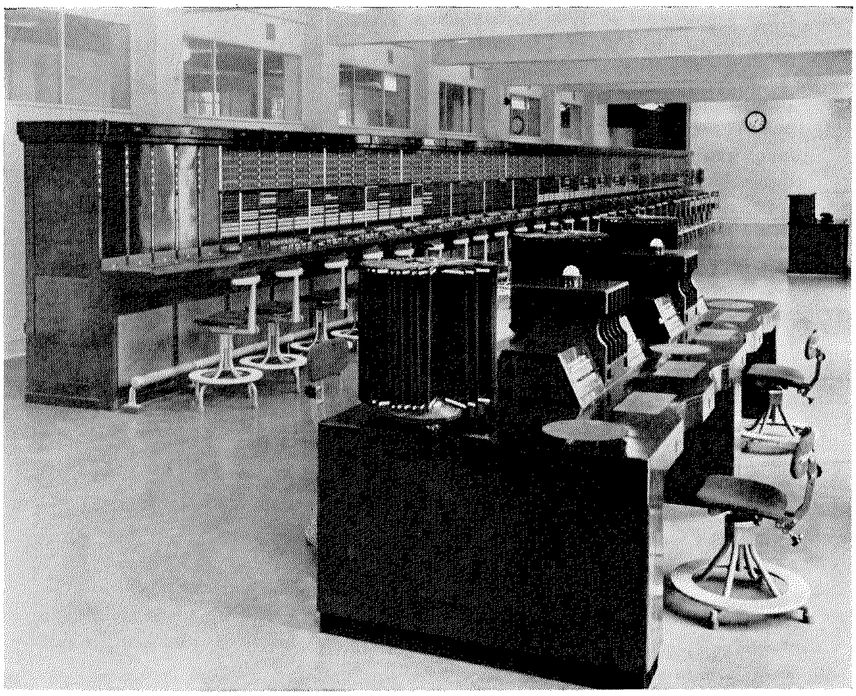
The magnitude of the striking departures from simple theory described and illustrated is to be viewed in relation to the magnitude of wall reflections in the anechoic room itself. The few results quoted serve to show the difficulty of setting up a locally uniform sound field and that great care must be taken in selecting the sound source appropriate to a particular experimental purpose.

Recent Telecommunication Development

Largest Private Telephone Exchange in Great Britain

THE LARGEST private telephone exchange in Great Britain serves the new government offices being erected in London on Victoria Embankment opposite to the site of the Festival of Britain to be held from May 3 to September 30.

This fully automatic exchange provides for 3000 internal extensions with 33 manual positions. Provision is made for 400 tie lines to allow direct dialling to other government offices in the London area, with special equipment for long-distance working.

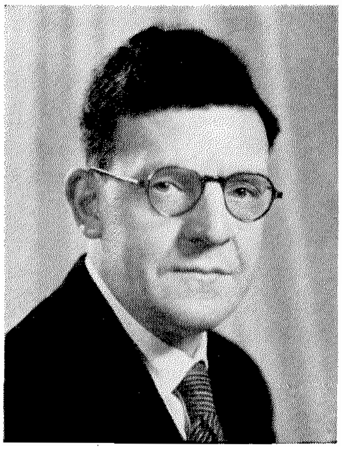


Although this new exchange has access to the public telephone network through Trafalgar exchange, all trunk and toll calls are controlled and timed by its own operators.

A unique feature is the specially designed directory and enquiry tables with their revolving drum files that are capable of containing some 12 000 records.

Standard Telephones and Cables, Limited, was responsible for the manufacture and installation of the equipment.

Contributors to This Issue



P. R. BARDELL

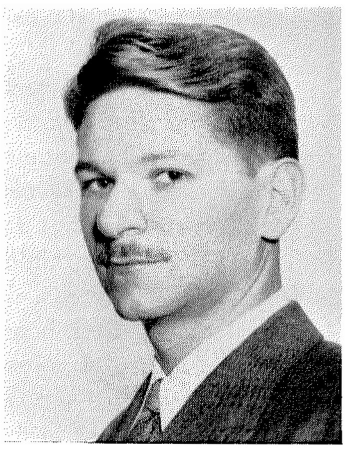
P. R. BARDELL was born at Hendon, England, on December 11, 1909. He received the B.Sc. degree from London University in 1932.

In 1930, he joined the research laboratories of General Electric Company as a student assistant and since 1932 has been a member of the scientific staff. Now in the telecommunications group, his work has primarily concerned the development and application of magnetic materials.

Mr. Bardell is an Associate Member of the Institution of Electrical Engineers and an Associate of the Institute of Physics.

• • •

E. M. BRADBURD was born in 1920 at Philadelphia, Pennsylvania. He received the B.S. and M.S. degrees in electrical engineering from Columbia University in 1941 and 1943, respectively.



E. M. BRADBURD

From 1941 to 1943, he served as an assistant in the electrical engineering department and in the Division of War Contracts at Columbia University.

In 1943, he joined the engineering staff of Federal Telecommunication Laboratories, where he has been engaged in work on aerial navigational aids and on television transmitters.

Mr. Bradburd received the Illig Medal for scholarship from Columbia University. He is a Member of the Institute of Radio Engineers and of Tau Beta Pi.

• • •

S. E. BUCKLEY

S. E. BUCKLEY was born at London on November 28, 1904. He received the B.Sc. degree from London University in 1924.

He joined the European engineering department of International Standard Electric Corporation in January, 1926, as a student engineer, and was transferred to the development and research laboratories at Hendon in 1929. He was again transferred in 1932 to Standard Telephones and Cables, Limited.

Associated with the development of Permalloy-type materials since 1927, he is at present responsible for development and engineering of magnetic materials including dust cores, loading coils, and telephone-cable accessories. Mr. Buckley is in charge of the chemical and metallurgical laboratories at the Woolwich plant.

He is an Associate Member of the Institution of Electrical Engineers.

• • •

KURT KONSTANTINOWSKY was born at Vienna in 1892. He received the Ph.D. and D.Sc. degrees from Vienna University and lectured there on physics.

From 1922 to 1939, he was associated With the Cable Manufacturing Company at Bratislava, serving finally as technical director.

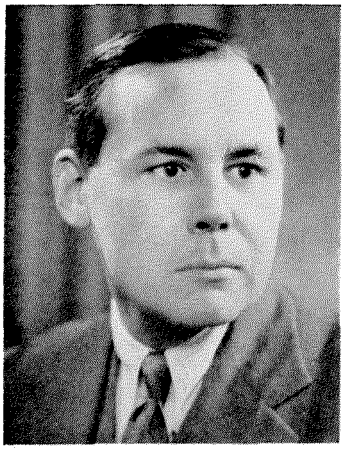
Dr. Konstantinowsky joined Standard Telephones and Cables, Limited, in 1940 and has been concerned with the design and production of rubber- and plastic-insulated cables and with research and development of power cables. At present, he is in charge of power-line engineering and development. He has done research work on the physics of submicroscopic particles, screened-type high-voltage cables, and dielectric losses at high voltages.

He is a Fellow of the Institute of Physics and a Member of the Institution of Electrical Engineers.

• • •

EDWIN C. LEE was born at London in 1907. He attended Whitgift School and Erith Technical College.

KURT KONSTANTINOWSKY



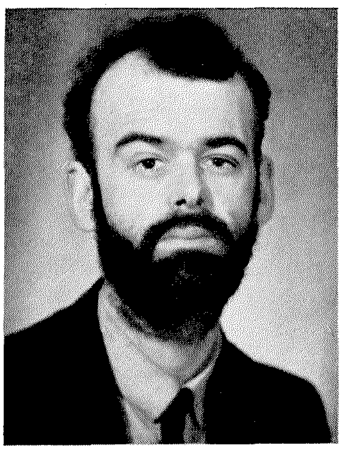
Edwin C. Lee

After experience in the manufacture of power cables, he joined the power cable division of Standard Telephones and Cables, Limited, in 1937. During the war, he was engaged in engineering and development of capacitors, loading cables, pulse transformers, molded polyethylene components, and other special products. Since 1945, Mr. Lee has been in charge of power cable engineering and development.

• • •

A. C. Lynch was born in 1914 at London, England. He received B.A. and M.A. degrees in 1935 and 1939, respectively, from Emmanuel College, Cambridge.

In 1936, he joined the staff of the Post Office engineering research station. His work has included studies of the electrical properties of insulating materials, powder cores, piezoelectric crystals, and the stability of oscillators.

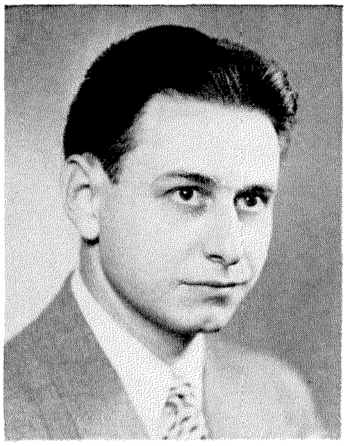


A. C. Lynch

• • •

Richard H. Myers was born on May 30, 1917, at West Pullman, Illinois. He received the B.S. degree in electrical engineering from Massachusetts Institute of Technology in 1939.

On leaving school, he joined the Federal Telegraph Company, becoming a project engineer in 1945. He has been active in the design of equipment for special communications for the armed forces, for industrial heating, frequency-modulation broadcasting, and for both high- and low-frequency loran.



Richard H. Myers

Mr. Myers is an Associate Member of the Institute of Radio Engineers.

• • •

C. E. Richards was born at Liverpool, England, in 1900. He studied at Liverpool Central Technical School and Sir John Cass Technical Institute.

He joined the Post Office in 1922, transferring to the research branch the following year. He is now in charge of the materials division of the research station at Dollis Hill, London. His work on materials has included the corrosion of cables and line plant, plastics, detection of dangerous gases, growth of piezoelectric crystals, and development of magnetic materials.

• • •

C. E. Richards

D. W. Robinson was born at London in 1920. He attended Imperial College of Science and Technology on a scholarship and received an honours degree in engineering in 1939.

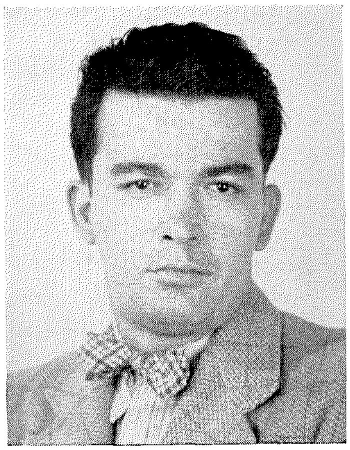
After a brief period with the Western Electric Company, he joined the Royal Air Force Volunteer Reserve, in which he served for five years.

In 1946, he entered the apparatus division of Standard Telephones and Cables, Limited, where he has been engaged on the development of telephone transmission-standard equipment and on other problems in applied acoustics. Recently, he has joined the physics division of the National Physical Laboratory at Teddington, England.

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

• • •

D. W. Robinson



William Sichak

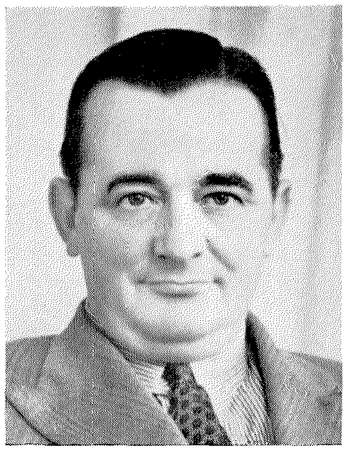
William Sichak was born on January 7, 1916, at Lyndora, Pennsylvania. He received the B.A. degree in physics from Allegheny College in 1942.

From May, 1942, to November, 1945, he was engaged in developing microwave radar antennas at the Radiation Laboratory, Massachusetts Institute of Technology. Since November, 1945, he has been with Federal Telecommunication Laboratories, working on microwave antennas and allied subjects.

• • •

C. Gordon White was born at Dublin, Eire. He received the Diploma in Engineering (A.R.C.Sc.I.) of the Royal College of Science for Ireland in 1919. He continued at the college in tutorial duties until 1923. In 1920, he obtained the B.Sc. (Eng.) degree of London University with Honours.

In 1924, Mr. White joined the Western Electric Company in London. After a year in the telephone circuit department at North Woolwich, he transferred to the circuit laboratory. In 1930, he



C. Gordon White

joined the remote-control department of Standard Telephones and Cables, Limited, where he has been engaged in technical sales, design, and planning of supervisory control and telemetering systems.

INTERNATIONAL TELEPHONE AND TELEGRAPH CORPORATION

Associate Manufacturing and Sales Companies

United States of America

International Standard Electric Corporation, New York, New York

Federal Telephone and Radio Corporation, Clifton, New Jersey

International Standard Trading Corporation, New York, New York

Capehart-Farnsworth Corporation, Fort Wayne, Indiana

Great Britain and Dominions

Standard Telephones and Cables, Limited, London, England

Creed and Company, Limited, Croydon, England

International Marine Radio Company Limited, Croydon, England

Kolster-Brandes Limited, Sidcup, England

Standard Telephones and Cables Pty. Limited, Sydney, Australia

Silovac Electrical Products Pty. Limited, Sydney, Austral a

Austral Standard Cables Pty. Limited, Melbourne, Australia

New Zealand Electric Totalisators Limited, Wel ington, New Zealand

Federal Electric Manufacturing Company, Ltd., Montreal, Canada

South America

Compañía Standard Electric Argentina, Sociedad Anónima, Industrial y Comercial, Buenos Aires, Argentina

Standard Electrica, S.A., Rio de Janeiro, Brazil

Compañía Standard Electric, S.A.C , Santiago, Chile

Europe and Far East

Vereinigte Telephon- und Telegraphenfabriks Aktiengesellschaft Cze ja, Nissl & Co., Vienna, Austria

Bell Telephone Manufacturing Company, Antwerp, Belgium

China Electric Co pany, Limited, Shanghai, China

Standard Electric Aktieselskab, Copenhagen, Denmark

Compagnie Générale de Constructions Téléphoniques, Paris, France

Le Matériel Téléphonique, Paris, France

Les Téléimprimeurs, Paris, France

C. Lorenz, A.G. and Subsidiaries, Stuttgart, Germany

Mix & Genest Aktiengesellschaft and Subsidiaries, Stuttgart, Germany

Süddeutsche Apparatefabrik Gesellschaft m.b.H., Nuremberg, Germany

Nederlandsche Standard Electric Maatschappij N.V., The Hague, Netherlands

Fabbrica Apparecchiature per Comunicazioni Elettriche, Milan, Italy

Standard Telefon og Kabelfabrik A/S, Oslo, Norway

Standard Eléctrica, Lisbon, Portugal

Compañía Radio Aérea Marítima Española, Madrid, Spain

Standard Eléctrica, S.A., Madrid, Spain

Aktiebolaget Standard Radiofabrik, Stockholm, Sweden

Standard Telephone et Radio S.A., Zurich, Switzerland

Telephone Operating Systems

Compañía Telefónica Argentina, Buenos Aires, Argentina

Compañía Telegráfico-Telefónica Comercial, Buenos Aires, Argentina

Compañía Telegráfico-Telefónica d l Plata, Buenos Aires, Argentina

Companhia Telefônica Nacional, Porto Alegre, Brazil

Compañía de Tel fonos de Chile, Santiago, Chile

Compañia Telefónica de Magallanes S.A., Punta Arenas, Chile

Cuban American Telephone and Telegraph Company, Havana, Cuba

Cuban Telephone Company, Havana, Cuba

Compañía Peruana de Teléfonos Limitada, Lima, Peru

Porto Rico Telephone Company, San Juan, Puerto Rico

Shanghai Telephone Company, Federal Inc. U.S.A., Shanghai, China

Radiotelephone and Radiotelegraph Operating Companies

Compañía Internacional de Radio, Buenos Aires, Argentina

Compañía Internacional de Radio Boliviana, La Paz, Bolivia

Companhia Radio Internacional do Brasil, Rio de Janeiro, Brazil

Compañía Internacional de Radio, S.A., Santiago, Chile

Radio Corporation of Cuba, Havana, Cuba

Radio Corporation of Porto Rico, San Juan, Puerto Rico

Cable and Radiotelegraph Operating Companies

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

The Commercial Cable Company, New York, New York[1]

Mackay Radio and Telegraph Company, New York, New York[2]

All America Cables and Radio, Inc., New York, New York[3]

Sociedad Anónima Radio Argentina, Buenos Aires, Argentina[4]

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

Laboratories

Federal Telecommunication Laboratories, Inc., Nutley, New Jersey

International Telecommunication Laboratories, Inc., New York, New York

Laboratoire Central de Télécom unications, Paris, France

Standard Telecommunication Laboratories, Limited, London, England