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Meetings at which technical papers are presented are held in the twenty-two cities in the United States and Canada listed on the inside front cover of this issue. A number of special meetings are held annually and include one in Washington, D. C., in co-operation with the American Section of the International Scientific Radio Union (U.R.S.I.) in April, which is devoted to the general problems of wave propagation and measurement technique, the Rochester Fall Meeting in co-operation with the Radio Manufacturers Association in November, which is devoted chiefly to the problems of broadcast-receiver design, and the Annual Convention, the location and date of which are not fixed.

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Synthetic Reverberation*

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Summary—An electrooptical method for producing reverberation synthetically is described together with a summary of the development work and the various models which have been built.

The basic principle consists in recording a fugitive sound pattern of the original program signal on the rim of a rotating phosphor-coated disk by means of a modulated light source and a simple optical system. The signal is picked up from the disk at later points through another simplified optical system and photocells. The logarithmical decay of the sound images on the phosphor as they pass the phototubes gives the required reverberation effect. This secondary signal is then mixed with the original program signal in any desired proportion.

The method of modulating a high-pressure mercury-vapor lamp as an integral part of the aforesaid development work is described, and also the method of modulating a high-pressure mercury-vapor lamp with audio frequencies, together with a simplified optical system of high efficiency.

An Electrooptical System for Controlling the Reverberation of Sound Signals

Discussions of the requirements for new studio facilities for sound broadcasting brought out the fact that it is desirable to have a method of adding artificial reverberation to certain types of programs. This also applies to motion-picture sound stages, television sound channels, and all types of recording studios. The reverberation time in such studios and auditoriums, after they are built, can be controlled only to a limited extent by the arrangement of draperies and furnishings.

However, there are many cases where a more-pronounced change is desirable. If, therefore, a practical device for producing reverberation synthetically were available, it would be advisable to build studios and auditoriums with a lower reverberation time than is normally required. Reverberation could then be added artifically to produce the desired effect.

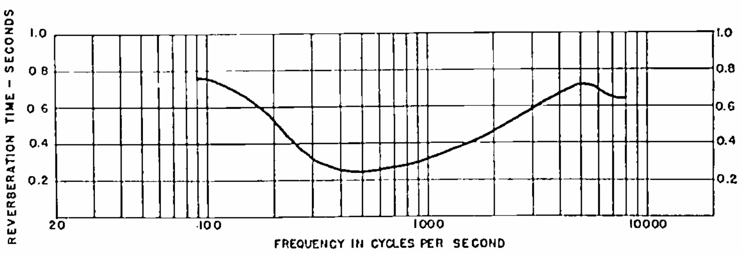

Fig. 1—Typical studio frequency-reverberation characteristic.
Length 34 feet Volume 9325 cubic feet
Width 21 feet Surface area 2918 square feet
Height 12 feet Treated area 1000 square feet

In addition to adjusting reverberation characteristics artificially there is the attractive possibility of adding brilliance to certain types of programs, besides producing the effect of a large auditorium when an orchestra or other large group must perform in the limited space of a regular studio.

The device to be described produces such artificial reverberation. The principle employed is based on the fact that the decay characteristic of phosphorescent substances excited by light or electronic bombardment is approximately logarithmic, similar to the decay of reverberant sound. This phenomenon is made use of by having the desired signal modulate a light source which is recorded through a suitable optical system on the rim of a phosphor-coated rotating disk. This fugitive signal is then picked up a number of times with decreasing amplitude on successive revolutions. By a proper choice of the size of the disk and its speed, the number of pickup tubes and their location together with a phosphor having an appropriate decay characteristic, it is possible to produce a large number of reverberation effects.

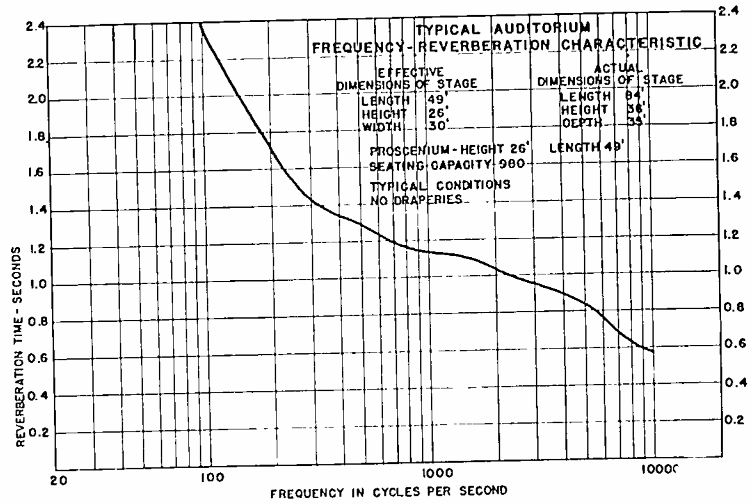

Fig. 2

Nature of Reverberation

Before going any further it might be well to look briefly into the nature of reverberation, which may be defined as the persistence of sound due to repeated reflections.

The phenomenon of reverberation is so common in everyday life that when familiar sounds are produced without it they may sound unnatural. Audiences have long been accustomed to hearing symphony concerts and soloists in auditoriums with considerable reverberation. If, therefore, a symphony orchestra should perform in a studio which was just large enough to accommodate the players with their instruments but relatively small compared to a concert hall, the result would be quite unnatural because of the dissimilarity of the reverberation characteristics.

A single echo is seldom heard except when reflected from a large surface such as a cliff or mountain at a distance. Reverberation indoors has a very complex sound structure because of the multiple reflections from many surfaces having different absorption coefficients and being at different distances.

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Experience has shown the approximate reverberation times which are desirable for typical studios and auditoriums. Fig. 1 shows a reverberation curve (reverberation time plotted against frequency) of a typical studio and Fig. 2 the reverberation characteristic of a typical auditorium.

Equipment has been developed with which it is possible to measure rapidly and accurately the rever-

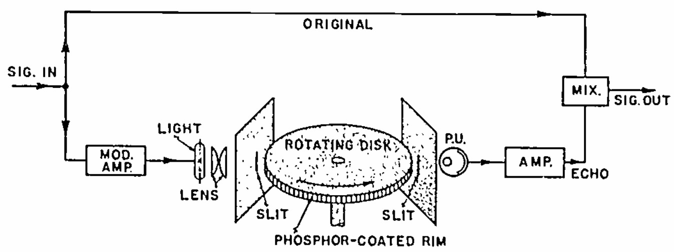

Fig. 3—Basic schematic diagram.

beration time of a studio or auditorium.[1] Measurements made with such equipment on many studios and auditoriums serve to show what reverberation time is most suitable for a given purpose. Reverberation time is defined as the time for a given sound to decay to an intensity of one millionth (or 60 decibels) of the original signal.

The device to be described, of which two different designs have been built and operated, provides a reverberation time of over 2.5 seconds, which is probably more than would be desired at any time in actual use. The artificial reverberation, once produced, is then mixed electrically with the original signal in the proper proportion to produce the desired effect.

It might be pointed out here for those not familiar with the nature of reverberation that any scheme which simply introduces a small time delay will not produce reverberation but only a single echo. In order to simulate reverberation it is necessary to have the echo repeated many times, perhaps 40 or more, with decreasing amplitude. The successive echoes must be frequent enough so that the individual impulses will not be noticeable.

Method Employed

Fig. 3 is a simple schematic diagram of the electro-optical system to be described and the manner in which it is used.

Because of the low luminous efficiency of phosphors, it was evident from the beginning that a powerful light source would be required. This eliminated the ordinary low-pressure ionized-gas lamp such as the neon and similar types. A search for something more powerful led to the newly developed mercury-pressure capillary-type lamp. First attempts at modulating the lamp were made by operating it with sound-modulated radio frequency. This scheme worked quite well but it was soon found that the lamp could be modulated just as well by operating it on direct current and modulating this as if it were supplying a radio-frequency generator tube. It was found that the lamp can be modulated to substantially 100 per cent, however, with some difficulties which will be brought out later.

With a powerful modulated light source focused through an $f=1:2$ cylindrical quartz lens and a slit onto a phosphor-coated disk, attempts were made to pick up a delayed signal from the disk through a slit similar to that at the modulating source. It was then focused onto a sensitive gas-type phototube by means of another $f=1:2$ lens.

The signal available, if any, was below the noise level of either the photocell or its coupling resistor. This was rather discouraging and success was not achieved until it was realized that the definition of the image projected on the disk need not be very sharp and that, therefore, the losses in the lenses might be avoided. The image from the disk was then transmitted to the cathode of the photocell through a slit acting as a lens, in the manner of a "pinhole" camera. The signal-to-noise ratio and frequency response were not very good at first and there were many other problems to be considered before satisfactory operation was achieved. The same principle of using a slit instead of a lens was later applied to the mercury-pressure lamp when projecting the modulation onto the disk. The dimensions of the slits and the spacings of the lamp, photocells, disk, and slits were arranged so as to take maximum advantage of the light available, as shown in the diagram, Fig. 4.

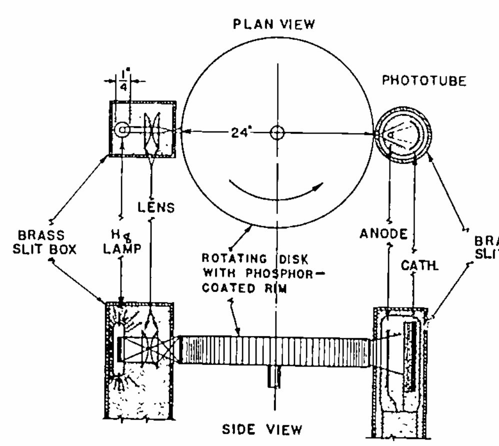

Fig. 4—Lamp, disk, and phototube optical system.

Phosphorescent Material

An important consideration was that of finding the most-suitable phosphor, taking into account the fact that a considerable portion of the light from the mercury-vapor high-pressure lamp is in the blue and ultraviolet region and that the maximum sensitivity of the most sensitive type of photoelectric cell is in the red end of the spectrum. Fortunately phosphorescent materials generally reradiate energy at a longer wavelength than that of the exciting source. The best compromise between decay time and light output was obtained with a material having a rather

[1] H. A. Chinn and V. N. James, "Apparatus for acoustic and audio measurements," *Jour. A.S.A.*, vol. 10, pp. 239–245; January, (1939).

slow decay time (several seconds) and giving a color in the yellow-orange region. Later, as the setup was improved, the light output was better utilized so that the choice of phosphor for a specific decay time was possible.

Lamp Operation and Modulation

The mercury-vapor pressure lamp selected is one of a type which is rated at 85 watts and forms an arc about 3/4 inch long and 1/16 inch in diameter, within a quartz tube about 1/4 inch in diameter. It was originally designed to operate from ordinary alternating-current lighting circuits with a reactive transformer. This has a no-load potential of about 450 volts (as required to start the lamp) which drops to about 20 volts after the lamp has ignited. The potential across the lamp gradually increases as it heats up; the pressure within the lamp increases to 20 to 30 atmospheres and the potential rises to about 250 volts after a few minutes. When the lamp is extinguished after it has become heated, it cannot be started again until it has cooled off.

When modulation was applied to the lamp while operating it near the rated input, using either radio frequency or direct current, it was likely to go out on any overmodulation peak, especially at low frequencies. This had to be remedied and several schemes were tried before a satisfactory operating condition was established.

Attempts were made to reduce somewhat the input power, but it was evident that the operating

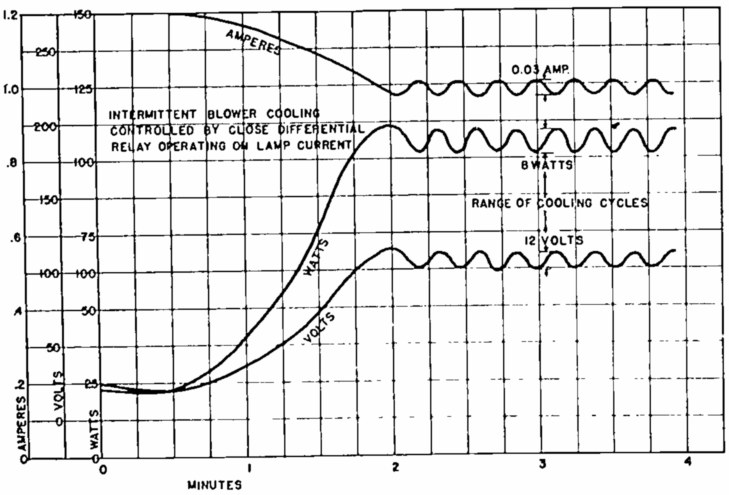

Fig. 5

voltage and current of the lamp were still rather critical. If the input was too low, the arc did not maintain itself properly and the intensity dropped to a value that provided insufficient light.

A scheme which proved to be fairly successful and at the same time improved the signal-to-noise ratio was the utilization of a triode valve in series with the lamp direct-current supply. The supply voltage was about twice the normal required and the resistance of the triode valve, consisting of a bank of low-plate-impedance triodes in parallel, was controlled by its bias. This bias was automatically adjusted to the average signal level by rectifying a part of the audio-frequency voltage and applying it to the valve grids through a suitable direct-current amplifier and resistance-capacitance network.

The difficulty with this scheme was that it was hard to find a combination of circuit constants which

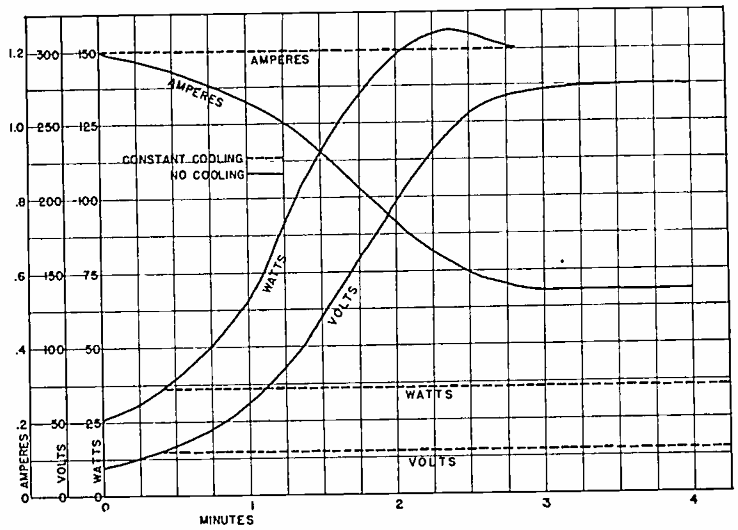

Fig. 6

could be relied on over a period of time. The lamp input would gradually drift to a lower or higher value, making reliable operation difficult.

Another satisfactory solution of the problem was relatively simple. The operation of the lamp, while being modulated, was investigated for a considerable time with various input voltages and currents. It was found that the most satisfactory condition exists when the lamp is operated in the unstable region of its resistance characteristic. This condition exists when the lamp current is about one ampere and the potential across it approximately 100 volts.

It was discovered that the operation of the lamp could be controlled successfully by controlling the temperature with a small stream of air. With a given supply voltage the current through the lamp varied according to the combined action of the room or equipment temperature and the modulating power. In order to maintain the lamp operating current at some fixed value, such as the 1.0-ampere condition indicated above, it was only necessary to have a cooling device which operated according to the lamp current or voltage. This was accomplished with a small blower which was turned on and off by a differential relay operating on the lamp current. This relay is of a type which drops out at 95 per cent of the pull in current, thus providing a rather fine cooling control.

Fig. 5 and Fig. 6 show the current and voltage characteristics of the lamp with and without cooling.

Choice of Phototubes

Choosing a suitable phototube presented an appreciable problem. Gas-type tubes were used because of their high sensitivity. The main difficulties involved were microphonic noises and the mechanical arrangement of the anodes, which two problems are more or less interconnected.

Microphonic noises cause trouble because the system requires that the cells with their slit tubes be maintained accurately spaced and close to the rotating disk. This means that they must be supported rigidly on the same structure which supports the

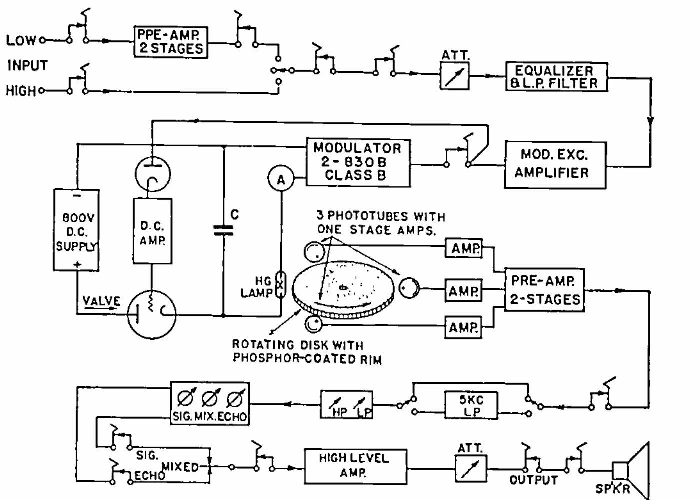

Fig. 7—Schematic of first complete model direct-control amplifier and tube valve.

rotating disk and the mercury lamp; therefore they are subject to any vibration which is developed. Anodes may be unsatisfactory because they obstruct the narrow slit which admits the signal light to the cathode. This applies particularly to the central-rod-anode type, as is shown in the diagram Fig. 4. Microphonic noises are generated within the tube because of a capacitance change due to a slight movement of the anode and cathode relative to each other. The interference of the anode can be overcome by using tubes having a rectangular wire-frame anode ("shadowless"). However, since the microphonics depend on the mechanical construction, it was found that the commercially available tubes with shadowless anodes were appreciably more microphonic than the corresponding rod anodes because they were not supported as well mechanically. This problem was finally eliminated when new cells of sufficient sensitivity and with short stub anodes, but otherwise of the same construction as the previous ones, were made available.

Phosphor Surface and Signal-to-Noise Ratio

The optimum signal-to-noise ratio is determined either by the thermal noise of the photocell coupling resistor or by the shot noise originated in the photocell by the unmodulated light source. Noise is also introduced by low-frequency "bumps" due to any unevenness of the phosphor coating or smudges on its surface. Considerable protection against touching the disk accidently was provided in the latest model of the apparatus by leaving small shoulders about 1/16 inch wide and 0.01 inch high (very slightly greater than the thickness of the phosphor coating) at the edges of the disk.

The maximum variation in the distance between the rim of the disk and the slit tubes is less than 0.005 inch. Such disks, with a variation in radius of not more than a few thousandths of an inch, can be machined without difficulty.

The phosphor binding material and method of applying the coating to the disk presented a difficult problem. Various kinds of binders were tried but considerable difficulty was experienced in getting a coating that was sufficiently uniform and at the same time adhered permanently. The use of sodium silicate as a binder gave a coating that was relatively easy to apply and was satisfactory for a time, but in warm weather with high humidity it apparently absorbed moisture and either blistered or crystallized.

A quite satisfactory and durable coating was finally achieved by thoroughly cleaning the metal-disk rim with acetone and lacquer thinner and then spraying on many coats of a mixture of the phosphor and a certain diluted lacquer.

Mechanical Construction

After the first experiments promised success, a rather elaborate model, including a number of pieces of test equipment to facilitate further development work, was built. Fig. 7 is a simplified diagram of this equipment. From the work done with this model it was decided that it would be possible to build into a cabinet rack of standard dimensions a satisfactory apparatus for commercial use.

Fig. 8 is a schematic diagram of the final apparatus. Note that it has been considerably simplified by combining a number of the amplifiers, equalizers, and filters into single units and by substituting a ballast resistor for the direct-current amplifier and the tube valve controlling the lamp.

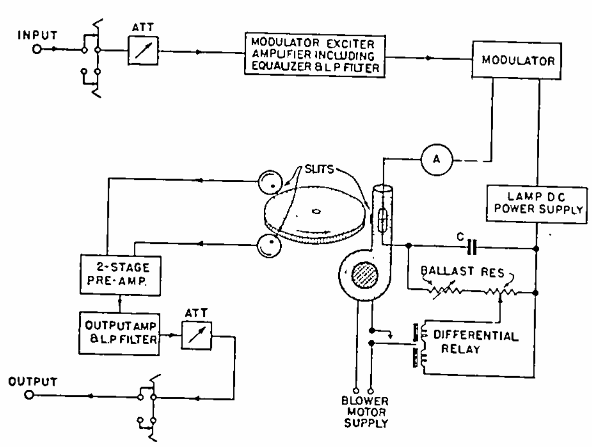

Fig. 8—Schematic of present model.

The temperature of the mercury-vapor lamp is regulated by means of a blower controlled by a differential relay operating on the lamp current. A large capacitor C across the ballast resistor R by-passes the audio-frequency modulating power and together with the ballast resistor serves to prevent the lamp from being extinguished by servere overmodulation peaks.

The photograph, Fig. 9, shows the front of the final model mounted in a standard cabinet rack. The input and output jacks and attenuators are on the fourth panel from the top. Above it is the power-control panel with the start-stop buttons on the left and a high-voltage switch on the right. The meter in the center reads the modulator filament voltage and also serves to indicate the line voltage. Above these there is, on the left, a direct-voltage meter with a switch to read lamp-supply voltage, lamp voltage, modulator-supply voltage, or the photocell-supply voltage. The direct-current ammeter in the center is permanently connected in the lamp circuit. The milliammeter on the right reads the class B modulator plate current normally and also serves as a modulation level indicator.

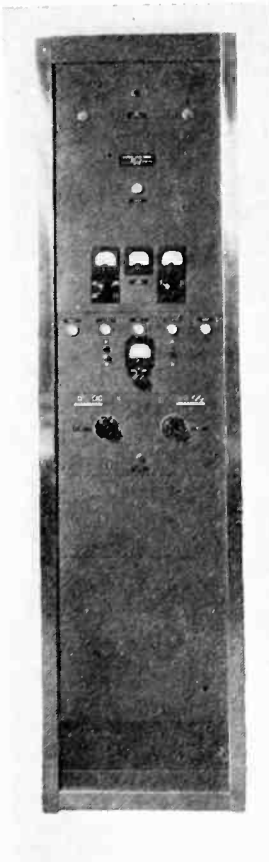

Fig. 9

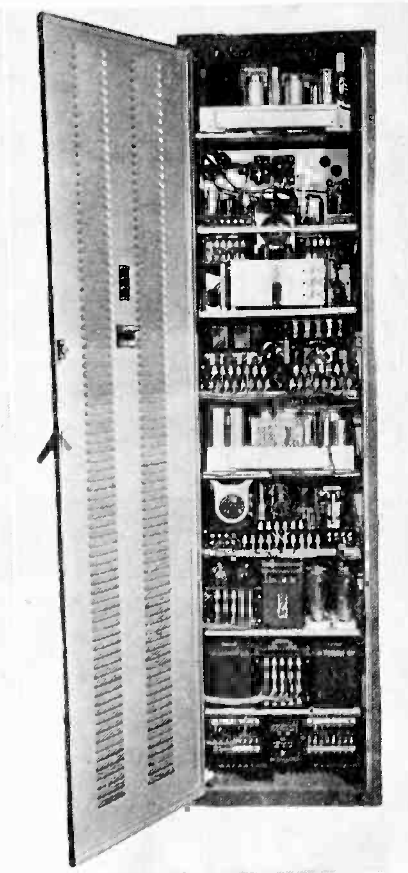

Fig. 10—Rear view of the single rack which contains the synthetic reverberation equipment in the latest model.

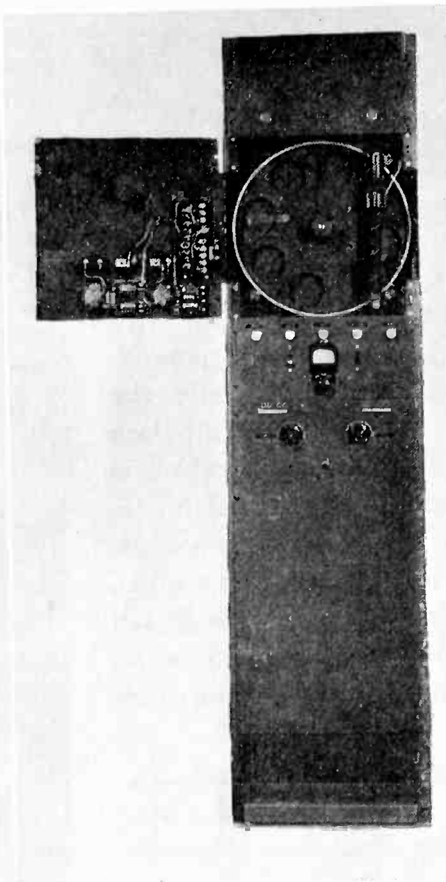

Fig. 11

Fig. 10 shows the rear of the rack with the door open. At the top is the output amplifier and directly below it the disk chassis. The disk is on the front panel side of a vertical partition, as shown in Fig. 11. At the center of this chassis is the disk shaft bearing with a centrifugal interlock switch to prevent damage to the disk coating if an attempt is made to operate the lamp without the disk running. The shelf at the shaft level supports the direct-current supply for the photocells. The lower shelf supports the two-stage preamplifier which connects to the pickup phototubes through a special low-capacitance shielded cable. Two phototube pickups are normally used in this setup, either one of which may be used alone if desired. The rear of one of the phototube housings, which shield the tube thoroughly and into which the optical slot is cut, is at the lower right of the chassis and the other one is diagonally opposite and just below the mercury-pressure-lamp housing. The fronts of these housings appear in Fig. 11 in the diagonally opposite corners.

The next chassis contains the power-control circuits and relays, including a time-delay relay to protect the mercury-vapor power rectifier tubes. It also contains the lamp-cooling blower and the disk-drive motor. The blower connects to the lamp mounting through a piece of flexible hose appearing at the right, in Fig. 11. An induction-type motor drives the disk with a small "V" belt running in grooved pulleys. Because of the fact that both the recording and the pickup occur on the same disk, it is unnecessary to maintain a very constant speed. The next lower chassis supports the modulator input exciter amplifier. On the front panel of this unit are mounted the input and output jacks and attenuators.

The class B modulator including input and output transformers is immediately below its exciter amplifier. The lamp series resistor and its by-pass capacitor are also on this modulator chassis along with the differential relay which may be seen to the right of the center. The modulator is protected against noload operation by the small underload relay at the left of the differential relay which also operates on the lamp current.

The next two chassis contain separate power supplies for the lamp and the modulator. The upper one contains the power transformers and rectifiers and the lower one the filter chokes, capacitors, and bleeders. All input and output connections for both signal and power circuits appear on the panel at the bottom of the rack. This panel also contains an overload circuit breaker which is in the main alternating-current power line.

Both the disk chassis panel and rear cabinet door have interlock safety switches for the high-voltage circuits. Terminals are provided for remote start, stop, and interlock connections.

Fig. 12 shows a remote start-stop and mixer-attenuator box which proved to be useful for some purposes. The attenuator on the right controls the output level. The unit at the left is a dual "T" type attenuator in which one section carries the original

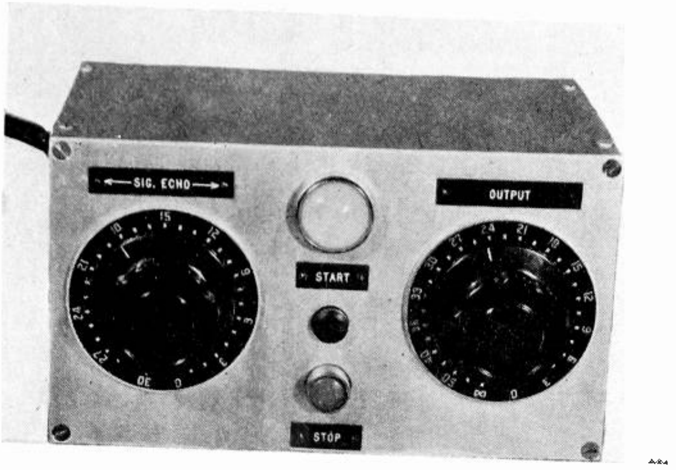

Fig. 12

signal and the other the reverberation signal. It is arranged so that when it is turned all the way counterclockwise only the original signal is passed, while when turned all the way clockwise only the reverberation signal passes. At any intermediate point the ratio of original signal to reverberation is proportional to the amount of rotation. The unit is so designed that when the outputs are fed to a common load having the proper terminating resistance, the overall signal level remains constant. Thus with one control knob it is possible to add any desired amount of reverberation without disturbing the over-all level of the program.

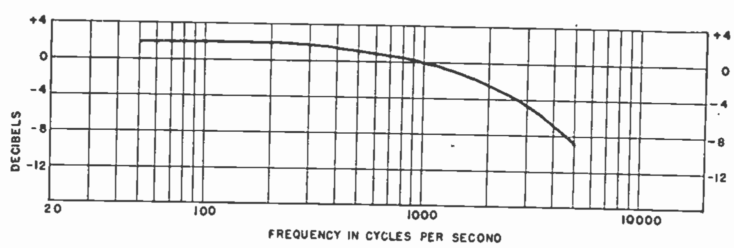

Fig. 13—Audio-frequency response of modulated mercury-pressure lamp without equalization.
Input 100 watts. (1 ampere at 100 volts)

Performance

The frequency response of the mercury-vapor lamp drops off toward the high-frequency end where at the same time the apparent impedance of the lamp increases. Equalization therefore becomes necessary and can easily be carried out if it is kept in mind that the power contained in sound programs is confined to the lower frequencies. Fig. 13 shows the lamp-output-versus-frequency characteristic before equalization.

From a practical operating viewpoint these facts mean that the modulation level at the higher frequencies can be increased, improving the over-all signal-to-noise ratio.

A frequency response for reverberation above 5000 cycles is hardly needed as proved by subsequent tests. Referring to Fig. 2 which shows the reverberation characteristic of a typical broadcast auditorium it can be seen that in such a space the reverberation

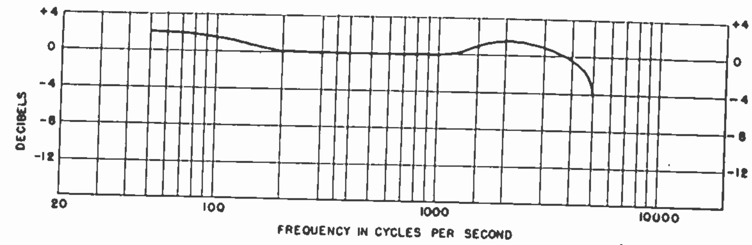

Fig. 14—Over-all frequency response of model C reverberation equipment.

time above 5000 cycles is negligible compared with that at lower frequencies.

The over-all frequency characteristic of the synthetic-reverberation device is shown in Fig. 14. Measurements showed that the total distortion is of the order of 2 to 5 per cent over-all at full modulation. The signal-to-noise ratio of reverberation only, picked up from the disk, is about 45 decibels at full modulation. However, since only a fraction of that signal is

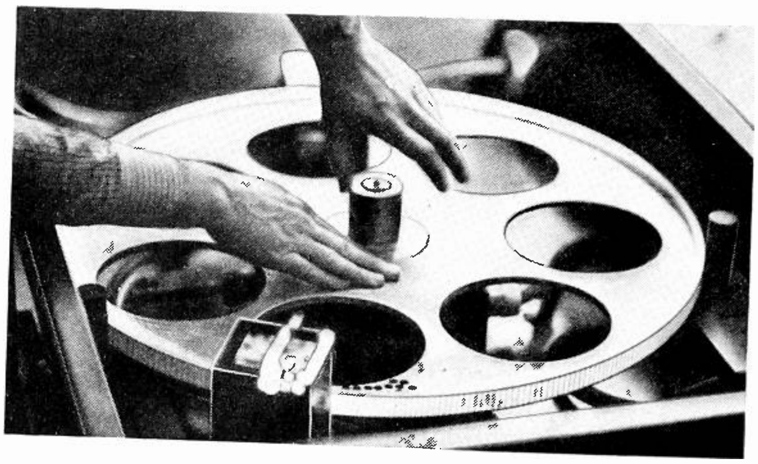

Fig. 15

added to the original sound, the over-all signal-to-noise ratio is appreciably better. A compression-type amplifier which reduces the volume range of modulation logarithmically is used in this model to improve further the signal-to-noise ratio at low modulation levels.

Fig. 15 shows a photo of the disk while the sound modulation is still visible.

Acknowledgment

We wish to express our appreciation to our associates at Columbia Broadcasting System—especially Messrs. Murphy, Dyer and Wilner—for assisting us with inspiration, advice and active work in this development.

The Scattering of Radio Waves in the Lower and Middle Atmosphere*

J. H. PIDDINGTON†, ASSOCIATE MEMBER, I. R. E.

Summary—*The evidence relating to the reflection of radio waves from levels below 80 kilometers is considered and apparatus used to investigate the reflection coefficients of these regions is described. The new experimental results here presented are not in agreement with those of earlier workers, but indicate that reflections from region B (below 10 kilometers) and region C (35 to 60 kilometers) are very weak and are due to scattering patches rather than reflecting strata.*

It is shown that reflections from region B are probably due to water molecules and that echoes with time delays corresponding to semipaths of 10–25 kilometers probably originate at scattering centers within the troposphere.

The equivalent reflection coefficient of region C is discussed and the mechanism of formation of this region of ionization is briefly considered in connection with atmospheric temperature gradients.

I. Introduction

During the past three years a number of publications have appeared relating to the reflection of radio waves from regions below the Kennelly-Heaviside layer (region E). These reflections appear to originate in two zones, the upper one extending from about 30 to 60 kilometers above ground level; the lower, as will be seen later, is probably coincident with the troposphere. We refer to these zones as regions C and B, respectively, region D being the absorbing layer which is thought to exist somewhere between 70 and 100 kilometers.

Using the well-known Breit and Tuve method of radio-pulse production, Colwell and Friend[1,2] claim to have observed echoes with group time delays corresponding to semipaths between 5 and 30 kilometers. Such reflections are referred to below as region-B echoes,[3] although it is not yet clear that they all arise within the troposphere. Colwell and Friend state that they "are led to believe that there is a third region at a height of 5 to 50 kilometers which strongly reflects radio waves." They also state[3] that the low-lying strata reflect so strongly on occasions that echoes from higher levels are noticeably weakened. They find correlations between region-B echo delays and magnetic and solar disturbances and suggest that the mean B-region height is steadily falling, owing to increased sunspot activity.

Using a very much more powerful pulse transmitter, Watson Watt and his associates[4,5] in England also found echoes returning from levels as low as 10 kilometers. Like Colwell and Friend they concluded that these were due to strongly reflecting, discrete layers. A third group of workers, in India, have also reported[6,7] fairly strong echoes from regions low down in the atmosphere. These, they state, are beyond doubt due to ionized layers.

The same phenomena have been more recently examined by Appleton and Piddington,[8] who made accurate measurements of equivalent reflection coefficients of the B region. Reference is made to this investigation in the following paper which is chiefly concerned with the nature of regions B and C.

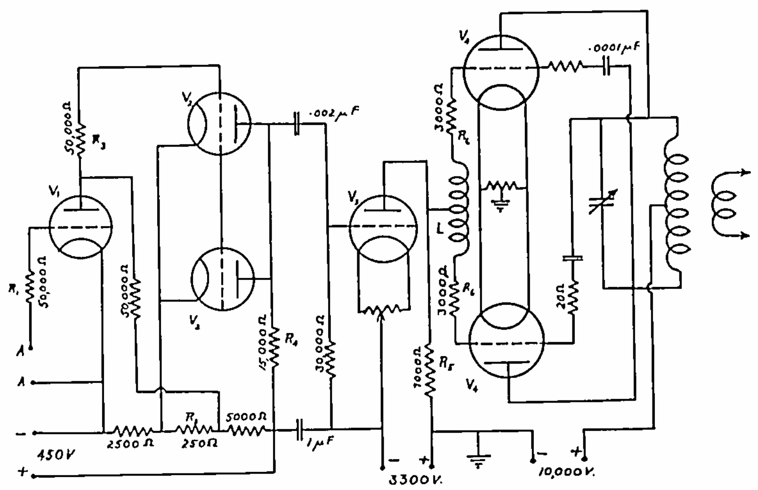

Fig. 1—The grid-modulated transmitter.

II. Apparatus and Results

The circuit diagram of a pulse transmitter which was used for investigating regions B and C is shown in Fig. 1. The modulator is not shown as it has been described elsewhere.[9]

The "negative" pulse is applied at *AA* (Fig. 1) and after amplification, causes the grid of V_3 to become negative, thus removing the 3000 volts grid bias from V_4V_4 and permitting oscillation for the duration of the pulse. The total emission of the valves V_4 (Mullard TZ2-250) was $2\frac{1}{2}$ amperes and the power input during oscillation about 15 kilowatts. The pulse duration was 20 microseconds.

* Decimal classification: R113.61. Original manuscript received by the Institute, October 10, 1938; abridgment received, August 30, 1939.

† Cavendish Laboratory, Cambridge, England, and Walter and Eliza Hall Research Fellow of the University of Sydney, Sydney, Australia.

[1] R. C. Colwell and A. W. Friend, "The D region of the ionosphere," *Nature*, vol. 137, p. 782; May 9, (1936).

[2] A. W. Friend and R. C. Colwell, "Measuring the reflecting regions in the troposphere," Proc. I.R.E., vol. 25, pp. 1531–1541; December, (1937).

[3] Colwell and Friend call these region-C echoes.

[4] R. A. Watson Watt, L. H. Bainbridge-Bell, A. F. Wilkins, and E. G. Bowen, "Return of radio waves from the middle atmosphere," *Nature*, vol. 137, pp. 866; May 23, (1936).

[5] R. A. Watson Watt, A. F. Wilkins, and E. G. Bowen, "The return of radio waves from the middle atmosphere-I," *Proc. Roy. Soc.*, ser. A, vol. 161, pp. 181–196; July 15, (1937).

[6] H. Rakshit and J. N. Bhar, "Some observations on the C region of the ionosphere," *Nature*, vol. 138, pp. 283–284; August 15, (1936).

[7] S. K. Mitra and J. N. Bhar, *Science and Culture*, vol. 1, p. 782; (1936).

[8] E. V. Appleton and J. H. Piddington, "The reflexing coefficients of ionospheric regions," *Proc. Roy. Soc.*, ser. A, vol. 164, pp. 467–476; February 18, (1938).

[9] G. Millington and S. W. Falloon, "An improved pulse transmitter," *Marconi Rev.*, no. 57, November, (1935).

The Receiver

The transmitter and receiver each used a horizontal half-wave dipole one-quarter wavelength above the ground and one kilometer apart.

The receiver was a superheterodyne with an overall frequency response band of 66 kilocycles per second. A pulse of 20 microseconds duration in the aerial was lengthened by 15 microseconds in the receiver, this increase corresponding approximately to a half period of the highest modulation frequency which could be passed by the receiver.

Pulse echoes with field strengths exceeding 2 microvolts per meter and with time delays corresponding to semipaths of 6 kilometers or more were observable.

Observations of echoes were made with the aid of a cathode-ray oscillograph, high-speed time base, and time-interval marker. This last-mentioned piece of apparatus makes small marks along the time base at intervals of 20 microseconds corresponding to echo semipath intervals of 18 kilometers.

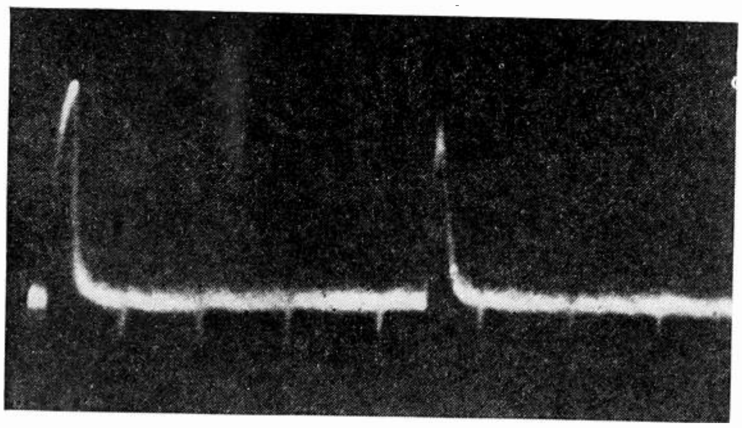

Fig. 2—The time base with height marks spaced 18 kilometers apart.

The time base with height marks 18 kilometers apart is shown in Fig. 2. The ground pulse is on the left and an echo of semipath just over 80 kilometers is on the right. It will be seen that echoes returned from distances as short as 7 kilometers can be separated from the ground ray, although none are visible in Fig. 2.

Results

In view of the strong reflecting properties attributed to region B by other investigators, it was thought desirable to make careful measurements of these properties. The equivalent reflection coefficient ρ of a stratum or patch is defined as the ratio of the intensity of the received echo to that of an echo returned by an infinite plane perfectly reflecting boundary at the same distance as the stratum or patch. The value of ρ for region-B echoes was determined by a comparison of their strength with those of the first and second region-F_2 reflections and also directly from the formula relating the field strength E volts per meter, due to a dipole radiating P kilowatts, after the energy has traveled γ kilometers and been reflected from a region of equivalent reflection coefficient ρ

$$E = \frac{0.3\rho\sqrt{P}}{\gamma}.$$

The value of ρ for region B in the southeast of England was found to be less than 0.0001 for echoes of semipath 10 kilometers or more and waves of frequency 6 megacycles per second. No individual steady echoes were found to exist but an irregular and unsteady pattern with ρ increasing steadily as the delay decreased and merging into the ground ray. For a semipath of 10 kilometers the maximum observed value of ρ at 6 megacycles per second was 0.00007.

III. The Scattering of Radio Waves in the Troposphere

As suggested by Appleton and Piddington,[8] region-B echoes are probably not due to reflection from continuous layers at all, but are signals of very low intensity reflected from numerous scattering centers. Watson Watt, Wilkins, and Bowen[5] have pointed out that in a typical snap photo of region-B echoes, the amplitude of what they term the fifth-order reflection is about 0.2 that of the first-order reflection. But if we cease to regard the gradually decaying echo system as multiple reflections from a small number of discrete layers, the obvious interpretation to be placed upon this fact is that the reflection coefficient of the scattering centers is everywhere of the same order, the gradual decay being caused by spatial attenuation. The agency causing reflection might, therefore, be in the form of clouds, and since these would be effective as reflectors when situated above a point some distance from the transmitter, it is seen that an echo which has a delay corresponding to a semipath of 20 kilometers might really be reflected from a region much below 20 kilometers in height.

The experimental technique of Watson Watt, Wilkins, and Bowen did not permit of the detection of echoes of semipath much below 10 kilometers. There is every indication from their records, however, and from those of other workers[2,6,7], that reflection takes place from centers situated well within the troposphere. Other experiments using aerials of different directivity, the details of which need not be given here, have given support to the view that the echoes may arrive at the receiver from directions at considerable angles to the vertical, and in the light of the subsequent theoretical investigation of the nature of the reflecting agency it is regarded as highly probable that all B-region echoes originate within the troposphere, even when the echo delay corresponds to a semipath as great as 20 to 25 kilometers.

It is difficult to reconcile the above-mentioned results with those of Colwell and Friend,[2] who obtain one or two distinct echoes originating within or near the troposphere instead of the whole pattern of echoes described above. Moreover, the power used by these workers, as far as can be estimated from an examination of the circuit diagram of their transmitter, is much too low to give detectable echoes unless the conditions under which they are working are very different to those experienced in England.

IV. The Agency Responsible for B-Region Echoes

The distribution and reflecting efficiency of the agency causing region-B echoes has now been dealt with and the question next arises as to the constitution and origin of this agency.

Ionization

Colwell and Friend[2] have adduced evidence to show that these echoes are due to ionization caused by solar radiation. They suggest correlations between the properties of B region and sunspot activity and auroral phenomena. Watson Watt, Wilkins, and Bowen naturally assumed ionization to be responsible for these reflections since no other agency could give the very large reflection coefficient which they found. It can, however, be shown that it is highly improbable that B-region echoes are due either to heavy ions or electrons.

The problem has been discussed by the writer elsewhere,[10] where it is shown that the reflection coefficient of a layer of ions in the troposphere is given by

$$\rho = \frac{\sigma}{2f}$$

where f is the wave frequency and will be taken as 2×10^6 cycles per second. The necessary value of σ (conductivity) to account, even for the very small values of ρ found by Appleton and Piddington[8] is of the order of 400 electrostatic units. This applies for both heavy ions and electrons.

A continuous record[11] was made of the conductivity of the atmosphere to a height well above the tropopause when the balloon *Explorer II* ascended in 1935. The highest value of conductivity found in the troposphere was about 2.5×10^{-3} electrostatic units and this is smaller by a factor of 160,000 than the least value necessary to account for observed reflection coefficients. Such considerations, therefore, lead us to reject the ionization hypothesis.

The Probable Process of Reflection

Since it would appear to be extremely improbable that B-region echoes are due to a process of reflection by free electrons or ions, we now proceed to consider the possibility that these echoes are due to reflections from discontinuities of the atmospheric dielectric constant, due to changes in composition. It is at once apparent that water-vapor molecules which have a large permanent dipole moment, make a very considerable contribution to the total dielectric constant of a moist atmosphere. If we write $K-1$ as the contribution to the dielectric constant due to any particular gas present in the atmosphere, then $K-1$ for air at normal temperature and pressure is 5.9×10^{-4} and for water in the form of droplets and density equal to its saturation vapor density at zero degrees centrigrade, $K-1$ is 3.9×10^{-4}. The distribution of water vapor in the atmosphere is much more irregular than that of the other common components and, in addition, the three states in which water exists have widely different dielectric constants at the frequencies under consideration. Water is, therefore, the most probable agency to account for the observed reflections.

A radio wave incident on a surface at which the dielectric constant changes will be partially reflected and if the transition is relatively sudden, the reflection coefficient may be written

$$\rho = \frac{n-1}{n+1}$$

where n is the refractive index on one side of the boundary and unity that on the other. In the atmosphere $n=K^{1/2}$ and, since $K-1$ is small we may write

$$\rho \doteq \frac{K-1}{4}.$$

In the case of water in its three states, K will vary irregularly throughout the atmosphere because of variations in the amount present per cubic centimeter and also because of variations in its state. The author has shown[10] that the contribution to the dielectric constant due to q grams per cubic centimeter of water is given by

$K-1=3q$ electrostatic units for ice

$K-1=80q$ electrostatic units for water

$K-1=12.7q$ electrostatic units for water vapor

the values applying at the wave frequencies in which we are interested at the moment. It is clear, therefore, that any change of state of water which occurs in the troposphere so as to form a boundary which is sharp compared to a wavelength may result in partial reflection of the electromagnetic waves.

Measurements made from pilot balloons sent up from Kew Observatory are used[10] to show that the contribution to the dielectric constant of water vapor at a height of 6 kilometers is often as high as 5×10^{-5}. If a discontinuity of state occurs, say from vapor to liquid, defining a plane boundary the reflection co-

[10] J. H. Piddington, "The origin of radio-wave reflections in the troposphere," *Proc. Phys. Soc.*, vol. 51, pp. 129-135; January, (1939).

[11] O. H. Gish and K. L. Sherman, "Information to be obtained from some atmospheric-electric measurements in the stratosphere," *Int. Assoc. Terr. Mag. and Elec.* (Edinburgh), September, (1936).

efficient will be about 10^{-5}; corresponding values of ρ at heights of two and four kilometers are 10^{-4} and 4×10^{-5}. If the surface of discontinuity is not an infinite plane, then higher values of ρ may occur. For instance, the value of ρ corresponding to a plane circular boundary normal to the ray, of diameter 540 meters and distant 5 kilometers is twice as high (for 30-meter waves) as that for an infinite plane sheet.

We conclude that the observed value of $\rho=2\times10^{-5}$ (at 9 megacycles per second) may be accounted for on a theory of reflection by water molecules. If so, a new method of investigating the distribution of water in the troposphere is available and further work with very high-powered pulse transmitters sending signals of duration less than 10 microseconds may be expected to add greatly to our knowledge in this field.

V. Reflections in the Middle Atmosphere

In addition to the B region, which appears to lie entirely within the troposphere, echoes from a second reflecting region (region C) lying between about 30 and 60 kilometers have been observed during the past few years. In 1935 Mitra and Syam[12] reported weak echoes from about 55 kilometers and this was followed by a similar report from Colwell and Friend.[1] Revised estimates of the strength of such reflections were made by Rakshit and Bhar[6] who claimed that on occasions they were as strong as echoes from regions E and F. Definite photographic evidence of the presence of weak, irregular reflections from levels between about 35 and 60 kilometers was produced by Watson Watt, Wilkins, and Bowen[4,5] in 1937. The strength of the echoes was of the same order as those from the B region and they were thought by the observers to indicate the presence of further strongly reflecting layers.

Using the apparatus described above, with a radiated power of about 3 kilowatts at a frequency of 8.8 megacycles per second the upper limit for the reflection coefficient of region C was found to be about 0.0005, which agrees with the results of Watson Watt, Wilkins, and Bowen, provided their records are reinterpreted, as suggested (for the case of region B) by Appleton and Piddington. Such a small value of ρ is not in agreement with results of Rakshit and Bhar or of Colwell and Friend, although in the case of the former workers the discrepancy might be due to the large difference in latitude of the points of observation.

The pressure and temperature at a height of 40 kilometers are probably about 2 millimeters and 300 degrees Kelvin, respectively, so that the collisional frequency ν of ions with neutral molecules is of the order of 10^8. The value of ν for electrons is higher so that for all wave frequencies less than about 5 megacycles per second the conductivity of a layer of ions or electrons is independent of the frequency. The reflection coefficient of a sharply bounded infinite layer is given by

$$\rho = \frac{\sigma}{2f}$$

that is, the reflection coefficient is inversely proportional to the wave frequency. Using very long waves (18.8 kilometers) Best, Ratcliffe, and Wilkes[13] have recently measured the level from which reflection takes place and found that it is probably about 74 kilometers during the day. We conclude that layers of ionization of sufficient density to reflect appreciably waves of frequency above 1 megacycle per second do not exist below 70 kilometers or they would strongly reflect these very long waves.

The value of ρ for region C (35 to 60 kilometers) may be estimated from the records of Watson Watt, Wilkins, and Bowen by a comparison with the strength of region-B echoes. It is found to be of the order 10^{-4} at a frequency of 6 megacycles per second. Even this low value of ρ should give appreciable reflection of the very long waves of Best, Ratcliffe, and Wilkes if the reflecting agency were in the form of a layer. Thus, as in the case of region B, it appears probable that region C consists, not of strata, but of scattering patches.

In a recent communication, Smith and Kirby[14] claim to have measured the critical penetration frequency of one of the low layers which were observed by Mitra and Syam[12] Colwell and Friend[1,2] and Watt and his associates[4,5]; that is, of regions B and C. It appears improbable that a critical-frequency phenomenon could be associated with the "conductivity" type of reflection which exists under the conditions of high collisional frequency obtaining in regions B and C. Also the values of field strength indicated by Smith and Kirby suggest a reflection coefficient of the order 0.1 which appears very high for regions B or C even at the very low angles of incidence if our picture of irregular scattering patches is correct.

The agency responsible for the formation of region-C ionization may be the same as that which causes the sudden appearance of patches of ionization in the E region.[8] These patches have been shown to extend down to 80 kilometers above England below which they are never found. If the solar or cosmic particles responsible penetrate below 80 kilometers then some modification in the state of the atmosphere just below this level is to be expected. Such a change might well be the positive downward temperature gradient

[12] S. K. Mitra and P. Syam, "Absorbing layer of the ionosphere at low height," *Nature*, vol. 135, pp. 953–954; June 8, (1935).

[13] J. E. Best, J. A. Ratcliffe, and M. V. Wilkes, "Experimental investigation of very long waves reflected from the ionosphere," *Proc. Roy. Soc.* ser. A, vol. 156, pp. 614–633; Sept., (1936).

[14] N. Smith and S. S. Kirby, "Critical frequencies of low ionosphere layers," *Phys. Rev.*, vol. 51, pp. 890–891; May 15, (1937).

suggested by Martyn and Pulley.[15] When this gradient inverts, as it must, the ensuing negative downward gradient is, as pointed out by the above authors, an atmospheric condition which favors the formation of reflecting regions of ionization when ionizing radiation is incident from above. Thus the C region may coincide with the lower slope of the temperature maximum of the middle atmosphere.

[15] D. F. Martyn and O. O. Pulley, "The temperature and constituents of the upper atmosphere," *Proc. Roy. Soc.*, ser. A, vol 154, pp. 455–486; April, (1936).

It may be seen, as in the case of region B, that those C-region echoes with the longest delays may originate at levels much below that corresponding to their semipath. Thus the upper limit of region C is undefined and may be as low as about 40 kilometers.

ACKNOWLEDGMENT

I wish to thank Professor E. V. Appleton for invaluable advice and help in connection with the above work.

The Electronic-Wave Theory of Velocity-Modulation Tubes*

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Summary—Following a brief discussion of the Hahn theory of velocity modulation in which there is explained the basic velocity-modulation tube phenomena by means of space-charge waves propagating along the electron beam, the wave theory is reformulated by means of the retarded potentials for the most important case, that of a magnetically focused electron beam. The use of the potentials is believed to result in sufficient simplification to merit consideration in choosing the best attack on the theory.

The electron beam is seen to be a medium for space-charge-wave propagation, the input signal serving to excite waves which propagate with beneficial change down the tube and induce output current in the output circuit. It is shown that important design constants for velocity-modulation tubes, such as optimum-drift tube length and the amount and phase of the transconductance, may be computed by use of the wave theory. Numerical values are given for a special case as an example.

INTRODUCTION

In a paper describing the velocity-modulation type of tube,[1] simple derivations were given of the transconductance, input impedance, and other characteristics of the new tubes. These derivations neglected important space-charge effects which have, however, been made the subject of complete studies by W. C. Hahn.[2] Hahn acquainted the author with this space-charge theory and showed how it was possible to utilize the theory to predict tube behavior and hence to form a basis for tube design. Also, it was evident that Hahn's theory had disclosed possibilities not originally envisioned in velocity modulation.

In commencing the study of velocity-modulation phenomena the author was accordingly influenced by previous experience with the Hahn theory. However, it was felt that at the present stage of the theory it would be well to seek first the most direct and convenient attack, both to simplify the presentation of the theory to others as well as to facilitate such new work as might be done on the theory. The theory for the case which turns out to be of the greatest practical importance, that of an electron beam focused by a very strong magnetic field, is therefore reformulated in this paper by the use of the retarded scalar electric and vector magnetic potentials. The use of the potentials is believed to lead to sufficient simplification to merit consideration in choosing the best attack on the theory. The value of the wave theory in explaining and predicting velocity-modulation-tube behavior is demonstrated by discussion and numerical examples.

THE HAHN THEORY

An analysis of the operation of a velocity-modulation tube in which space charge is to be considered would appear to require the inclusion of displacement currents and the variation of fields, charge, and current densities with beam cross section, length, and time. The problem is thus a natural one for attack by Maxwell's equations. A major difficulty arises at the outset, even if small-signal theory is all that is asked for, in choosing a set of assumptions that not only will be reasonably close to the conditions met in practice but that will not actually violate Maxwell's equations.

For example, one cannot simply assume that in the absence of signal the velocity-modulation tube consists of a beam of electrons of uniform charge density and uniform velocity drifting down the axis of a cylindrical conducting tube unless additional qualifications (which will be introduced later) are made. Such a situation immediately contradicts the equations which ultimately, it is hoped, will give the solution, unless of course space charge and hence the problem

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† General Engineering Laboratory, General Electric Company, Schenectady, N. Y.

[1] W. C. Hahn and G. F. Metcalf, "Velocity-modulated tubes," PROC. I.R.E., vol. 27, pp. 106–116; February, (1939).

[2] W. C. Hahn, "Small signal theory of velocity-modulated electron beams," *Gen. Elec. Rev.*, vol. 42, pp. 258–270; June, (1939).

itself is neglected. To arrive at a more proper, zero-signal, steady-state solution it is possible to consider what distribution of charge and velocity must actually exist in the absence of signal in a velocity-modulation tube, but such distributions will be so dependent upon the geometry of the beam, the conducting tube down which it drifts, and terminal conditions at each end of the drift tube that such a solution, once obtained for one case, would be of doubtful application to another case.

Hahn[2] proposed what might be called a "separation of assumptions." He suggested that a mathematically exactly soluble counterpart be substituted for the velocity-modulation tube. Then, in interpreting the results and in applying them to design, the differences between the ideal tube and the particular practical tube under consideration would be considered.

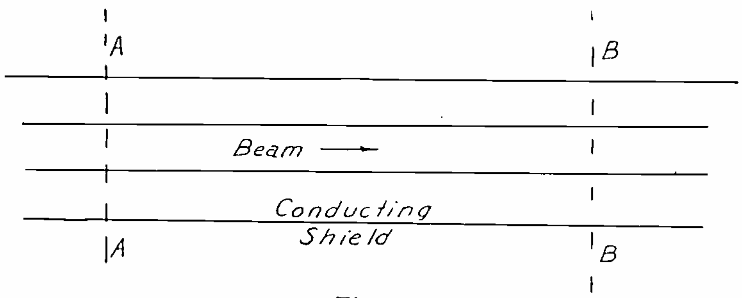

Fig. 1

The tube proposed by Hahn is shown diagramatically in Fig. 1. A beam consisting of a uniform density of both positive ions and electrons drifts with uniform velocity down the axis of a cylindrical conducting tube. The electrons and positive ions are supposed to have acquired their uniform axial velocity before entering the tube which is assumed infinitely long. The mass of the positive ions is assumed to be infinite so that the velocity modulation will apply to the electrons only. The positive ions will thus not enter into any part of the wave solution but they aid in establishing convenient direct-current conditions. Now, the steady or direct-current charge density contributed by the positive ions is supposed to neutralize exactly the direct-current charge density due to the electrons so that, in the absence of signal there is no steady electric field, no net direct-current charge density, no direct current, nor any steady magnetic field due to the beam. Thus the situation postulated is a completely consistent one. Furthermore, this drifting ionized region may be exactly analyzed for the effect of small-signal velocity modulation applied at some point along its length.

It should be understood that this idealized beam has only its zero-signal or direct-current space charge nullified. The heavy positive ions will not depart from their steady drift positions and thus as soon as a signal is applied to the beam there is immediately a space-charge effect due to the high-frequency motion of electrons. Thus while the direct-current conditions are those of no space charge, the high-frequency phenomena which we wish to study will include the important space-charge effects.

The Electronic-Wave Concept

If the beam is disturbed at some point, say A-A in Fig. (1), this disturbance will in general propagate down the tube since the electrons are constantly in motion. Thus, for example, if at A-A a small increment is added to the velocity of electrons, we should expect to observe some effects in the nature of changed motion of the beam as it passes a later point B-B. Basically, the usefulness of the velocity-modulation tube arises from the ability of the beam to receive a disturbance in the form of voltage input at some point, propagate it with beneficial change down the length of the tube, and finally make that disturbance available to an output circuit in the form of an induced output current.

The study of the tube then consists actually of the study of the exciting, the propagating, and the withdrawing of waves in the tube. We have now to ask what kinds of waves, once started at A-A, will propagate down the tube. Before entering into the mathematical search for these waves it may be helpful to discuss the significance of the results which later will be obtained for the case of a beam perfectly focused by a magnetic-focusing field.

In the analysis which follows we shall find that the important phenomena of the velocity-modulation tubes are largely explained by the possible existence of a pair of slow space-charge waves. We find, in other words, that if either of these two space-charge waves be started in some manner or other at any point in the beam, then it will propagate down the beam without attenuation (assuming a perfectly conducting shield). These two waves are termed slow space-charge waves because they have velocities differing only slightly from that of the beam itself. One of the pair of space-charge waves has a velocity slightly greater and the other a velocity slightly less than that of the beam. This is one way in which the two space-charge waves are distinguished.

Another way in which the two slow waves differ is in the phase angle that exists for each between its velocity modulation V_z and its conduction-current modulation ψ_z. We shall see later that the amplitude of the ratio ψ_z/V_z is practically the same for each of the slow waves. However the phase of this ratio is zero for the wave which is slower than the beam and 180 degrees for the wave which travels faster than the beam. In other words, if we express the velocity modulation of the former wave as

$$V_{z,s} = V_s \sin(\omega t - \gamma_s z - \alpha_s) \qquad (1)$$

then the accompanying conduction-current modulation of this wave is

$$\psi_{z,s} = gV_s \sin(\omega t - \gamma_s z - \alpha_s) \qquad (2)$$

in which the subscript letter s indicates that the wave travels "slower" than the beam, ω is the angular frequency, γ is the propagation constant in the assumed z direction of propagation, α is an arbitrary phase-angle constant, and g is a positive constant whose value depends upon tube parameters and will be determined later. Similarly, we can express the wave which travels "faster" than the beam by

$$V_{z,f} = V_f \sin(\omega t - \gamma_f z - \alpha_f) \tag{3}$$

and

$$\psi_{z,f} = gV_f \sin(\omega t - \alpha_f - \gamma_f z - \pi). \tag{4}$$

It will be more convenient to substitute for γ_s and γ_f their values in terms of γ_0, the propagation constant of the beam,[3] since it has been stated that γ_s and γ_f are nearly equal to γ_0. As will be seen later, for most cases it is sufficiently accurate to write

$$\begin{aligned} \gamma_s &= \gamma_0[1 + \delta] \\ \gamma_f &= \gamma_0[1 - \delta] \end{aligned} \tag{5}$$

where δ is a small fraction depending upon tube parameters and will be determined later.

The common way in which waves are introduced in practice is pictured in Fig. 2 which shows a gap in the conducting tube of Fig. 1. The voltage of the electrons passing the gap is changed according to the signal applied across the gap. Since the gap is short, so that the changes in velocity take place in a very short distance of travel, there is little time for the electrons of different velocities to drift apart in position while yet near the gap. Consequently, if the beam enters the gap with no conduction-current modulation, it will leave the gap with added velocity modulation but essentially no conduction-current modulation as yet. This tells us that some of each of the two slow waves are started at such a gap; moreover the relative magnitudes and the phase angle between the waves is determined.

If the conduction current modulation is to be zero at A-A then the two waves must have their ψ_z's of equal magnitude but opposite in phase at A-A. Thus the velocity modulations will be equal and in phase at A-A because of relations (1) to (4). The expressions for the waves started at A-A and received at B-B may be written thus:

$$\text{waves started at } A\text{-}A \begin{cases} V_{z,s} = V_s \sin \omega t \\ V_{z,f} = V_s \sin \omega t \\ \psi_{z,s} = gV_s \sin \omega t \\ \psi_{z,f} = gV_s \sin(\omega t - \pi) \end{cases}$$

$$\begin{array}{l}\text{waves received} \\ \text{at } B\text{-}B\text{, a} \\ \text{distance } l \\ \text{from } A\text{-}A\end{array} \begin{cases} V_{z,s} = V_s \sin[\omega t - \gamma_0(1 + \delta)l] \\ V_{z,f} = V_s \sin[\omega t - \gamma_0(1 - \delta)l] \\ \psi_{z,s} = gV_s \sin[\omega t - \gamma_0(1 + \delta)l] \\ \psi_{z,f} = gV_s \sin[\omega t - \gamma_0(1 - \delta)l - \pi]. \end{cases}$$

[3] For beam velocity v_0 and angular frequency ω, $\gamma_0 = \omega/v_0$.

The total conduction current available at B-B is

$$\begin{aligned} [\psi_{z,f} + \psi_{z,s}]_{B\text{-}B} &= gV_s\{\sin[\omega t - \gamma_0(1 + \delta)l] \\ &\quad + \sin[\omega t - \gamma_0(1 - \delta)l - \pi]\} \\ &= 2gV_s \sin(\gamma_0 \delta l) \sin[\omega t - \gamma_0 l - \pi/2]. \end{aligned}$$

This result shows two very important characteristics of velocity-modulation tubes: first, that there are certain distances between input and output points on the tube for which the ratio[4] of conduction-current modulation in the beam at the output point to the velocity modulation injected into the beam at the input point becomes maximum. These lengths are called "optimum drift-tube lengths" and occur for $(\gamma_0 \delta l)$ equal to an odd multiple of $\pi/2$. Furthermore, the phase of the wave transconductance may

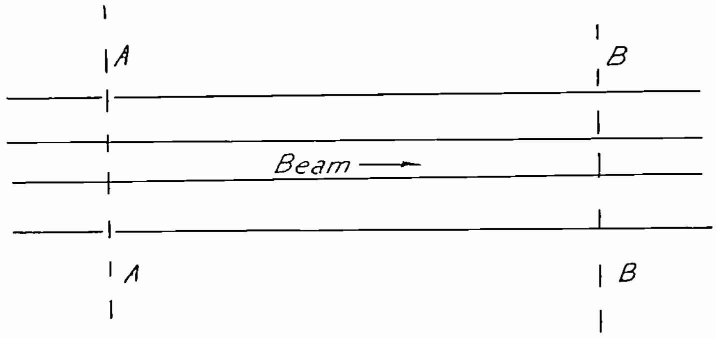

Fig. 2

be any amount depending upon the value of $\gamma_0 l$. Thus we may express the wave transconductance in polar vector form by

$$\begin{aligned} G_W &= \frac{[\psi_{z,f} + \psi_{z,s}]_{B\text{-}B}}{[V_{z,f} + V_{z,s}]_{A\text{-}A}} \\ &= g \sin(\gamma_0 \delta l) \underline{/ -\gamma_0 l - \pi/2} \end{aligned} \tag{6}$$

in which it is shown that the output current at B-B lags the voltage applied at A-A by an angle of $(-\gamma_0 l - \pi/2)$. The above equation also shows that a grid of length $\pi/2$ (i.e., $\gamma_0 l = \pi/2$) has negative resistance characteristics since $(-\gamma_0 l - \pi/2) = -\pi$ for this length of grid.

Further discussion of the limitations and applications of the above concept will be given in later sections of this paper with the derivations.

The Wave Equations

To derive the wave characteristics utilized for the preceding explanation it is necessary to show that the electromagnetic equations, giving the relations existing among the electric and magnetic fields and the charge and current densities, and the equations of mechanics, giving the acceleration of the electrons in terms of the field forces, lead to wave equations which possess the solutions already described. These solutions must be symmetrical about the axis of the beam, they must represent waves propagating in the

[4] This ratio will be termed "wave transconductance" and will be denoted by G_W.

z direction, and they must satisfy the boundary conditions.

The beam will be assumed to be of circular cross section and coaxial with a perfectly conducting cylinder as shown in Fig. 3. In the absence of waves the beam is assumed to possess a uniform distribution of electrons of charge density ρ_0 all traveling in the axial direction with the same constant velocity v_0. The presence of waves will be assumed to cause modulation in the instantaneous velocity of electrons in only the axial direction. In other words, the radial

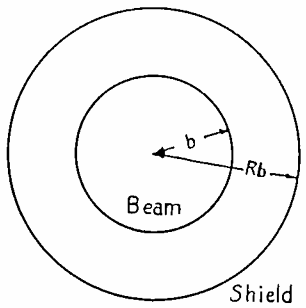

Fig. 3

and azimuthal velocities will be assumed to be zero at all times so that this portion of the following theory applies strictly only to the case in which a very strong axial magnetic-focusing field is employed. At the present time this case is of the greatest importance.

To apply the retarded potentials we need only recall that if in Maxwell's equations the electric and magnetic fields are replaced by the potential functions by means of the following relations:[5]

$$\left.\begin{aligned} \bar{E} &= -\nabla\Phi - \frac{1}{c}\frac{\partial \bar{A}}{\partial t} \\ \bar{H} &= \nabla \times \bar{A} \end{aligned}\right\} \quad (7)$$

in which $\bar{E}$ is the electric-field vector, $\bar{H}$ is the magnetic-field vector, Φ is the electric scalar potential, and $\bar{A}$ is the vector magnetic potential, then the equations reduce to the well-known wave equations[5] for Φ and $\bar{A}$

$$\begin{aligned} \left[\nabla^2 - \frac{1}{c^2}\frac{\partial^2}{\partial t^2}\right]\Phi &= -\rho \\ \left[\nabla^2 - \frac{1}{c^2}\frac{\partial^2}{\partial t^2}\right]\bar{A} &= -\frac{\rho\bar{v}}{c} \end{aligned} \quad (8)$$

in which c is the velocity of light, ρ and $\bar{v}$ are charge density and velocity, respectively, and Heaviside-Lorentz or rational units are used throughout. Equations (8) imply that the divergence of $\bar{A}$ has been determined by[5]

$$\nabla\cdot\bar{A} = -\frac{1}{c}\frac{\partial\Phi}{\partial t}. \quad (9)$$

[5] See for instance chapter 21 of "Introduction to Theoretical Physics," a text by J. C. Slater and N. H. Frank, McGraw-Hill Book Company, Inc., 1933.

Since we are concerned only with wave components which propagate in the z direction with axial symmetry, it will be convenient to denote the scalar electric potential, the charge density, and the velocity by

$$\Phi_1\epsilon^{i(\omega t-\gamma z)}, \quad \rho_0 + \rho_1\epsilon^{i(\omega t-\gamma z)}, \quad v_0 + v_z\epsilon^{i(\omega t-\gamma z)},$$

respectively. Φ_1, ρ_1, and v_z are then functions of r alone. Then using cylindrical co-ordinates the first of equations (8) becomes

$$\frac{\partial^2\Phi_1}{\partial r^2} + \frac{1}{r}\frac{\partial\Phi_1}{\partial r} + (k^2 - \gamma^2)\Phi_1 = -\rho_1 \quad (10)$$

in which $k = \omega/c$.

It is easy to express ρ_1 in terms of Φ_1, for from the continuity equation,

$$\nabla\cdot(\rho\bar{v}) = -\frac{\partial\rho}{\partial t} \quad (11)$$

there is the relation $\omega\rho_1 = \gamma(\rho_0 v_z + v_0\rho_1)$ or

$$\rho_1 = \frac{\gamma\rho_0}{\omega - \gamma v_0}v_z. \quad (12)$$

This neglects modulation cross products of ρ and $\bar{v}$ and thus limits us to small-signal theory. Now

$$m\frac{d\bar{v}_z}{dt} = eE_z \quad (13)$$

in which E_z is the amplitude of the z modulation component of $\bar{E}$, e is the charge, and m is the mass of the electron. Again

$$\frac{dv_z}{dt} = \frac{\partial v_z}{\partial t} + \frac{\partial v_z}{\partial z}\frac{dz}{dt} = i(\omega - \gamma v_0)v_z \quad (14)$$

for small signals and from (7) and (9) E_z is seen to be given by

$$E_z = i[\gamma\Phi_1 - kA_z] = i\left[\gamma - \frac{k^2}{\gamma}\right]\Phi_1 \quad (15)$$

if it is noted that only the z component of $\bar{A}$ can be present since only the z component of $\bar{v}$ is present. Equations (12), (13), (14), and (15) give

$$\rho_1 = \frac{e\rho_0}{m}\frac{\gamma^2 - k^2}{(\omega - \gamma v_0)^2}\Phi_1 \quad (16)$$

so that (10) now becomes

$$\frac{\partial^2\Phi_1}{\partial r^2} + \frac{1}{r}\frac{\partial\Phi_1}{\partial r} + \left[\frac{e\rho_0}{m}\frac{\gamma^2 - k^2}{(\omega-\gamma v_0)^2} + k^2 - \gamma^2\right]\Phi_1 = 0 \quad (17)$$

which is a form of Bessel's equation[6] for functions of zero order whose solution is[7]

[6] See "Bessel Functions for Engineers," N. W. McLachlan, a text. Oxford University Press, London and New York, (1934).

[7] Functions of the second kind are not included in (18) because they become infinite at the origin. J_0 is chosen for (18) while I_0 and K_0 are chosen for (20) with judgment based on the fact that γ, for those slow waves, is expected to be close to γ_0, the propagation constant of the beam. For $\gamma \sim \gamma_0$, T and τ will be appropriately real.

$$\Phi_1 = BJ_0(Tr) \tag{18}$$

where

$$T = \sqrt{(\gamma^2 - k^2)\left[\frac{c\rho_0}{m(\omega - \gamma v_0)^2} - 1\right]} \tag{19}$$

and B is an arbitrary constant.

In the space between the beam and the conducting boundary the charge density is zero so that we may write directly for the electric scalar potential in this region[8]

$$\Phi_2 = C[I_0(\tau r) + DK_0(\tau r)] \tag{20}$$

in which I_0, and K_0 are modified[8] Bessel functions, and

$$\tau = \sqrt{\gamma^2 - k^2}. \tag{21}$$

The constant D is determined by applying the condition at the conductor where $r = bR$ (Fig. 3). Here the tangential-electric field must equal zero, a requirement satisfied by making Φ_2 equal to zero at this radius. Thus

$$D = -\frac{I_0(\tau bR)}{K_0(\tau bR)}\cdot \tag{22}$$

Two boundary conditions remain to be applied at the beam's surface where $r = b$ (Fig. 3). These two conditions will serve to determine the ratio C/B and also the value of γ in terms of the given parameters. Continuity of the tangential electric fields is attained by continuity of the potentials. This gives

$$\frac{C}{B} = \frac{J_0(Tb)}{I_0(\tau b) + DK_0(\tau b)}\cdot \tag{23}$$

For continuity of tangential magnetic field, (7) discloses that since only z components of $\overline{A}$ exist then the only component of $\overline{H}$ is the azimuthal component. Continuity of this component requires continuity of $\partial A_z/\partial r$ which, by (9), leads to continuity of $\partial\Phi/\partial r$. Hence

$$\frac{C}{B} = -\frac{T}{\tau}\,\frac{J_1(Tb)}{I_1(\tau b) - DK_1(\tau b)}\cdot \tag{24}$$

A comparison of (23) and (24) yields

$$-(Tb)\frac{J_1(Tb)}{J_0(Tb)} = (\tau b)\frac{I_1(\tau b) - DK_1(\tau b)}{I_0(\tau b) + DK_0(\tau b)} \tag{25}$$

or more conveniently

$$f_1(Tb) = f_2(\tau b).$$

Propagation Constant

There are probably many systematic ways to arrive at the values of γ which will satisfy (25). Hahn[2] gives a method which has great advantages for design purposes since the procedure is reduced to the handling of probably the fewest number of design parameters. His method also distinguishes between the dependence of the results on tube geometry, beam voltage and density, and frequency. A somewhat different procedure will be followed here because the purpose of the present paper is more to demonstrate the existence of the waves and their characteristics than to indicate design procedure. Accordingly, in seeking the space-charge waves for which $\gamma \approx \gamma_0$ it is

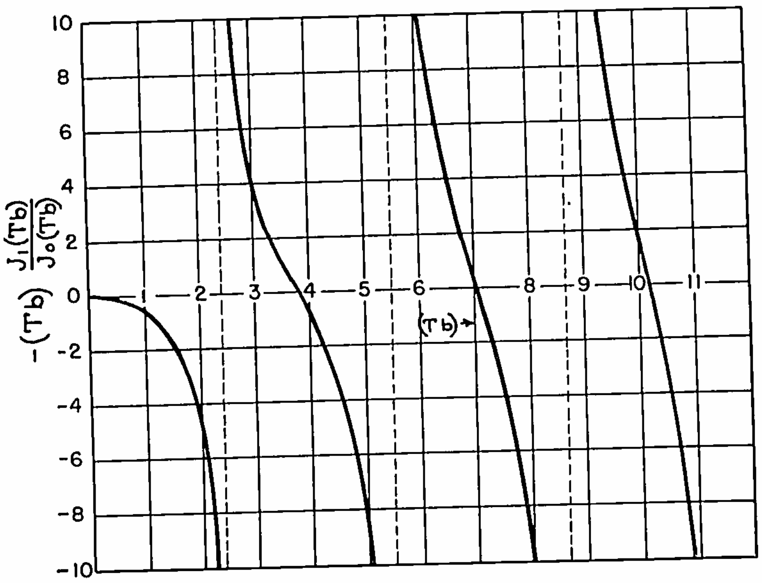

Fig. 4

profitable to note that whereas τ is not very responsive to a small change in γ near γ_0, T varies very rapidly indeed near $\gamma = \gamma_0$ because of the term $(\omega - \gamma v_0)^2$ appearing in a denominator. It would thus appear that for values of γ lying close to γ_0, $\gamma = \gamma_0$ could be substituted in $f_2(\tau b)$ without substantial loss of accuracy in the final determination of γ. Thus, once a given set of tube parameters are chosen, $f_2(\tau b)$ may be regarded as known and it is only necessary to plot $f_1(Tb)$ to find the values of T that will satisfy (25).

An important consideration arises from the fact that if $f_1(Tb)$ is plotted against (Tb), a curve similar to that of Fig. 4 is obtained. $f_2(\tau b)$, which turns out to be a negative quantity, has an infinite number of intersection points with this curve. For each of these values of (Tb) we have from (19) that

$$\frac{T^2}{\gamma^2 - k^2} = \frac{c\rho_0}{m(\omega - \gamma v_0)^2} - 1.$$

Recalling that $\gamma \approx \gamma_0$ for these space-charge waves and that $\gamma^2 \gg k^2$ unless the beam velocity is quite huge, this equation becomes

$$(\omega - \gamma v_0)^2 = \frac{c\rho_0\gamma^2}{m[T^2 + \gamma_0^2]}$$

and finally,

$$\gamma = \frac{\omega}{v_0}(1 \pm \delta) = \gamma_0(1 \pm \delta) \tag{26}$$

in which

$$\delta = \sqrt{\frac{c\rho_0}{m(\omega^2 + T^2 v_0^2)}} \tag{27}$$

[8] See chapter 7 of footnote reference 6.

and will be found to be so small compared to unity for a large portion of practical cases that the assumption made in its derivation ($\gamma \approx \gamma_0$) may be considered as completely justified.

As the value of ρ_0 increases, δ increases and the approximation $\gamma \approx \gamma_0$ becomes poorer. For a more precise solution appropriate to these cases (25) must be re-examined and the substitution of γ_0 for γ in $f_2(\tau b)$ cannot be made. It may be shown[2] that for these cases the space-charge waves have the same general characteristics but the mean velocity of the two waves of the pair departs from the value γ_0 as ρ_0 grows larger.

Wave Characteristics

It is now possible to write the ratio of conduction-current-modulation density to the velocity modulation. Denoting the former by ξ_z, (11) and (26) yield

$$\frac{\xi_z}{v_z} = -\frac{\rho_0}{\pm\,\delta}\,. \qquad (28)$$

Equation (28) shows that the faster wave of the pair, whose velocity exceeds that of the beam ($-\delta$), has its ξ_z and its v_z in time phase while its mate, whose velocity is slightly less than that of the beam, has these two modulations out of phase by π radians. The phase of the conduction-current modulation in the beam is not, however, the same as that of the induced current in the external circuit. This induced current is that on the inner surface of the shielding conducting cylinder which is always equal in magnitude and opposite in phase to the beam-modulation current. (This statement neglects the displacement current which is very small compared to the conduction current. The true total current can be found either by adding the integrated displacement current or by evaluating $\oint \overline{H}\cdot \overline{dl}$ around a circle enclosing the beam and of radius equal to that of the tube since this integral is equal to the total current enclosed.)

It should be clear that since there are an infinite number of values for T there are an infinite number of pairs of waves each pair of which will have the properties previously described. One of the waves of each pair has a velocity of propagation slightly greater than the velocity of the beam while its mate will have a velocity an equal amount less than that of the beam. The pair of waves having the largest departure from the mean velocity is evidently that corresponding to the smallest value of T, for then δ in (27) is maximum. It is these waves in which interest is centered at the present.

To explain the place which the other possible space-charge waves,[9] corresponding to the higher values of T, occupy in the velocity-modulation-tube performance, it will be necessary to summmarize the distinguishing features of the higher-rank waves as compared to the zero-rank wave. Since (Tr) is the parameter of the Bessel function, $J_0(Tr)$, which gives the variation of velocity modulation and conduction-current modulation over the beam cross section, one important difference between high- and zero-rank waves is in their distribution over the beam's area. It will be evident from Fig. 4 that the zero-rank wave is the only one for which $J_0(Tr)$ has no roots between $r=0$ and $r=b$. The higher-rank waves will have consecutively 1, 2, and 3, etc., roots between $r=0$ and $r=b$. Thus the zero-rank wave will be fairly uniform over the cross section while the higher-rank waves will actually fluctuate over the cross section.

Whether a pair of zero-rank waves or higher-rank waves is to propagate down the beam depends of course on whether such waves are started. The means used to start waves at the present is to apply an accelerating voltage gradient over a very short length of the beam by means of a voltage difference applied across a gap in the conducting cylinder. This means of producing waves must impart a velocity modulation to the beam which is quite uniform over the cross section. At least this is certainly true for beams appreciably smaller in radius than the cylinder. It is for this reason that it is believed that only relatively small amplituder of the higher-rank waves are started by this prccess.

Other important differences in the waves are in "optimum-drift tube length" and in "wave transconductance." Since δ is smaller for high values of T the two waves of each higher-rank pair will have velocities closer to the mean than have the two zero waves. Consequently the optimum drift-tube length will go up with T. Equation (28) indicates a higher ratio of ξ_z/v_z for the higher values of T.

Numerical Example

The total conduction-current modulation in the beam is

$$\psi_z = \xi_z\Big|_{r=0}\int_0^b J_0(Tr)2\pi r dr \qquad (29)$$

$$= \frac{1}{\sqrt{4\pi}\times 3\times 10^9}\,\xi_z\Big|_{r=0} 2\pi b^2\,\frac{J_1(Tb)}{Tb}\ \text{(amperes)} \qquad (30)$$

(in which ξ_z remains, of course, in rational units). For the total transconductance in mhos we shall want the ratio of the current in amperes to the velocity modulation in volts V_z. If V_0 is the average beam velocity in volts and v_z and v_0 remain in centimeters per second, then for small signals

$$\frac{v_z}{v_0} = \frac{1}{2}\,\frac{V_z}{V_0}\,. \qquad (31)$$

Thus the optimum total transconductance[10] is

[9] These will be called "higher-rank," waves starting with the lowest T pair as "zero-rank" waves.

[10] Note that this is the ratio of total current over the cross section to the velocity modulation at the center of the beam.

$$G=\left|\frac{\psi_z}{V_z}\right|$$

$$=\frac{v_0}{2V_0}\cdot\frac{1}{\sqrt{4\pi}\times 3\times 10^9}\left.\frac{\xi_z}{v_z}\right|_{r=0}\frac{2\pi b^2 J(Tb)}{(Tb)}\text{ (mhos)} \quad (32)$$

which by the aid of (28) reduces to

$$G=\frac{\rho_0 v_0 \pi b^2}{\sqrt{4\pi}\times 3\times 10^9 V_0}\frac{J_1(Tb)}{\delta(Tb)}\text{ (mhos)}. \quad (33)$$

But $\rho_0 v_0 \pi b^2/\sqrt{4\pi}\times 3\times 10^9 = I_0$, the average beam current in amperes. So G may be written

$$G=\frac{I_0}{V_0}\frac{J_1(Tb)}{\delta(Tb)}\text{ (mhos)}. \quad (34)$$

Approximate values of the needed parameters for a particular receiver amplifier tube are

diameter of conducting cylinder, $2Rb$	½ inch
diameter of beam, $2b$	¼ inch
total beam current, I_0	10 milliamperes
drift-tube voltage, V_0	1500 volts
operating frequency, $\omega/2\pi$	1000 megacycles

Substitution of these values into the equations which have been derived results in $(Tb)=2.4$ and $\delta=0.03$. J_1 $(2.4)=0.52$, so that

$$G=\frac{10\times 10^{-3}}{1.5\times 10^3}\frac{0.52}{3\times 10^{-2}\times 2.4}=48\times 10^{-6}\text{ (mhos)}.$$

The optimum drift-tube length is

$$l_0=\frac{\pi}{2\gamma_0\delta}=19.2\text{ centimeters}.$$

The transconductance computed above is that associated with the wave only. The over-all tube transconductance is this figure multipled by three other quantities:

1. The number of gaps used in series as an input grid. For a single grid of π electrical degrees length this figure would be 2.
2. The number of gaps associated with the output grid. For a single π output grid this figure would also be 2.
3. The "coefficient" of the gaps which will be somewhat less than unity because not all the voltage applied across the gap will be received by the electrons. The coefficient approaches unity as the transit-time of electrons in traversing the region of influence of the gap decreases to zero. (This statement neglects the fact that small amounts of higher-rank waves are in general also excited at the gap and "use up" some of the input voltage, a portion of this apparent loss perhaps being made up by the subsequent contribution to output current of these higher-rank waves.)

An Experimental Investigation of the Characteristics of Certain Types of Noise*

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Summary—*The results of an investigation of the effect of the band width on the effective, average, and peak voltages of several different types of noise are given for band widths up to 122 kilocycles. For atmospheric noise and that due to the thermal agitation of electric charge in conductors, both of which consist of a large number of overlapping pulses, the peak, average, and effective voltages were all proportional to the square root of the band width. For very sharp, widely separated, clean, noise pulses, the average voltage was independent of the band width and the peak voltage was directly proportional to the band width. For noise of a type falling between these two the effect of the band width depended upon the extent of the overlapping.*

The ratio of the peak to effective voltage of the noise due to the thermal agitation of electric charge in conductors was measured and found to be 4. The ratio of the average to effective voltage of this type of noise was found to be 0.85.

The experiments showed that when a linear rectifier, calibrated by a continuous-wave signal having a known effective voltage, is used to measure the effective voltage of this type of noise the measurements should be increased by ½ decibel to obtain the correct result.

Introduction

NOISE, as defined by the dictionary, is sound, especially sound without agreeable musical quality. However, in the parlance of the communication engineer the term gradually came to be used to designate, not sound itself, but those electrical currents which caused undesired sounds to appear at the output of a telephone system, and, hence, is now used to designate those currents which cause interference in any communication system. It is in the latter sense that the word is used throughout this paper. Thus, noise may vary all the way from a single-frequency continous-wave signal to a series of completely random discontinuous disturbances such as those due to the thermal agitation of electric charge in conductors.[1]

The characteristics of a single-frequency signal, or of one composed of any given number of definite frequencies, are well known and will not be discussed here. This paper will be confined to a study of those types of noise which are more or less discontinuous

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[1] Hereafter referred to as "thermal noise."

and which, therefore, have a more or less infinite and continuous spectrum.

Those characteristics of noise which are readily measurable and which, accordingly, are most frequently used for measuring purposes, are the peak voltage obtained over a given time interval, the effective or root-mean-square voltage, the average voltage, and, occasionally, the frequency of occurrence of voltage peaks above a specified minimum value. When the noise has a more or less infinite and continuous spectrum, the values obtained for these voltages will depend to a large extent upon the band width of the equipment through which the noise has passed before reaching the point where the measurement is made. It is the purpose of this paper to present the results of an investigation of this effect of the band width on these various voltages. This investigation was carried on intermittently over a period of several years at the Holmdel Radio Laboratories of Bell Telephone Laboratories, Inc.

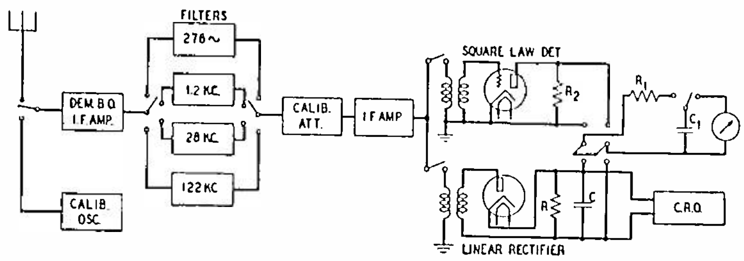

Fig. 1—Schematic diagram of the measuring equipment.

Apparatus and Experimental Procedure

Fig. 1. shows a block diagram of the apparatus used. It consisted of a short-wave, double-detection, measuring set of conventional design except for the following feature. By the use of four separate intermediate-frequency filters, two of which contained crystal elements, four different intermediate-frequency band widths were made available. The effective band widths of these filters were about 276 cycles, 1.2 kilocycles, 28 kilocycles, and 122 kilocycles, all centered on a frequency of about 2.0 megacycles. These band widths were determined by dividing the total area under the energy response curve (output current squared plotted against frequency) of the receiver by the height of the curve at resonance, thus obtaining the width of the equivalent rectangular band. The curves given in Fig. 2 are the four response curves of the receiver.

The noise being measured was picked up on an antenna at some frequency between 10 and 18 megacycles, reduced to the intermediate frequency by heterodyning with a local oscillator, and then detected. No carrier was used when the measurements were made. A linear rectifier or a square-law detector could be used either separately or simultaneously as desired. Relative gains of the different branches were measured before and after each series of measurements, by means of a signal obtained from a local oscillator. With this equipment the relative peak, average, and effective voltages obtained with the different band widths were measured for several different types of noise.[2]

By effective voltage is meant that voltage defined by the equation

$$E_{rms} = \sqrt{\frac{1}{T}\int_0^T e^2 dt} \tag{1}$$

where E_{rms} is the effective voltage acting over the time interval T and e is the instantaneous value of the intermediate-frequency voltage. Relative effective voltages were measured by connecting the resistance R_1 and the condenser C_1 (see Fig. 1) in series and across the resistance R_2 in the plate circuit of the square-law detector. After a definite length of time, short in comparison with the time constant R_1C_1, this condenser was discharged through a ballistic galvanometer the deflections of which were, then, proportional to the square of the effective voltage.

By average voltage is meant that voltage defined by the equation

$$E_a = \frac{1}{T}\int_0^T \sqrt{e^2}\, dt \tag{2}$$

where E_a is the average voltage and T and e have the same meaning as before. To measure the relative average voltages, the resistor-and-condenser combination R_1C_1 was connected across the resistor R in the rectifier circuit. After a definite time interval, again short in comparison with the time constant R_1C_1, the charge on the condenser C_1 was measured with the ballistic galvanometer as before, the deflec-

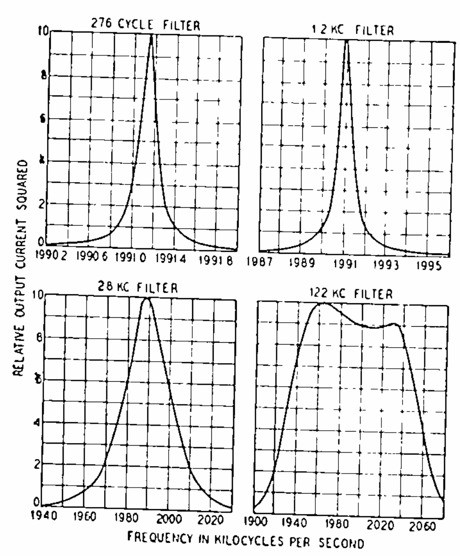

Fig. 2—Characteristics of intermediate-frequency filters.

[2] V. D. Landon in "A study of the characteristics of noise," Proc. I.R.E., vol. 24, pp. 1514–1521, November, (1936), has given data on the variation with the band width of the peak voltage of noise due to the thermal agitation of electric charge and of that due to a sharp voltage impulse, the measurements being made with two different band widths.

tions of which, this time, were proportional to the average voltage.[3]

By peak voltage is meant the peak value reached by the instantaneous intermediate-frequency voltage e. Relative peak voltages were measured by means of a small cathode-ray oscilloscope, the vertical deflecting plates of which were connected across the resistor R.

In making the measurements of the effective and average voltages, it was necessary to use the ballistic galvanometer for the reason, that, when the wider band widths were used to measure noise for which the individual pulses occurred at relatively infrequent intervals, the power was insufficient to give a readable indication with the usual thermocouple- or rectifier-type meter. However, such instruments are entirely suitable for measuring steady noise such as thermal noise.

The noise studied was obtained from the following sources:

1. The thermal agitation of electric charge in the early circuits of the receiver.
2. Atmospherics.
3. A 1000-cycle buzzer.
4. A sharp voltage impulse.
5. The ignition system of an automobile.

For the measurements on atmospheric noise the receiver was tuned to some frequency around 10 megacycles which, at the time, was free of any unwanted signals.

To obtain the "buzzer noise" a 1000-cycle buzzer was placed within a few feet of the antenna.

The sharp voltage impulse was obtained by discharging a condenser two times a second by a mercury-in-vacuum switch, through an inductance which was coupled to the first circuit of the receiver. The condenser and the inductance were of such a size that when connected together the resonant circuit so formed was tuned to the same frequency as the first circuit of the receiver.

To obtain the automobile-ignition noise, an automobile was driven to within a few yards of the receiving antenna and the throttle set so that the motor ran at a speed corresponding to a car speed of about 25 miles per hour.

In this paper no attempt will be made to go into a theoretical analysis of the results to be expected for the different types of noise beyond pointing out that, for a sharp voltage pulse applied to a simple, series-tuned circuit such as would be obtained by the discharge of a condenser through such a circuit, the peak voltage should be directly proportional to the band width, the effective voltage should be proportional to the square root of the band width, and

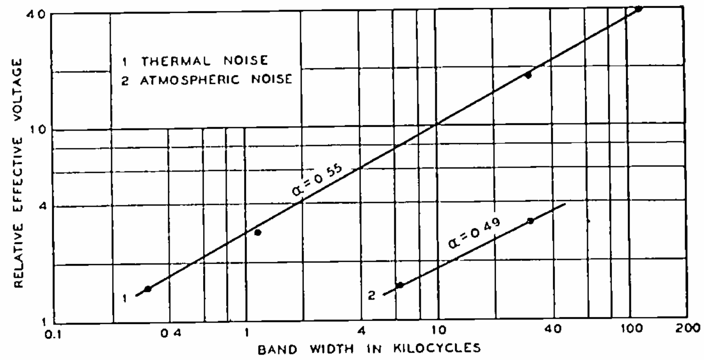

Fig. 3—Effect of band width on effective voltage of noise.

the average voltage should be independent of the band width.[4]

Results

Effective Voltage

The variation of the effective voltage with the band width was studied for thermal noise and for atmospheric noise. The results are given in Fig. 3. The data given for atmospheric noise were obtained in 1934 with apparatus having band widths of 6.4 and 31.8 kilocycles. Relative effective voltages are plotted as ordinates and band widths as abscissas. If the relationship between the band width B and the effective voltage E_{rms} is expressed by the equation

$$E_{rms} = k_1 B^{\alpha} \tag{3}$$

then, since the abscissas and ordinates are plotted on logarithmic scales, the value of α for a particular type of noise is given by the slope of the curve for that noise. The values obtained are given in the figure. The difference between the slopes of the two curves in this case is not considered significant.

Because of the limitations of the apparatus, it was impossible to measure the relative effective voltages for those types of noise which consisted of sharp, widely separated pulses. Thus, no curves are given for ignition noise or that obtained from the buzzer or from the sharp voltage impulse. However, the band widths were measured in such a way that α should have been approximately equal to 0.5 for all cases, since the energy received from this type of noise over the limited range of band widths used is approximately proportional to the band width.

Peak Voltage

The effect of the band width on the peak voltage was studied for all the different types of noise. The results are given in Fig. 4 where, as before, band widths are plotted as abscissas and, this time, rela-

[3] For these measurements and for those of the peak voltage described later, the time constant RC of the rectifier circuit was made small enough so as not to distort the shape of the noise pulse, but, to keep the intermediate-frequency currents flowing in the leads to the oscilloscope as small as possible, it was necessary that the capacitance of C be fairly large. This resulted in a time constant for RC which was such that the voltage across R was more nearly equal to the envelope of the intermediate frequency than to the average voltage as defined above. However, since the narrowest noise pulse studied was still wide enough to contain several cycles of the intermediate frequency, it is believed that the average rectifier output was proportional to the average voltage.

[4] See Appendix.

tive peak voltages as ordinates. If the relationship between the band width B and the peak voltage E_p is expressed by the equation

$$E_p = k_2 B^\beta \tag{4}$$

then the value of β for any particular noise is given by the slope of the curve for that noise. The values obtained are given in the figure.

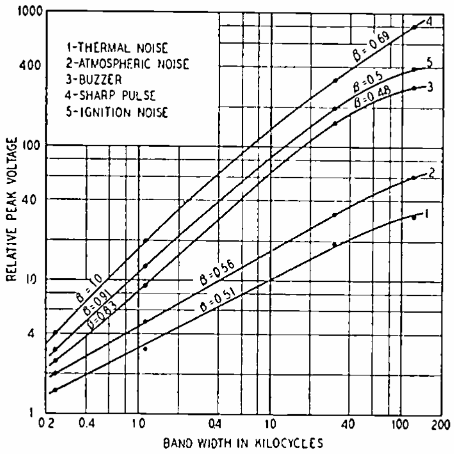

Fig. 4—Effect of band width on peak voltage of noise.

Experience gained in the taking of the data showed that the results were definitely affected by the spacing of the individual noise pulses and by the width and shape of the generated pulse in relation to the band width of the receiver. Thus, if the spacing was at random and so small that there was considerable overlapping of the intermediate-frequency wave trains, the value of β was found to be around 0.5. For larger spacings, the value of β was greater, and for spacings so large that there was no overlapping, even when the narrowest filter was used, the value of β was found to be approximately 1. On the other hand it is evident that if any particular noise pulse were wide enough and of such a shape as to be passed without alteration by any particular filter, any further widening of the band would not affect the peak voltage. The curve of peak voltage plotted against band width for this particular noise would, then, gradually bend over and become parallel to the axis of band widths. Theoretically, for a perfectly discontinuous noise pulse having an infinite spectrum this condition would never be reached, but actually, as will be seen later, the conditions under which some of the noise pulses were generated were apparently such that the range of the component frequencies was definitely limited.

Thermal noise and atmospheric noise are good examples of those types for which the spacing is so small that the overlapping is considerable. For both, the curves in Fig. 4 are very nearly straight lines, the slopes of which are approximately 0.5. The slight departure from linearity could be due to the fact that, because of the increasing sharpness of the peaks of the pulses as the band width was increased, it was increasingly difficult to determine exactly the height of the pulse on the screen of the cathode-ray tube, the tendency being to record values which were somewhat low.

As far as could be determined visually the pulses obtained from these two types of noise were always as narrow as could be passed by the filter being used, indicating that even with the widest filter the width of the band was still a limiting factor as regards the pulse height and width.

The pulses generated by the discharge of a condenser were separated to such an extent that there was no overlapping regardless of the filter used. The curve for this noise has a slope approximately equal to 1.0 for the narrower band widths, but for the wider widths the slope is much less. Although some of this reduction in the slope of the curve could be due to the difficulty in measuring exactly the height of the pulses, in addition, for this noise and for that generated by the buzzer and by the automobile-ignition system, the pulses as they appeared on the face of the cathode-ray tube when the 122-kilocycle filter was being used were slightly wider than the narrowest pulse which could, theoretically, be passed by that filter indicating that the band width was no longer the limiting factor. This, no doubt, accounts to a large extent for the reduction in the slope of the curves for these three types of noise at the wider band widths.[5]

The noise pulses from the buzzer occurred in spurts at the rate of 1000 spurts per second. Each spurt consisted of one large pulse accompanied by several much smaller ones. Although there was considerable overlapping of these small pulses and the associated large pulse, the effect on the slope of the curve should not have been great because of the difference in size. However, when the 1.2-kilocycle filter was used, each of the large pulses was broadened out to such an extent that they overlapped slightly, and when the 276-cycle filter was used this overlapping was considerable. The curve for this noise, then, should have a relatively low slope for the narrower band widths and a greater slope for the wider band widths. Curve 3 of Fig. 4 does show a comparatively small slope for the narrower band widths and a slight increase for the middle portion, but, in accordance with the explanation given above, the slope is much less for the wider band widths.

The automobile-ignition noise was very similar to that generated by the buzzer in that it consisted of periodically occurring spurts made up of one large pulse and several smaller ones, but the time interval

[5] As is shown in the Appendix, for the case of a simple series-tuned circuit and a perfectly discontinuous pulse, β should be equal to 1. The 276-cycle and 1.2-kilocycle filters have frequency characteristics which are very nearly the same as would be obtained with such a circuit. That of the 28-kilocycle filter is slightly different, but that of the 122-kilocycle filter is considerably different. Although this change in the frequency characteristic should have no effect on the value of α (equation (3)), it may well affect the value of β obtained at the wider band widths.

between the spurts was such that there never was any overlapping of the larger pulses. It would be expected, therefore, that the curve for this noise would have a slope similar to that for the noise from the buzzer for the wider band widths but a greater slope for the narrower widths. Inspection of curve 5 of Fig. 4 shows that just such a curve was obtained for this noise.

Average Voltage

The effect of the band width on the average voltage was studied for all of the different types of noise except ignition noise. The results are given in Fig. 5. If the relationship between the band width B and the average voltage E_a is expressed by the equation

$$E_a = k_3 B^{\gamma} \tag{5}$$

then the value of γ is given by the slope of the curves. The values obtained are given in the figure.

Here, as before, the extent of the overlapping of the pulses affected the results. For thermal noise and for atmospheric noise, γ is approximately equal to 0.5 whereas the slope of the curve for the artificially generated pulse is zero which is just what the theory shows should be obtained in the case of a perfectly discontinuous pulse applied to a simple, series-tuned circuit.

The data obtained with the buzzer are especially interesting. As mentioned before, the spacing of the pulses was such that with the 122- and 28-kilocycle filters there was no overlapping, with the 1.2-kilocycle filter there was slight overlapping, and with the 276-cycle filter the pulses were drawn out to such an extent that several pulses overlapped at any given instant. For the narrow band widths the curve has an appreciable slope, but for the wide band widths, where there is no overlapping, the curve levels off and becomes nearly parallel to the axis of abscissas like that for the artificially generated pulses.

Miscellaneous Measurements

Thermal Noise

Since, for the band widths normally used, and for any appreciable period of observation, the peak voltage, average voltage, and effective voltage obtained from thermal noise are practically constant and independent of the length of the period of observation, then there should be a definite relationship existing among these voltages. Measurements made of these relationships gave the value 4 for the ratio of the peak to effective voltage[6] and the value 0.85 for the ratio of the average to effective voltage.

In many cases it is desirable to know the error involved when a linear rectifier is used to measure the effective voltage of this type of noise. If the measuring equipment is calibrated with a continuous-wave signal the effective voltage of which is taken as the calibrating voltage, then, since the ratio of the average voltage to the effective voltage of a continuous-wave signal is 0.9 while the ratio of the average voltage to the effective of this type of noise is 0.85, the readings for the noise will be 0.90/0.85, or $\frac{1}{2}$ decibel, too low and must be increased by this amount to give the correct effective voltage.

Unpublished experiments by C. B. Feldman of these laboratories made at audio frequencies with a full-wave rectifier gave the same result.

Atmospheric Noise

Observations on the wave form of atmospheric noise have shown that there is considerable overlapping of the individual discharges even for ultra-short-wave noise received through the 122-kilocycle filter. Accordingly, this noise would be expected to act very much like thermal noise as far as the relations between the various voltages and the band width are concerned. The data given above have shown that such is the case. The spasmodic character of atmospheric noise, however, would cause a wide range of values to be obtained for the ratio of peak to effective voltage, the value depending upon the nature of the atmospherics during the period of observation and the length of that period. For this reason it would not be expected that the ratio of the average to effective voltage would have a constant value either as is the case for thermal noise. As a matter of interest, however, a few measurements were made of this ratio. They gave results varying from 0.55 to 0.8.

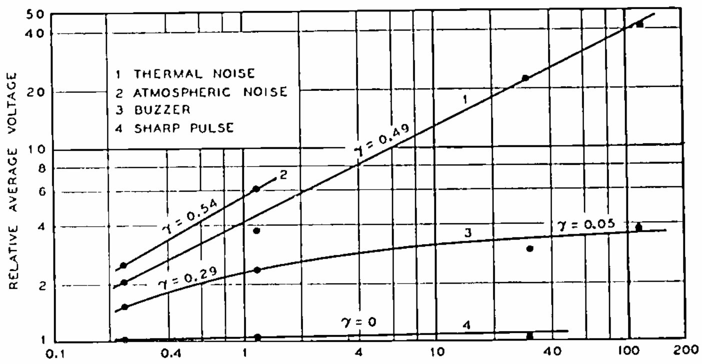

Fig. 5—Effect of band width on average voltage of noise.

Since the characteristics of atmospheric noise, are so very similar to those of thermal noise, it would seem likely that the interfering effect of the former could be determined by a comparison between it and the latter, the interfering effect of which can be very easily determined. Thus, if the carrier-to-peak-thermal-noise ratio needed for a good circuit is known, then, the same ratio in the case of atmospheric noise should always give a satisfactory signal. This ratio would be required when the atmospheric noise is very steady and continuous, but when it is very intermittent with only an occasional loud crash, the ratio could be made considerably lower (it could even be made less than one) without unduly impair-

[6] Landon (footnote 2) states that he obtained the value 3.4 for this ratio whereas a coworker obtained 4.47.

ing the intelligibility. But, the ratio of the peak to effective (or average) voltage tells us much about the nature of the noise and together with the carrier-to-peak-noise ratio may be all the data needed to give accurately the interfering effect of atmospheric noise.

Conclusion

Data have been given above on the relationships between the effective, peak, and average voltages of various types of noise and the band width of the

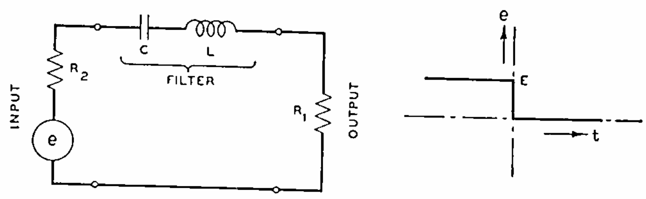

Fig. 6—Series-tuned circuit and form of applied voltage.

receiver for band widths up to 122 kilocycles. It was shown that these relationships depend to a large extent upon the shape and width of the noise pulse as it is generated and upon the overlapping of the individual pulses either before or after passing through the receiving filter. It should be mentioned that the relationships found for this limited band-width range may not hold for much wider band widths such as will be used for television purposes. However, in the case of thermal noise and atmospheric noise at least, it does not seem likely that even these wider bands will be of sufficient width to pass the noise pulses without some alteration, or to separate the individual noise pulses to such an extent that the overlapping will be negligible.

Appendix

The effective band width B of a filter consisting of a series-tuned circuit, such as is shown at (a) in Fig. 6, is given by the equation

$$B = \frac{\int_0^\infty I_f^2 df}{I_0^2} \tag{1}$$

where I_f is the root-mean-square current which flows in the circuit in response to a fixed voltage of frequency f, and I_0 is the root-mean-square current which flows in the circuit in response to the same voltage when the frequency is the resonant frequency of the circuit.

Now[7]

$$\int_0^\infty I_f^2 df = \int_0^\infty \frac{E^2 df}{R^2 + \left(2\pi f L - \dfrac{1}{2\pi f C}\right)^2} = \frac{E^2}{4RL}$$

and

$$I_0^2 = E^2/R^2$$

therefore

$$B = R/4L \text{ cycles per second} \tag{2}$$

[7] D. Bierens de Haan, "Nouvelles Tables D'Integreles Definies." P. Engils, Leide, 1867. Table 20, equation (7).

where R and L are the total resistance and inductance of the circuit.

If now the voltage applied to the circuit by the generator is of the form shown at (b) Fig. 6, then the instantaneous current flowing in the circuit is given by the well-known equation

$$i = \frac{-E}{I\omega} e^{-(R/2L)t} \sin \omega t \tag{3}$$

where $R = R_1 + R_2$

and

$$\omega = \sqrt{\frac{1}{LC} - \frac{(R)^2}{(2L)^2}} \doteq \frac{1}{\sqrt{LC}}.$$

To determine the effect of the band width on the peak, effective, and average voltages, let R_1, R_2, and ω be kept constant and the band width varied by varying L and C. The peak voltage appearing across the output circuit is

$$E_p = R_1 i_{\max} = \frac{R_1 E}{L\omega} = \frac{4R_1 E}{R\omega} B \tag{4}$$

which is directly proportional to the band width.

The effective voltage appearing across the output circuit is given by the equation

$$E_{rms} = \sqrt{\frac{1}{T}\int_0^T R_1^2 i^2 dt}$$

where T is the period over which the effective voltage is measured. If T is made large with respect to $2L/R$, then

$$E_{rms} = \frac{R_1 E}{\omega\sqrt{2RLT}} = \sqrt{\frac{2}{T}} \frac{R_1 E}{R\omega} \sqrt{B}. \tag{5}$$

The effective voltage is, therefore, proportional to the square root of the band width.

The average voltage was defined by the equation

$$E_a = \frac{1}{T}\int_0^T \sqrt{e^2}\, dt$$

but it was also pointed out that, under the conditions of the experiments, this was proportional to the envelope of the intermediate frequency. For this case, then, the average voltage would be given by the equation

$$E_a = k \frac{1}{T}\int_0^T \frac{R_1 E}{L\omega} e^{-(R/2L)t} dt.$$

Again, if T is made large with respect to $R/2L$

$$E_a = \frac{2kR_1 E}{T\omega R} \tag{6}$$

and is independent of the band width.

Biconical Electromagnetic Horns*

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AND J. J. JANSEN†, NONMEMBER, I.R.E.

Summary—*Horn antennas for radiating uniformly in a plane are described. The propagation of waves in the biconical-shaped horn, and their radiation from it, are treated analytically. An experimental investigation at a wavelength of 8.3 centimeters shows the detailed behavior of this unique horn, which should find applications to services employing "broadcast" radiation at ultra-high frequencies.*

Introduction

THIS paper deals mainly with electromagnetic-horn "antennas" that are particularly adapted for radiating or absorbing uniformly in a plane and which appear to have possible applications to radio services utilizing broadcast transmission at ultra-high frequencies. Simplicity of construction and adjustment and ability to operate over a broad band of frequencies are outstanding features, as they are generally of antennas of the electromagnetic-horn type.

A description of the essential elements of these horns was published in 1936.[1] It comprises a rotationally symmetrical structure having means for sending or receiving at the center between the top and bottom members, as illustrated in Fig. 1A. These members have smooth metallic surfaces that are close together at the center but that flare with increasing distance from the center to a spacing of several wavelengths at their outer edge; one half of the cross-sectional profile resembles that of a simple horn, and the horns of this rotational shape may be thought of as generated by rotating a simple horn through 360 degrees about the vertical axis. The shape of the profile may be curved or straight, depending on the requirements of the problem at hand. We have termed this general type of horn "biconical," thereby giving a sufficiently broader significance than is usual to the term biconical to include all shapes like those of Fig. 1. The gently flaring shapes of B and C can provide a relatively wide frequency response and would be particularly applicable to television and multiband operation (for example, the latest RCA television antenna[2] bears a certain resemblance to C), although they may not give as sharp radiation patterns as horns of straight profiles of equal aperture. It is not essential that the two members be similar; at E and F, modifications are illustrated that have dissimilar members and that concentrate the radiation on a conical surface turned downward in E and upward in F. The simplest profile from both constructional and theoretical viewpoints has straight sides, as shown in Fig. 1D, E, and F. The two members are conical in this case. The analysis and the experiments reported in this paper are confined to horns of straight-line profile having two similar members, Fig. 1D.

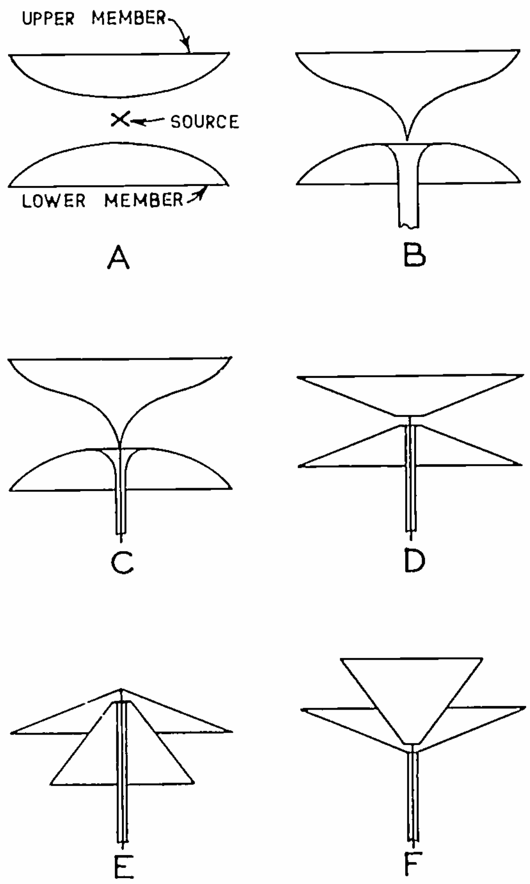

Fig. 1—Profiles of biconical horns.

As in the case of other electromagnetic horns, here also a number of distinct types of waves may exist. The excitation of a particular wave is accomplished by appropriate exciting means near the center or "throat" of the horn. Such means may be actual generating apparatus disposed between the two members, or it may be a system of exciting rods with a transmission-line connection of either hollow-pipe or conventional two-conductor kind, as illustrated in Figs. 1B and C, respectively, which are reproduced from the 1936 paper. For practical applications, the two lowest-order waves,[3] the transverse electromagnetic TEM (or $E_{0,0}$) and the transverse electric $TE_{0,1}$ (or $H_{0,1}$), respectively, appear to be of main importance. The former provides vertically polarized

* Decimal classification: R111.2. Original manuscript received by the Institute, August 15, 1939. A part of this research was conducted as a thesis in the Department of Electrical Engineering at M.I.T. by J. J. Jansen, May, 1939.
† Massachusetts Institute of Technology, Cambridge, Mass.
[1] W. L. Barrow, "Transmission of electromagnetic waves in hollow tubes of metal," Proc. I.R.E., vol. 24, pp. 1298-1328; October, (1936).
[2] N. E. Lindenblad, "Television transmitting antenna for Empire State Building," *RCA Rev.*, vol. 3, pp. 387-408; April, (1939).
[3] The nomenclature suggested by Schelkunoff will be used in this paper. See S. A. Schelkunoff, "Transmission theory of plane electromagnetic waves," Proc. I.R.E., vol. 25, pp. 1457-1492; November, (1937).

radiation and the latter horizontally polarized radiation when the principal axis of the horn is vertical. The experiments to be reported here concern the *TEM* wave only.

In this paper, we shall first present the results of an analysis of the transmission of waves within the horn and of its external radiation characteristics. This analysis is quite parallel to that previously given[4,5] for the sectoral horn. In addition to its bearing on the biconical horn, this analysis may be considered as the direct solution from Maxwell's equations for transmission on a coaxial line whose inter-conductor spacing varies directly with its length, i.e., a special case of a "tapered" coaxial line. Next, we shall present the more important results of an experimental investigation of a biconical horn made at a wavelength of about 8.3 centimeters. Finally, we shall conclude with some remarks on further modifications and applications.

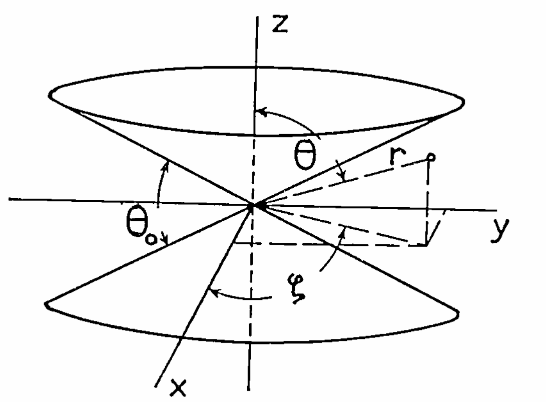

Fig. 2—Co-ordinate system used for the biconical horn.

Theory of Biconical Waves

Fig. 2 shows a spherical co-ordinate system r, θ, ζ in relation to the biconical-horn structure. The principal axis of the horn coincides with the Z axis, and the sides of the cones, if extended to their apexes, would meet at the origin. In a practical horn, the exciting means are situated at or near the origin, which is a singular point in the analysis. This region is not included in the present analysis. Waves are excited near the origin and propagate outward in the space between the cones. The transmission properties of the waves[6] within the horn may be obtained by making the appropriate solutions of Maxwell's equations in spherical co-ordinates satisfy the boundary conditions for perfect conductors on the inner surfaces of the cones.

Horn waves[4] of the transverse electric type *TE* (or *H*) have no radial component of electric intensity. They may be obtained by the use of a *magnetic* Hertzian vector[7] that has only a radial component, $\Pi = i_r \cdot \Pi_r$. The electric intensity E and the magnetic intensity H are then found by means of the relations

$$
\begin{aligned}
H &= \omega^2\mu\epsilon\Pi + \operatorname{grad}\frac{\partial}{\partial r}\Pi_r \\
E &= -j\omega\mu \operatorname{curl}\Pi.
\end{aligned}
\qquad (1)
$$

The MKS system of units is used, where H is in amperes per meter, E in volts per meter, μ (permeability of space within the horn) in henrys per meter, ϵ (dielectric constant of space within the horn) in farads per meter, $\omega = 2\pi$ times the frequency in cycles per second, $j = \sqrt{-1}$, and i_r is the unit vector in the radial direction. For air or vacuum, $\mu = 4\pi \times 10^{-7}$ and $\epsilon = 1/(36\pi \times 10^9)$. The Hertzian vector must satisfy the wave equation

$$
\frac{\partial^2 \Pi_r}{\partial r^2} + \frac{1}{r^2 \sin\theta}\frac{\partial}{\partial\theta}\sin\theta\frac{\partial}{\partial\theta}\Pi_r + \frac{1}{r^2\sin^2\theta}\frac{\partial^2}{\partial\zeta^2}\Pi_r + k^2\Pi_r = 0 \qquad (2)
$$

whose solution is as follows:

$$
\Pi_r = \begin{matrix}\cos\\ \sin\end{matrix}(m\zeta)\sqrt{r}\,K_p(kr)L_l^m(\cos\theta) \qquad (3)
$$

where

m and l are real numbers
K_p = Hankel function of the second kind
$L_l^m(\cos\theta) = AP_l^m(\cos\theta) + BQ_l^m(\cos\theta)$
P_l^m = associated Legendre's function of the first kind[8]
Q_l^m = associated Legendre's function of the second kind
$p = [l(l+1)+1/4]^{1/2}$
A and B are amplitude constants
$k = \omega\sqrt{\mu\epsilon}$.

The Hankel function of the second kind must be used to represent a wave that is propagated outwardly from the apex. The values of A, B, l, and m are determined by the conditions at the boundary. From (1) and (3) we find the field to be given by

$$
\begin{aligned}
H_r &= \begin{matrix}\cos\\ \sin\end{matrix}(m\zeta)L_l^m(\cos\theta)(p^2 - \tfrac{1}{4})r^{-3/2}K_p(kr) \\
H_\theta &= -\begin{matrix}\cos\\ \sin\end{matrix}(m\zeta)\frac{\partial}{\partial\theta}L_l^m(\cos\theta) \\
&\qquad [(p - \tfrac{1}{2})r^{-3/2}K_p(kr) - kr^{-1/2}K_{p-1}(kr)] \\
H_\zeta &= \pm m\begin{matrix}\sin\\ \cos\end{matrix}(m\zeta)\frac{1}{\sin\theta}L_l^m(\cos\theta) \\
&\qquad [(p - \tfrac{1}{2})r^{-3/2}K_p(kr) - kr^{-1/2}K_{p-1}(kr)] \\
E_r &= 0 \\
E_\theta &= \pm j\omega\mu m\begin{matrix}\sin\\ \cos\end{matrix}(m\zeta)\frac{1}{\sin\theta}L_l^m(\cos\theta)r^{-1/2}K_p(kr) \\
E_\zeta &= j\omega\mu\begin{matrix}\cos\\ \sin\end{matrix}(m\zeta)\frac{\partial}{\partial\theta}L_l^m(\cos\theta)r^{-1/2}K_p(kr).
\end{aligned}
\qquad (4)
$$

[4] W. L. Barrow and L. J. Chu, "Theory of the electromagnetic horn," Proc. I.R.E., vol. 27, pp. 51–64; January, (1939).
[5] L. J. Chu and W. L. Barrow, "Electromagnetic horn design," *Elec. Eng.*, vol. 58, pp. 333–338; July, (1939).
[6] Since completing this work, a paper by Schelkunoff has discussed certain of the theoretical aspects of this problem. See S. A. Schelkunoff, "Transmission of spherical waves," Bell System Monograph, B-1092, (1939).
[7] A. Sommerfeld in Rieman-Weber, vol. II, p. 496, 7th edition, F. Vieweg und Sohn, Braunschweig, (1927).
[8] E. W. Hobson, "Spherical and Ellipsoidal Harmonics," Cambridge University Press, London, England, (1931).

The boundary conditions for the symmetrical biconical horn require the vanishing of E_ζ at $\theta=(\pi\pm\theta_0)/2$, where θ_0 denotes the flare angle between the cones, thus:

$$\frac{\partial}{\partial\theta}L_l^m(\cos\theta) = 0 \quad \text{at} \quad \theta = (\pi \pm \theta_0)/2. \tag{5}$$

This equation is satisfied by choosing proper values of l for given values of m. For unsymmetrical horns, the boundary conditions must be satisfied at the values of θ corresponding to the angles of the cones.

Horn waves of the transverse magnetic TM (or E) type have no radial component of magnetic intensity. They may be obtained by the use of an *electric* Hertzian vector[7] that has only a radial component. The electric and magnetic intensities are then found from the relations

$$H = -j\omega\mu \,\mathrm{curl}\,\Pi$$

$$E = \omega^2\mu\epsilon\Pi + \mathrm{grad}\frac{\partial}{\partial r}\Pi_r. \tag{6}$$

The same wave equation (2) is valid. The expressions for the field of the E waves are found to be given by

$$H_r = 0$$

$$H_\theta = \mp j\omega\epsilon m \begin{matrix}\sin\\ \cos\end{matrix}(m\zeta)\frac{1}{\sin\theta}L_l^m(\cos\theta)r^{-1/2}K_p(kr)$$

$$H_\zeta = -j\omega\epsilon \begin{matrix}\cos\\ \sin\end{matrix}(m\zeta)\frac{\partial}{\partial\theta}L_l^m(\cos\theta)r^{-1/2}K_p(kr)$$

$$E_r = \begin{matrix}\cos\\ \sin\end{matrix}(m_\zeta)L_l^m(\cos\theta)(p^2-\tfrac{1}{4})r^{-3/2}K_p(kr) \tag{7}$$

$$E_\theta = -\begin{matrix}\cos\\ \sin\end{matrix}(m\zeta)\frac{\partial}{\partial\theta}L_l^m(\cos\theta)\left[(p-\tfrac{1}{2})r^{-3/2}K_p(kr) - kr^{-1/2}K_{p-1}(kr)\right]$$

$$E_\zeta = \pm m \begin{matrix}\sin\\ \cos\end{matrix}(m\zeta)\frac{1}{\sin\theta}L_l^m(\cos\theta)\left[(p-\tfrac{1}{2})r^{-3/2}K_p(kr) - kr^{-1/2}K_{p-1}(kr)\right].$$

The boundary conditions for the symmetrical biconical horn require that

$$\frac{1}{\sin\theta}L_l^m(\cos\theta) = 0 \quad \text{at} \quad \theta = (\pi \pm \theta_0)/2. \tag{8}$$

The factor $\begin{matrix}\cos\\ \sin\end{matrix}(m\zeta)$ in these solutions defines the variation in the equatorial plane. An integral value of m is required. The constant m specifies the number of full periods of sinusoidal variation of the field about the principal axis from $\zeta=0$ to $\zeta=2\pi$. The functions $P_l^m(\cos\theta)$ and $Q_l^m(\cos\theta)$ influence the field distribution in the meridian planes. We shall use the symbol n, which is an integer, to indicate the number of maxima in the field distribution in the meridian planes. Thus, we get the subscripts m and n to specify the order of both transverse magnetic and transverse electric waves, i.e., $TM_{m,n}$ and $TE_{m,n}$.

Since the flare angle of a horn is an independent quantity fixed by the construction, the constant l is so chosen that the boundary conditions are satisfied on the surfaces of the horn. By taking the nth root of $L_l^m(\cos\theta)$ or of $(\partial/\partial\theta)L_l^m(\cos\theta)$ from $\theta=\pi/2$, employing asymptotic expansions for P_l^m and Q_l^m, and adjusting the constants A and B so that L_l^m has the proper symmetry about the equatorial plane, we have the flare angle for the several lowest-order waves. These relations are presented in the curves of Fig. 3.

The lowest-order or dominant wave is found to be the transverse electromagnetic TEM, and the next lowest-order wave the transverse electric $TE_{0,1}$. In both cases ($m=0$) the fields are independent of ζ. It is this uniform field distribution of these two waves

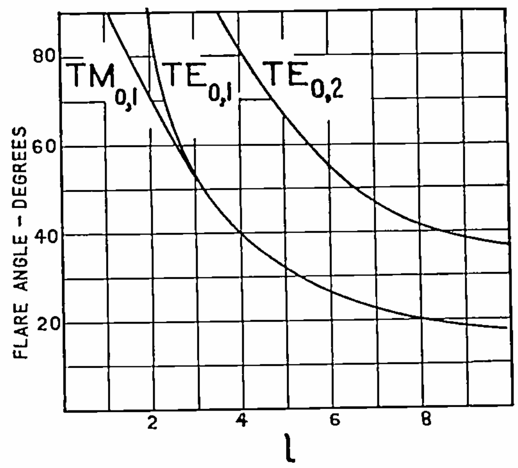

Fig. 3—Relations between the flare angle and the order l of the associated Legendre function for three types of waves.

that gives particular practical interest to biconical horns.

The field of the $TE_{0,1}$ wave is given by the following expressions:

$$H_r = L_l(\cos\theta)(p^2-\tfrac{1}{4})r^{-3/2}K_p(kr)$$

$$H_\theta = L_l^1(\cos\theta)\left[(p^2-\tfrac{1}{2})r^{-3/2}K_p(kr) - kr^{-1/2}K_{p-1}(kr)\right]$$

$$E_\zeta = -j\omega\mu L_l^1(\cos\theta)r^{-1/2}K_p(kr). \tag{9}$$

A sketch of the field configuration is shown in Fig. 4. The electric lines of force form concentric circles parallel to the equatorial plane on spherical surfaces about the origin, and the magnetic lines form closed loops in meridian planes. At distances remote from the apexes, a horizontally polarized field concentrated in the horizontal plane and uniform about the principal axis obtains. The transmission of the waves outward is conveniently described by the propagation constant[4] $\gamma=\alpha+j\beta$, which is found to be given by

$$\alpha = \frac{2p+1}{2r} - Re\left[kK_{p-1}(kr)/K_p(kr)\right]$$

$$\beta = -Im\left[kK_{p-1}(kr)/K_p(kr)\right]. \tag{10}$$

Fig. 5 shows curves of α and β. For large values of

$(\omega/c)r$, the phase constant β approaches the value ω/c of any transverse electromagnetic wave and the attenuation constant α approaches the value $1/r$ of a

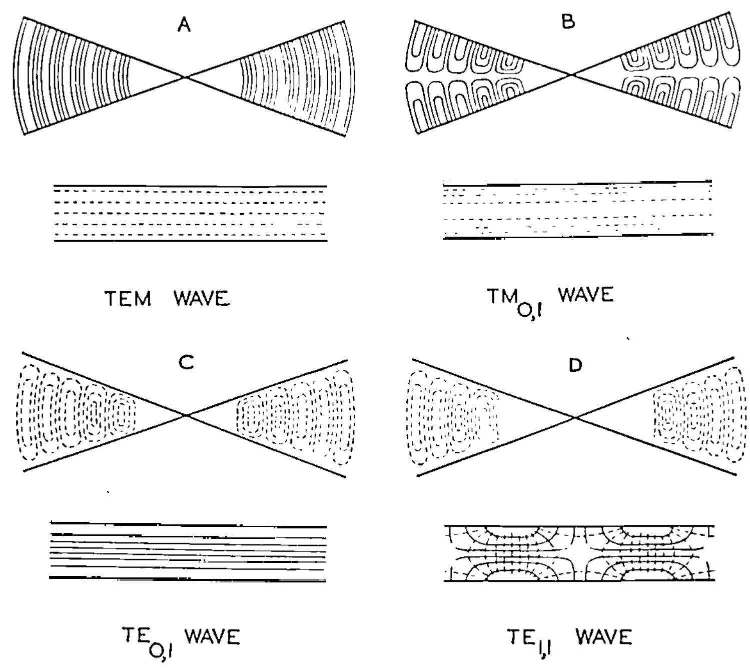

Fig. 4—Field configurations of four types of waves inside biconical horns. The upper portion of each sketch shows a cross section passing through the axis of the cones. The lower portion shows a developed section of a spherical surface. The solid lines indicate lines of electric intensity and the dotted lines indicate lines of magnetic intensity.

spherical wave. The characteristic impedance Z_0, defined as the ratio of the transverse electric field to the transverse magnetic field, is also shown in Fig. 6. For

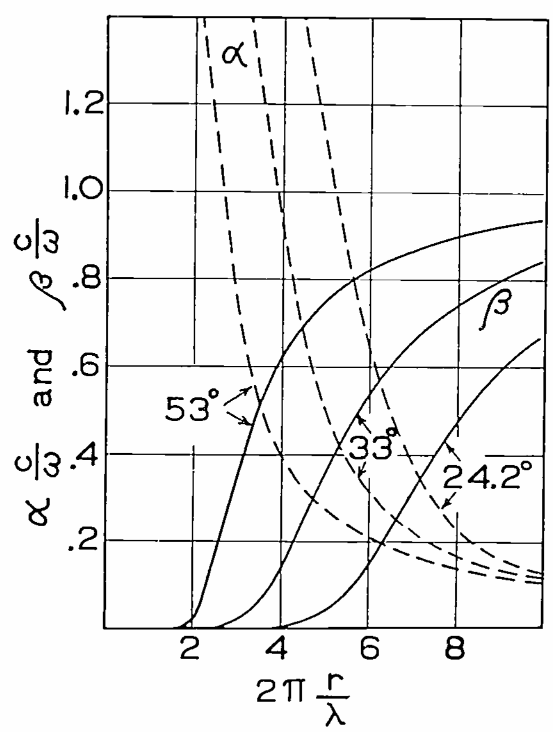

Fig. 5—The variations of attenuation constant α and phase constant β of the $TE_{0,1}$ wave with the radial distance. The numerals indicate the corresponding flare angle of the horn.

large values of $(\omega/c)r$, Z_0 approaches the value $\sqrt{\mu/\epsilon}$ for a transverse electromagnetic wave, indicating that the horn waves may be perfectly matched to free space by proper horn design. A consideration of α, β, and Z_0 shows that the $TE_{0,1}$ wave is in the "complementary-wave" category. A distinct cutoff characteristic and a dependence of the transmission properties on the cross-sectional dimensions establish this fact. The higher-order TE waves have larger attenuations than that of the $TE_{0,1}$ wave. Appropriate design of the horn enabled us to eliminate higher-order TE waves by virtue of this attenuation relationship. The power P transmitted through the horn may be calculated by integrating the power density flowing in the radial direction over a closed surface r = constant between the two cones. For the $TE_{0,1}$ wave, it is given by

$$P = -\int_0^{2\pi}\int_{\pi-\theta_0/2}^{\pi+\theta_0/2} \tfrac{1}{2} E_\zeta H_\theta^* r^2 \sin\theta\, d\theta d\zeta = \omega\mu\theta_0, \qquad (11)$$

where H_θ^* is the conjugate of H_θ. This formula is

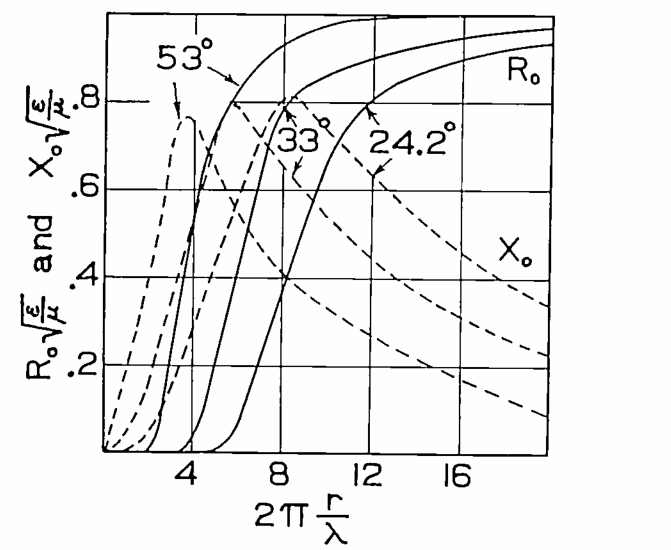

Fig. 6—The variations of the characteristic impedance ($Z_0 = R_0 + jX_0$) of the $TE_{0,1}$ wave with the radial distance. The numerals indicate the corresponding flare angle of the horn.

based upon the first approximation of Legendre's function

$$L_l^1(\cos\theta) \cong \cos\left[\frac{\pi}{\theta_0}\left(\theta - \frac{\pi}{2}\right)\right]\Big/\sqrt{\sin\theta}. \qquad (12)$$

The field of the TEM wave, which is a special case of the TM wave, is given by the following simple expressions:

$$E_\theta = -\sqrt{\frac{2k}{\pi}}\,\frac{1}{r\sin\theta}\,e^{-jkr}$$

$$H_\zeta = \sqrt{\frac{\epsilon}{\mu}}\,E_\theta. \qquad (13)$$

The field configuration shown in Fig. 4 has only two components both of which are transverse to the direction of propagation. Since E_θ is always perpendicular to the conical surfaces, the boundary conditions are automatically satisfied in any biconical horn, whether symmetrical or unsymmetrical, except only that a flare angle of 180 degrees is prohibited. The propagation constant and characteristic impedance are readily obtained as

$$\alpha = \frac{1}{r}, \qquad \beta = \omega\sqrt{\epsilon\mu}, \qquad Z_0 = \sqrt{\frac{\mu}{\epsilon}}. \tag{14}$$

These values, which are the same as those for the radiation field of a dipole, do not manifest cutoff properties. This wave has the lowest attenuation among the biconical-horn waves. The *TEM* wave belongs to the category of "principal waves." Excitation may take place at arbitrarily low frequencies. The present analysis, which assumes a horn of infinite extent, would naturally become invalid for a horn of finite radial length excited at very low frequencies. It is evident, however, that a mode of excitation corresponding to the *TEM* wave would exist even in finite horns. The net power P transmitted through the horn may be calculated as follows:

$$P = \frac{8\pi}{\lambda}\sqrt{\frac{\epsilon}{\mu}} \log \tan\left(\frac{\pi + \theta_0}{4}\right). \tag{15}$$

The field of the $TM_{0,1}$ wave is given by the following expressions:

$$E_r = L_l(\cos\theta)(p^2 - \tfrac{1}{4})r^{-3/2}K_p(kr)$$

$$E_\theta = L_l^1(\cos\theta)\left[(p - \tfrac{1}{2})r^{-3/2}K_p(kr) - kr^{-1/2}K_{p-1}(kr)\right]$$

$$H_\zeta = j\omega\epsilon L_l^1(\cos\theta)r^{-1/2}K_p(kr). \tag{16}$$

The value of l is to be so chosen that L_l $(\cos\theta)$ has its first zero from $\theta=\pi/2$ at the value $\theta=(\pi\pm\theta_0)/2$. The field configuration of the $TM_{0,1}$ wave is shown in Fig. 4 and the relation between l and the flare angle θ_0 is given in Fig. 3.

Radiation Theory of Biconical Horns

The radiation properties of biconical horns may be calculated by following the general procedures already presented.[4] The assumption is made that the field distribution between the edges of the cones is that which would exist were the horn not finite in length. This assumption is subject to experimental verification and is verifiable in horns whose length is several wavelengths and which are properly operated. The magnitude of the electric intensity in space over a sphere of radius great compared to both the wavelength and the dimensions of the horn comprises the radiation pattern. The Stratton-Chu[9] formulation of Kirchhoff's principle has been used in this investigation.

We shall present here the radiation patterns of the $TE_{0,1}$ and the *TEM* waves only. Since the fields of these waves have circular symmetry, the radiated wave must also have circular symmetry, and we need consider but one meridian plane. The radial length of the horn will be designated by r_0 and the flare angle, measured from cone to cone, by θ_0. The point of observation is located at (r', θ', ζ'), θ' being measured from the equatorial plane. By applying the Stratton-Chu formula to the fields for the *TEM* and the $TE_{0,1}$ waves, given, respectively, by (13) and (9), we find the following expressions for the electric intensity in any meridian plane:

TEM Wave.

$$E_\theta(\theta') = j\,\frac{\pi}{2r'\sqrt{\lambda}\cos\theta'}\,e^{jk(r_0+r')}\left[\frac{\theta+\theta'}{j\pi^2}\sqrt{\frac{\lambda}{r_0}}\cos\left[2\pi\frac{r_0}{\lambda}\cos(\theta+\theta')-\frac{\pi}{4}\right] + e^{j[2\pi(r_0/\lambda)-(\pi/4)]}F(v)\right]_{\theta=-(\theta_0/2)}^{\theta=(\theta_0/2)}. \tag{17}$$

$TE_{0,1}$ Wave.

$$E_\zeta(\theta') = -\frac{\pi}{4r'}\sqrt{\frac{\mu}{\epsilon\lambda\cos\theta'}}\,e^{jk(r_0+r')}\left[\frac{(\lambda/r_0)^{3/2}}{8\theta_0}\cos\left\{2\pi\frac{r_0}{\lambda}\cos(\theta-\theta')-\frac{\pi}{4}\right\}\sin\left(\frac{\pi}{\theta_0}\theta\right)+e^{j\pi^3\lambda/32r_0\theta_0{}_2}\left\{e^{j(\pi/\theta_0)\theta'}F(v_1)+e^{-j(\pi/\theta_0)\theta'}F(v_2)\right\}\right]_{\theta=-(\theta_0/2)}^{\theta=(\theta_0/2)} \tag{18}$$

where

$$v = kr_0(\theta_0+\theta)^2 4/\pi^2$$

$$F(v) = \int \tfrac{1}{2}\left[J_{-1/2}(v) - jJ_{1/2}(v)\right]dv$$

$$v_1 = kr_0\frac{4}{\pi^2}\left[\theta - \theta' - \frac{\pi^3}{8kr_0\theta_0}\right]^2$$

$$v_2 = kr_0\frac{4}{\pi^2}\left[\theta + \theta' - \frac{\pi^3}{8kr_0\theta_0}\right]^2.$$

The corresponding magnetic fields can be calculated from Maxwell's equations.

The general radiation characteristics of the biconical horn are similar to those of horns of other shapes. The sharpness of the beam, when a pure single-wave type obtains in the horn, depends only upon the length r_0 and the flare angle θ_0. For a fixed length, there is an optimum flare angle that gives the sharpest beam. The optimum flare angle decreases with an increase of the length, with a corresponding sharpening of the beam. An interesting comparison may be made with the radiation patterns of sectoral horns which have been evaluated numerically in a previous paper.[5] The $TE_{0,1}$ wave in biconical horns has similar properties to those of the $TE_{0,1}$ or $H_{0,1}$ wave in sectoral horns, and the *TEM* wave in biconical horns compares to the $TE_{1,0}$ or $H_{1,0}$ wave in the sectoral horns.

The power gain of a horn may be defined as the ratio of the power radiated from a Hertzian dipole

[9] J. A. Stratton & L. J. Chu, "Diffraction theory of electromagnetic waves," *Phys. Rev.*, vol. 56, pp. 99–107, July 1, (1939).

to that radiated from the horn to produce, in each case, the same electric intensity at a fixed remote point in the direction of principal transmission. For symmetrical biconical horns, that point lies in the equatorial plane. The expressions (17) and (18) and the well-known dipole formula permit the calculation of the power gain giving the results:

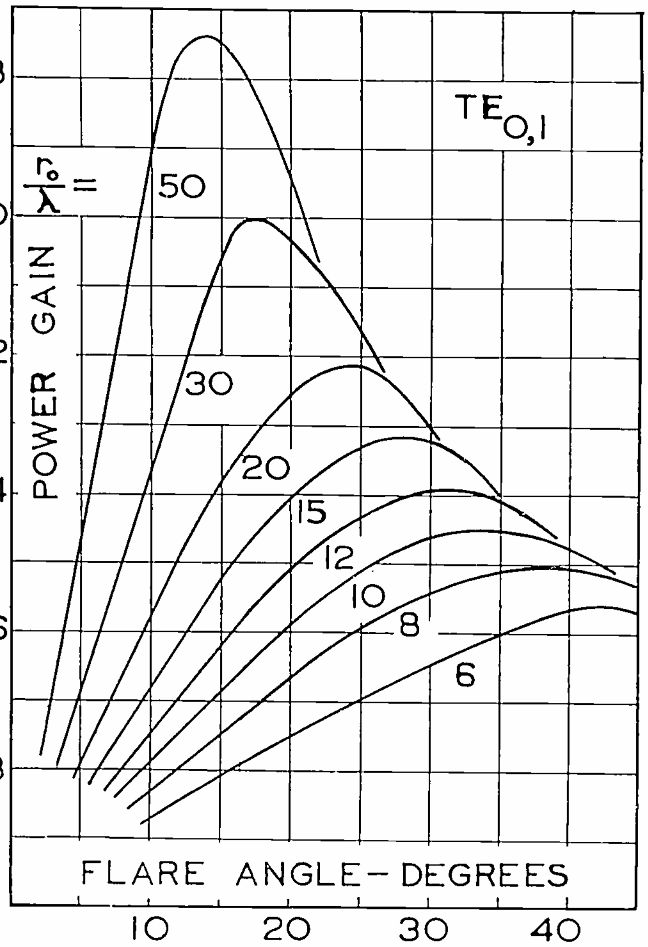

Fig. 7—Power-gain curves for the $TE_{0,1}$ wave.

TEM Wave.

$$\text{Power gain} = \frac{\lambda r'^2 \,|\, E_{\theta'} \,|^2_{\theta'=0}}{6 \log \tan \left(\dfrac{\pi + \theta_0}{4}\right)} \cdot \qquad (19)$$

$TE_{0,1}$ Wave.

$$\text{Power gain} = \frac{2r'^2 \epsilon \lambda}{3\theta_0 \mu} \,|\, E_{\zeta'} \,|^2_{\theta'=0}. \qquad (20)$$

The power gains are plotted against the flare angle in Fig. 7 for the $TE_{0,1}$ wave and in Fig. 8 for the *TEM* wave. We find in both sets of curves pronounced maxima. These maxima, taken together, define a family of horn dimensions that provide the smallest and most economical horn construction for any given power gain. These optimum data are plotted separately in Fig. 9.

The curves contained in Figs. 7 to 9 provide design information from which biconical horns may be built to satisfy given specifications, particularly when the radiation patterns from (18) and (19) are considered simultaneously.

A consideration of Figs. 7, 8, and 9 will show that the power gain becomes greater as the flare angle is decreased, if the separation between the edges of the cones is held constant. The limiting case of two parallel disks produces theoretically the largest power gain and the sharpest beam. Practically, however, considerable difficulty is presented to the excitation of a single wave, say the *TEM* wave, to the exclusion of higher-order waves in this limiting case. It is precisely the flaring feature of the horns that gives them superiority in this respect.

Experimental Tests of the Biconical Horn

The experimental work here reported was directed towards an exploration of the practical features of this new horn and towards a comparison of the developed theory with experiment. The transverse electromagnetic wave, having a simple radiation pattern with vertical polarization, is of immediate interest for practical applications. Also, it may be excited comparatively easily and therefore it affords a convenient experimental approach. An experimental symmetrical biconical horn has been constructed and its operation thoroughly tested.

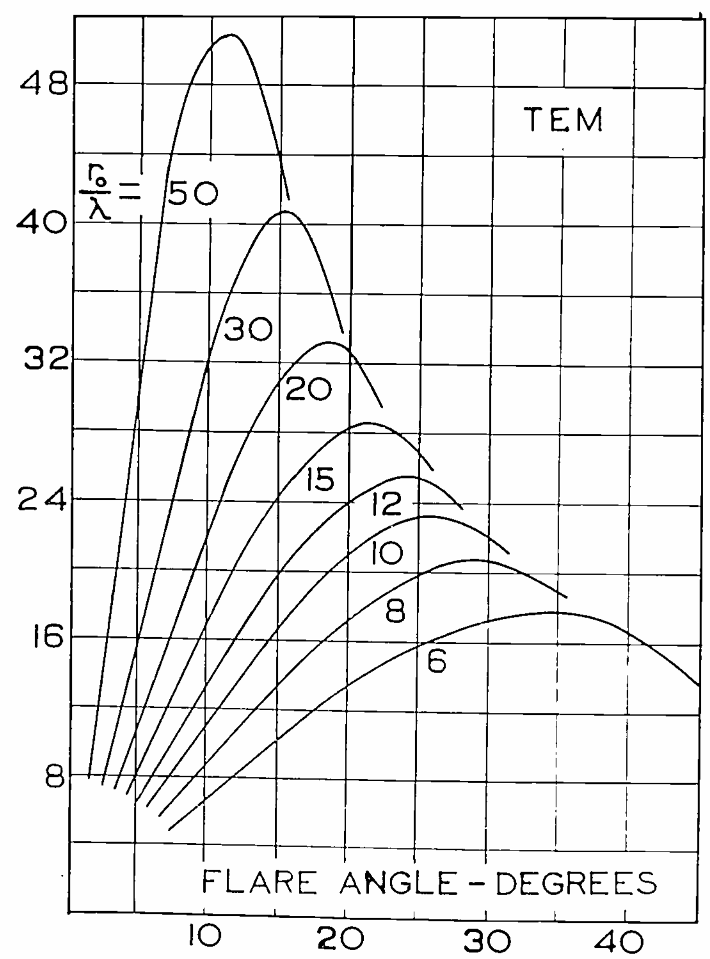

Fig. 8—Power-gain curves for the *TEM* wave.

The construction of the experimental horn necessitated consideration of such practical matters as wavelength, size, support of the cones, and details of the exciting system. The wavelength of the magnetron generator used in the tests was 8.3 centimeters. It was desirable to keep the greatest dimension of the horn less than about 3 feet and, at the same time, to utilize the optimum design features previously described. A radial length of 46 centimeters or 5.6λ was chosen with the aid of the optimum design curves of Fig. 9, which also gave a flare angle of 35 degrees. Both cones were held at their edges by 3 vertical wooden supports, which were made adjustable to permit a variation of the separation between the cones. A straight rod between the apexes served as exciting means. It was connected to the apex of the top cone

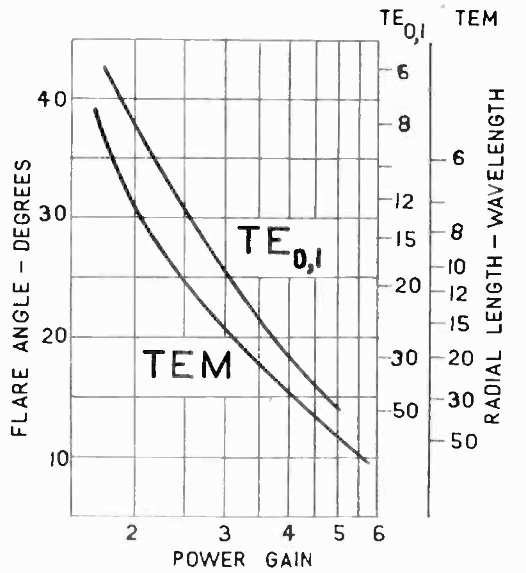

Fig. 9—Optimum-design curves, power-gain basis.

at one end and to the center conductor of a coaxial line that projected through the vertex of the bottom cone at the other end. The sketch of Fig. 10 and the photograph of Fig. 11 show the details and dimen-

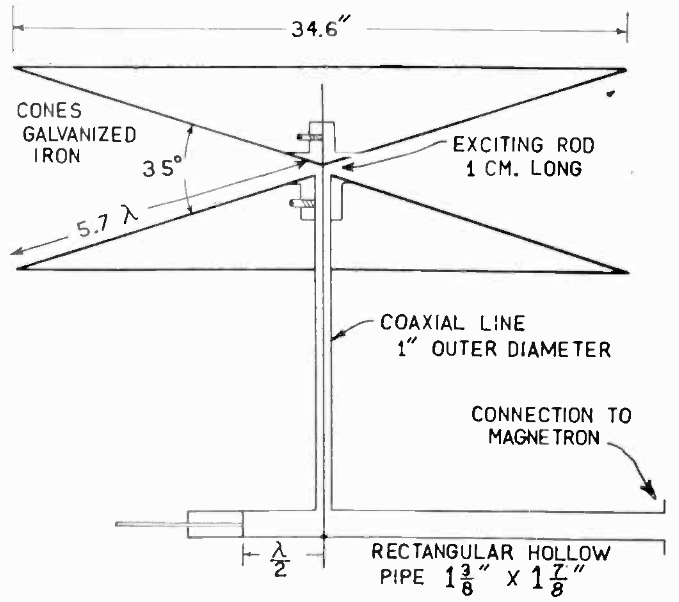

Fig. 10—Sketch showing constructional details of the biconical horn used in the experiments.

sions of the experimental horn. The center conductor which was conductively connected to the bottom side of the hollow pipe and to the apex of the top cone, required no insulating spacers. The rectangular hollow pipe connecting the coaxial line to the magnetron source was supplied with an adjustable reflector for obtaining maximum energy transfer from the hollow pipe to the coaxial line and thence to the horn. The spacing between this reflector and the extension of the center conductor was about one-half of the hollow-pipe wavelength. The magnetron gen-

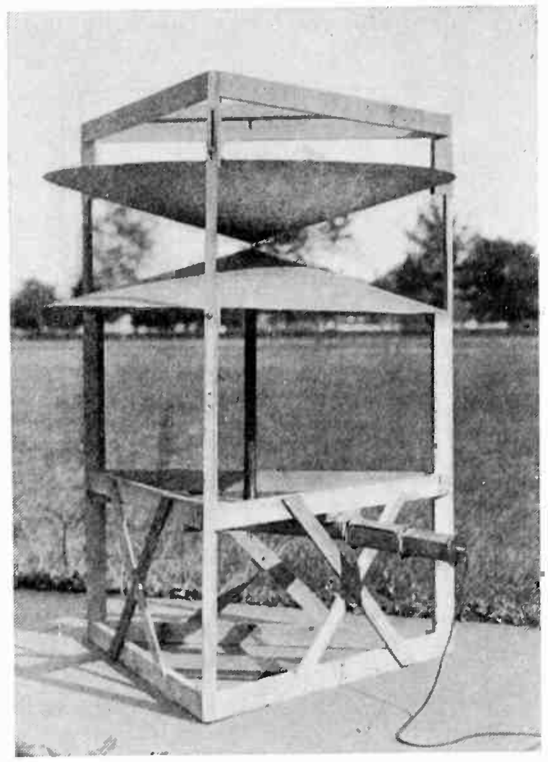

Fig. 11—Photograph of the experimental biconical horn.

erator and other related ultra-high-frequency equipment will be described in greater detail elsewhere.

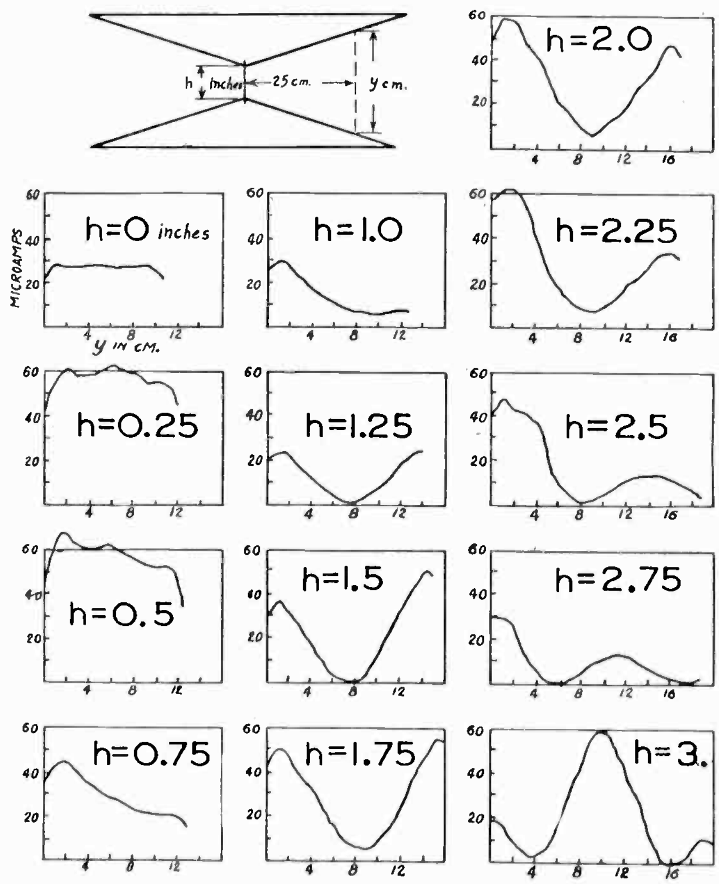

Fig. 12—Effect of the spacing between the cones on the waves within the horn.

Measurements of electric-field intensity were made with a minute crystal detector (galena) probe and a

50-microampere meter. The response of this combination followed substantially a square law.

From a practical standpoint, it is necessary to know how exact the construction must be in order that the desired field characteristics may be realized. Preliminary tests showed that the inner surfaces of

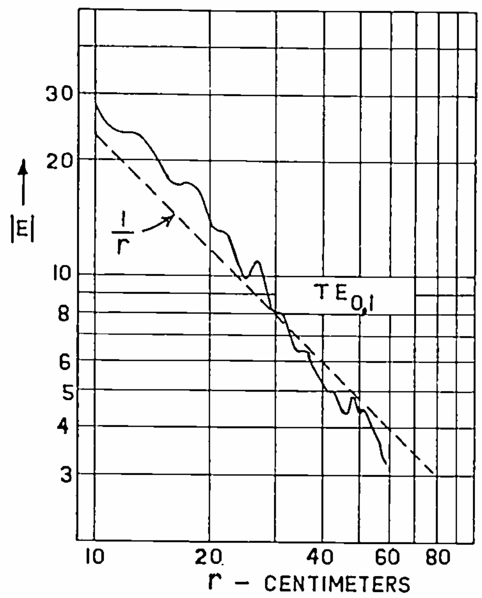

Fig. 13—Variation of electric intensity along a radial line. The dotted line is the theoretical curve.

the cones had to be held to rather close tolerances for at least the first wavelength from the exciting rod to produce a symmetrical distribution in the horn. The outer portions of the cones have a much

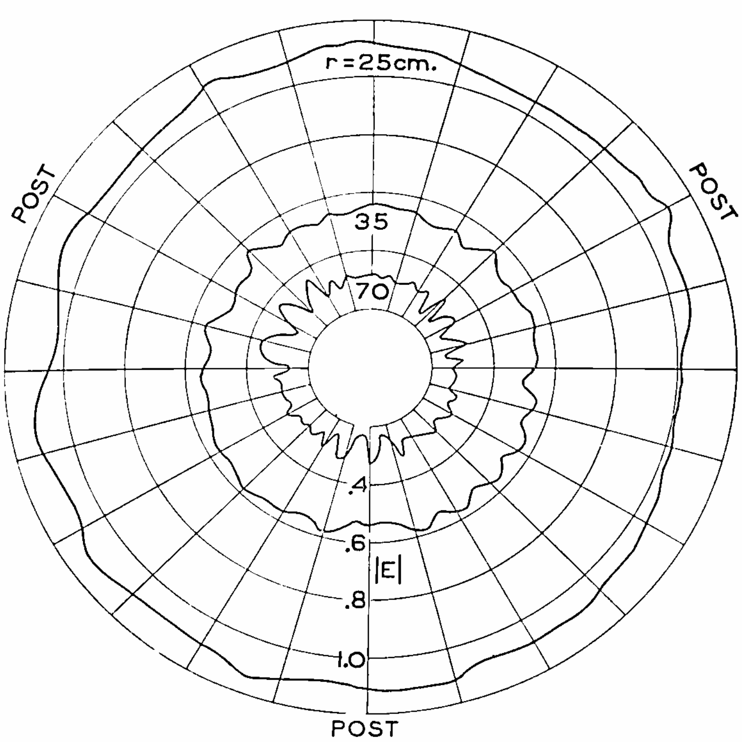

Fig. 14—Variation of electric intensity in the equatorial plane along circles of radii 25, 35, and 70 centimeters, respectively.

smaller effect. The excitation rod must be in good alignment and all electrical contacts between cones and lines should be complete. If these conductors were poor, some of the energy would escape or "leak" through the top or the bottom of the cones, making it difficult or impossible to realize the desired radiation pattern.

Effect of Cone Spacing

The two cones do not necessarily have to be disposed with their apexes coinciding, as was assumed in the analysis. In one feasible construction, the small ends of the cones would be cut off and replaced by parallel disks. The separation between the disks could then be adjusted, preferably by changing the spacing between the truncated cones. In the construction followed in the experimental horn, as described above, substantially the same result was obtained by making the spacing of the cones alone adjustable. The spacing influences the impedance of the exciting rod and thereby the energy transfer from coaxial line to horn. In addition, the spacing influences directly the type of wave that may be excited. In general, higher-order waves are easier to excite with greater separations of the cones.

A set of measurements was made to determine the effect of the spacing on the waves within the horn and are reproduced in Fig. 12. The axial distance in inches between apexes is denoted by h. For $h=0$ the exciting rod still had a nonzero length, because the extreme vertex of the lower cone was cut away to permit the coaxial line to enter. Measurements were taken for successive separations differing by one quarter of an inch.

A consideration of the curves of Fig. 12 shows that the separations of $h=0$ and $h=\frac{1}{4}$ inch produce the field distribution for the TEM wave. Then, there is a gradual transition with increasing spacing to the $TM_{0,1}$ wave, which is complete for a spacing of about $1\frac{1}{2}$ inches, corresponding to a half wavelength. The $TM_{0,1}$ pattern, in turn, gives way to that of the $TM_{0,2}$ wave as the spacing is still further increased to 3 inches, corresponding to a full wavelength. On comparing the curves for $h=0$ and $h=\frac{1}{4}$ inch, we observe that their shape is substantially the same but the

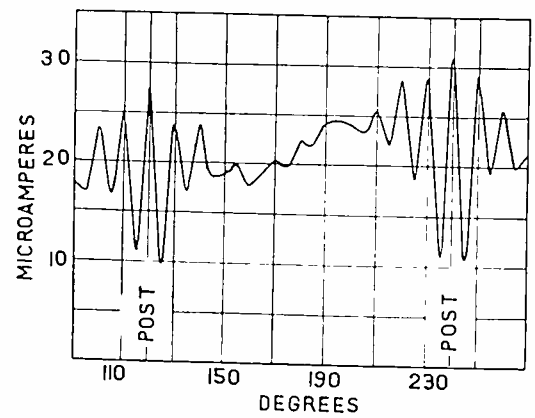

Fig. 15—Effect of the posts on the field distribution in the equatorial plane.

magnitude is greater with $h=\frac{1}{4}$ inch, indicating a better transfer of energy to the horn. This spacing was therefore selected as substantially optimum and was maintained throughout the remainder of the experiments. It was not only an efficient operating condition but also preserved a true TEM wave.

This experiment illustrates an important principle of horn design. The current distribution along the exciting rod is by no means constant and may not be symmetrical. Such distributions tend to produce not only the TEM wave but also the higher-order $TM_{0,n}$ waves. The higher-order waves are more greatly attenuated than those of lower order, but the attenuation of all waves decreases as the spacing of the cones is increased. By increasing the spacing in steps, the higher-order $TM_{0,n}$ waves successively become free to propagate outward. Thus, as shown in Fig. 11 for $h=3$ inches, the distribution is actually the result of a superposition of TEM, $TM_{0,1}$, and $TM_{0,2}$ waves; the predominance of the $TM_{0,2}$ wave is probably caused by the full-period sinusoidal current distribution in the exciting rod.

Field Inside the Horn

The experiments we shall now discuss deal with the field distribution inside and just outside of the horn in the equatorial plane.

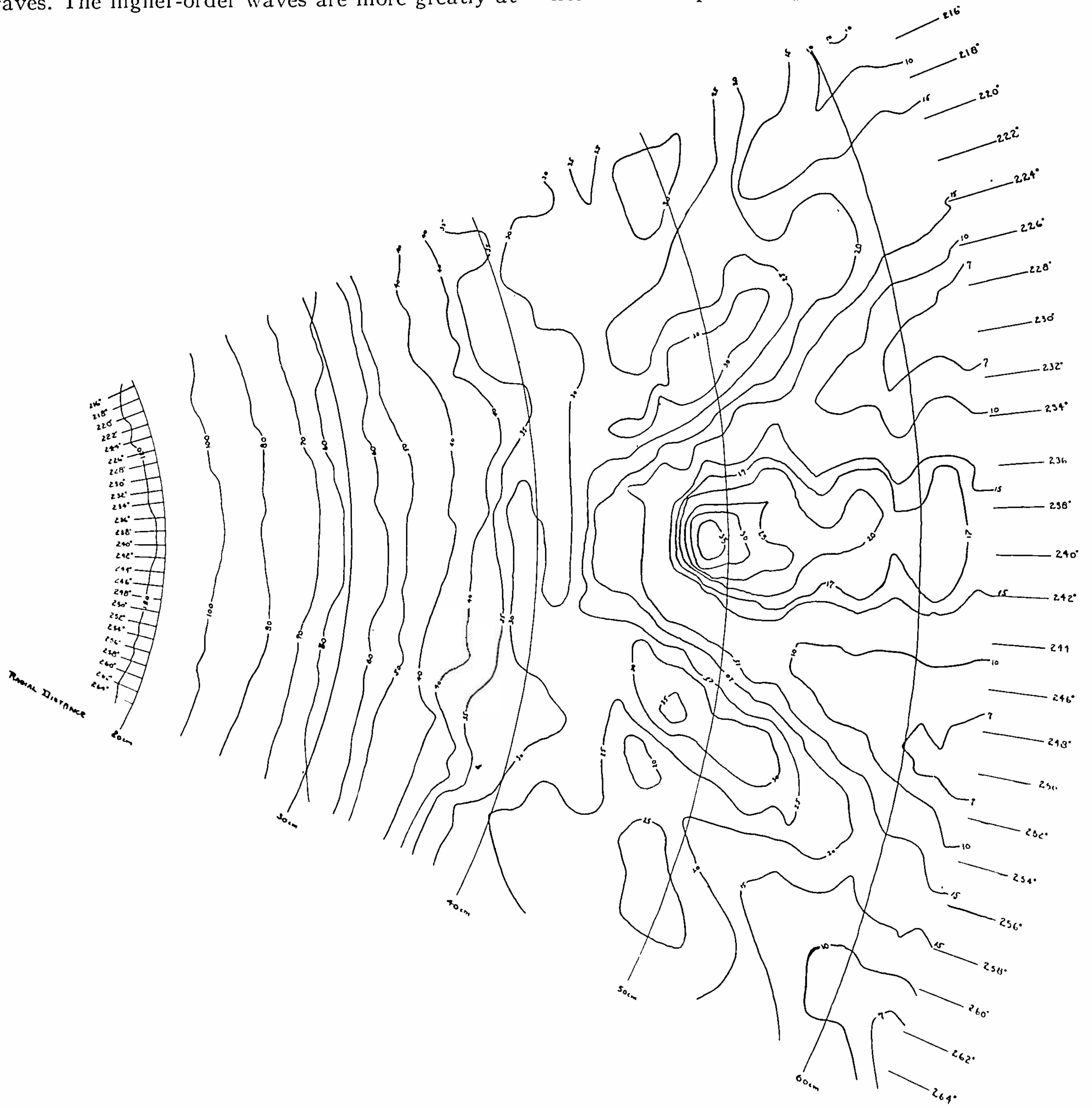

Fig. 16—Contour line of constant electric intensity in the $\theta=90$ degrees plane in the vicinity of a wooden post, $\frac{3}{4}$ inch by $1\frac{3}{4}$ inches, located at $\phi=240$ degrees. The number associated with each line is the value of the received current in microamperes.

Fig. 13 shows the variation of the electric intensity along a radial line. On mathematical grounds, a variation as $1/r$ was predicted which is included in the figure as a dotted curve. Near the center, the experimental curve drops off more rapidly than $1/r$, and it also has observable ripples superposed on it, which may be caused by waves of higher order than 0,1 but of relatively small magnitude.

Fig. 14 shows the variation of electric intensity in the equatorial plane along circles of three different radii. In each case, the pattern is essentially circular or uniform. These curves also give some idea as to how the field drops off at increasing distances from the center, but more particularly they show that three distinct ripples become superimposed on the essentially circular pattern at radial distances approaching that of the edge of the cones. These ripples are centered on the vertical wooden supports. Redrawing a portion of one of these polar patterns in rectangular co-ordinates (the meter deflection in microamperes is plotted versus the angle ζ as in Fig. 15, shows the ripples in an even more striking way. The similarity between this curve and that for the diffraction of light through a slit comes to mind at once. The supports in this case were wooden and were $\frac{3}{4}$ inch wide (facing center) and $1\frac{3}{4}$ inches long (in radial direction), and they are responsible for the ripple by diffracting or scattering the wave in its progress outward. A contour map of lines of constant field intensity in the vicinity of a support was constructed from 1400 individual measurements. Fig. 16 shows the contour map, which includes a sector 25 degrees on either side of a support and a radial extent from 18 to 64 centimeters in the equatorial plane. The edge of the cones and the location of the post are indicated in the figure. It is evident that the field lines are symmetrically disposed about a radial line passing through the center of the post. The above diffraction patterns can be closely correlated with the field map and a clearer picture obtained of the effect of the posts. Further measurements were made to determine the approximate relation between the magnitude of the disturbance of the otherwise circular pattern and the size and material of the post. The disturbance depends more on the cross-sectional size of the post than on the material, so long as wood, glass, bakelite, and other readily available dielectrics are employed. Using the maximum distance u between maxima and minima of curves similar to those of Fig. 15 as a criterion, it appeared that an empirical relation of form $u=(a+bx)y$ fitted experimental data fairly well, where x=depth of post, y=width of post, and a,b=constants which probably depend on the material. Obviously, the supports should be made as small as is consistent with mechanical stability. The final size was $x=\frac{1}{2}$ inch, $y=\frac{3}{4}$ inch, which reduced the ripple to an inconsequential magnitude.

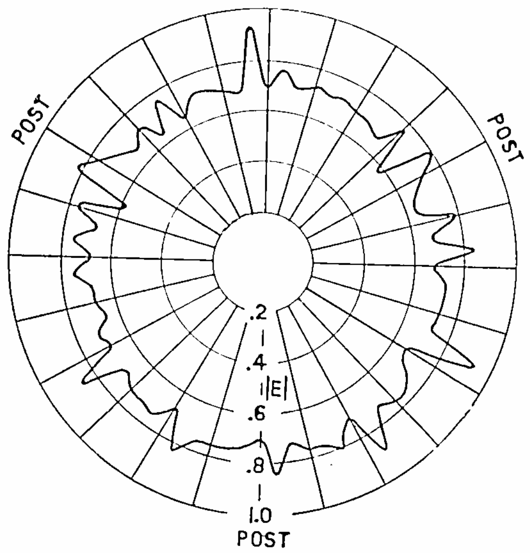

Fig. 17—Radiation pattern in the equatorial plane.

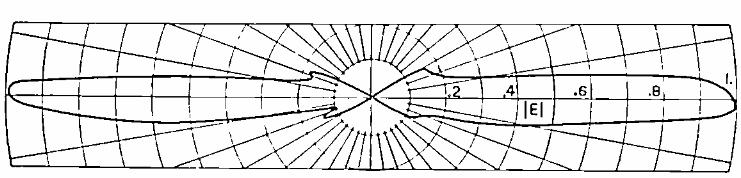

Fig. 18—Radiation pattern in the meridian plane.

Radiation Patterns

The field at a great distance from the horn, although directly dependent on that within the metal surfaces, is different from it and comprises perhaps the most important performance feature. The measurement of such radiation patterns is more difficult with the biconical horn, because of its uniform equatorial-plane pattern, than with horns having beam characteristics. The following two patterns for our experimental horn are thought to be sufficiently accurate for most purposes. They show in a clear way that the general over-all behavior predicted from the theory has been realized.

The equatorial or horizontal pattern is shown in Fig. 17. The ripple is still to be seen, but the position of maximum variation has shifted to a location midway between the supports. The pattern is nevertheless fairly circular. The maximum variation from uniformity is ± 2.2 decibels, which is of no practical consequence, as far greater variations are introduced by terrain, buildings, etc.

The meridian, or vertical, pattern is shown in Fig. 18. The strong concentration of energy in the equatorial or horizontal plane is evident. Two secondary lobes are present; they may be reduced by a reduction of the flare angle of the horn.

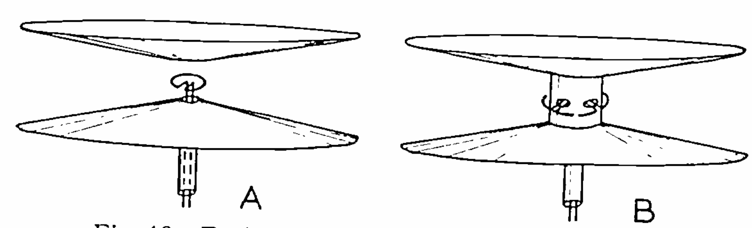

Fig. 19—Exciting means for the $TE_{0,1}$ wave in a biconical horn.

Concluding Discussion

The preceding sections have included discussion of certain details and modifications, as well as more basic material. We wish to present a few additional ideas in this concluding section.

The $TE_{0,1}$ wave may find application because, with a vertical principal axis, it radiates horizontally polarized waves. It may be excited by a small current-carrying loop (magnetic dipole) lying in the equatorial plane at the center. Such an exciting system is illustrated in Fig. 19A. It is essential that the phase and the amplitude of the current be constant about the loop, which may be realized by making its length substantially less than a half wavelength. The radiation effectiveness (resistance) of this system is low, however, and improved ones may be readily designed. An example of one such system is illustrated at B in Fig. 19. A number of antennas are bent in a circle and disposed about the center in the equatorial plane. They are fed in such a manner that the currents in all have the same angular sense about the circle. The diameter of the circle is best made an odd multiple of a half wavelength. The balanced-wire feed prevents the excitation of TM waves. In A and B, and in any system, it is difficult to have the exciting system perfectly symmetrical. The spacing of the cones provides the most practical means of eliminating the higher-order waves, and this spacing should be appropriately adjusted.

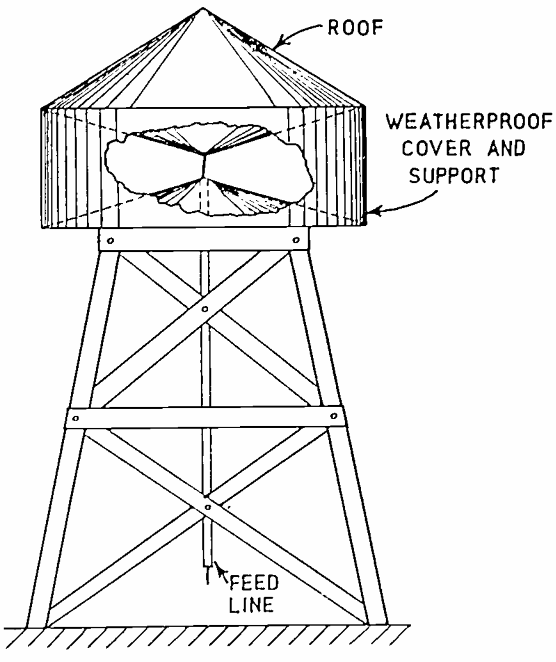

Fig. 20—Weatherproof biconical horn that also eliminates the use of posts for supporting the upper cone.

Let us next consider the support of the horn structure. The effect of diffraction around the supports can be reduced to a very small amount by the use of supports of small cross section and preferably of low dielectric constant. Another means of support that completely avoids this diffraction effect and at the same time affords a weatherproof arrangement comprises a cylindrical sheet of dielectric material fastened along its two edges to the rims of the upper and lower members, as illustrated in Fig. 20. This figure also illustrates a protective roof, which may be a conductor or a dielectric, and an elevated tower support for the entire horn to locate it above near-by obstacles.

If a uniformly radiating horn is located high above the ground, say atop a large building, it may be necessary or desirable to use a nonsymmetrical horn of the general character of that of Fig. 1E in order to provide a signal in the neighborhood directly below and near to the horn location. Either vertical (TEM wave) or horizontal ($TE_{0,1}$ wave) polarization can be employed.

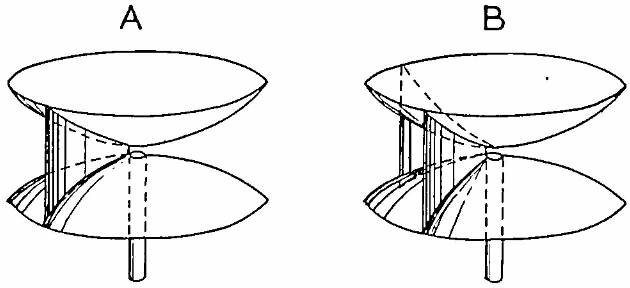

Fig. 21—Septate biconical horn.

When a noncircular radiation pattern in the equatorial plane is desired, waves of the type $TE_{1,1}$ (see Fig. 4D) may be employed. They may be excited in horns like those of Fig. 1 by a suitable exciting system. Another horn construction, however, leads to a simpler exciting means. This horn may be termed a septate biconical horn, because of the septa or partitions dividing the horn interior, as illustrated in Fig. 21. One partition and an exciting system for the $TE_{0,1}$ wave are shown at A in the figure; two partitions are illustrated at B. It is interesting to observe that the configuration for the $TE_{0,1}$ wave is such that the boundary conditions are automatically satisfied on the septum of Fig. 21, or on any number of similar septa. It is therefore feasible to employ two septa as at B and to send waves mainly through one sector, keeping the other closed or partially closed for waves. In this way, the radiation pattern in the equatorial plane may be varied over the extreme limits from a circular pattern to a sharp beam. Applications of this horn may be made where the density of receiving stations is not uniform about the transmitter to provide a more efficient coverage. In some cases, it might be desirable to make the septa rapidly adjustable, thereby obtaining a horn of variable radiation pattern.

Atmospherics in Radio Broadcast Reception at Calcutta*

S. P. CHAKRAVARTI†, MEMBER I.R.E., P. B. GHOSH†, NONMEMBER, I.R.E., AND H. GHOSH†, NONMEMBER, I.R.E.

Summary—*This paper relates to investigations extending from January to August, 1938, (including winter, summer, and rainy seasons) on atmospheric disturbance in medium- and short-wave bands (0.6 to 6 megacycles) prevalent in the eastern part of India. A suggestion has been made for broadcast transmission standards to be adopted in India on the basis of atmospheric field-strength measurements. Effective service areas of 1.5 kilowatts, 370 meters and 5 kilowatts, 235 meters, medium-wave broadcast transmissions have been estimated on the standards suggested.*

MEASUREMENTS have been carried out at the Kanodia Electrical Communication Engineering Laboratories, University of Calcutta (22° 40′ N., 88° 30′ E.), as follows:

(1) Direction of arrival of maximum disturbance (Figs. 1 and 2).

(2) Classification of atmospherics observed on 0.6 megacycle according to their field strength (Fig. 3).

(3) (a) Total number of atmospherics per minute on frequencies from 0.6 to 6 megacycles (Figs. 4 and 5).
(b) Total number of atmospherics per minute at different hours of the day (Fig. 6).

(4) (a) Peak field strength of atmospherics on frequencies from 0.6 to 6 megacycles (Fig. 7).
(b) Peak field strength of atmospherics at various hours of the day (Fig. 8).

Observations were carried out simultaneously as far as possible to obtain information on all aspects of the disturbance at a particular moment. The equipment set up consists of a cathode-ray direction finder[1,2] with a crossed loop supplemented by a receiver (connected to a separate outdoor aerial) to distinguish on the screen of the oscilloscope the deflections caused by man-made static, for observing the direction of arrival of maximum disturbance, equivalent field strength, and total number per minute on 0.6 megacycle; and a specially designed receiver connected to an aerial of known characteristics and containing a linear detector to obtain relative field strengths of the atmospherics and the number of atmospherics per minute on several frequencies from 0.6 to 6 megacycles in accordance with Moullin's theoretical paper.[3]

Fig. 1 shows the direction of arrival of maximum disturbance on 0.6 megacycle from January to August, 1938. It must be noted that there is an ambiguity of 180 degrees in the determination of the direction of arrival. Propagation of disturbance on 0.6 megacycle has been that of a normally polarized ground wave, since the trace on the cathode-ray oscilloscope screen has invariably been a straight line. In January, the direction was almost constant at 20 degrees east of north and in February it varied between 20 and 40 degrees east of north indicating that the sources are either in northeast Bengal or southwest in the Madras Presidency. In March, the direction was almost 90 degrees from north (i.e., east-west) indicating that the sources were somewhere in East or West Bengal. April, May, and June were months of more frequent thunderstorms all around Calcutta and hence a day-to-day variation in direction of arrival can be accounted for. July and August were rainy months and the direction indicates that the sources are somewhere in northeast Bengal or on the border of Madras Presidency, possibly in the former region as it was known to experience severe thunderstorms and rain during these two months. Fig. 2 shows these directions plotted on a map of eastern India.

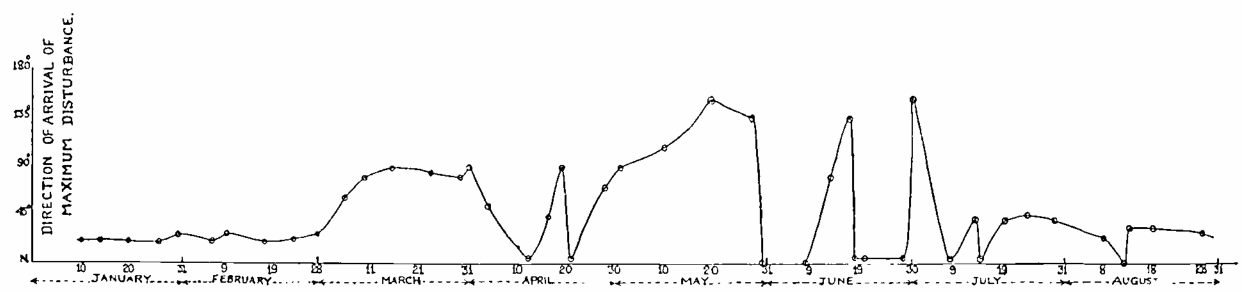

Fig. 1

The atmospherics observed on 0.6 megacycle have

* Decimal classification: R114. Original manuscript received by the Institute, February 6, 1939; revised manuscript received, May 26, 1939.
† Kanodia Electrical Communication Engineering Laboratories, University of Calcutta, Calcutta, India.
[1] Watt and Herd, *Jour. I.E.E.* (London), vol. 64, p. 611, (1926).
[2] Watt, Herd, and Bainbridge-Bell, H.M. Stationery Office Publication, 47–96, pp. 145–172, (1933).
[3] Moullin, *Jour. I.E.E.* (London), vol. 62, p. 353, (1924).

been placed in six classes, A, B, C, D, E, and F according to their field strength (see Fig. 3). In clear weather, classes C, D, and E and D, E, and F mainly contribute to the total number during summer (and rains) and winter, respectively; in cloudy weather, classes B, C, and D and C, D, and E mainly con-

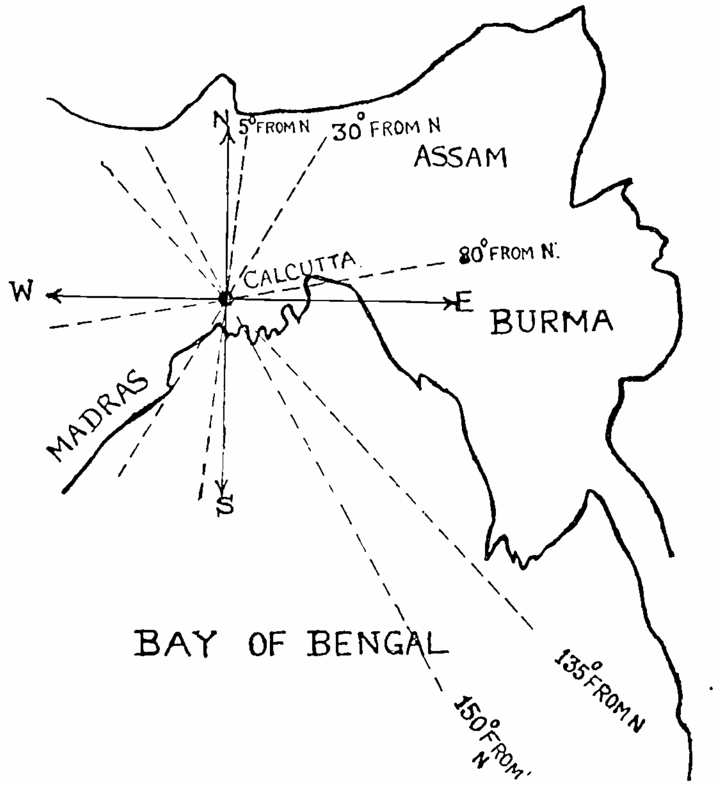

Fig. 2

tribute to the total number during summer (and rains) and winter, respectively; and in thundery weather, classes A, B, and C make up the total number during all seasons. Fig. 3 shows the observations for clear and cloudy weathers.

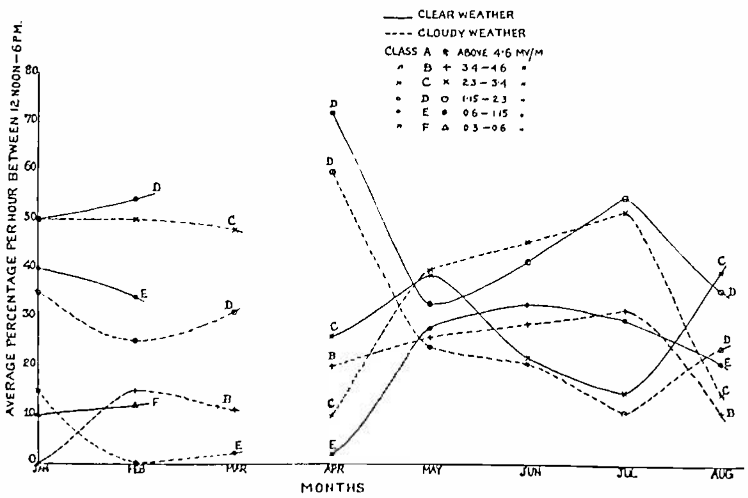

Fig. 3

Fig. 4 shows the typical variation of the total number per minute with frequency observed at afternoon and night during cloudy and thundery weathers in summer. The nature of variation during clear weather is similar to that of cloudy, only the total number per minute has generally been found less. An explanation for this mode of variation may be given as follows: A single source of disturbance should give the same number of impulses on all frequency components at the receiving point, only the field strength would vary with frequency. In thundery weather, there are thunderstorm sources near by and other sources in action at greater distances. On frequencies from 0.75 to 1.5 megacycles, all impulses received per minute during daylight hours are from near-by sources being propagated as ground waves and are therefore almost the same; while the total number per minute received during night hours are partly from some near-by sources being propagated

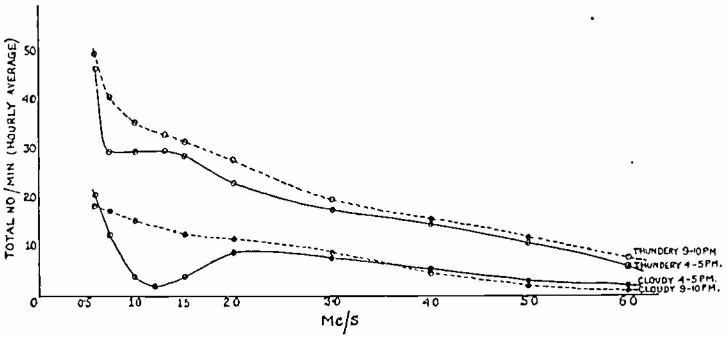

Fig. 4

as ground waves and partly from distant sources propagated as sky waves through the E layer. The total number per minute on frequencies lower than 0.75 megacycle has at all times been greater, as they are always the sum total of those received from near-by and distant sources both being propagated as ground waves. The lower the frequency, the more distant the sources which can be taken to contribute. The total number per minute on frequencies higher than 2 megacycles can be greater or less than that on the medium-wave band according as the contribution is partly from near-by sources received as ground waves and partly from distant sources received through

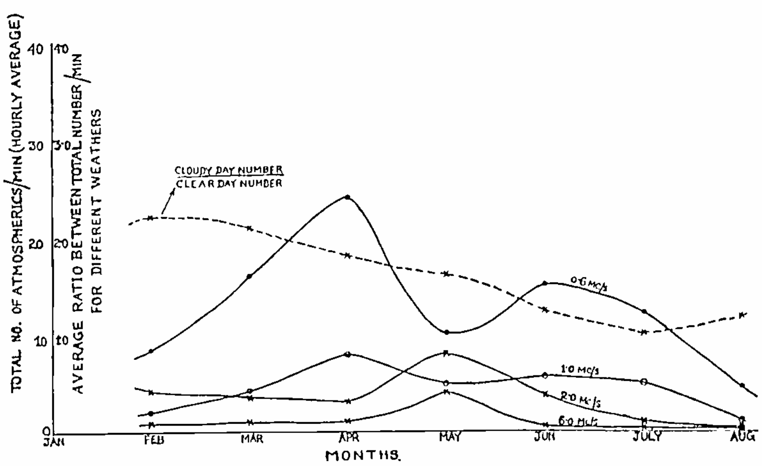

Fig. 5

the F layer, or, the contribution is entirely from distant sources the receiving point being in the skip distance for these frequency components from near-by sources. In cloudy or clear weather, there are no near-by sources and reception from distant sources only are to be considered at all times.

Fig. 5 shows the hourly average of total number of atmospherics per minute for clear afternoon conditions, and the average ratios of (cloudy day/clear day) numbers during each month on frequencies from 0.6 to 6 megacycles.

Fig. 6 shows the typical variation of the total number of atmospherics per minute with the hour

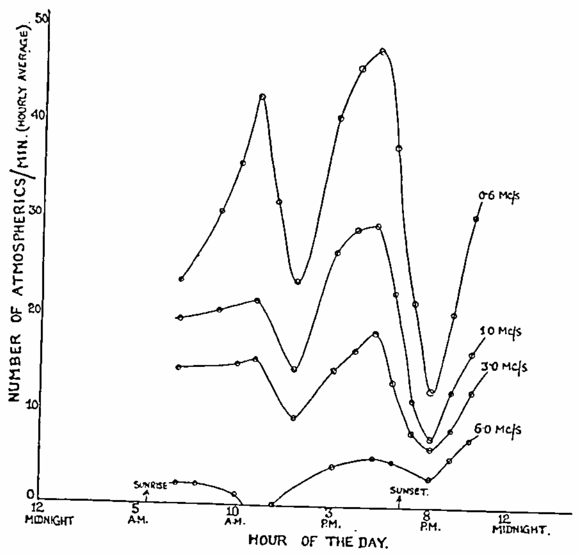

Fig. 6

of the day. It will be observed that on all frequencies except 6.0 megacycles the number per minute reaches high values between 10 to 12 A.M., then decreases to low values at about 1 P.M., then increases to the highest values between 3 to 6 P.M., then decreases again after sunset and subsequently increases again at night.

Fig. 7 shows the peak values of the field strength of the most powerful atmospherics observed for clear afternoon conditions, and the average ratios of (cloudy day/clear day) and (thundery day/clear day) field strengths during each month on frequencies

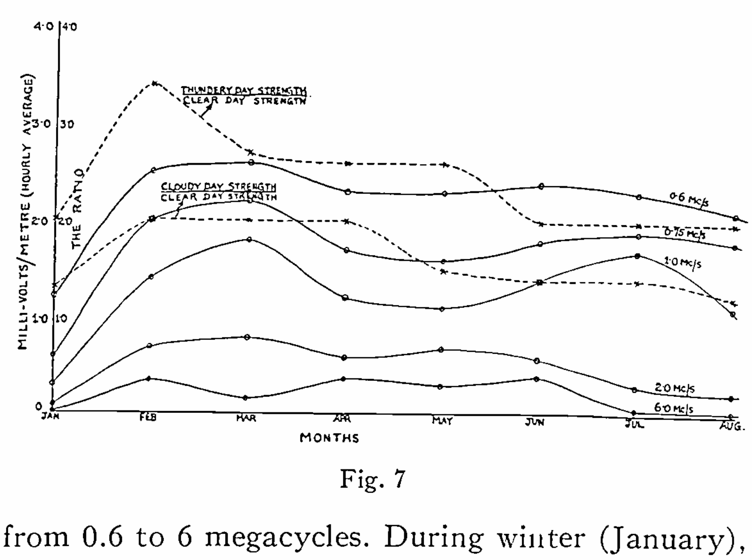

Fig. 7

from 0.6 to 6 megacycles. During winter (January), the field strength was the lowest, during the summer months (February to June) it increased reaching highest values in March and during the rainy months (July and August) it decreased appreciably. The approximate law of variation of atmospheric strength with wavelength has been found to be $s\alpha\lambda$ and $s\alpha\lambda^2$ in winter and rainy seasons, respectively; but in summer the law has been $s\alpha\lambda$ during clear and cloudy weathers and of the form $s=A\lambda+B\lambda^2$ during thundery weather, where s = atmospheric strength, λ = wavelength, and A and B are constants. In winter, as well as in clear and cloudy days of summer, the disturbance from distant sources is propagated through the ionized layers and the strength will be directly proportional to the wavelength; in the rainy season with storm sources near by the disturbance is mostly propagated as ground wave and variation of strength will be much greater with wavelength (since strength will be a function both of wavelength and effective ground conductivity which depends upon wavelength); and in thundery days of summer (unaccompanied by rain) a mixed law of variation will hold.

Fig. 8 shows the typical variation of peak values of field strength with the hour of the day. It will be observed that the strength is high before 6 A.M., reaches low values between 9 A.M. and 1 P.M., then rises to highest values in the afternoon between 1 to

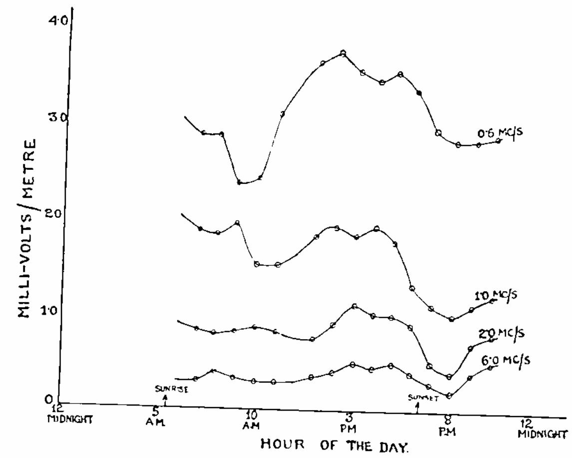

Fig. 8

6 P.M., then falls to very low values after sunset and subsequently rises to high values at night.

In the absence of a definite broadcast transmission standard in India, the standards adopted in the United States of America[4,5] and Great Britain[6] have been examined and a suggestion is given for the standard to be adopted in India on the basis of atmospheric field-strength measurements. In India, the atmospheric disturbance is much more severe than in most parts of the United States and Great Britain. The signal field strength which will be at least 20 decibels above the peak strength of the worst atmospheric is shown in Table I.

It is suggested that the effective service area of a medium-wave broadcast station in India should be divided into three zones; viz, (1) first-class zone, in which the minimum signal strength is such that good reception is possible almost throughout the year even

[4] Fifth Annual Report of the Federal Radio Commission, U.S.A., (1931), p. 30.

[5] Seventh Annual Report of the Federal Radio Commission, U.S.A., (1933), p. 19.

[6] Ashbridge, Bishop, and Maclarty, *Jour. I.E.E.* (London), vol. 77, p. 437, (1935).

TABLE I

Season	Signal field strength in millivolts per meter which is 20 decibels above atmospheric peak strength	
	Medium-Wave Band (0.6–1.5 megacycles)	Short-Wave Band (3–30 megacycles)
(1) Winter (October–January)	16–3	2–0.3
(2) Summer (February–June)	(a) 80–30 (b) 50–25	(a) 15–4 (b) 10–3
(3) Rainy Season (July–September)	40–8	2.5–0.8

(a) Thundery weather.
(b) Weathers other than thundery.

in thundery weather; (2) second-class zone, in which the minimum signal strength is such that good reception is possible only during 80 to 90 per cent of the days in the year; and (3) third-class zone, in which minimum signal strength is such that good reception is possible during 60 to 70 per cent of the days in the year.

Table II shows the minimum signal strength desirable for each zone in the case of city and rural areas.

TABLE II

	Minimum signal strength (millivolts per meter) in zones					
	First Class		Second Class		Third Class	
Megacycles	1.5	0.75	1.5	0.75	1.5	0.75
City Area	30	65	20	35	10	20
Rural Area	20	40	12	25	6	12

Two cases of medium-wave transmitters have been taken to show the limits of the various zones as given in Table III.

TABLE III

Transmitter	Limit of the zone from station in miles					
	City Area			Rural Area		
	First	Second	Third	First	Second	Third
(1) 1.5 kilowatts, 370-meter transmitter. Aerial current, 18 amperes; effective height, 20 meters.	2	6	12	6	8	20
(2) 5 kilowatts, 235-meter transmitter. Aerial current, 35 amperes; effective height, 30 meters.	10	16	32	12	20	40

The height of the ionized layer is comparatively lower in India and other tropical countries and therefore the effect of the sky wave reflected from the upper atmosphere becomes prominent even at 60 to 65 miles from the station. It will be noted from Table III that the whole of the effective service area will have "fading-free" service in each of the above cases.

Acknowledgment

The first author desires to thank Mr. H. L. Kirke of the British Broadcasting Corporation for suggesting observation of atmospherics on several frequencies in medium- and short-wave bands during his visit in March, 1936, and Mr. C. W. Goyder of All-India Radio for discussions with him from time to time between January and August, 1938. The authors desire to thank very heartily Professor P. N. Ghosh for giving them all workshop and construction facilities.

Measurements of Currents and Voltages Down to a Wavelength of 20 Centimeters*

M. J. O. STRUTT†, NONMEMBER, I.R.E., AND K. S. KNOL†, NONMEMBER, I.R.E.

Summary—*The more common circuits in which diode voltmeters are used are discussed in Section I. The properties of these circuits are analyzed on the basis of the diode characteristics, with special reference to the input impedance. In Section II, two devices are described for current measurements, a hot-wire air-expansion device and thermocouples of special construction. Consideration of the requirements in the calibration of these devices in the short-wave range is followed by a layout which is described in detail, and with which calibrations down to a 20-centimeter wavelength could be carried out with an accuracy of within 1 per cent; one type of thermocouple proved exceptionally suitable for absolute current measurements.*

In Section III, it is shown how a diode voltmeter can be calibrated in the short-wave range with the calibrated current-measuring devices. Of the two arrangements described, one can be used down to approximately 3 meters and the other down to a wavelength of 75 centimeters. Diode voltmeters with diodes of special design exhibit at this wavelength a maximum deviation of 2 per cent from the calibrated values obtained with longer waves.

I. Voltmeters for Use in the Ultra-Short-Wave Range

A large number of instruments have been devised for measuring voltages in the ultra-short-wave range. Discussion in the present paper will be limited to diode voltmeters, the usual circuits of which are sufficiently known. A special circuit is shown in Fig. 1, which can be used, if the voltage must be measured between two points (1 and

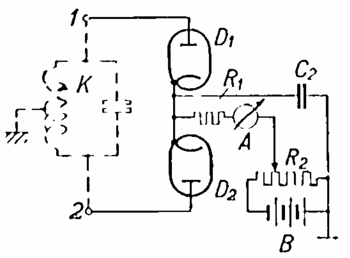

Fig. 1—Push-pull diode voltmeter circuit.
D_1 and D_2 = diodes
A = microammeter
R_1 = leak resistance approximately 0.1 megohm for direct current
C_2 = blocking condenser (mica) approximately 1000 micromicrofarads
R_2 = potentiometer
B = battery.

* Decimal classification: R243.1×R242.1. Original manuscript received by the Institute, March 31, 1939; abridgment received, July 24, 1939.

† Natuurkundig Laboratorium der N. V. Philips' Gloeilampenfabrieken, Eindhoven, Holland.

2) which have the same impedance with respect to earth, for instance if 1 and 2 are the terminals of a push-pull resonant circuit.

It is of great importance to keep the damping caused by the diode voltmeter as small as possible. For that purpose we have to know, in the first place,

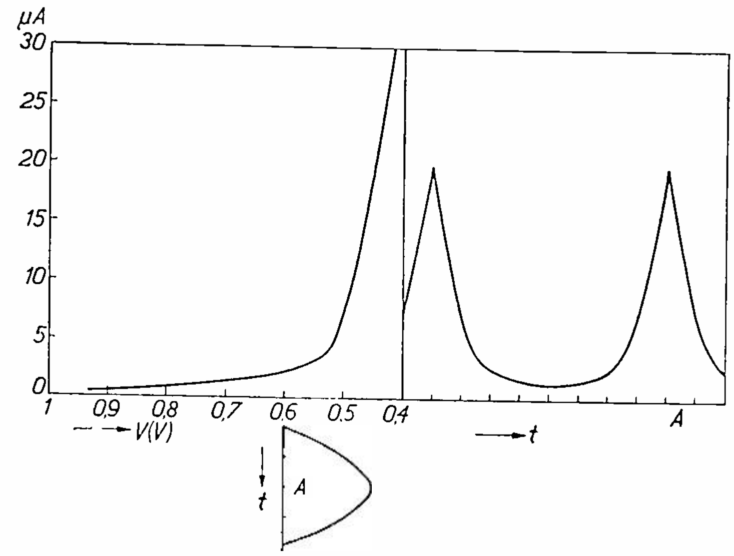

Fig. 2—Left: Characteristic of diode. Ordinate, microamperes and abscissa, voltage. Below: Half cycle of an alternating voltage as a function of the time t. Right: Resultant current through the diode as a function of the time at a direct voltage of -0.6 volt and an alternating-voltage amplitude of 0.15 volt.

the input resistance of the diode under the conditions of use. A theory for the diode with exponential current-voltage characteristic is given by Aiken.[1] We shall outline very shortly a simplified method to get an idea of the order of magnitude of the input resistance. In Fig. 2 (left-hand upper corner) the direct-voltage characteristic of a diode is given. When upon a direct voltage an alternating voltage is superposed (left bottom), the curve on the right of Fig. 2 will be obtained. This curve can be approximately represented by triangles (Fig. 3), and the fundamental

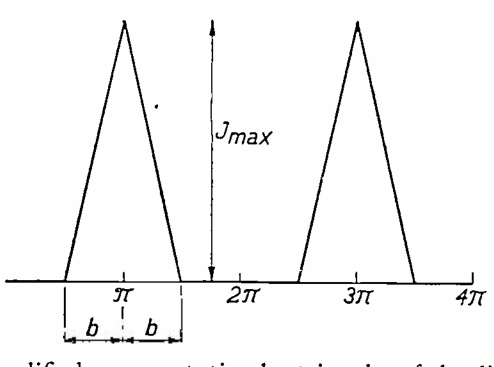

Fig. 3—Simplified representation by triangles of the diode current on the right of Fig. 2 plotted as a function of the time.

component of the resolved Fourier series of the current as a function of the time can be readily calculated for these triangular figures; the following expression is then obtained for the amplitude of the alternating current

$$\frac{2}{\pi} I_{\max} \cdot \frac{1 - \cos b}{b} \cos \omega t \tag{1}$$

[1] Refer to Bibliography.

where ω is the angular frequency of the alternating voltage. The resulting direct current is, according to Fig. 3,

$$I_g = \frac{b I_{\max}}{2\pi}. \tag{2}$$

From (1) and (2) we get for the alternating current

$$4 I_g \cdot \frac{1 - \cos b}{b^2} \cos \omega t. \tag{3}$$

In these equations and in Fig. 3, b/π determines the portion of a period during which current flows. If $b/\pi = 1$, the factor $4(1 - \cos b)/b^2$ becomes equal to 0.81 and for $b \to 0$ we get the value 2.

The ratio of the amplitude E of the alternating voltage of the source to the alternating-current amplitude may be termed the effective alternating-current resistance or impedance R_i of the diode. With high alternating voltages R_i is equal to $E/2I_g$ and for low values is approximately $E/0.8\ I_g$. This determines the order of magnitude of R_i for known values of E and I_g. The impedance of the diode to an alternating voltage may be represented by this resistance R_i in

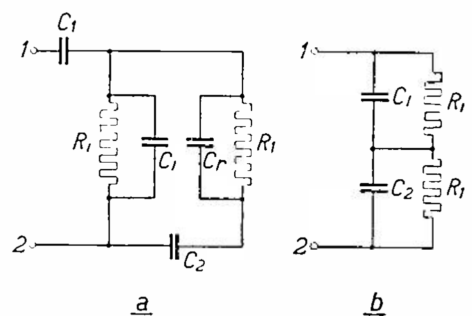

Fig. 4—Equivalent circuits of two diode voltmeter circuits, pertaining to the alternating-current components. Symbols described in the text.

parallel with a capacitance C_i. Actually C_i is not constant during a period of the alternating voltage, but varies in a similar way to R_i. With a low diode direct current I_g, C_i may be assumed equal to the capacitance of the diode measured with a cold cathode.

In order to estimate the damping of the diode voltmeter two possible circuits are drawn in Figs. 4 (a) and 4 (b). In Fig. 4 (a) the diode voltmeter is connected by means of a capacitance C_1. Here R_i and C_i are, respectively, input resistance and input capacitance of the diode, R_1 is the resistance leak, and C_r the parallel capacitance associated with every resistance leak. C_2 is a blocking condenser, the reactance of which is very small compared to R_1 at the measuring frequency and so can be replaced by a short circuit. We can replace the whole circuit of Fig. 4 (a) by a resistance R_e in parallel with a capacitance C_e. We thus have

$$\left.\begin{aligned} R_e &= R_1 \frac{1 + \omega^2 (C_i + C_r + C_1)^2 R_1^2}{\omega^2 C_1^2 R_1^2} \\ C_e &= C_1 \frac{1 + \omega^2 (C_i + C_r)(C_i + C_r + C_1) R_1^2}{1 + \omega^2 (C_i + C_r + C_1)^2 R_1^2} \end{aligned}\right\}. \tag{4}$$

In discussing these equations we may postulate two different conditions. It can be stipulated that the same alternating voltage shall exist between the diode electrodes as between 1 and 2. In this case the reactance of C_1 must be small compared with the impedance in the parallel circuit of R_i with C_i and C_r. We then get from equation (4) for short waves that R_e is roughly equal to R_i and C_e equal to C_i+C_r. But it may also be stipulated that R_e be made as large as possible and C_e as small as possible, in order that the diode circuit shall cause the minimum possible interference in the other parts of the measuring system between points 1 and 2. It is obviously advantageous in obtaining high values of R_e to make $\omega C_1R_1<1$. If, for instance, we take $\omega=10^8$ (roughly 20 meters wavelength in air) and $R_1=20$ kilohms, this condition becomes: 2 C_1 (micromicrofarads) <1; e.g., $C_1=0.2$ micromicrofarads. The numerator of the expression for R_e is then approximately 5 and R_e is roughly equal to 100 R_1. In this example, the capacitance becomes nearly equal to C_1. For higher frequencies C_1 may be given roughly the same value, for we have approximately

$$R_e = R_1(C_i + C_r + C_1)^2 \cdot C_1^{-2}.$$

Consideration of the circuit diagram in Fig. 4 (b) need not occupy much space. Capacitance C_2 has such a high value that its reactance is small compared with R_1. Therefore, we may dispense with R_1 altogether and we obtain the following expressions for the input resistance R_e and the input capacitance C_e between 1 and 2, the latter being again assumed to be in parallel with R_e:

$$\left.\begin{aligned} R_e &= R_i\,\frac{1+\omega^2(C_i+C_2)^2R_i^2}{\omega^2C_2^2R_i^2} \\ C_e &= C_2\,\frac{1+\omega^2C_i(C_i+C_2)R_i^2}{1+\omega^2(C_i+C_2)^2R_i^2} \end{aligned}\right\}. \qquad (5)$$

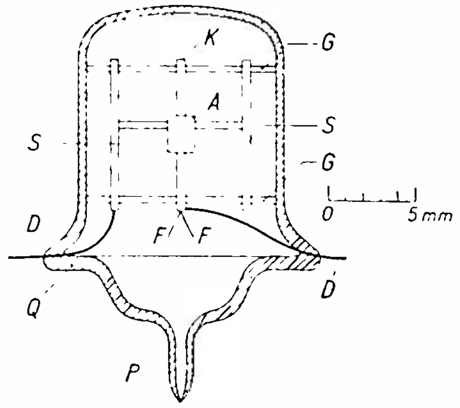

Fig. 5—Sketch of a short-wave measuring diode for use in the circuits in Fig. 4.

A = anode cylinder
K = cathode cylinder
F = filament
S = supporting rod
G = small mica plate
P = exhaust tube
Q = pinch
D = electrode leads through pinch Q

For example, take $C_2=10$ micromicrofarads, $\omega=10^8$, and $R_i=1$ megohm. R is then roughly equal to R_i and C_e equal to C_i. While R_e and C_e are here of the same order of magnitude as with a small C_1 in the circuit of Fig. 4 (a), circuit 4 (b) is the more advantageous as practically the whole of the alternating-voltage amplitude, impressed between terminals 1 and 2, is applied to the diode electrodes.

In the short-wave range, it is extremely important to keep all conductors and leads as short as possible, since they constitute self-inductances which at the

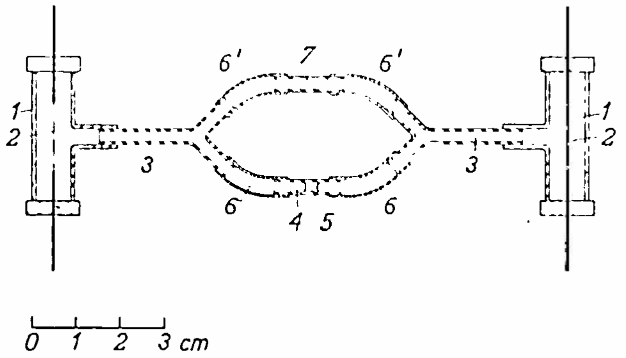

Fig. 6—Sketch of a compensation hot-wire milliammeter.
1 = polystyrol tube with airtight connections
2 = constantan wire, 10–20 microns diameter
3 = glass tube with branch
6 and 6' = rubber tubing (valve tubes)
4 and 7 = glass capillaries, 7 being narrower than 4 (4 and 7 are drawn equally in the figure)
5 = droplet of colored liquid

frequencies in question are equivalent to impedances and can then no longer be neglected in comparison to the other impedances. Taking into account the indispensable leads, the diode impedance should be made as high as possible; i.e., C_i is given a small value and R_i a high value. On the other hand, these values are directly determined by the sensitivity of the diode; i.e., by the direct current I_g when a given alternating-voltage amplitude E is impressed on the diode. We have had quite efficient diodes made with C_i approximately 0.5 micromicrofarad and R_i approximately 1 megohm at an I_g value of roughly 0.5 microampere. One of these special diodes is shown in Fig. 5.

II. Ammeters for Measuring Current in the Decimeter-Wave Range and Their Calibration

The first instrument, which we investigated, consisted of a sealed air-filled tube enclosing a hot wire. When this wire becomes heated by the passage of current, the air in the tube expands and displaces a drop of liquid in a connected capillary. The practical application of this simple principle is shown in Fig. 6. The whole arrangement consists of two exactly similar halves; the current under measurement is passed through one hot wire 2, while a known direct current is passed through the other hot wire 2. These hot wires are so thin, being made of constantan about 20 microns in diameter, that no disturbing skin effect (the increase in resistance is less than 3 per cent) occurs even with the very shortest waves, e.g., 20-centimeter waves. The hermetically sealed tubes 2 are made of polystyrol, a material with a very low thermal conductivity, and they are moreover encased in asbestos wool to obtain maximum thermal insulation. The attached glass tubes 3 have a branch connected by rubber tubing 6 and 6′ to the capillaries 7 and 4. Capillary 4 is wider than 7 and contains a drop of a colored liquid 5, which can be viewed through a reading microscope with scale. The currents passed through hot wires 2 are switched on and off simultaneously. The direct current through

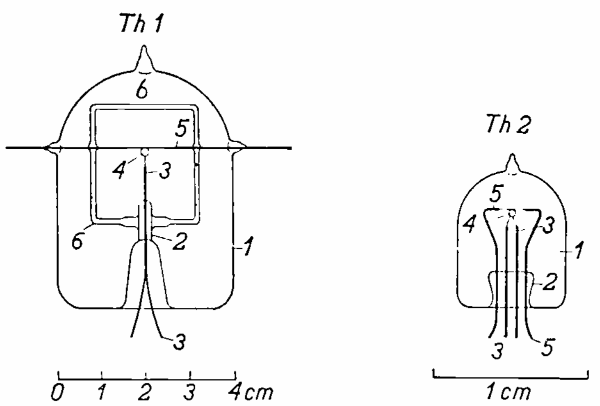

Fig. 7—Two high-vacuum thermocouples for current measurements.
5 = hot wire, 10–20 microns diameter
4 = bead of insulating material
3 = two welded thermowires
2 = pinch
1 = glass envelope
6 = glass arm for supporting filament 5 in thermocouple *Th*1

one of the hot wires is regulated until the drop of liquid 5 remains stationary on switching on the currents. The narrow capillary 7 serves for slow compensation of the differences in pressure in the two tubes 1. With this arrangement, which can be calibrated with direct current, currents of a few milliamperes can be measured with an accuracy within about 1 per cent. The resistance of the hot wires is of the order of 20 ohms. Hence, with a current of 2 milliamperes, a power value of $4 \cdot 10^{-6} \cdot 20 = 8 \cdot 10^{-5}$ watt can be measured with an accuracy within about 2 per cent.

The other measuring arrangements which we used consisted of high-vacuum thermocouples, of which two specially satisfactory designs are shown side by side in Fig. 7. The hot wire 5 in the couples *Th*1 and *Th*2 is made of very thin wire which exhibits no skin effect down to the very shortest waves. Using a suitable millivoltmeter for measuring the thermovoltage, e.g., (Cambridge Instrument Company's Unipivot) currents of some milliamperes through the hot wire can be measured with an error within approximately 1 per cent when the hot-wire resistance is of the order of 20 ohms. The arrangement in *Th*1 with the straight leads to the hot wires is better than the standard squash type *Th*2, in that the hot wire has a lower capacitance and a lower mutual inductance towards the thermocouple wires and their leads. On the other hand the construction of *Th*1 is more complicated than *Th*2.

We shall now deal with the calibration of the measuring arrangements briefly outlined above. One of the main difficulties here was to satisfy the condition that the same alternating current must be passed through the two components under comparison with each other, or alternating currents which are in a simple known ratio to one another. The arrangement shown diagrammatically in Fig. 8 was devised to arrive at this equality. A parallel-wire system 1 is coupled with a transmitter *Tr*, these conductors being arranged as symmetrically as possible with respect to surrounding apparatus, while the coupling itself is also made as symmetrical as possible. Two similar high-vacuum wire fuses 2 are connected to two thermocouples 3 of type *Th*1 (as shown in Fig. 7); these couples being made as closely equal to each other as possible. The ends of the couple wires are connected to the surrounding housing (earth) by blocking condensers 4 and connect up with the millivoltmeters 5. Unit 6 is the thermocouple to be calibrated, while unit 7 is one of the two tubes 1 in Fig. 6. The resistances of the hot wires in 6 and 7 have again been chosen as closely equal to each other as possible (their deviation is much below 1 per cent). The geometrical center of the bridge between the conductors of the parallel-wire line is earthed at 8. With this arrangement to which the maximum degree of symmetry has been imparted, the center of couple 6 is exactly at the same distance from 8 as the center of hot wire 7. A small, and in this arrangement unavoidable asymmetry is caused by the earthing condensers 4, which are contained in unit 6 but not in 7. These condensers 4, as well as the careful screening against external influences of the whole arrangement by the provision of a sheet-copper enclosure, as shown in Fig. 9, were found to be necessary to avoid any hand-capacitance effect, i.e., an alteration in the deflection obtained on the meters by the approach of the observer.

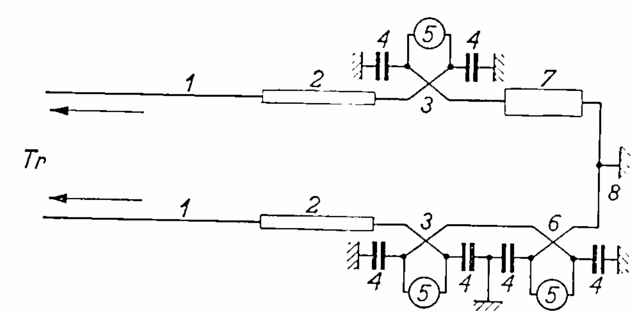

Fig. 8—Diagrammatic sketch of arrangement for comparison of current-measuring devices at a wavelength of 20 centimeters.
Tr = short-wave transmitter
1 = parallel-wire line
2 = high-vacuum hot-wire fuses
3 = two exactly equal thermocouples
4 = blocking condensers (mica condensers)
5 = millivoltmeters (direct voltage)
6 = comparison thermocouple with a hot-wire resistance exactly equal to that in tube 7 (according to Fig. 6)
8 = earth connection at geometrical center of line.

The arrangement shown in Fig. 6 was calibrated by passing known direct currents through the two hot wires. As all sources of error have been eliminated as far as possible in this instrument, we regard it as a standard ammeter for waves down to wavelengths of

20 centimeters and have compared the thermocouples with this standard. The thermocouples were also calibrated with direct current. The currents through the two thermocouples 3 could be adjusted to well within 1 per cent, and the following values were obtained.

Wavelength centimeters	Thermocouple 6 (Fig. 9) milliamperes	Instrument 7 (Fig. 9) milliamperes	Error of 6 per cent
114	6.72	6.65	+1
114	7.60	7.55	+0.7
90	6.04	6.00	+0.7
90	7.30	7.20	+1.4
50	5.66	5.68	−0.3
50	5.34	5.50	−3
50	5.38	5.34	+0.7
22.5	5.09	5.10	−0.2
22.5	5.86	5.95	−2
22.5	6.06	5.95	+2

On arriving at the values in this table, the corresponding milliampere values were determined both for thermocouple 6, which was of type *Th*1 as in Fig. 7, and for unit 7 (Fig. 6), from the readings on the measuring instruments using the direct-current calibration values. When a thermocouple *Th*2 (Fig. 7) was used at 6 in Fig. 9, this couple already gave an error of approximately 2 per cent at a wavelength as great as 150 centimeters. This is in agreement with the statement above that this type of couple is more susceptible to disturbance from mutual inductance and capacitance of the hot wire towards the thermowires than type *Th*1.

For waves shorter than about 90 centimeters a disturbing effect occurred, viz., the parallel-wire lead could not be adjusted to a balanced condition with respect to the enclosure. Therefore an improvement

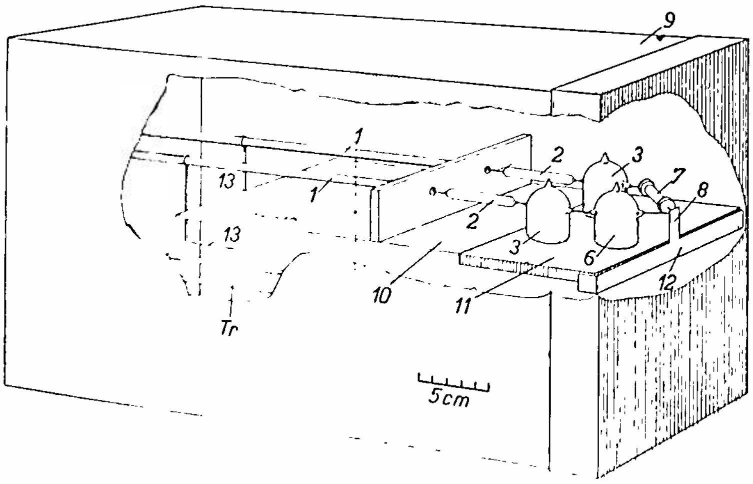

Fig. 9—Practical form of the arrangement shown in Fig. 8. A housing made of 1-millimeter sheet copper is divided into two compartments by the sheet-metal partition 10. The lower compartment contains the transmitter *Tr* which at 13 is connected with line 1 in the upper compartment. Numerals 2, 3, 6, 7, 8 indicate the same units as in Fig. 8. 11 is a polystyrol plate on which the thermocouple and the unit 7 are fixed; 12 is a strip of sheet copper for earthing the center of the line.

of the arrangement of Fig. 9 was used for these waves. In this layout the screening of the transmitter against the parallel-wire line was made more effective, while the thickness of the wires and their mutual distance were reduced considerably. So the effect, mentioned above, could be avoided. It was, however, difficult to balance exactly the currents through the two comparison thermocouples 3 (see Fig. 8). Differences between these currents of at most 5 per cent occurred. This is caused by some remaining asymmetries in the system. It may, however, be assumed, that the currents through one of thermocouples 3 will be proportional to those through the instruments behind it.

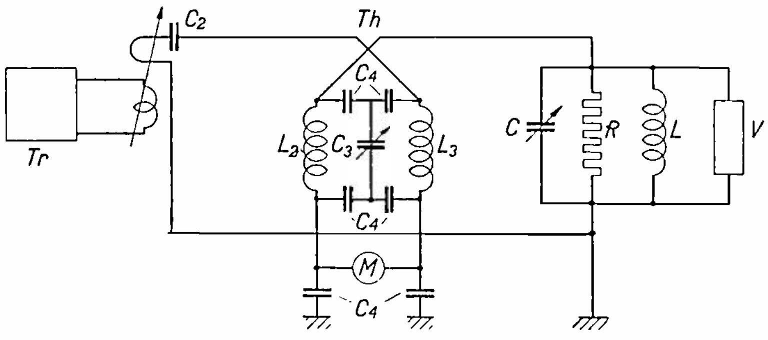

Fig. 10—Circuit for absolute calibration of a diode voltmeter (V) using a calibrated thermocouple *Th*.
Tr = transmitter
C_2 = small condenser
C_4 = blocking condensers
L_2 and L_3 = two small self-inductances, which with C_3 form a resonant circuit tuned to the frequency under measurement
M = millivoltmeter
CL = resonant circuit with calibrated variable condenser C
R = impedance of this circuit when tuned

It may be concluded from these measurements that, provided the necessary precautions are taken, thermocouples of type *Th*1 enable alternating currents of a wavelength of 20 centimeters to be measured with an error of less than about 2 per cent.

III. Absolute Calibration of Voltmeters

In short-wave measurements an absolutely calibrated voltmeter is essential for certain applications.

A circuit as shown in Fig. 10 has been used for the absolute calibration of diode voltmeters and thermocouples against each other. The two equivalent self-inductances L_2 and L_3 are tuned to the frequency under measurement by means of a condenser C_3. Together they constitute a resonant circuit inserted between the thermocontact of the thermocouple and earth. The impedance of this circuit when tuned is R_t, while the impedance of circuit CL on similar adjustment is R. Impedance R_t must be large compared with R, which follows from the following considerations: As shown in Fig. 7, there is no direct contact between the hot wire and the leads to the couple, although there is capacitance and mutual inductance between them. This coupling may be represented by a capacitance C_t of several tenths of a micromicrofarad for ordinary layouts, which at a 2-meter wavelength corresponds to an impedance of the order of several thousand ohms. This capacitance is in series with R_t and constitutes an impedance of a magnitude of $(R_t^2+1/\omega^2C_t^2)^{1/2}$, which is in parallel with R. If an alternating voltage is impressed on the circuit CL, the current will be split in the thermocouple. To divert as little current as possible to earth, C_t must be made as small as possible and R_t as large as possible. A large value of R_t can be readily obtained in the short-wave range by inserting a symmetrical quarter-wavelength line between the thermowire connections of the thermocouple and the earthed meter terminals. We make C_2 so small that the current indicated by the couple remains constant,

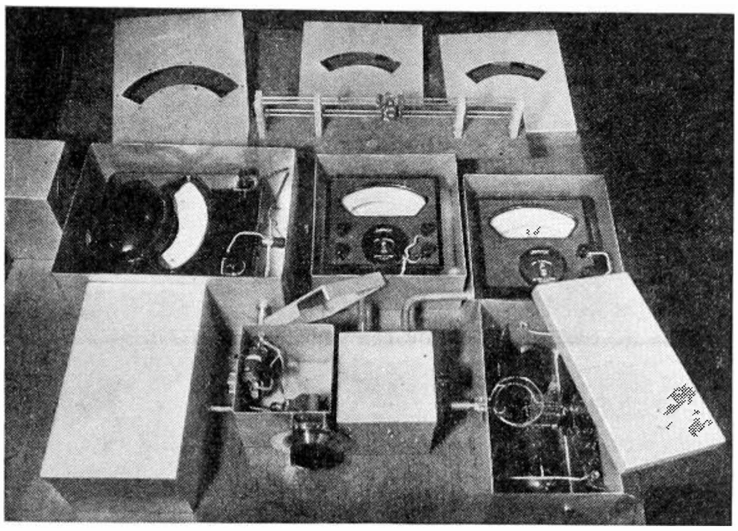

Fig. 11—Apparatus according to circuit shown in Fig. 10 for a wavelength of 4 meters. The sheet-metal enclosures in the foreground contain, from right to left, the transmitter Tr, thermocouple Th with circuit $L_2L_3C_3$, the diode voltmeter V and the circuit CL with calibrated capacitance C (scale) and batteries of the diode voltmeter. In the rear, meters and a small push-pull transmitter for a wavelength of 40 to 80 centimeters.

despite alterations in the impedance formed by C and L. Then the impedance R of this circuit when tuned can be measured by adjusting the calibrated variable condenser C and plotting the resonance curve with the diode voltmeter V. If the thermocouple has been absolutely calibrated, we know the alternating-current amplitude through R and hence the alternating-voltage amplitude at the terminations of R, which serves for the calibration of V. A measuring device for a wavelength of approximately 4 meters designed on this principle is shown in Fig. 11, and showed that the diode voltmeter, when connected up according to Fig. 4 (a) and when using diodes as shown in Fig. 5, gives, at a wavelength of 4 meters, the same volt readings within 1 per cent as at much lower frequencies, e.g., at a wavelength of 200 meters.

A parallel-wire line was used for the absolute calibration of diode voltmeters at wavelengths of about 1 meter. The principle of measurement may be gathered from Fig. 12. A symmetrical parallel-wire line L is fed from an alternating-voltage source of nearly zero internal impedance, and in series with a resistance R_0 which is equal to the surge impedance of the line. A thermocouple Th with a hot wire having a known resistance at this wavelength is connected to the termination of the line. As the hot wire and the leads still have self-inductance, small variable condensers C are provided for compensating these self-inductance values. A diode voltmeter D is mounted on a base of insulating material (polystyrol) which can be made to slide along the line. The distance a should be one-half wavelength. The conductors of the line have such a low resistance (copper tubes of about 1 centimeter diameter) that the attenuation of the line can be neglected. The adjustment of condensers C is altered until a minimum voltage amplitude is obtained on the line for a distance a of one-half wavelength. The complete diode voltmeter had a capacitance C_d of approximately 0.5 micromicrofarad between the conductors of the line (diode, see Fig. 5). The corresponding impedance at a 1-meter wavelength is approximately 1000 ohms. The surge impedance R_0 of the parallel wire is approximately 300 ohms. The diode voltmeter has only a slight effect on the characteristics of the line as the impedance between points 1 and 2, as viewed from 3, is $(1/R_0^2+\omega^2C_d^2)^{-1/2}$, which for the values stated is only 5 per cent less than the impedance R_0. As a is one-half wavelength the voltage amplitude between points 1 and 2 is the same as the voltage amplitude at the terminations of the hot wire 3, and is equal to the resistance of the hot wire multiplied by the current amplitude measured with the thermocouple. Measurements with this circuit indicated that at a wavelength of 1 meter the diode voltmeter gives the the same voltage reading within about 2 per cent as with much longer waves (e.g., 200 meters). The hot-wire resistance was about 30 ohms, the current amplitude approximately 10 milliamperes, and hence the voltage amplitude approximately 0.3 volt.

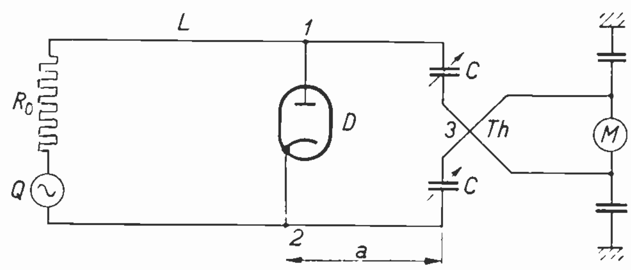

Fig. 12—Circuit of an arrangement for the absolute calibration of a diode voltmeter (D) at a wavelength of about 1 meter Q is an alternating-voltage source without internal impedance, which is in series with the resistance R_0, equal to the surge impedance of the line L.

D = diode voltmeter C_1 = blocking condensers
C = small variable condensers M = millivoltmeter
Th = calibrated thermocouple

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Characteristics of the Ionosphere at Washington, D.C., October, 1939, with Predictions for January, 1940*

T. R. GILLILAND†, ASSOCIATE MEMBER, I.R.E., S. S. KIRBY†, ASSOCIATE MEMBER, I.R.E., AND N. SMITH†, NONMEMBER, I.R.E.

DATA on the critical frequencies and virtual heights of the ionosphore layers during October are given in Fig. 1. Fig. 2 gives the monthly average values of the maximum usable frequencies for undisturbed days, for radio transmission by way of the regular layers. The F_2 and F layers ordinarily determined the maximum usable frequencies during the day and night, respectively. Fig. 3 gives the distribution of hourly values of F and F_2 data about the undisturbed average for the month. Fig. 4 gives the expected values of the maximum usable frequencies for radio transmission by way of

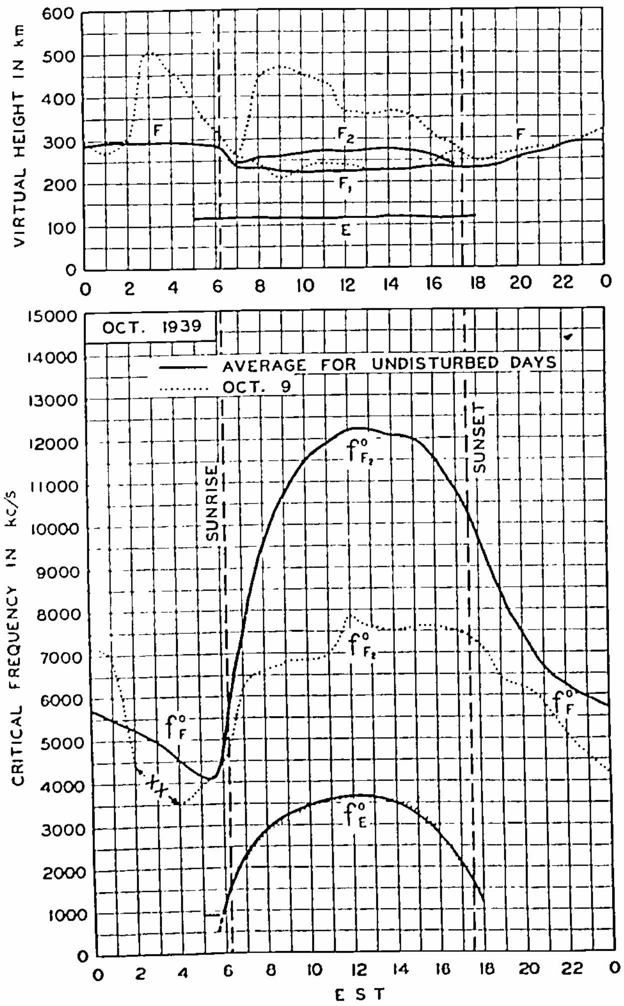

Fig. 1—Virtual heights and critical frequencies of the ionospheric layers, October, 1939. The solid-line graphs are the averages for the undisturbed days; the dotted-line graphs are for the ionospheric storm day of October 9. The crosses represent the times on October 9 when the F-layer reflections were so diffuse that the critical frequencies could not be determined.

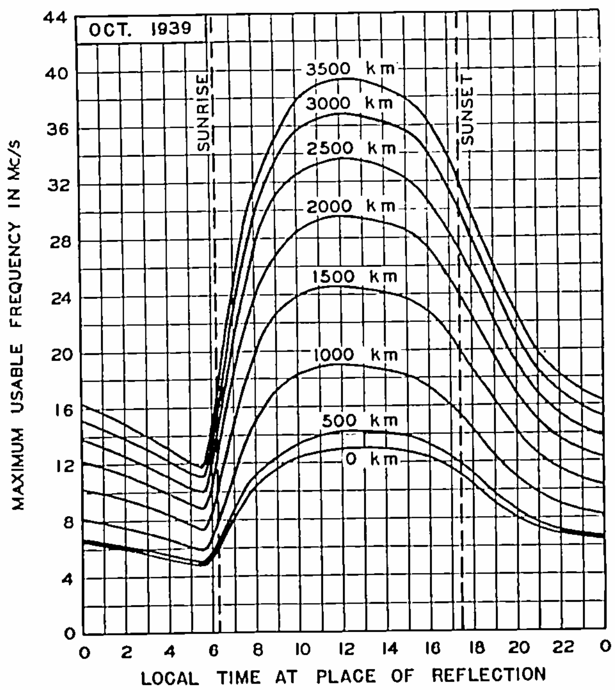

Fig. 2—Maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days for October, 1939.

* Decimal classification: R113.61. Original manuscript received by the Institute, November 10, 1939. These reports have appeared monthly in the Proceedings starting in vol. 25, September, (1937). See also vol. 25, pp. 823–840, July, (1937). Publication approved by the Director of the National Bureau of Standards of the U. S. Department of Commerce.
† National Bureau of Standards, Washington, D.C.

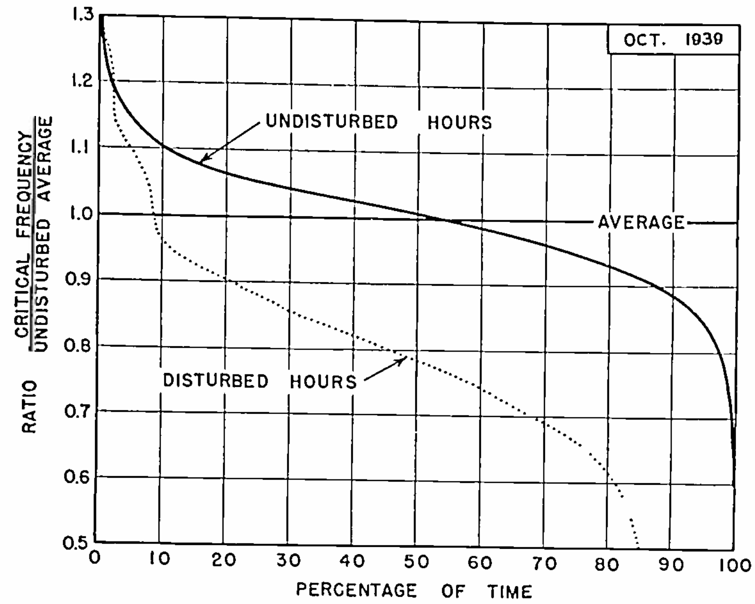

Fig. 3—Distribution of F_2 and F_2-layer ordinary-wave critical frequencies (and approximately of maximum usable frequencies) about monthly average. Abscissas show percentage of time for which the ratio of the critical frequency to the undisturbed average exceeded the values given by the ordinates. The solid-line graph is for 500 undisturbed hours of observation; the dotted graph is for 143 disturbed hours of observation listed in Table I.

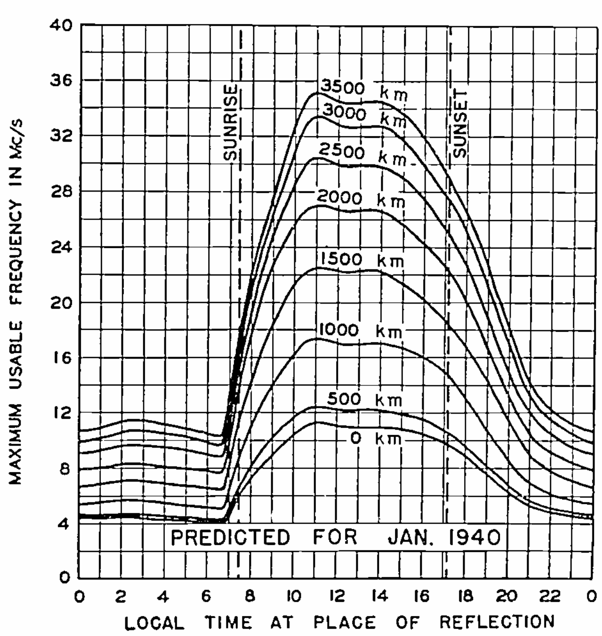

Fig. 4—Predicted maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days, for January, 1940.

the regular layers, average for undisturbed days, for January, 1940. Ionospheric storms and sudden ionospheric disturbances are listed in Tables I and II,

TABLE I

IONOSPHERIC STORMS (APPROXIMATELY IN ORDER OF SEVERITY)

Day and hour E.S.T.	h_F before sunrise (km)	Minimum $f_F{}^0$ before sunrise (kc)	Noon $f_{F_2}{}^c$ (kc)	Magnetic character[1] 00–12 G.M.T.	Magnetic character[1] 12–24 G.M.T.	Ionospheric character[2]
October						
13 (after 1400)	—	—	—	1.5	1.7	1.6
14 (to 1000)	386	diffuse	—	1.7	1.2	1.3
14 (after 2000)	—	—	—	1.7	1.2	1.4
15	324	3400	8300	1.6	0.5	1.3
16 (to 1200)	338	2600	11000	0.6	0.8	0.5
8 (after 2200)	—	—	—	0.2	0.3	0.3
9	404	3500	7800	0.8	0.5	1.5
10 (to 0600)	322	2400	—	0.0	0.1	0.6
3 (after 0400)	345	4800	8300	0.7	1.3	1.2
4 (to 0600)	362	2900	—	1.6	0.7	1.2
5 (after 2100)	—	—	—	0.4	0.7	0.5
6 (to 0600)	368	diffuse	—	1.1	0.6	1.2
18 (after 2000)	—	—	—	0.5	0.7	0.3
19 (to 0800)	364	3800	—	0.7	0.2	0.7
13 (0100 to 0600)	358	4200	—	1.5	1.7	.4
For comparison: Average for undisturbed days	291	4050	12200	0.2	0.3	0.0

[1] American magnetic character figure, based on observations of seven observatories.

[2] An estimate of the severity of the ionospheric storm at Washington on an arbitrary scaie of 0 to 2, the character 2 representing the most severe disturbance.

respectively. During October few strong vertical-incidence sporadic-E reflections were observed. Pro-

TABLE II

SUDDEN IONOSPHERIC DISTURBANCES

Day	G.M.T. Beginning	G.M.T. End	Locations of transmitters	Relative intensity at minimum[1]
October				
5	1902	1930	Ohio, Ont., D. C.	0.1
18	1838	1940	Ohio, Ont., Mass., D. C.	0.02
20	1401	1420	Ohio, Ont., Mass., D. C.	0.02
21	1934	2130	Ohio, Ont.	0.1
22	2153	2300	Ohio, Ont., Mass.	0.1

[1] Ratio of received field intensity during fade-out to average field intensity before and after; for station WLWO, 6060 kilocycles, 650 kilometers, distant.

longed periods of low-layer absorption occurred for several hours during the middle of the day on October 20 and 23.

Institute News and Radio Notes

RAYMOND A. HEISING
President, 1939

Raymond A. Heising was born August 10, 1888, at Albert Lea, Minnesota. He received the E.E. degree from the University of North Dakota in 1912 and the M.S. degree from the University of Wisconsin in 1914. Since 1914 he has been a member of the technical staff of of the Engineering Department of the Western Electric Company and of its successor, the Bell Telephone Laboratories.

Mr. Heising designed and operated the Arlington radiotelephone transmitter which in 1915 was heard in Paris, Honolulu, and Darien. He has been closely identified with the development of modulation systems and invented the constant-current method on which broadcasting is based.

During the World War, Mr. Heising was engaged in numerous radio projects of a military nature and also instructed technical men assigned by the War Department to the Western Electric laboratories. Since the war he has continued in research and development work. In 1930, he was made supervisor of the department which carries on piezoelectric research and part of the ultra-high-frequency and crystal research. He has published numerous papers in the PROCEEDINGS *and other journals, and holds over one hundred patents applying to practical radio developments.*

He joined the Institute as an Associate in 1920, transferring to Fellow in 1923. The Morris Liebmann Memorial prize was awarded to him in 1921. Mr. Heising has served as a member of the the Board of Directors for several years maintaining, as well, an active participation in committee work.

Board of Directors

The regular monthly meeting of the Board of Directors held on November 1, 1939, was attended by R. A. Heising, president; H. H. Beverage, Ralph Bown, F. W. Cunningham, Alfred N. Goldsmith, Virgil M. Graham, L. C. F. Horle, C. M. Jansky, Jr., I. J. Kaar, F. B. Llewellyn, Haraden Pratt, B. J. Thompson, H. M. Turner and H. P. Westman, secretary.

Thirty-eight applications for Associate, one for Junior, and two for Student were approved.

The Tellers Committee report was accepted and L. C. F. Horle was declared elected President for 1940; F. E. Terman, Vice President for 1940; and Austin Bailey, H. M. Turner, and H. A. Wheeler directors to serve during 1940–1942.

An invitation from the Boston Section for the holding of our Fifteenth Annual Convention in that city was accepted. The convention will be held on June 27, 28, and 29 with headquarters at the Statler Hotel. W. L. Barrow, Chairman of the Section, was appointed Chairman of the Convention Committee.

European Journals and the War

The nonreceipt by a subscriber of any European scientific journal seriously needed as research material should be promptly reported to the American Documentation Institute.

The Cultural Relations Committee of the American Documentation Institute, which co-operates closely with the cultural Relations Division of the Department of State, is working on this problem, and hopes to be able to surmount such war obstacles as interrupted transportation, embargoes, and censorship, which so grievously affected the progress of research during the last war.

The principle should be established, if possible, that the materials of research having no relation to war shall continue to pass freely, regardless of the countries of origin or destination.

Reports, with full details of where a subscription was placed and the name and address of the subscriber, volume, date, and number of last issue received, should be addressed to:

American Documentation Institute
Bibliofilm Service
c/o U. S. Department of Agriculture Library
Washington, D. C.

Committees

Board of Editors Co-ordinating Committee

The Co-ordinating Committee of the Board of Editors met in the Institute office on October 18 and reviewed a number of manuscripts that were submitted for publication. Those present were Alfred N. Goldsmith, chairman; R. R. Batcher, L. E. Whittemore, and H. P. Westman, secretary.

Tellers Committee

The Tellers Committee, consisting of W. M. Goodall, chairman; H. F. Dart,

PEDER OLUF PEDERSON
Vice President, 1939

Peder Oluf Pedersen was born in Sig, Denmark, on June 19, 1874. The King of Denmark became interested in his education and made it possible for him to secure the necessary preparatory education and to enter the Royal Technical College, where he was graduated with honor in civil engineering in 1897.

Soon after his graduation he became interested in electrical research work and in 1899 became associated with Valdemar Poulsen in his development work on the telegraphone. He later aided in the development of the Poulsen arc system for continuous-wave wireless telegraphy and telephony.

In 1909 he was appointed Assistant Professor in Telegraphy, Telephony, and Radio at the Royal College at Copenhagen, becoming Professor in 1912, which chair he still holds. In 1922 he was appointed principal of that college. He received a Ph.D. degree from the University of Copenhagen in 1929.

Doctor Pedersen has contributed a great number of papers on scientific matters in electrophysics and electrotechnics, mainly on experimental researches carried out by himself.

In 1907 he received the Gold Medal of the Royal Danish Society of Sciences and in 1927 he was awarded the C. H. Oersted Medal. The Institute Medal of Honor was presented to him in 1930.

He is a member of numerous technical societies and has served several as president. Professor Pedersen has been a Fellow of the Institute since 1915, and its Vice President during 1939.

D. G. Fink, and H. P. Westman, secretary, met in the Institute office on October 27 and counted the ballots cast in the election of officers.

Electronics Conference

The Electronics Conference Committee met on October 17 and on November 3. Both meetings were attended by F. R. Lack, chairman; F. B. Llewellyn, R. W. Sears, B. J. Thompson, and H. P. Westman, secretary.

The earlier meeting was to complete the plans for the Conference which was held on October 20 and 21 and the later meeting was devoted to an analysis of the Conference.

Subcommittee on Tube Noise

On September 22, there was held a meeting of the Subcommittee on Tube Noise of the Electronics Conference Committee to prepare the program for that part of the Conference for which the subcommittee was responsible. B. J. Thompson, chairman and acting secretary; R. L. Freeman, F. B. Llewellyn, J. R. Nelson, and D. O. North were present.

Electronics

P. T. Weeks, chairman and acting secretary; R. S. Burnap, E. L. Chaffee, K. C. De Walt, Ben Kievit, Jr., F. R. Lack, F. B. Llewellyn, and J. R. Wilson attended a meeting of the Electronics Committee held at the Hotel Pennsylvania on September 22.

Reports of the several subcommittees on their standardization, annual review, and other activities were heard and discussed.

Subcommittee on Ultra-High Frequencies

F. B. Llewellyn, chairman; R. L. Freeman, L. S. Nergaard, A. L. Samuel, and H. P. Westman, secretary; attended a meeting of the Subcommittee on Ultra-High Frequencies of the Electronics Committee in the Institute office on November 6. The meeting was devoted chiefly to a discussion of specific arrangements for the preparation of the annual review. Standardization matters were also considered.

Television

The Technical Committee on Television met in the Institute office on October 10 to prepare for the writing of a standards report and to arrange for the preparation of an annual review of its field. Thos present were: E. K. Cohan, chairman; R. R. Batcher, R. B. Dome (representing I. J. Kaar), A. B. DuMont, E. W. Engstrom, D. E. Foster, P. C. Goldmark, T. T. Goldsmith, Jr., A. G. Jensen, George Lewis, H. M. Lewis, R. E. Shelby, and H. P. Westman, secretary.

Sections

Chicago

Harner Selvidge, assistant professor of electrical engineering at Kansas State College, presented a paper on "Television Cable and Transmission-Line Problems."

Attenuation characteristics of various types of transmission lines such as twisted pair, shielded ignition cable, shielded rubber-covered wire, and coaxial cables using various ceramics for insulation were discussed. Their use for ultra-high-frequency transmission was treated. The paper was closed with a short discussion of phase shift in these circuits.

September 29, 1939, V. J. Andrew, chairman, presiding.

Cincinnati

"Automotive Receiver Development and Design" was the subject of a paper by Roger Daugherty, automotive radio engineer for the Crosley Corporation.

A short history of the field was first presented. The first receivers were converted household sets but were not widely used. The introduction of the vibrator-type power supply gave great impetus to the field and the automobile manufacturers then became interested in it.

Statistics presented indicated the typical automobile set to be a six-tube superheterodyne using both metal and glass tubes. An intermediate frequency of 455 kilocycles is used. A preselector stage is included. The audio-frequency power output is about 4.7 watts and is fed to a six-inch loud speaker. The set requires about 6.5 amperes from the automobile battery. Push-button tuning is used. The cost varies between $20.00 and $70.00.

The automobile set must be from two to three times as sensitive as a household set. The long roof-type antenna is most effective, the under-car type next in performance, and the vertical type gives the least satisfactory results.

Car receivers must operate with battery voltages which vary from 5 to 9 volts and over a temperature range from -10 to $+170$ degrees Fahrenheit. Frequency drift caused by temperature change can be corrected by using compensating condensers. Humidity effects may be avoided by mounting the oscillator coil in an evacuated tube.

At low car speeds, about one watt of audio-frequency output is satisfactory but between 5 and 10 watts are required at 60 miles per hour. A 5000-cycle upper limit is satisfactory. An increase in output at about 70 cycles gives a pleasing effect.

Push-button tuning systems of both the padder-switching and main-condenser-tuning types were next covered. The operation of mechanical and solenoid systems was described and models of both types demonstrated.

The paper was closed with a description of the design of an experimental multiband automobile receiver. The performance characteristics of the set operating on frequencies up to 18 megacycles were described.

October 24, 1939, H. J. Tyzzer, chairman, presiding.

Cleveland

The "Communication System of the Cleveland Police," was the subject of a paper by L. N. Chatterton, radio engineer for the Department of Public Safety of the City of Cleveland.

A brief history of police communication methods was first presented.

Public-address systems are located at precincts, police bureaus, and railroad stations. A private teletype system connects precincts and bureaus at twelve locations. In addition, the regular Bell System teletype gives nation-wide contact.

In Cuyahoga County there are over 300 radio-equipped police cars, 93 of which can handle two-way traffic. Thirty-two of these cars are basic patrol units on duty 24 hours a day. Transmission to the cars is under the control of dispatchers who are constantly informed of their location. Although the cars are generally restricted to definite zones, 30 radio-equipped motorcycles operate generally.

Three channels in the 30-megacycle region are used for transmission from the cars. Three receivers at one precinct station deliver the signals over a telephone line to the dispatchers in the central police station.

A 10-channel 500-watt Bendix transmitter and a 100-watt 33.5-megacycle transmitter are located at the same precinct house in which the receivers are installed. The ultra-high-frequency transmitter is used for communication with the cars and is operated remotely from the central police station. The 10-channel transmitter provides a 2458-kilocycle channel to cars and three sets of three frequencies each for telegraph communication with the Ohio State Patrol and 12 city and state police organizations outside of Ohio. Reply telegraphic signals are picked up by a multiwave receiver at the central police station.

At the central police station transmitters operating at 33.1 megacycles and at 2458 kilocycles are operated.

Between 1000 and 1500 calls are made daily and on the average less than three minutes elapses between the receipt of a complaint and the arrival of a car on the scene. In the case of major crimes, the car arrives usually within a minute. In some cases, the complainant is connected directly with the squad car which will answer his call.

A mobile transmitter and loud speaker connected to the telegraph facilities and the microphone at the dispatcher's desk were set up for demonstration purposes. Following the paper, an inspection of the regular communication equipment in operation was made.

September 28, 1939, S. E. Leonard, chairman, presiding.

Emporium

Four papers were presented at the third Annual Summer Seminar.

"Methods and Apparatus for Measuring Phase Distortion in Television" was presented by C. E. Brigham, technical director of Kolster Brandes, Ltd. (England). This paper was written by M. Levy of the Paris Laboratories of Le Materiel Telephonique.

After indicating numerous methods of measuring phase distortion, the paper discussed one providing quick and accurate results when dealing with 4-terminal networks. The apparatus could also trace on the screen of a cathode-ray tube a

Nyquist diagram for the network being measured.

I. R. Weir of the engineering department of the General Electric Company (Schenectady), discussed "Frequency Modulation." This paper was summarized in the March, 1939, PROCEEDINGS in the report on the Connecticut Valley Section.

"Comments on European Radio Developments" was presented by R. M. Wise, chief radio engineer of the Hygrade Sylvania Corporation. His views were the result of a recent European trip and he pointed out particularly the interest being shown in tubes similar to the loctal type.

M. A. Acheson of the engineering department of the Hygrade Sylvania Corporation, discussed the "Ratings and Characteristics of 1.4-Volt Tube Types." Considerable attention was given to the minimum allowable power output and the desirability of using lower plate voltages on these tubes.

A picinic concluded the meeting.

July 28–29, 1939, R. K. McClintock, chairman, presiding.

D. G. Fink, managing editor of *Electronics*, presented a paper on "Recent Progress in Television Technique." The paper covered only the 441-line transmissions. The three main characteristics of a television image, detail, brightness, and contrast, were defined and discussed. A description of some of the latest television equipment concluded the paper. In addition, the speaker presented some brief comments on his experience with frequency-modulated-wave receivers.

October 12, 1939, R. K. McClintock, chairman, presiding.

Los Angeles

A "Symposium on Frequency Modulation" resulted in the fundamentals being discussed by B. M. Oliver of the California Institute of Technology, the latest technique in transmitter design being treated by G. W. Downs, Jr., of the William Miller Corporation, and Edward Simmons of the California Institute of Technology presenting material on receivers with emphasis on the fundamental differences between those for amplitude-modulated-wave signals and frequency-modulated-wave signals. Additional comments were contributed by Frank Kennedy of the Don Lee Broadcasting System, J. N. A. Hawkins of the Walt Disney Studios, and Samuel Waite of the Yankee Network (Boston).

The consensus of opinion of those present seemed to be that frequency modulation has certain advantages over amplitude modulation.

September 19, 1939, F. G. Albin, chairman, presiding.

Philadelphia

"Modern Microphones" was the subject of a paper by H. F. Olson, director of acoustical research of the RCA Manufacturing Company (Camden).

Dr. Olson described methods of securing undistorted sound pickup under various conditions met in practice.

Descriptions were then presented of ultradirectional, unidirectional, and bidirectional velocity microphones and their use in discriminating against reflected or otherwise undesired noise.

While sound waves'can be focused by a parabolic reflector, waves of different frequencies do not focus at the same point and it is impracticable to focus the lower frequencies. Other methods for obtaining directional pickup than those depending on focusing were then described. One involves the combining of the response characteristics of velocity and pressure microphones. Another method depending on the phase relation of high- and low-frequency waves involves the transmission of the waves to the microphones through pipes of different lengths.

The author described the laws governing sound pickup with different forms of microphones and how they are made to reject undesired sounds.

October 5, 1939, R. S. Hayes, chairman, presiding.

Pittsburgh

Robert Shelby, television engineer of the National Broadcasting Company, presented a "Demonstration of Television Equipment."

This paper covered many phases of television and described equipment which was used to demonstrate the transmission and reception of images. The demonstration equipment was also made available for inspection by the audience.

The meeting was held jointly with the Physical Society of Pittsburgh and the Engineering Society of Western Pennsylvania.

October 17, 1939, Joseph Baudino, chairman, presiding.

Portland

"A New Development for Measuring Impedance at Radio Frequencies" was presented by M. T. Smith of the General Radio Company.

September 29, 1939, H. C. Singleton, chairman, presiding.

San Francisco

"The Voder, Voice Mirror, and Auditory Test at the Bell Exhibit, Treasure Island" was the subject of a paper by Julian Edwards, an engineer for the Pacific Telephone and Telegraph Company. It was devoted to the devices installed at the Bell exhibit at the Golden Gate International Exposition.

The Voder makes synthetic speech from two types of sound, one a buzzer-like tone and the other a hiss corresponding, respectively, to the vocal-cord tone and the breath tones of normal speech. Properly controlled variations in duration and intensity of these two sound streams produce intelligible artificial speech.

The voice mirror utilizes a magnetic tape recorder which permits one to hear the sound of his own voice over a telephone handset.

The paper was concluded with a demonstration of the Voder over a special wire circuit from the Exposition.

October 18, 1939, H. E. Held, chairman, Papers Committee, presiding.

Seattle

A. V. Eastman of the department of electrical engineering of the University of Washington, presented a paper on a "Study of Cross Modulation in the City of Seattle."

Professor Eastman first reviewed briefly the theory of cross modulation, showing that the impression of two or more voltages on a nonlinear circuit in which $I=K+AE+BE^2+CE^3+\cdots$ will produce not only the familiar double, sum, and difference frequencies because of the second-order term of the series, but will produce another set of cross-modulation frequencies for each of the higher-order terms present. It was shown that there were nine cross-modulation frequencies of the third-order term ranging from 570 to 1620 kilocycles produced in Seattle by Stations KOMO, KJR, and KOL. Two other stations in Seattle also combine to produce cross-modulation frequencies in the broadcast band. Since KOMO and KJR use the same antenna which, in turn, is close to the antenna of KOL, and because both antennas are within the city, the magnitude of these interfering third-order frequencies has in many actual cases risen to a high value.

The author then described a series of experiments which L. C. F. Horle and he conducted to determine the causes and magnitudes of the cross-modulation signals. Some of the findings of these tests made both in the city and at Puget Sound were: (1) the use of a single antenna by KOMO and KJR did not in itself cause cross modulation since comparable effects were produced by the proximity of the antenna of KOL; (2) Nonlinear re-radiating structures such as electric power wiring were the chief cause of the trouble; (3) many radio receivers, including some of recent manufacture, were found to produce cross modulation; and (4) even though tests were made away from shore by a receiver causing no cross modulation, weak cross-modulation signals were picked up and were probably caused by the antennas.

As a conclusion, Professor Eastman suggested the desirability of spreading farther apart the stations serving a community or, if they must be closely adjacent, they should be remotely located from the community they serve. Broadcast-receiver design and manufacture should include greater consideration of the problem of cross modulation.

October 13, 1939, R. O. Bach, chairman, presiding.

Toronto

"Fabricated-Plate Capacitors" was the subject of a paper by B. V. K. French of the engineering department of the P. R. Mallory Company.

The original electrolytic condensers used plates of pure aluminum foil on which an aluminum-oxide coating was formed electrolytically. The electrolyte acts as one plate of the condenser, the aluminum plate as the other, and the aluminum oxide as the dielectric.

Etching of the aluminum foil before

forming the oxide increased the capacitance by two or three times. Even though the etching was done with a high degree of uniformity, the high current density necessary for the formation of the aluminum oxide resulted in a leveling off of the etching and a smaller capacitance.

Deep etching made possible a capacitance increase of about five times over that obtained without etching but was also subject to the same difficulties as ordinary etching

If a fabric is sprayed with a zinc spray gun and the plates formed with their coating of aluminum oxide before being placed in the complete assembly, the burning at the edge of the electrolyte may be avoided since the high current densities do not occur as a result of preforming.

The necessity of specifying ripple current or voltage which will be applied to the condenser as well as the direct voltage was pointed out. A condenser connected at the output of a filter circuit need not be as adequately protected as if it were at the input where the amount of ripple is so much greater.

The standardization of electrolytic condenser systems was then discussed. It was pointed out that formerly receiver manufacturers demanded that the condensers be made of almost any shape that would happen to fit some open space in the radio set. This resulted in a large number of designs and uneconomical manufacture. By reducing the number of sizes of cans to a minimum, the price of these condensers has been greatly reduced.

The paper was closed with a discussion of the special problem of the use of electrolytic condensers in voltage-doubler circuits.

October 16, 1939, G. J. Irwin, chairman, presiding.

Washington

I. F. Byrnes, chief engineer of the Radiomarine Corporation of America, presented a paper on the "Development and Design of Auto-Alarm Equipment for Shipboard Service."

Introductory remarks were made by E. M. Webster, of the Federal Communications Commission staff, regarding the circumstances leading to the development of auto-alarm equipment in this country. The apparatus used on ships of other nations and the features which the Federal Communications Commission felt should be incorporated in equipment for American ships were outlined

Mr. Byrnes traced the development of the present equipment from its original conception to the present commercial product. Circuits employed and the operating procedure were described in detail. A complete auto-alarm equipment was available for inspection and was demonstrated to show the operation of the equipment under various conditions. Devices to minimize the possibility of false alarms and those giving visual and aural indications of power-supply failure were also described and demonstrated.

October 9, 1939, Gerald C. Gross, chairman, presiding.

Personal Mention

The following members have recently informed us of changes in their company affiliations or titles to those given below.

Gilman, George W.; American Telephone and Telegraph Company, 195 Broadway, New York, N. Y.

Kahn, Louis; Engineer, Aerovox Corporation, New Bedford, Mass.

Overbeck, Wilcox P.; Research Associate, Massachusetts Institute of Technology, Cambridge, Mass.

Robinson, E. B.; Lieutenant, U.S.N.; U.S.S. *Enterprise*, San Francisco, Calif.

Sherman, Warren K.; Lieutenant Commander, U.S.N.; c/o U. S. Navy Purchasing Office, Shanghai, China.

Snyder, Graves H.; Lieutenant, France Field, Panama Canal Zone.

Membership

The following indicated admissions to membership have been approved by the Admissions Committee. Objections to any of these should reach the Institute office by not later than December 30, 1939.

Admissions to Associate (A) and Student (S)

Abrahams, I. C., (S) 127 Fay Ave., Lynn, Mass.
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Cafferata, H., (A) "Knotty Ash," Greenways, Bloomfield Rd., Chelmsford, Essex, England.
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Incorrect Addresses

Listed below are the names and last-known addresses of fifty-six members of the Institute whose correct addresses are unknown. It will be appreciated if anyone having information concerning the present addresses of the persons listed will communicate with the Secretary of the Institute.

Adams, James J., 5656 W. Race Ave., Chicago, Ill.
Adams, Ralph E., Apt. 204, 1029 Second St., Santa Monica, Calif.
Adams, Robert L., 598 Williams St., Atlanta, Ga.
Asthana, R. P., Resident Engineer, Power House, Ujjaim, India.
Aylor, Raymond P., Jr., Hampton Rd., Broadcasting Corp., Newport News, Va.
Bergstrom, Raymond, 1332 Termaine Ave., Los Angeles, Calif.
Blasier, Herbert E., 2802 West Ave. 32, Los Angeles, Calif.
Booker, Eugene R., Box 531, Route 1, San Jose, Calif.
Brohl, Earl M., E. Falls Church, Va.
Chittick, K. A., 120 Wayne Ave., Haddonfield, N. J.
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Congdon, Carl L., U. S. Naval Radio, Cavite, Philippine Islands
Daniels, Thomas E., KUT Broadcasting Co., Norwood Bldg., Austin, Texas.
Darwin, Fred A., 5514 Blackstone Ave., Chicago, Ill.
Decker, Joseph L., 20 Austin Ave., Albany, N. Y.
De Forest, M. J., 4524 Wrightwood Ave., Chicago, Ill.
Eddy, Myron Fish, Stewart Tech., 253 W. 64th St., New York, N. Y.
Engel, Francis H., RCA Manufacturing Co., Radiotron Div., Harrison, N. J.
Erlandson, Paul M., 104 Mt. Pleasant Ave., Wyoming, Ohio.
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Gerstle, John, 235 Palm Dr., Oakland, Calif.
Grenly, Mark, 113, Brim Hill, Hampstead Garden Suburb, N.2., London, England.
Halligan, Clair, W., 150 Waverly Place, New York, N. Y.
Hiehle, Ernest M., 2527 Hope St., Walnut Park, Calif.
Hilgedick, W. C., 250 Federal Office Bldg., National Park Service, San Francisco, Calif.
Hopkins, Nelson S., Phenix Aircraft Products Co., 5565 Main St., Williamsville, N. Y.
Howe, Roger M., 71 Fulton St., Medford, Mass.
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Miller, Jack A., 128 E. 5th St., Emporium, Pa.
Murray, Donald A., 246 U. S. Court House, Chicago, Ill.
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Phillips, Frank Arthur, c/o G. P. O., Sydney, N.S.W., Australia.
Rahn, Ernest, Frohnerstr. 7, Berlin, Germany.
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THE INSTITUTE OF RADIO ENGINEERS

(Incorporated, May 13, 1912)

Constitution

Adopted at the First Meeting of the Institute of Radio Engineers May 13, 1912. Amended, November 2, 1914; December 5, 1915; October 7, 1931; and March 1, 1939

ARTICLE I

NAME AND OBJECT

SEC. 1—The name of this organization shall be The Institute of Radio Engineers, Incorporated.

SEC. 2—Its objects shall be the advancement of the theory and practice of radio and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members. Among the means to this end shall be the holding of meetings for the reading and discussion of professional papers and the publication of papers, discussions, communications, and such other matters as may be appropriate for the fulfillment of its objects.

ARTICLE II

MEMBERSHIP

SEC. 1—The membership of the Institute shall consist of:

a. Fellows, who shall be entitled to all rights and privileges of the Institute.

b. Members, who shall be entitled to all rights and privileges of the Institute except the right to hold the offices of President and Vice President.

c. Associates, who shall be entitled to attend all meetings and to receive copies of all Institute publications. In addition, Associates of record at the time of adoption of this Constitution shall be voting members so long as a continuous membership is maintained.

d. Juniors, who shall be entitled to attend all meetings and to receive copies of all Institute publications.

e. Students, who shall be entitled to attend all meetings and to receive copies of all Institute publications.

SEC. 2—Fellow: For admission or transfer to the grade of Fellow, a candidate shall be at least thirty-two years of age and shall be either:

a. A radio engineer or radio scientist. As such he shall have attained distinction in his profession and shall be eminently qualified to take responsible charge of important radio work. He shall have been in the active practice of his profession for at least ten years, and shall have had responsible charge of important radio work for at least three years.

When the candidate holds, in a principal national society of an allied branch of engineering or science, membership in a grade for which the qualifications indicate a standing equal to that required for the grade of Fellow herein, such membership shall be considered equivalent to three years of the required ten years of active practice of the radio profession.

b. A professor of electrical engineering or of physical science. As such he shall have attained special distinction as an expounder of the principles of radio engineering or of radio science. He shall have had at least ten years experience as a teacher of electrical or physical subjects and shall have had responsible charge, for three years, of a radio course in a school of recognized standing.

c. A person who has done notable original work contributing to the advancement of radio engineering which has given him a recognized standing at least equivalent to that required for Fellow under paragraph "*a*" or "*b*."

d. A person regularly engaged in radio work for at least ten years, who, by invention or by contributions to the advancement of radio engineering or radio science, or to technical radio literature has attained a standing at least equivalent to that required for Fellow under paragraph "*a*" or "*b*."

SEC. 3—Member: For admission or transfer to the grade of Member, a candidate shall be at least twenty-six years of age and shall be either:

a. A radio engineer or radio scientist. As such he shall have performed and taken responsibility for important radio engineering or scientific work and shall have been in the active practice of his profession for at least four years.

b. A teacher of radio or closely allied subjects for at least four years in a school of recognized standing.

c. A person regularly employed in radio or closely allied work for at least four years, who by invention or by contributions to the advancement of radio engineering or radio science, or to technical radio literature, has attained a standing equivalent to that required for Member under paragraph "*a*."

d. An executive of a radio enterprise who, for at least six years, has had under his direction, important radio engineering or research work and who is qualified for direct supervision of the technical or scientific features of such activities.

SEC. 4—Associate: For admission or transfer to the grade of Associate, a candidate shall be at least twenty-one years of age and shall be interested in the theory or practice of radio communication or of the closely related arts and sciences.

SEC. 5—Junior: For admission to the grade of Junior, a candidate shall be at least eighteen and not more than twenty-one years of age and shall be interested in the theory or practice of radio communication or of the closely related arts and sciences.

A Junior shall be transferred to the Associate grade on reaching the age of twenty-one years.

SEC. 6—Student: For admission to the grade of Student, a candidate shall be devoting a major proportion of his time as a registered student in a regular course of study in engineering or science in a school of recognized standing. Membership in this grade shall not extend more than one and one-half years beyond the termination of his student status described above.

SEC. 7—The expression "school of recognized standing" is interpreted as applying to schools of college grade providing an engineering or scientific curriculum of not less than four years and granting degrees.

SEC. 8—In all cases, graduation from a radio or electrical course of a school of recognized standing shall be accepted in lieu of one year's radio experience

SEC. 9—The time requirements for admission to any grade of membership may be satisfied by applying *pro rata* the experience of the candidate under the various alternative requirements for that grade.

SEC. 10—The terms "member" and "membership" when printed without an initial capital where used in this Constitution and By-Laws includes all grades.

SEC. 11—The term "voting member" where used in this Constitution and Bylaws means a member entitled to vote on Institute matters.

ARTICLE III

ADMISSIONS, TRANSFERS, AND EXPULSIONS

SEC. 1—Admission or transfer to Fellow grade shall be by invitation by the Board of Directors only.

SEC. 2—Applications for admission or transfer to any grade of membership, except Fellow, shall be submitted to the Board of Directors. An affirmative vote of at least two thirds of the Board members voting shall elect or transfer an applicant to any grade.

SEC. 3—A reapplication for admission or transfer may be made after the expiration of one year from the date of a rejection.

SEC. 4—The admission fee and dues are payable on notification of election and if not received within six months from notification, the election shall be considered void.

SEC. 5—A member in good standing may resign by submitting a written resignation to the Secretary.

SEC. 6—Subject to the approval of the Board of Directors, a resigned member may resume his membership upon payment of current dues.

SEC. 7—When a member's dues become three months in arrears his membership shall be considered terminated. Subject to the approval of the Board of Directors, such membership may be resumed on payment of a new entrance fee and current dues or by the payment of all dues in arrears.

SEC. 8—To initiate action toward expulsion of a member, a written complaint must be submitted to the Board of Directors, which if it deems the reason sufficient, shall notify the accused by letter of the charges against him and of the place and date of the hearing, which shall be at least twenty days away. The accused may present his defense in person, in writing, or by an authorized representative. There shall be a majority of the members of the Board of Directors present at the hearing and the votes cast must be unanimous in order to expel. The action of the Board of Directors shall be final and conclusive.

ARTICLE IV

ENTRANCE FEES AND DUES

SEC. 1—The entrance fees, transfer fees, and annual dues shall be as follows:

Entrance Fees

Fellow	$10.00
Member	5.00
Associate	3.00
Junior	1.00
Student	

The transfer fee from one grade of membership to another shall be the difference between the corresponding entrance fees except that there shall be no fee when transferring immediately from Student to Associate membership.

Annual Dues

Fellows	$10.00
Members	10.00
Associates	6.00
Juniors	4.00
Students	3.00

SEC. 2—The annual dues shall be payable in advance on the first day of January.

SEC. 3—Under exceptional circumstances, the payment of fees and dues may be deferred or waived in whole or in part by the Board of Directors.

ARTICLE V

OFFICERS

SEC. 1—The governing body of the Institute shall be the Board of Directors and shall consist of the President, Vice President, Secretary, Treasurer, Chairman of the Board of Editors, nine elected Directors, five appointed Directors, and the two most recent past Presidents.

SEC. 2—Except for the elected Directors, the terms of all officers shall be for one year each.

SEC. 3—The terms of the elected Directors shall be for three years each.

SEC. 4—The terms of the appointed Directors shall be for the current calendar year.

SEC. 5—No officer shall receive, directly or indirectly, any salary, compensation, or emolument from the Institute, either as such officer, or in any other capacity, unless authorized by a vote of a majority of the entire Board of Directors, except as authorized by the Bylaws.

ARTICLE VI

MANAGEMENT

SEC. 1—The President shall be the regular presiding officer at meetings of the Board of Directors and at meetings of the Institute. He shall be an *ex officio* member of each committee.

The Vice President shall assume the duties of the President in the absence or incapacity of the President.

In the event that neither the President nor the Vice President can personally act, the Board of Directors may elect a chairman from its membership who is authorized to perform the presidential duties during the period of the incapacity of the President and Vice President. The tenure of such temporary chairman shall be at the discretion of the Board of Directors.

SEC. 2—The Board of Directors shall manage the affairs of the Institute. An annual report shall be made to the members on the activities and finances of the Institute.

Six members of the Board of Directors shall constitute a quorum.

SEC. 3—The Board of Directors may make, amend, or revoke Bylaws to this Constitution. The proposed changes and reasons therefore shall be mailed to all members of the Board at least twenty days before the stipulated meeting at which the vote shall be taken. Two thirds of all votes received at the stipulated meeting shall be required to approve any new Bylaw, amendment, or revocation.

SEC. 4—The Treasurer, under the control of the Board of Directors, shall have general supervision of the fiscal affairs of the Institute.

The Institute shall secure a surety bond on the treasurer.

SEC. 5—The Secretary shall attend all meetings of the Board of Directors and principal meetings of the Institute and prepare the business and record the proceedings thereof. He shall have charge of the books of account of the Institute, and shall furnish from them such information as is requested by the Board of Directors. He shall conduct the correspondence of the Institute and keep full records thereof.

The Institute shall secure a surety bond on the Secretary. An annual audit of the affairs of the Institute shall be made by certified public accountants and submitted to the Board.

SEC. 6—All funds received by the Institute shall be deposited in an account requiring the signatures of at least two of the following for withdrawal: President, Vice President, Treasurer, Secretary, and Chairman of the Board of Editors. Funds from this account shall, in general, be deposited in a second account which shall never exceed an amount specified by the Board of Directors and shall be withdrawable on the signature of the Secretary alone for current disbursements. Before funds are transferred from the first-mentioned account to the other, the Secretary shall submit a statement of the disposition of the previously expended funds to the Treasurer.

SEC. 7—All standing committees shall be appointed by the incoming President with the consent of the Board of Directors, at the annual meeting of the Institute. Additional committees may be established by the Board of Directors.

SEC. 8—The fiscal year of the Institute shall end with the thirty-first day of December.

ARTICLE VII

NOMINATION AND ELECTION OF PRESIDENT, VICE PRESIDENT, AND THREE DIRECTORS, AND APPOINTMENT OF SECRETARY, TREASURER, CHAIRMAN OF THE BOARD OF EDITORS, AND FIVE DIRECTORS

SEC. 1—On or before July first of each year, the Board of Directors shall submit to qualified voters a list of nominations containing at least one name each for the office of President and Vice President and at least six names for the office of elected Director and shall call for nominations by petition.

Nominations by petition may be made by letter to the Board of Directors setting forth the name of the proposed candidate and the office for which it is desired he be nominated. For acceptance a letter of petition must reach the executive office before August fifteenth of any year and shall be signed by at least thirty-five voting members.

Each proposed nominee shall be consulted and if he so requests his name shall be withdrawn. The names of proposed nominees who are not eligible under the Constitution shall be withdrawn by the Board.

On or before September first, the Board of Directors shall submit to the voting members as of August fifteenth, a list of nominees for the offices of President, Vice President, and elected Director, the names of the nominees for each office being arranged in alphabetical order. The ballots shall carry a statement to the effect that the order of the names is alphabetical for convenience only and indicates no preference.

Voting members shall vote for the candidates whose names appear on the list of nominees, by written ballots in plain sealed envelopes, enclosed within mailing envelopes marked "Ballot" and bearing the member's written signature. No ballots within unsigned outer envelopes shall be counted. No votes by proxy shall be counted. Only ballots arriving at the executive office prior to October twenty-fifth shall be counted. Ballots shall be checked, opened, and counted under the supervision of the Tellers Committee between October twenty-fifth and the first Wednesday in November. The result of the count shall be reported to the Board of Directors at its first meeting in November and the nominees for President and Vice President and the three nominees for Director receiving the greatest number of votes shall be declared elected. In the event of a tie vote the Board shall choose between the nominees involved.

SEC. 2—The Secretary, Treasurer, and Chairman of the Board of Editors, shall be appointed by the Board of Directors at its annual meeting to serve until the next annual meeting.

SEC. 3—The Board of Directors is authorized to fill a vacancy occurring in the governing body.

ARTICLE VIII

MEETINGS

SEC. 1—There shall be an annual meeting of the Board of Directors during January of each year at which newly elected officers shall begin their terms of service, and the Board shall make necessary appointments. There shall be a meeting of the Board of Directors in November on or after the first Wednesday to receive the report of the Tellers Committee.

SEC. 2—There shall be an annual meeting of the Institute as soon as practicable after the annual meeting of the Board of Directors at which general reports of the Secretary and Treasurer shall be presented.

SEC. 3—Meetings of the Board may be held at such times as are necessary to carry out the provision of this Constitution and shall be held at such other times as any five members of the Board may determine, but only on notice to all members of the Board.

ARTICLE IX

INSTITUTE SECTIONS

SEC. 1—Sections of the Institute may be authorized by the Board of Directors.

SEC. 2—The Board of Directors may at any time terminate the existence of any section when in its judgment the interests of the Institute makes such action desirable.

ARTICLE X

AMENDMENTS

SEC. 1—Amendments to this Constitution may be proposed by means of a resolution adopted by the Board of Directors or by means of a petition signed by at least thirty-five voting members. Such proposed amendment or amendments shall be submitted to legal counsel by the Board of Directors, and, if in the opinion of such counsel, they are in accordance with the laws under which the Institute is organized, a copy shall be mailed with a letter ballot to each member.

SEC. 2—Constitutional amendment ballots shall be mailed to the voting members at least sixty days before the date appointed for counting the ballots and the ballots shall carry a statement of the time limit for their return to the executive office. The Tellers Committee shall count such votes and report to the Board of Directors at its next meeting. If the total vote be at least twenty per cent of the total voting membership and if at least seventy-five per cent of all votes cast shall be favorable, the proposed amendment or amendments shall become part of this Constitution.

SEC. 3—Amendments shall take effect thirty days after their adoption, but officers and officers-elect of the Institute at the time any amendment becomes effective shall continue in office until the end of the terms for which they were elected.

SEC. 4—Copies of the amendments shall be distributed to the members as soon as practicable after adoption.

SEC. 5—A complete history of amendments shall be kept in the files of the Institute.

Bylaws

Article VI, Section 3, of the Institute Constitution provides for Bylaws as follows:

"The Board of Directors may make, amend, or revoke Bylaws to this Constitution. The proposed changes and reason therefore shall be mailed to all members of the Board at least twenty days before the stipulated meeting at which the vote shall be taken. Two thirds of all votes received at the stipulated meeting shall be required to approve any new Bylaw, amendment, or revocation."

MEMBERSHIP

Sec. 1—Institute members are authorized to use the following abbreviations or symbols indicating their grade of membership:

Fellow—F.I.R.E.
Member—M.I.R.E.
Associate—A.I.R.E.

Sec. 2—The emblem of the Institute is copyrighted and shall be reproduced only in connection with official business of the Institute.

Sec. 3—Applicants for membership shall furnish names of sponsors as follows:

For Member—five Fellows or Members.

For Associate—three Fellows, Members, Associates, or other responsible individuals.

For Junior—three Fellows, Members, Associates, or other responsible individuals.

For Student—a member of the faculty of his school.

Sec. 4—When the work or location of an applicant for Member grade is such as to make impracticable compliance with Section 3, the Admissions Committee may waive that section upon obtaining other suitable references.

Sec. 5—The names of applicants for admission to the Institute, after approval by the Admissions Committee, shall be posted in the PROCEEDINGS.

Sec. 6—Objection to the admission of a candidate must include reasons for such objection and must reach the office of the

Institute by the first day of the month following posting in the PROCEEDINGS. All such statements shall be treated as confidential.

Sec. 7—Transfer of an Associate to Member grade may be proposed by any member acting as sponsor, in which case the sponsor shall fill in the application blank and provide letters of reference for submission to the Admissions Committee. If the application is favorably acted on, the sponsor shall secure the candidate's signature to a duplicate application blank after which the application shall be submitted to the Board of Directors.

Sec. 8—The Membership Committee may recommend for transfer to higher grade those members who they think are qualified.

Sec. 9—Each year, the Awards Committee shall recommend to the Board of Directors nominees for Fellow grade. A citation summarizing the accomplishments of the nominee shall be a part of each recommendation.

Sec. 10—Diplomas shall be presented to the newly elected Fellows. If practicable, this presentation shall be made by the President at the next Annual Convention.

Sec. 11—A member whose dues are more than two months in arrears shall be notified by the Secretary and informed that, in accordance with Article III, Section 7, of the Constitution, should his dues become three months in arrears, he loses the right to vote or to receive the publications of the Institute.

Sec. 12—The mailing of bills or statements to the last known address of a member shall be considered a valid notice of indebtedness.

Sec. 13—On resuming membership and paying dues in arrears, a member may receive available copies of the PROCEEDINGS during the period covered by the back dues. A rebate of 25 cents per copy will be made in lieu of copies of the PROCEEDINGS not available.

BOARD OF DIRECTORS

Sec. 14—Unless otherwise set, meetings of the Board of Directors are held on the first Wednesday of each month except in July and August in the office of the Institute in New York, N. Y. Minutes of all meetings of the Board of Directors shall be sent to each member of the Board of Directors.

Sec. 15—The Secretary, appointed as prescribed in Article VII, Section 2, of the Constitution, shall be paid a salary determined by the Board of Directors.

SECTIONS

Sec. 16—A petition for the formation of a Section shall be signed by not fewer than twenty-five (25) Fellows, Members, and Associates residing within the proposed territorial limits.

Sec. 17—The territory of a Section shall be specified by the Board of Directors.

Sec. 18—All Sections shall accept and conform to a "Constitution for Sections" provided by the Institute Board of Directors.

Sec. 19—For Section maintenance, fifty cents shall be paid by the Institute to each Section for each Fellow, Member, and Associate residing within the territory of the Section at the end of the fiscal year, namely, December 31, plus ten ($10.00) dollars for each meeting up to and including the tenth meeting held during the year.

Sec. 20—Sections shall have no authority to contract debts for, pledge the credit of, or in any way bind the Institute.

Sec. 21—Section Secretaries shall forward to the Secretary of the Institute a report of each meeting held by the Section for the presentation or discussion of papers, and during January of each year a financial statement for the preceding year.

Sec. 22—A Section of the Institute may co-operate with other organizations in the holding of joint meetings and may invite members of such organizations and the public to its meetings.

Sec. 23—Failure of a Section to maintain the required activities, which shall include the holding of at least five meetings each year, shall place the Section on probation. All members of the Section shall be informed of the probation by the Secretary of the Institute who shall also call to their attention the requirements for maintaining the Section.

If the delinquency continues for a second year, a second notification to the Section membership shall be made by the Institute Secretary and the Board of Directors shall be informed of the probationary status of the Section.

If the delinquency continues for a third year, the Section shall, thereupon, be dissolved. The Secretary shall so report to the Board of Directors and so inform the Section membership.

COMMITTEES

Sec. 24—The standing committees, each of which shall normally consist of five or more persons, shall include the following:

Admissions	Nominations
Annual Review	Papers
Awards	Publicity
Board of Editors	Radio Receivers
Constitution and Laws	Sections
Electroacoustics	Standards
Electronics	Symbols
Facsimile	Television
Membership	Tellers
New York Program	Transmitters and Antennas
Wave Propagation	

These committees shall be advisory to the Board of Directors on those matters which are reasonably described by the committee names.

Sec. 25—The Membership Committee shall include the Secretary of each Section, *ex-officio*.

Sec. 26—The Sections Committee shall include the Chairman of each Section *ex-officio*.

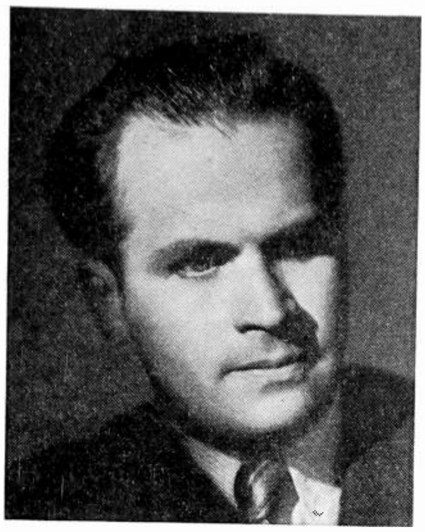
REPRESENTATIVES ON OTHER BODIES

Sec. 27—The Board of Directors may appoint representatives of the Institute on joint committees, boards, and other local, national, and international bodies.

PUBLICATIONS

Sec. 29—The Secretary is authorized to receive annual subscriptions to the monthly PROCEEDINGS at the rate of ten ($10.00) dollars per annum with an extra postage charge when the bulk rate of postage does not apply. A discount of fifty per cent from the subscription price of ten ($10.00) dollars will be allowed to colleges and public libraries upon direct subscription to Institute headquarters. Members, publishers, and subscription agencies may be allowed a discount of twenty-five per cent.

Contributors



Peter C. Goldmark

Peter C. Goldmark (A'36-M'38) was born on December 2, 1906, at Budapest, Hungary. He received the B.Sc. degree in 1930 from the University of Vienna and the Ph.D. degree in physics in 1931. Dr. Goldmark was in charge of the Television Department of Pye Radio, Limited, Cambridge, England from 1931 to 1935; consulting engineer in New York City, 1933 to 1935. Since 1935 he has been Chief Television Engineer at the Columbia Broadcasting System.

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Karl G. Jansky (A'28-M'34) was born on October 22, 1905, at Norman, Oklahoma. He received the A.B. degree in 1927, and the M.A. degree in 1936 from the University of Wisconsin. Since 1928 he has been with the Bell Telephone Laboratories.

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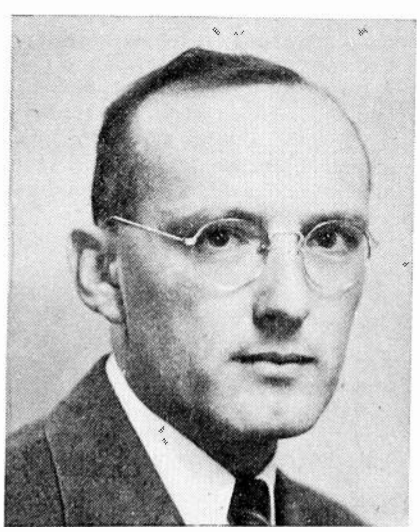
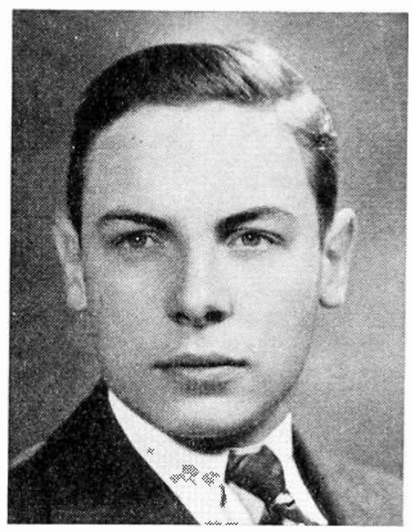


Karl G. Jansky

J. Jan Jansen was born in the Netherlands on August 5, 1916. He received the S.M. and S.B. degrees from the Massachusetts Institute of Technology in June, 1939. He is an Associate member of Sigma Xi and a Student member of the American Institute of Electrical Engineers.

❖

Paul S. Hendricks (A'26) was born at Souderton, Pennsylvania, on April 18, 1901. From 1918 to 1920 he was a Laboratory Assistant at Leeds and Northrup, and from 1920 to 1928 he was engaged in amateur, commercial, and broadcast station operation, maintenance, equipment design and construction. Mr. Hendricks was Development Program Assistant at the



J. Jan Jansen

American Radio Relay League in 1928; Communications Research Assistant at the Massachusetts Institute of Technology from 1929 to 1932; and engaged in high-frequency communication equipment design with Hendricks and Harvey and A. H. Ross and Company, 1932 to 1934. From 1934 to 1936 he was in the Ultra-High-Frequency Development and General Engineering Departments of the Columbia Broadcasting System, and since 1936 he has been in the Television Engineering Department.

❖

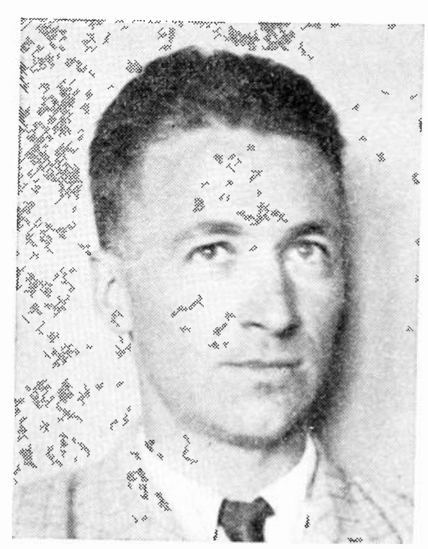
Jack Hobart Piddington (A'35) was born on November 6, 1910, at Wagga, New South Wales, Australia. He received the B.Sc. degree from Sydney University in 1932, and the B.E. degree with first-class honors and University Medal in 1934. The same year he was a Science Research Scholar at Sydney University.

Paul S. Hendircks

Since his appointment in 1935 Mr. Piddington has been with the Radio Research Board, Commonwealth Council for Scientific and Industrial Research. He was a Walter and Eliza Hall Fellow, University of Sydney 1936–1939; Cavendish Laboratory, Cambridge, 1936–1938, receiving the Ph.D. degree in Physics in 1938; Commonwealth Government Research Fellow, 1939; and Research Officer, Radiophysics Section, Council for Scientific and Industrial Research, 1939.

❖

For biographical sketches of W. L. Barrow, L. J. Chu, T. R. Gilliland, S. S. Kirby, and Newbern Smith, see the Proceedings for January, 1939; for Simon Ramo, September, 1939; and for M. J. O. Strutt, March, 1939.



Jack Hobart Piddington

Proceedings *of the* I·R·E

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GENERAL INFORMATION

The Institute

The Institute of Radio Engineers serves those interested in radio and allied electrical-communication fields through the presentation and publication of technical material.

Membership has grown from a few dozen in 1912 to more than five thousand. Practically every country in the world is represented in our roster of membership, with approximately a quarter of it located outside of the United States. There are several grades of membership, depending on the qualifications of the applicant, with dues ranging from $3.00 per year for Students to $10.00 per year for Members.

PROCEEDINGS, Standards Reports, and any other published material are sent to members without further payment.

The PROCEEDINGS

The PROCEEDINGS has been published without interruption from 1913 when the first issue appeared. Over 1800 technical papers have been included in its pages and portray a currently written history of developments in both theory and practice. The contents of every paper published in the PROCEEDINGS are the responsibility of the author and are not binding on the Institute or its members. Material appearing in the PROCEEDINGS may be reprinted or abstracted in other publications on the express condition that specific reference shall be made to its original appearance in the PROCEEDINGS. Illustrations of any variety may not be reproduced, however, without specific permission from the Institute.

Subscriptions

Annual subscription rates for the United States of America, its possessions, and Canada, $10.00; to college and public libraries when ordering direct, $5.00. Other countries, $1.00 additional.

Back Copies

The Institute endeavors to keep on hand a supply of back copies of the PROCEEDINGS for sale for the convenience of those who do not have complete files. However, some issues are out of print and cannot be provided.

Standards

In addition to the material published in the PROCEEDINGS, Standards on Electroacoustics, Electronics, Radio Receivers, and Radio Transmitters and Antennas were published in 1938. These are available to members free of charge as long as the supply lasts; other may purchase them for fifty cents each.

CONTENTS OF VOLUME 27—1939*

VOLUME 27, NUMBER 1, JANUARY, 1939

VOLUME 27, NUMBER 2, FEBRUARY, 1939

* A cumulative index of the same type as this but covering the PROCEEDINGS from its start to the end of 1936 is available at $1.00 per copy. The 1937 and 1938 Indexes will be found in the December, 1937, and December, 1938, issues of the PROCEEDINGS, respectively.

VOLUME 27, NUMBER 3, MARCH, 1939

VOLUME 27, NUMBER 4, APRIL, 1939

Volume 27, Number 10, October, 1939

Volume 27, Number 11, November, 1939

Volume 27, Number 12, December, 1939

AUTHOR INDEX

Numbers refer to the chronological list. **Bold-face** type indicates papers, light-face type indicates discussions, and *italics* refer to books and book reviews.

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NONTECHNICAL INDEX

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Biographical Notes

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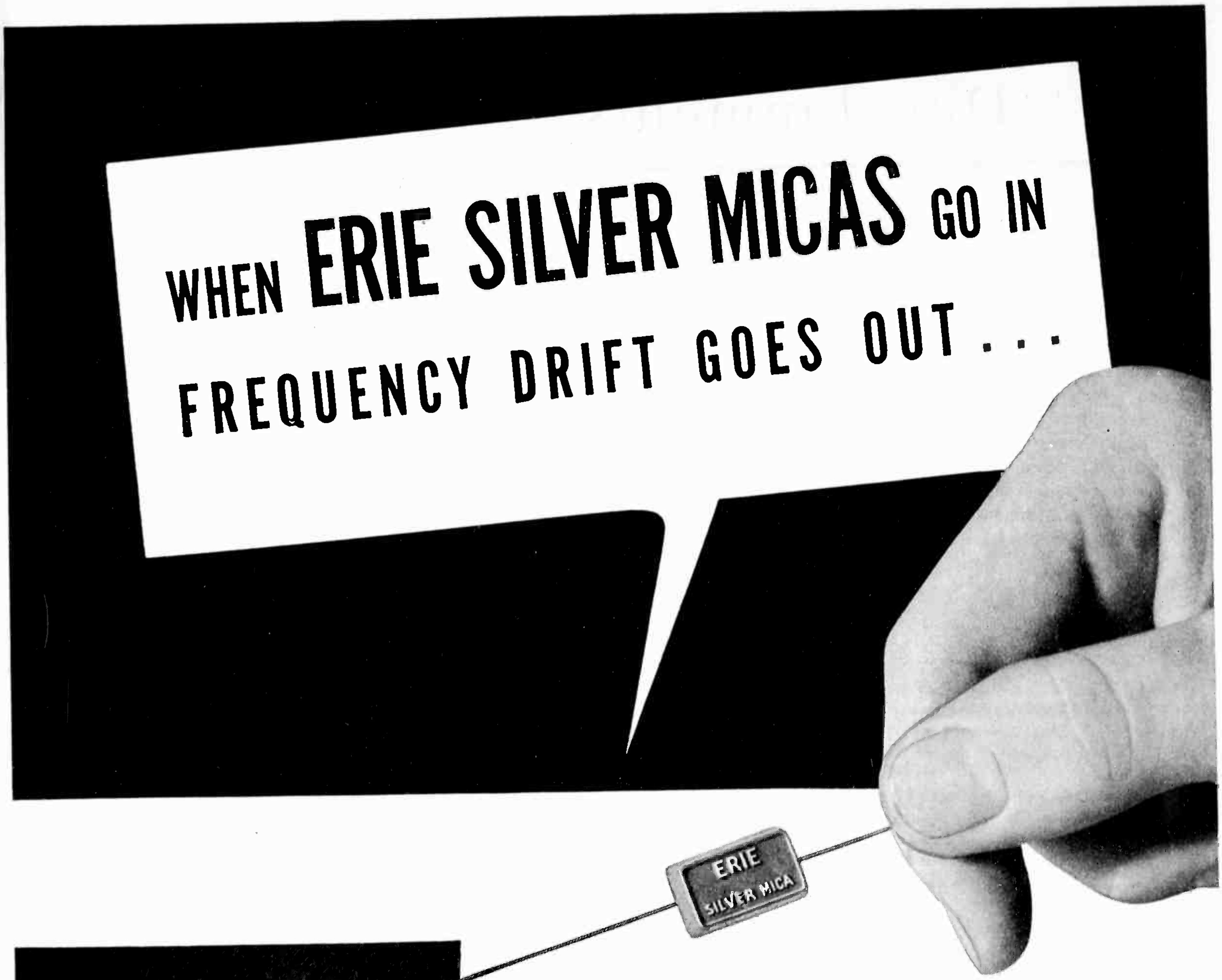
Report of Secretary

Representatives on Other Bodies



When used in place of ordinary mica condensers, the new molded Erie Silver Micas will prevent the possibility of frequency drift caused by temperature change in condensers.

Because their capacity change is only approximately 25 parts per million per °C. and power factor is less than .030%, these dependable condensers will provide an unusually high degree of tuning stability. Their molded construction makes them practically impervious to humidity changes. After 100 hours in 100% relative humidity at 104 °F., maximum capacity change is less than .05%.

In addition to the improved operation of oscillators brought about by the above characteristics, performance of your receivers can be still bettered by using tolerances of 1% or smaller at critical points. The unique construction of Erie Silver Micas makes possible their production in these close tolerances without excessive cost.

A letter to us will bring samples and complete engineering data to you. Write today.

ERIE RESISTOR CORPORATION, *Erie, Pa.*

TORONTO, CANADA • LONDON, ENGLAND • PARIS, FRANCE-J.E.CANETTI CO.

MANUFACTURERS OF RESISTORS • CONDENSERS • MOLDED PLASTICS

Commercial Engineering Developments

These reports on engineering developments in the commercial field have been prepared solely on the basis of information received from the firms referred to in each item.

Sponsors of new developments are invited to submit descriptions on which future reports may be based. To be of greatest usefulness, these should summarize, with as much detail as is practical, the novel engineering features of the design. Address: Editor, Proceedings of the I.R.E., 330 West 42nd Street, New York, New York.

●

Resistors

There is now available a new wire wound resistor based upon engineering developments* which are said to overcome certain basic limitations in prior resistor technique.

Power wire-wound resistors have been manufactured heretofore by winding single layers of space-wound bare resistance wire and covering the winding with vitreous enamels or cements to insulate the winding and provide mechanical protection for the wires.

This past construction has been used because there has been no satisfactory wire insulation which could be applied to resistance wires which had the required electrical and mechanical properties at the temperatures reached in resistor operation. The enamels and cements used, having the required mechanical properties, have been relatively poor insulators at resistor operating temperatures, and it has been necessary to insulate carefully the resistor surface from chassis or from other current carrying parts.

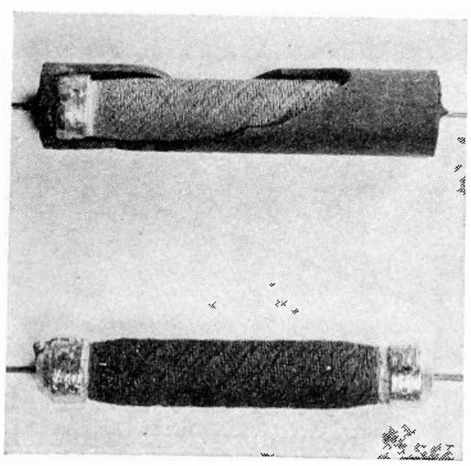
A more serious limitation has been the necessity for the use of very fine resistance wires. Since the physical size of the resistor structure is primarily determined by wattage rating and permissible temperature rise, it has been necessary to use progressively finer resistance wire to produce the desired higher resistance values in a given winding space. One mil diameter wire has been very widely used, with unsatisfactory field results. The very fine resistance wires have not been able to safely carry the currents involved in power resistor operation.

These limitations have resulted in the necessity for using much larger physical sizes of resistor than are actually necessary for the wattage dissipation involved in order to allow safe wire sizes to be used. They also place restrictions on the potential gradient across the resistor because of the low dielectric strength of the coating at elevated temperatures.

*Sprague Products Company, North Adams, Massachusetts

Resistor designs are based upon a new resistance wire insulation having characteristics not previously available. The wire is continuously insulated, by a special process, with a uniform, concentric, layer of an inert and completely inorganic insulating material. The insulation is sintered on the wire at temperatures of the order of 1000 degrees Centigrade and is subsequently resistant to heat, moisture, and mechanical abrasion, It is a good dielectric and yet is flexible to permit the winding of the wire. The wire insulation has a dielectric strength of the order of 350 volts per mil at 400 degrees Centigrade and can be operated at red heat without harm.

This insulated resistance wire can be wound turn against turn and in layers to produce a high space factor and permit the winding of much higher resistance values in a given winding space than could previously be produced with bare wires of satisfactory diameter. This permits the



use of wire of sufficiently large cross section to carry safely the currents involved at any resistance value and wattage rating. Extremely high density, "patterned" windings are used for the high resistance values to reduce the voltage gradients in the windings to negligible values.

No secondary insulations, such as cements or enamels, are needed on the winding, as the wire itself is insulated.

The resistor windings are on solid ceramic cores. Contact is made between the external terminals and the windings, through alloy castings which encircle the core ends and embed the terminals and windings ends.

The resistors are enclosed in cylindrical ceramic shells, which form a rigid mechanical protection for the windings and provide a completely insulated surface which can be mounted in direct contact with chassis or live parts.

Physical dimensions and materials received careful attention, in order to produce a relatively small temperature rise. The enclosing ceramic shell is dark brown in color, has an "etched" surface, and is made of a high heat conductivity material

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(Continued from page ii)

to obtain as rapid dissipation of heat as possible. A 10-watt unit has a temperature rise of only 185 degrees Centigrade at the hottest spot for full wattage dissipation, whereas a 250-degree Centigrade rise is common for resistors of this size and is permitted by Underwriters' standards.

The use of insulated resistance wire has permitted the design of a truly non-inductive unit. Ayrton-Perry windings, with one turn carrying current going in the opposite direction from the turn next to it and cancelling its magnetic field, are used. Measurements, by parallel resonance methods, at frequencies up to 60 megacycles, show the non-inductive units to have a negligible residual inductance. The distributed capacitance is 2.5 micromicrofarads. The resistors show a declining impedance with frequency due to the effect of the distributed capacitance.

Special resistors having a closely controlled value of resistance and residual inductance can be produced in this construction. Relatively large values of inductance, with small distributed capacitance, can be obtained with the "patterned" windings which are possible with the insulated wire.

All of the new resistors are marked on their surface with a temperature indicator. which is a spot of orange color which changes to dark brown when the resistors are overloaded 25 per cent in wattage, returning to orange color when the overload is removed. This indicator is entirely for the convenience of the user, as the resistors can safely handle heavy overloads.

Current Literature

New books of interest to engineers in radio and allied fields—from the publishers' announcements.

A copy of each book marked with an asterisk (*) has been submitted to the Editors for possible review in a future issue of the Proceedings of the I. R. E.

* Cathode-Ray Tubes. By Manfred von Ardenne, Radio Engineer, Lichterfelde Ost, Germany. New York: Pitman Publishing Corporation; London: Sir Isaac Pitman & Sons, Ltd., 1939. xiii +519+10 index pages, illustrated, 6×8½ inches, cloth. $12.50.

* Radios (No. 25 of Better Buymanship Series). By Albert R. Hodges. Chicago: Household Finance Corporation, 1939. 34 pages, 6×9 inches, paper. 2½ cents.

* Servicing by Signal Tracing. By John F. Rider. New York: John F. Rider, 1939. xi+360 pages, illustrated, 6×8½ inches, cloth. $2.00.

* The Radio Amateur's Handbook (Seventeenth Edition). By Headquarters Staff of the American Radio Relay League. West Hartford: American Radio Relay League, 1939, 448+8 index+120 page catalog, illustrated, 6½×9½ inches, paper $1.00, cloth $2.50.

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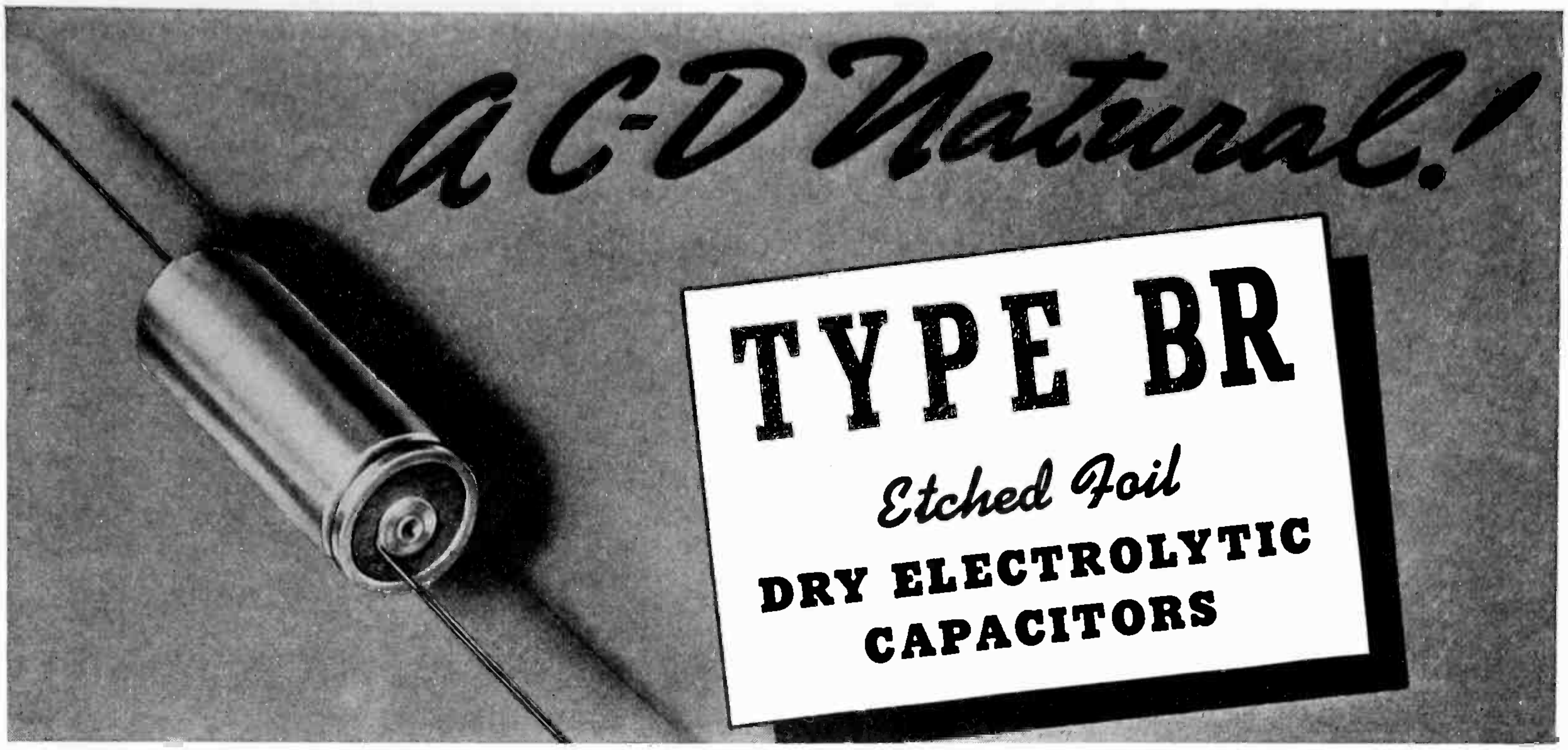

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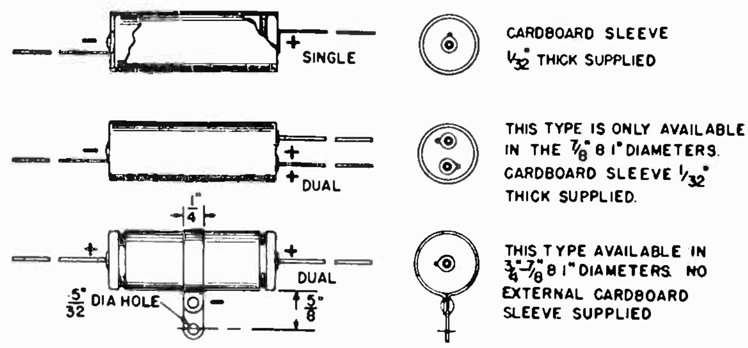

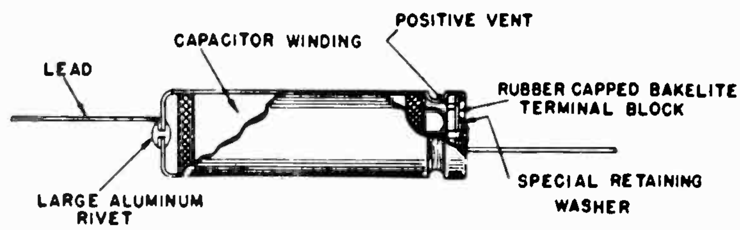

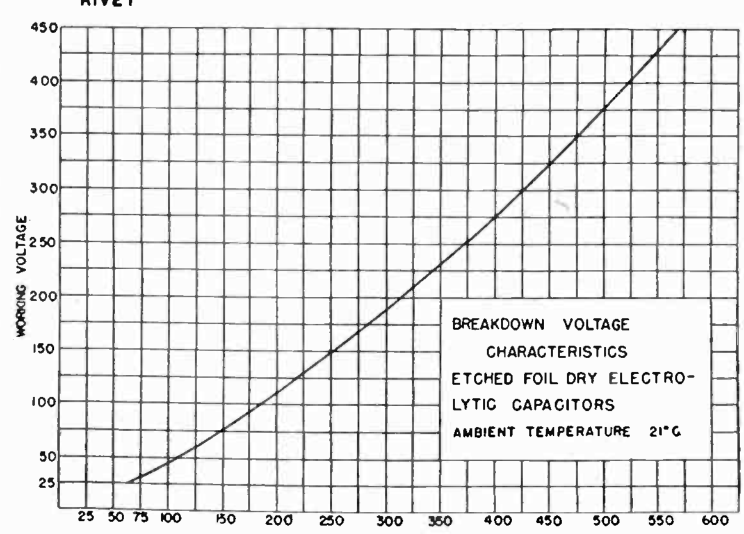

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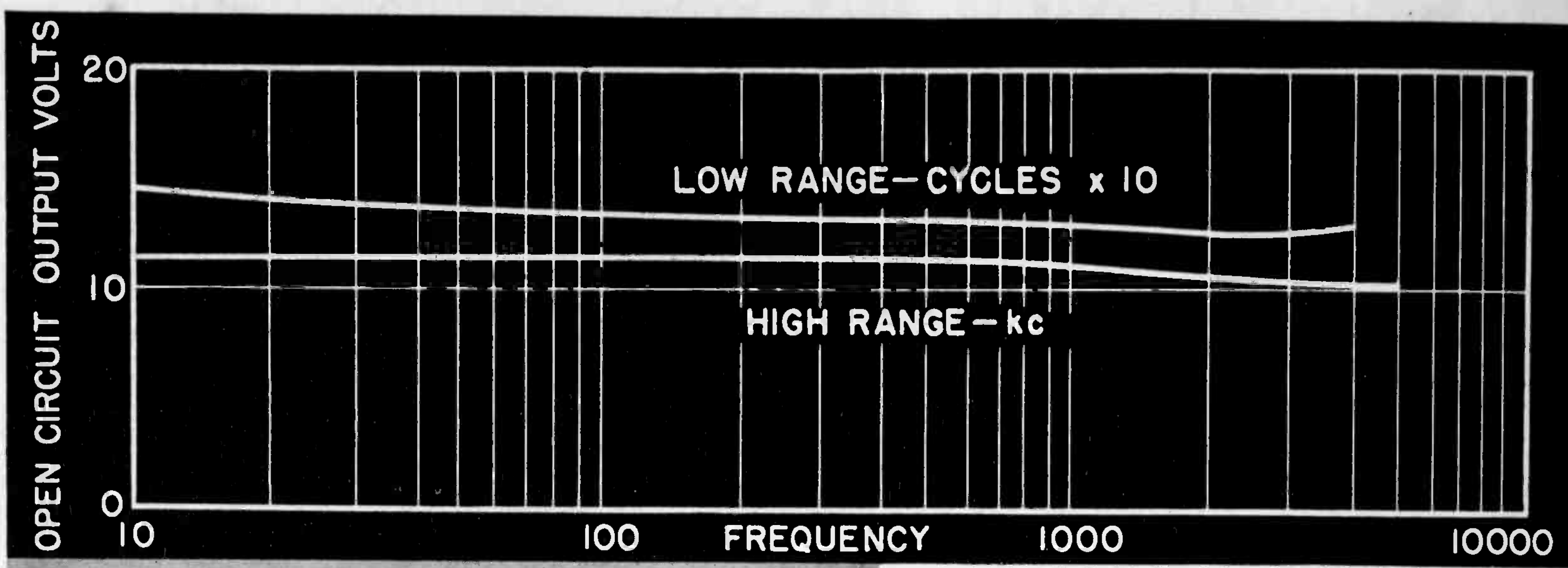

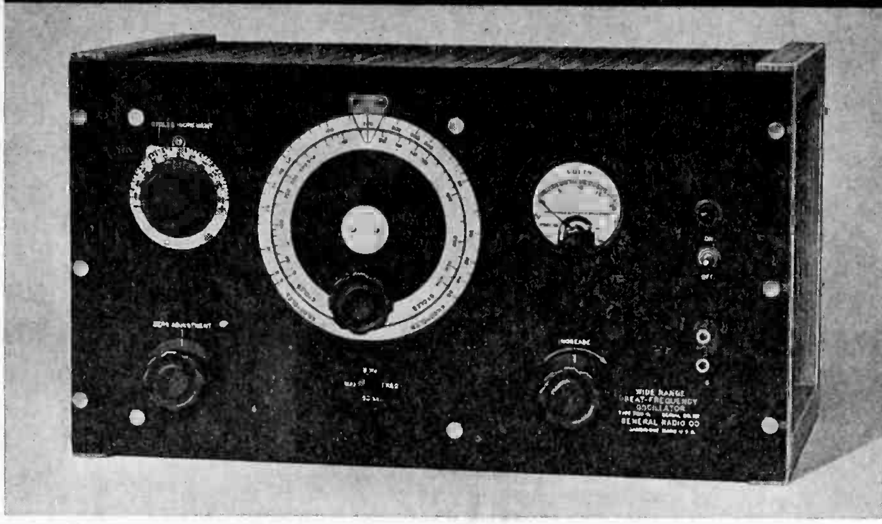

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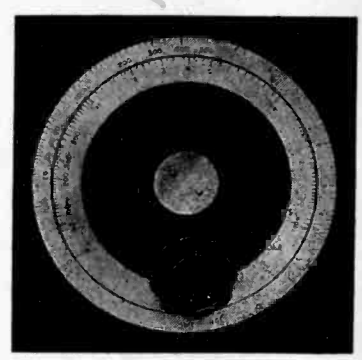

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