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PROCEEDINGS OF

The Institute of Radio Engineers

VOLUME 24 September, 1936 NUMBER 9

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The Institute of Radio Engineers

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Proceedings of the Institute of Radio Engineers
Volume 24, Number 9 *September, 1936*

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George Ashley Campbell was born on November 27, 1870, at Hastings, Minnesota. He received a Bachelor of Science degree from Massachusetts Institute of Technology in 1891 and from Harvard University a Bachelor of Arts in 1892, Master of Arts in 1893, and Doctor of Philosophy in 1901. He joined the American Telephone and Telegraph Company in 1897 and was retired from Bell Telephone Laboratories to which he was transferred the year before, in 1935.

The Institute Medal of Honor for 1936 was voted to Dr. Campbell for his contributions to the theory of electrical networks and received by him during the Institute's eleventh annual banquet in Cleveland, Ohio, on May 12.

In addition to his development of the electrical wave filter, Dr. Campbell was responsible for pioneering research in connection with the loading of long communication lines, cross talk, four-wire repeater circuits, side-tone reduction, inductive interference, antenna arrays, maximum output networks, Fourier integrals, and electrical units.

Proceedings of the Institute of Radio Engineers
Volume 24, Number 9 *September, 1936*

TECHNICAL PAPERS

A NEW HIGH EFFICIENCY POWER AMPLIFIER FOR MODULATED WAVES*

By
W. H. Doherty
(Bell Telephone Laboratories, Inc., Whippany, New Jersey)

Summary—This paper introduces a new form of linear power amplifier for modulated radio-frequency waves. Plate circuit efficiencies of sixty to sixty-five per cent independent of modulation are obtained by means of the combined action of varying load distribution among the tubes and varying circuit impedance over the modulation cycle.

The theory of operation is developed and detailed observations on the behavior of tubes in the new circuit are given in the paper. The use of stabilized feedback in connection with this circuit is discussed and significant measurements on a laboratory model of a fifty-kilowatt transmitter are shown.

THE trend toward increasingly higher power levels in broadcasting in the last few years has attached new importance to the matter of more economical operation of radio transmitters. Most of the opportunity for improvement in this direction lies in increasing the efficiency of the high power stages to reduce the cost of power, the size of high voltage transformers and rectifier, and the water-cooling requirements. With power levels of fifty kilowatts and higher these items account for an important part of the operating expense of a broadcast station, and the development of practical methods for increasing the efficiency should provide considerable stimulus to the use of higher power.

Methods hitherto employed for reducing power consumption include the high level class B modulation system, such as is used at WLW,[1] and the ingenious method of "outphasing modulation"[2] invented by Chireix and employed in a number of European installations.

* Decimal classification: R355.7 Original manuscript received by the Institute, June 5, 1936. Presented before Eleventh Annual Convention, Cleveland, Ohio, May 13, 1936.

[1] Chambers, Jones, Fyler, Williamson, Leach, and Hutcheson, "The WLW 500-kilowatt broadcast transmitter," Proc. I.R.E., vol. 22, p. 1151; October, (1934).

[2] Chireix, "High power outphasing modulation," Proc. I.R.E., vol. 23, p. 1370; November, (1935).

The development of these schemes was occasioned by the fact that the linear radio-frequency power amplifier, in the form in which it has been used for years in radio transmitters, may not be operated at an efficiency of more than about thirty-three per cent, for unmodulated carrier, if it is to supply the peak power output of a completely modulated wave. With this efficiency the direct-current power input to a fifty-kilowatt amplifier, for example, is 150 kilowatts, of which 100 kilowatts must be dissipated at the anodes of the water-cooled tubes.

The new form of linear power amplifier to be described in this paper removes this limitation of the conventional circuit, permitting efficiencies of sixty to sixty-five per cent to be realized, while retaining the advantages which account for the widespread use of linear amplifiers in broadcasting. These advantages include, notably, the elimination of high power audio equipment, since modulation may be accomplished at a low power level; and the ease with which linear amplifiers may be added to an existing transmitter to increase its power output. Linear amplifiers, moreover, are suitable not only for the carrier and double side-band signal employed in present-day broadcasting, but for any other type of transmission, such as the single side-band system now in use in the transoceanic radiotelephone circuit and frequently suggested as a remedy for the congestion in the broadcast spectrum.

A brief consideration of the mode of operation of the conventional linear power amplifier will show the reason for its low average efficiency and will afford a clew as to how, by the application of a new principle in power amplifier design, this efficiency may be approximately doubled.

Operation of Conventional Linear Power Amplifiers

The tubes are usually biased nearly to the cutoff point, so that the plate current flows in a series of pulses having approximately the shape of half sine waves, as shown in Fig. 1. The output circuit is antiresonant to the fundamental and has a low impedance to the harmonic components of the plate current, so that the plate voltage wave is nearly sinusoidal and opposite in phase to the plate current and grid voltage.

With a peak value of plate current of $i_{\max}$ and a peak amplitude of $e_{\max}$ in the plate voltage wave, the power output of the tube is

$$P_{\text{out}} = \frac{e_{\max} i_{\max}}{4}. \tag{1}$$

Since the average value of the half sine wave of plate current is $1/\pi$

times the maximum value, we have for the direct-current input power to the tube

$$P_{\text{in}} = \frac{E_B i_{\max}}{\pi}. \tag{2}$$

The efficiency is accordingly

$$\text{Eff.} = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{\pi}{4} \frac{e_{\max}}{E_B}. \tag{3}$$

If one were able to utilize a value of $e_{\max}$ equal to the direct plate voltage, the efficiency as given by (3) would be $\pi/4$, or 78.5 per cent. In practice, with a tube working close to its full output capacity, the plate swing is usually limited to a value of 0.85 to 0.9 times E_B,

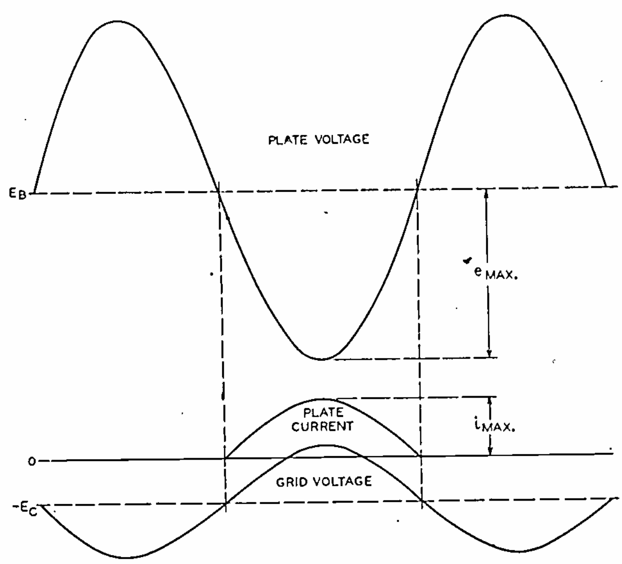

Fig. 1—Operating conditions in a linear power amplifier.

since the output will be decreased if the plate potential swings down to a value lower than the maximum grid voltage. Under these conditions, and allowing for a five per cent loss in the tuned output circuits of the amplifier, expression (3) will give a value of sixty-three to sixty-seven per cent as the maximum over-all plate circuit efficiency obtainable.

Now with a modulated wave applied to the grid this efficiency is obtained only at the maximum instantaneous output of the amplifier, and since the amplitude of the radio-frequency plate voltage wave, in a transmitter capable of 100 per cent modulation, is only half as great for the unmodulated condition as for the peaks of modulation, the efficiency with zero modulation does not exceed thirty-three per cent. Even during complete modulation the effective efficiency over the whole audio cycle is only fifty per cent, and for the average percentage modulation of broadcast programs the all-day efficiency is scarcely in excess of the value for unmodulated carrier.

The weakness, then, of the conventional method of amplifying a modulated wave is that the amplitude of the radio-frequency plate voltage is too small during most of the operating time, and in order to improve the situation it is necessary to devise a system in which a larger amplitude is employed.

High Efficiency Operation

The method of attack on this problem is to consider an amplifier operating, arbitrarily, with a high plate voltage swing and consequently high efficiency at the carrier output, and then to find what must be done to permit an increase in output. We shall see that a simple and fundamental means is available for doing this.

A tube will operate at high efficiency at any desired output power,

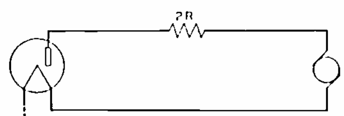

Fig. 2—Insertion of a hypothetical source of additional voltage.

however small, provided the alternating plate voltage is high; i.e., provided the load impedance is high enough to require a large voltage output from the tube. To take a concrete example, a tube capable of delivering 100 kilowatts at high efficiency into an impedance of R ohms will deliver fifty kilowatts into $2R$ ohms at the same voltage output and consequently the same efficiency. We should find, however, upon modulating the radio-frequency grid voltage, that the tube could not respond to the upward swings of modulation because the alternating plate voltage had already reached its maximum value at the fifty-kilowatt output.

Suppose now that an additional source of voltage could be inserted in series with the load, as represented by the generator of Fig 2. If the voltage of this generator increases from zero to a value equal to the output voltage of the original tube, we shall obtain the necessary increased voltage for modulation peaks. The current in the circuit will increase to twice its original value and the power in the load, which was originally fifty kilowatts, will increase to the necessary peak power of 200 kilowatts, or four times the carrier power, 100 kilowatts being furnished by the tube and the other 100 kilowatts by the generator. The sensation experienced by the original tube as the added generator comes into play is a gradual lowering of the impedance into which it works, since its output current increases without any increase in its output voltage, and when the added generator has a voltage equal to that of the tube this impedance has effectively been reduced from $2R$ to R ohms. The increase in current occasioned by the activity of the generator of course tends to reduce the output voltage of the original tube because of the greater internal drop, but since the grid excitation on the tube is continuing to increase in accordance with the modulation the tube is able to maintain its output voltage in spite of the increase in load current.

We now have the problem of replacing this added generator with a tube. Obviously we cannot replace it directly, because while a generator would offer no impedance to the flow of current from the original tube, an inactive tube would offer an infinite impedance. The solution is to interpose, as shown in Fig. 3, a network having a certain property;

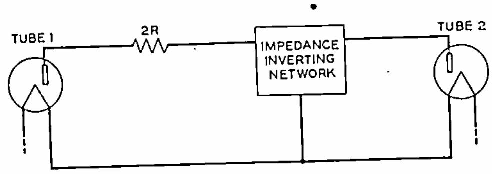

Fig. 3—Fundamental form of a high efficiency circuit.

namely, that the impedance at the sending end is inversely proportional to the terminating impedance. This is a familiar property of quarter-wave transmission lines and their equivalent networks. As long as the second tube does not conduct, the network is terminated in an open circuit. Its input impedance is therefore zero and the first tube works into an impedance of $2R$ ohms. When the second tube is permitted to conduct, the terminating impedance of the network is reduced, and since the grid excitation on the tube causes the plate current to be opposite in phase to the plate potential, this terminating impedance provided by the tube is a negative shunt resistance;[3] that is, the tube delivers power to the circuit. As the contribution of the second tube increases, lowering the negative terminating resistance of the network, the input impedance of the network, which was originally zero, increases. This input impedance is a negative series resistance which reduces the impedance presented to tube No. 1 from its original value of $2R$ ohms, and when, at the peak of modulation, tube No. 2 is contributing half the total power, the load impedance to No. 1 is R ohms, and No. 1 is able to supply twice the carrier power at the same radio-frequency plate potential as at the carrier.

We may revert to the generator analogy by recalling an associated

[3] Crisson, "Negative impedances and the twin 21-type repeater," *Bell Sys. Tech. Jour.*, vol. 10, p. 485; July, (1931).

property of impedance-inverting networks; namely, that any definite current at one pair of terminals is associated with a definite coexisting voltage at the other pair, entirely without regard to the terminating impedances. The supplying of current to the circuit by tube No. 2 at the far end of the network is, accordingly, identical in its effect to the injection of a voltage at the near end, in series with the voltage of tube No. 1, after the fashion of our original hypothetical generator.

Considering now the dynamic characteristic of the amplifier as a whole, the operation is as follows: The grids of both tubes are excited by the modulated output of the preceding stage, but for all instantaneous outputs from zero up to the carrier level tube No. 2 is prevented by a high grid bias, or some other means, from contributing to the output, and the power is obtained entirely from tube No. 1, which is working into twice the impedance into which it is to work when delivering its peak output. In consequence, the radio-frequency plate voltage on this tube at the carrier is nearly as high as is permissible and the efficiency is correspondingly high. Beyond this point the dynamic characteristic of tube No. 1, unassisted, would flatten off very quickly because the plate voltage swing could not be appreciably increased. The second tube, however, is permitted to come into play as the instantaneous excitation increases beyond the carrier point. In coming into play the second tube not only delivers power of itself, but through the action of the impedance-inverting network causes an effective lowering of the impedance into which the first tube works, so that the first tube may increase its power output without increasing its plate voltage swing, which was already a maximum at the carrier point. At the peak of a 100 per cent modulated wave each tube is working into the impedance R most favorable to large output and delivering twice the carrier power, so that the total instantaneous output is the required value of four times the carrier power. Thus, the required tube capacity is the same as in a conventional linear power amplifier.

What we have, then, is a two-tube amplifier in which the contribution of the second tube is delayed until the first tube has reached an efficient operating condition; whereupon there ensues a supplementary action between the tubes, to which the impedance-inverting network is a necessary adjunct.

The arrangement of Fig. 3 is one of the two fundamental forms of the high efficiency circuit. The other form is shown in Fig. 4. In this case the physical load impedance used is $R/2$, which is the same as would be employed if the tubes were to be connected in parallel in the conventional type of amplifier. The impedance-inverting network is then interposed between the load and tube No. 1, which is to deliver the carrier power. As long as tube No. 2 is inactive the network is terminated in $R/2$ ohms, and the network is so designed that the impedance presented to tube No. 1 under this condition is $2R$ ohms, the impedance necessary for attaining high efficiency at the carrier out-

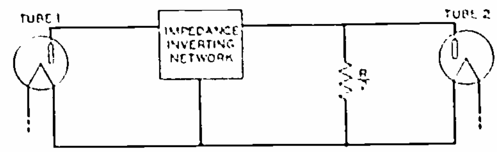

Fig. 4—Second fundamental form of high efficiency circuit.

put. As the second tube comes into action in parallel with the load it raises the effective terminating impedance of the network, with a consequent lowering of the impedance presented to tube No. 1; and again we have each tube, at the instantaneous peak of modulation, working into the desired effective load impedance of R ohms.

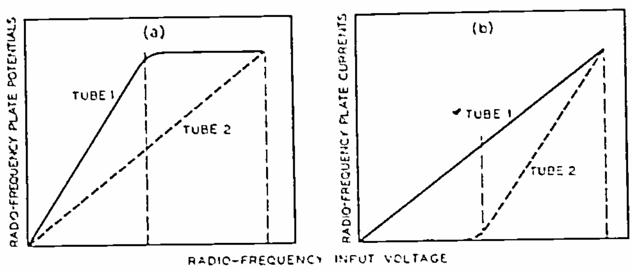

Fig. 5—Voltage and current relations in the two tubes.

Fig. 4 may be said to show a shunt connected load, as contrasted with the series connected load of Fig. 3. The shunt connection appears to be more advantageous for most practical purposes because the load circuit is grounded, while in the series arrangement the load is neither grounded nor balanced to ground.

General Observations

Voltage and Current Relations

The voltage and current relations in the two tubes as the amplitude of the grid excitation is varied are shown by Figs. 5(a) and 5(b), and the corresponding shapes of the envelopes of radio-frequency plate currents and voltages during complete modulation are shown in Fig. 6.

If we denote by k the ratio of the instantaneous amplitude of the envelope to the peak amplitude reached during 100 per cent modulation, then for amplitudes between zero and the carrier point, where

$k=1/2$, the total output of the amplifier comes from tube No. 1, and is given by the expression

$$\begin{aligned} P_{\text{total}} &= P_1 = (2kE_{\max})\,(kI_{\max}) \\ &= 2k^2E_{\max}I_{\max} \end{aligned} \tag{4}$$

where $E_{\max}$ and $I_{\max}$ are the root-mean-square values of radio-frequency plate voltage and plate current which are to exist when the

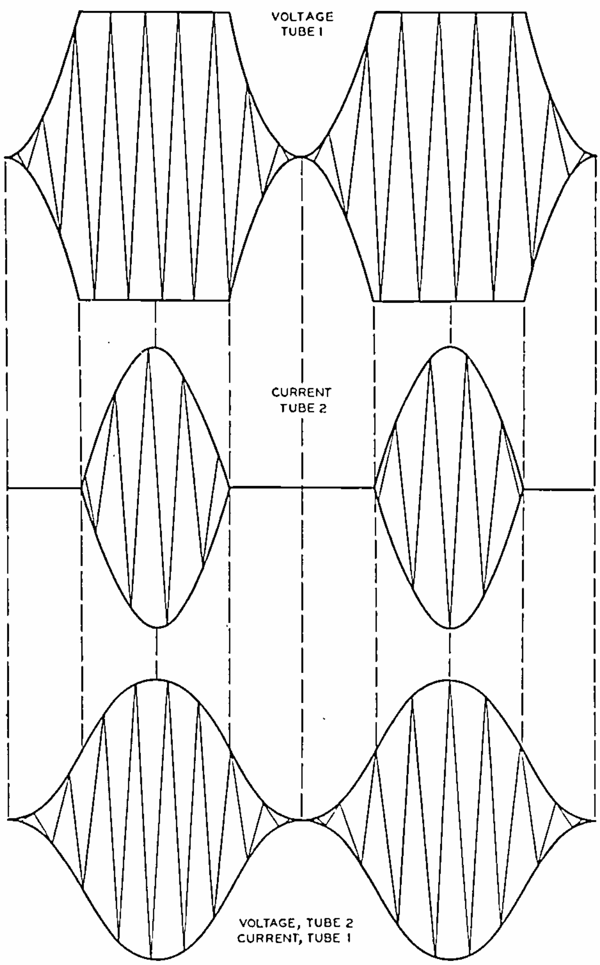

Fig. 6—Envelopes of the radio-frequency plate currents and voltages during complete modulation.

tube is delivering its maximum power; i.e., when $k=1$. The factor 2 above is required because the voltage on tube No. 1 reaches the value $E_{\max}$ when $k=1/2$.

Between $k=1/2$ and $k=1$ the voltage on No. 1 remains at $E_{\max}$ volts while the current in No. 1 and the voltage across No. 2 continue to rise linearly; meanwhile the current in No. 2 commences and rises twice as fast in order to reach the value I_{max} at $k=1$. The total power between these two values of k is the sum of the outputs of the two tubes:

$$\begin{aligned} P_{total} &= P_1 + P_2 \\ &\doteq E_{max}(kI_{max}) + (kE_{max})\,(2k - 1)I_{max} \\ &= 2k^2E_{max}I_{max} \end{aligned} \tag{5}$$

which is the same as expression (4) above, showing that the current and voltage relations of Fig. 5 are consistent with continuity in the dynamic characteristic of the amplifier.

By assuming k to vary sinusoidally about its carrier value of 1/2, in accordance with the modulation, and integrating the values of P_1 and P_2 as given by (4) and (5) over the appropriate half cycles of modulation, the average output of each tube during modulation may be determined. This integration gives for the average output of tube No. 1 during 100 per cent modulation a value of 0.93 times the carrier output, and for tube No. 2 a value of 0.57 times the carrier output, these two figures adding up to the factor 1.5 by which the average total output is increased when complete modulation occurs.

The sharp transitions shown at the carrier amplitude for this hypothetical case would necessitate extremely careful adjustment to insure that they were simultaneous. In practice neither transition is so abrupt, and the adjustment becomes, as will be shown, an operation requiring only reasonable care.

Since the direct plate current of a power amplifier is closely proportional to the radio-frequency plate current, the audio-frequency component of the plate current of tube No. 1 is very nearly sinusoidal with sinusoidal modulation, and the average or direct plate current of this tube is unchanged from its carrier value, while the audio-frequency plate current of No. 2 is a half sine wave. The average or direct-current value of a half sine wave existing during the positive half of the modulation cycle is $1/\pi$ times the peak value. This peak value is determined by the peak efficiency of tube No. 2, and by virtue of the smaller pulse width of the radio-frequency plate current of this tube, resulting from its being biased well beyond cutoff, this peak efficiency is somewhat higher than can be obtained in tube No. 1, whose pulse width is about a half cycle. It is easy to show, for the hypothetical conditions of Fig. 5, that if the average plate current of No. 2 at 100 per cent modulation is to be half that of No. 1, so that the over-all efficiency of the amplifier may be the same as with unmodulated carrier, the peak efficiency of tube No. 2 must be $4/\pi$ times that of No. 1, or about eighty per cent for a combined efficiency of sixty-three per cent.

Impedance-Inverting Networks

It will be useful at this point to note the structure and properties of the impedance-inverting networks which are employed to obtain the impedance variations necessary for high efficiency operation.

In Fig. 7 are shown two networks having the interesting and easily verifiable property that a voltage E applied or appearing at either end of the network is associated with a coexisting current at the other end having an amplitude E/X and a phase differing by ninety degrees from the phase of E. This current is independent, in amplitude and phase, of the nature of the impedance through which it flows at the

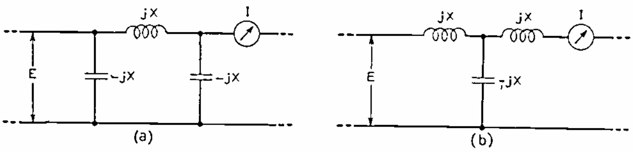

Fig. 7—Typical impedance-inverting networks.

terminals of the network. When this terminating impedance is a resistance R_1 and a voltage E is applied across the other terminals, the above property leads immediately to the following relations:

(a) The voltage phase retardation introduced by the network is ninety degrees regardless of the value of R_1; this indicates that no phase modulation will be occasioned by the variation in terminating impedance of the network when the second tube comes into action.

(b) The input impedance of the network is a resistance inversely proportional to R_1; it is equal to X^2/R_1.

(c) The voltage across R_1 due to E is proportional to R_1; it is equal to $-jE \cdot R_1/X$.

The above properties are likewise possessed by a quarter-wave transmission line whose characteristic impedance is X ohms, and by certain other network configurations.

If the coils in the above networks are replaced by condensers and the condensers by coils the same properties hold except that the phase is advanced ninety degrees instead of retarded.

By using the π-network of Fig. 7(a) a very simple output circuit for the amplifier may be arrived at. The fundamental relations are shown in Fig. 8(a). A load circuit is designed having a shunt resistance of $R/2$ ohms, as if the two tubes were to be connected directly in parallel. R, as before, is the impedance into which each tube is to work when delivering its peak power. The ninety-degree network is then interposed between tube No. 1 and this load circuit. The constants of this network must be such that when tube No. 2 is inactive (i.e., when the network is terminated in $R/2$ ohms) the alternating voltage across No. 2 is half the voltage across No. 1, so that if the voltage on No. 1 has reached the maximum permissible value when the carrier power is being delivered, the voltage on No. 2 will have reached only half this value and may therefore rise sufficiently to accommodate the peaks of modulation. By relation (c) above, this requires that the three elements of the ninety-degree network should have reactances of R ohms.

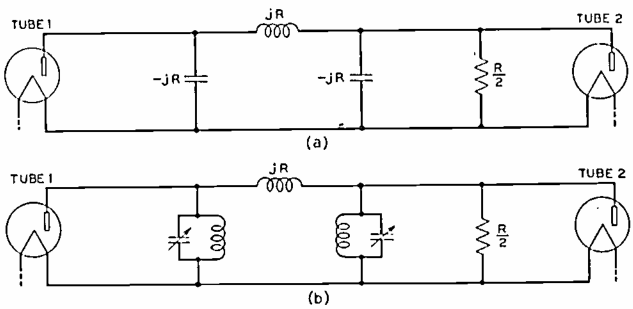

Fig. 8—Evolution of the output circuit.

Next it is desirable, as in any radio-frequency power amplifier, to connect across each tube an antiresonant circuit to provide a path of fairly low impedance for the harmonic components of the radio-frequency plate currents. By detuning these antiresonant circuits sufficiently to obtain negative shunt reactances of R ohms, the shunt condensers of the ninety-degree network may be eliminated, and the complete output circuit is reduced to the simple form shown in Fig. 8(b).

The effective shunt load $R/2$ would in most cases be obtained by coupling the antenna circuit to the necessary extent into the tuned circuit across tube No. 2. The intertube coupling coil designated jR may be used to carry the direct plate current from one tube to the other, so that only one plate choke is necessary.

The process of tuning the circuit is very simple. The proper reactance for the intertube coupling coil having been selected and the load having been coupled in so as to obtain the proper shunt resistance $R/2$, the condenser across tube No. 2 is adjusted under power to give a ninety-degree phase relation between the plate potentials on the two tubes, as indicated on a cathode-ray oscilloscope by an elliptical pattern with axes at right angles. The condenser across tube No. 1 is then adjusted for minimum plate current on this tube, in accordance with the usual method of tuning a power amplifier, and the plate circuit tuning is complete.

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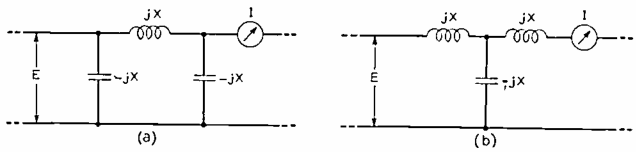

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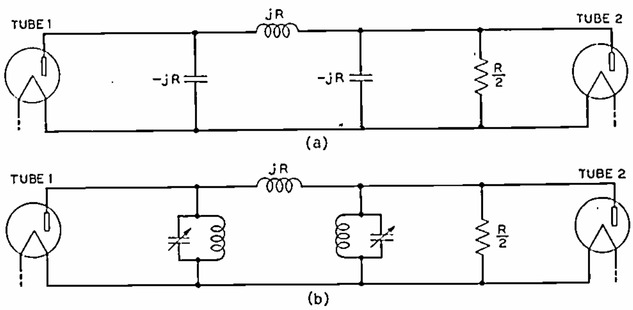

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An interesting feature of this circuit is that when the carrier is unmodulated the radio-frequency harmonic output is considerably reduced because the power is coming entirely from tube No. 1 and the harmonics generated in this tube are attenuated by the intertube coupling circuit.

The Input Circuit

The high efficiency system being described is unique in being purely a method of amplification and not a modulation scheme, and therefore requiring only a modulated wave from a lower powered

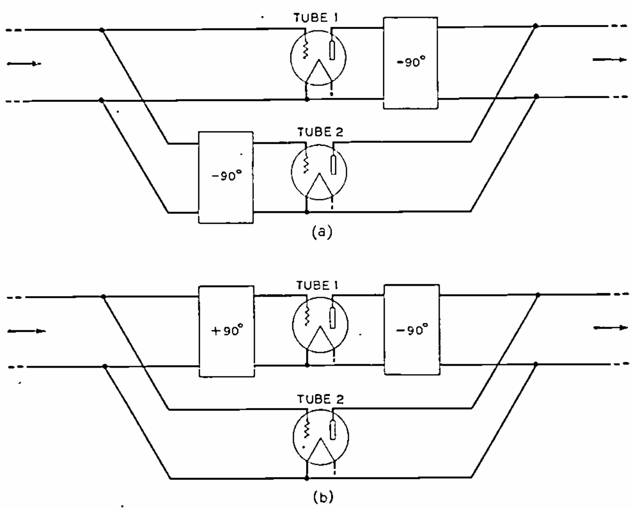

Fig. 9—Alternative arrangements of the grid phase shifting network.

transmitter for exciting the grids. In applying this modulated input to the amplifier it is necessary to have the voltages on the two grids ninety degrees apart in phase in order that each may be opposite in phase to the related plate potential, as is necessary in any power amplifier.

The ninety-degree phase relation may be obtained by using a network of the general type shown in Fig. 7, and the adjustment is accomplished under power, using a cathode-ray tube, in the same way as has been described for the plate circuit.

With the appropriate ninety-degree network inserted in the grid circuit, the amplifier assumes one of the forms indicated in Fig. 9. This figure illustrates clearly the fundamental simplicity of the new system. *Obviously the operation is essentially the same as that of the conventional linear amplifier at any time that the load is equally divided between the tubes, as is the case for an instant at the peak of a completely modulated wave; but merely by the introduction of two phase shifts it becomes possible* *to make use of a variable load distribution to establish the necessary conditions for high efficiency at outputs much smaller than the peak output.*

The simplest and most obvious way of keeping the second tube inactive until the carrier point is reached is by the use of a higher grid bias on this tube, with sufficient excitation so that at the peak of modulation the required output is obtained from the tube in spite of the higher bias.

The first tube, being biased nearly to cutoff, behaves like a conventional linear amplifier from zero excitation up to the carrier point. Beyond this point, because of the changing load impedance, the excitation is not required to rise to twice its carrier value. In most of the tests it has been found that the effective excitation on the first tube is required to rise only about forty per cent instead of 100 per cent on the positive half of the modulation cycle. A further increase in drive would carry the instantaneous grid potential up to a point where too

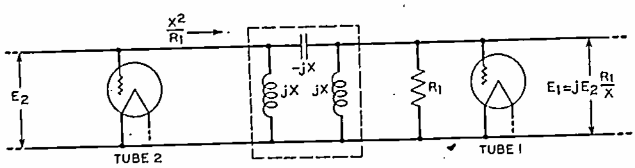

Fig. 10—Grid exciting circuit.

many of the electrons would go to the grid instead of the plate, causing a very rapid increase in positive grid current. A number of ways immediately suggest themselves for obtaining the required limiting action. One of the simplest ways is by the use of a grid leak in the bias supply to the first tube; another useful method is to permit the grid current to limit the drive by its shunting effect on the exciting circuit. In Fig. 10 the two grids are shown connected to a phase shifting network consisting of reactances of X ohms and a resistance R_1. When the grid of tube No. 1 is not conducting, the voltage E_1 obtained on this grid with a voltage E_2 applied to the network is $jE_2 \cdot R_1/X$. When grid current flows in tube No. 1, the effective reduction in R_1 causes a diminution in the ratio of E_1 to E_2. The proper value of R_1 to give the desired limiting effect is easily calculated from the approximate maximum grid current. The proper resistance is not at all critical because of the rapidity with which the grid current increases with drive at power outputs close to the maximum. This value of R_1 is considerably higher (i.e., the required driving power E_1^2/R_1 is lower) than in the case of the conventional linear amplifier, where R_1 must be made low enough so that the grid current will not cause appreciable diminution in the drive.

As R_1 is effectively lowered by the rapidly increasing grid conductance, the input impedance X^2/R_1 of the grid phase-shifting network is increased, compensating to a large extent for the shunting effect of the grid current which flows in tube No. 2 as it approaches its peak output, so that the driving stage is assisted in maintaining the proper drive on No. 2.

On account of these effects the grid driving power required for this type of amplifier is actually less than for the conventional linear amplifier, and in the experimental work a driving stage has been used in most cases having only about half the power capacity customarily employed for exciting a conventional amplifier of the same output.

Linearity of Amplification

For high quality transmission it is important to obtain in power amplifiers a linear relation between grid exciting voltage and output circuit voltage. In the high efficiency circuit the high impedance used for tube No. 1 over the lower half of the modulation envelope causes the dynamic characteristic to be quite straight in this region. To obtain linearity from the point where curvature begins on the first tube, up to the point representing the peaks of modulation, is a matter involving both the point at which the second tube comes into action, and the rate at which its contribution increases with drive. For securing a satisfactory adjustment we have available two variables; namely, the bias on the second tube and the amplitude of the excitation on this tube. A higher bias requires a higher exciting voltage to overcome it, but the rate at which it is overcome is greater. By a reasonably careful selection of these two quantities the amplifier may be made to operate with low distortion.

Careful adjustment of excitation and bias to obtain low distortion is no longer necessary when the feed-back principle[4] due to Black is employed. One application of stabilized feedback to reduce distortion has been described[5] in connection with a fifty-kilowatt transmitter. This application involved feeding back to the audio circuits a sample of the rectified output. With transmitters employing low level modulation and linear radio-frequency amplifiers we have available not only this type of feedback but also the opportunity for using a radio-frequency feedback from the final power amplifier to one of the earlier stages of linear radio-frequency amplification. Such feedback may be

[4] H. S. Black, "Stabilized feedback amplifiers," *Elec. Eng.*, January, (1934); *Bell Sys. Tech. Jour.*, vol. 13, pp. 1–18; January, (1934).

[5] Poppele, Cunningham, and Kishpaugh, "Design and equipment of a fifty-kilowatt broadcast station for WOR," Proc. I.R.E., vol. 24, pp. 1063–1081; August, (1936).

applied independently of any audio-frequency feedback, and has an important advantage over the latter in that it corrects not only distortion and noise but carrier shift as well.

In applying feedback to a transmitter it is found that a stabilization or improvement in linearity is obtained which corresponds to the decibel reduction caused by the feedback in the over-all gain. It is found, moreover, that each individual noise component arising in the amplifier as a result of power supply ripples is automatically reduced by an amount corresponding to the reduction in gain.

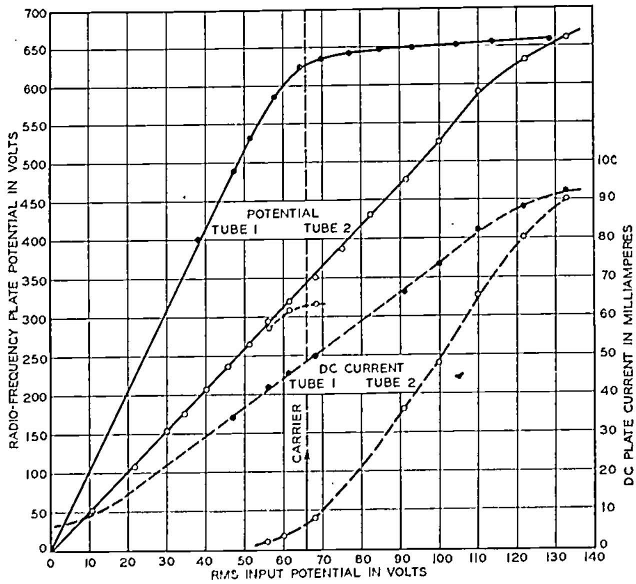

Fig. 11—Typical experimental results.

Because of the gain reduction it is necessary to increase the level ahead of the point at which the feedback is introduced. This is usually an easy matter because the power at this point is small in comparison with the output of the final stage. Moreover, the intermediate stages included in the feed-back loop may be designed for smaller output than is normally required, since the distortion caused by their deficiency in output will be taken care of by the feedback itself. The use of feedback, therefore, requires little or no increase in the tube complement of a transmitter.

Experimental Results

The first test of the high efficiency circuit was made with a pair of small tubes, and was intended chiefly to permit a study of the action of the tubes under these new conditions of variable load distribution and to show whether the amplifier as a whole would behave in the manner predicted. No feedback was used in this first test. Fig. 11 shows the observed variation of the radio-frequency plate potentials and direct plate currents of the two tubes as the excitation on the amplifier was increased. The radio-frequency plate potential of tube No. 2 is the potential across the load circuit and is required to be linear with excitation. The short dotted portion halfway up on this characteristic shows the curvature that would be obtained if the second tube were not allowed to come into action. With proper adjustment of the bias and excitation on the second tube this effect is eliminated and the characteristic continues to rise up to the desired peak amplitude.

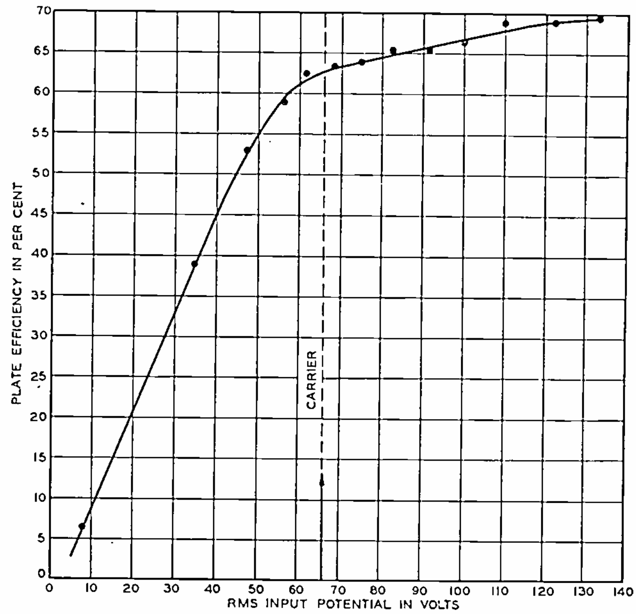

Fig. 12—Instantaneous plate efficiency for the test of Fig. 11.

The radio-frequency plate voltage of tube No. 1 is seen to be twice that of No. 2 up to the point where curvature begins, and then to increase only slightly between the carrier output and peak output. The plate current of No. 2 commences just before the carrier point is reached and rises twice as rapidly as the plate current of No. 1. The equality of plate currents and radio-frequency plate potentials on the two tubes at the peak of modulation indicates that the tubes are contributing about equally to the instantaneous output at this point.

The observed efficiency is plotted against excitation in Fig. 12. It is seen to be sixty-three per cent at the carrier level. It continues to rise even after the second tube comes in, because the plate voltage swing on the first tube is still rising slightly. The initial efficiency of tube No. 2 is not zero, of course, but about forty per cent, since the plate voltage swing on this tube is already half its maximum value and its plate current pulse width is very small; there is little sacrifice in efficiency, therefore, in permitting this tube to contribute a little power at the carrier point.

By integrating the direct plate currents of Fig. 11 over a complete cycle of modulation the effective average efficiency has been calculated for various percentages of modulation. This calculation shows the efficiency to be practically independent of modulation, being sixty-three per cent at 100 per cent modulation as well as with unmodulated carrier, and having a minimum value of sixty-one per cent at about seventy per cent modulation. With this constant efficiency the direct plate current increases fifty per cent at full modulation, as does the radio-frequency output power. Chireix,[2] who obtains approximately

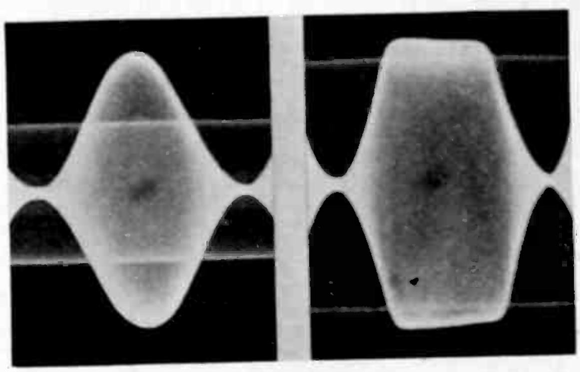

Fig. 13—Typical cathode-ray oscillograms of the plate potentials of tube No. 2 (left) and tube No. 1 during complete modulation.

the same efficiency as above for unmodulated carrier, gives a value of seventy-six per cent as the increase in direct plate current at full modulation with his system, with a corresponding reduction in efficiency to about fifty-four per cent for a fully modulated output. The reduction is due to the fact that over most of the modulation cycle the tubes in the Chireix system work into loads which are not purely resistive. In the system now being described there is no such detuning effect and the load impedances are purely resistive at all times.

Fig. 13 shows typical cathode-ray oscillograms of the radio-frequency plate potentials across the two tubes in a high efficiency setup at 100 per cent modulation. The faint horizontal lines are for the unmodulated carrier condition, the carrier potential on tube No. 1 being twice as great as on No. 2.

Fig. 14 gives the results of 400-cycle distortion measurements at the input and output of a two-stage amplifier having fourteen decibels of radio-frequency feedback and with the second stage delivering a five-kilowatt carrier at an over-all plate efficiency for the tubes and output circuit of sixty-two per cent, the actual tube efficiency being about sixty-eight per cent. The first stage was a 100-watt linear power amplifier of the conventional type. The distortion in the output is seen to differ only slightly from the distortion present in the source of modulated power used to excite the amplifier. This result is typical of what would be obtained in adding a high efficiency amplifier to a transmitter already existing, and indicates that high fidelity transmission may be achieved if the driving source has good quality.

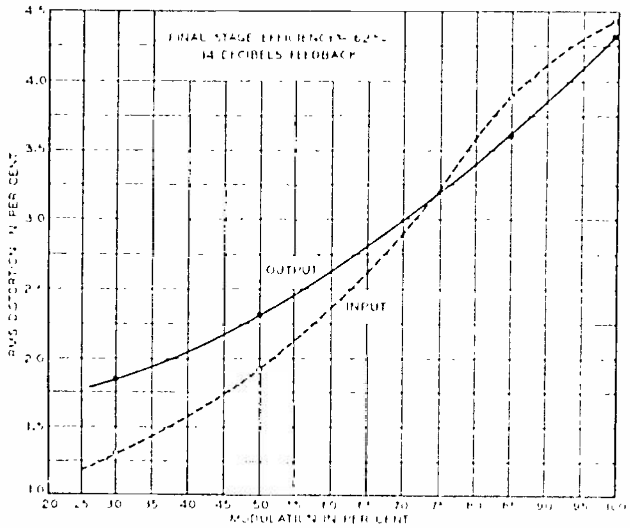

Fig. 14—Distortion measurements on a five-kilowatt amplifier with radio-frequency feedback.

The carrier shift measured in this test was zero. The use of feedback rendered the amplifier remarkably free from the necessity for critical adjustments of bias or relative excitations on the tubes, and the output circuit could be thrown far out of tune with no appreciable effect on the output current and no visible effect on the wave shape.

A difficulty encountered in applying feedback to radio transmitters is that introduced by cumulative phase shifts in the successive stages within the feed-back loop, which not only limit the amount of feedback obtainable, but may cause the distortion at the higher modulating frequencies to be increased rather than reduced. So serious is this difficulty, particularly at the lower broadcast frequencies where the band width of the tuned circuits is least, that when an attempt is made to introduce feedback in a transmitter not designed for it, there usually results only a few decibels improvement in noise and in distortion at low modulation frequencies and in many cases increased distortion at modulating frequencies even as low as 1000 or 2000 cycles. Although the percentage modulation in broadcast programs is normally very small at the higher audio frequencies, it nevertheless behooves the designer to do everything possible to reduce this effect. Fig. 15, which gives the results of distortion measurements on a complete fifty-kilowatt high efficiency transmitter built in the laboratory, shows what may be accomplished by careful design of the various stages to reduce phase shifts to a minimum. This transmitter was operated at the low-frequency end of the broadcast spectrum in order that the most unfavorable conditions in the matter of band width might be encountered. Audio-frequency feedback was employed to the extent of twenty-eight decibels, with a resulting distortion level less than one per cent at any frequency between fifty and 1000 cycles, and increasing

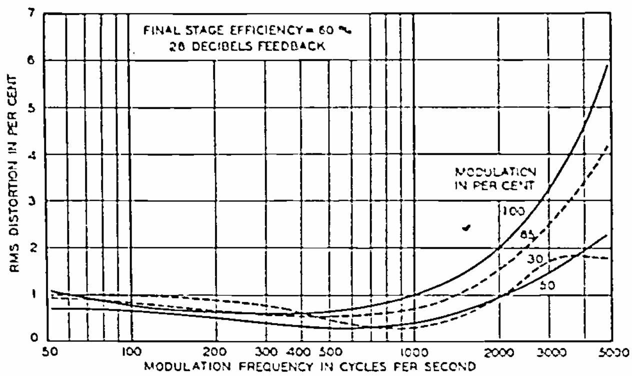

Fig. 15—Distortion measurements on a complete fifty-kilowatt transmitter with audio-frequency feedback.

only slowly until the high audio frequencies are reached where the modulation in an actual program of course rarely exceeds a few per cent.

The frequency characteristic was flat to within 0.25 decibel from thirty to 15,000 cycles.

The large amount of feedback employed permitted the use of alternating-current filament supply to all of the tubes with a noise level sixty-six decibels down unweighted and seventy-six decibels down as measured on a standard program noise weighting network, with no devices employed for hum suppression except for the feed-back action, which of course is entirely automatic.

The efficiency of the final stage with its output circuits was sixty per cent with the carrier unmodulated and sixty-three per cent at 100 per cent modulation. This high efficiency operation of the final stage, together with the reduced grid driving power permitted by the new system, resulted in a reduction in the total power consumption of a fifty-kilowatt transmitter, including all auxiliaries, from about 230 kilowatts, required by the conventional type, to approximately 135 kilowatts in the new system with normal program modulation. It is interesting to note that the total power input in this case is actually less than the power dissipated in the water-cooling system in the usual fifty-kilowatt installation.

Conclusion

As compared to the conventional linear amplifier the new system affords a power saving of nearly fifty per cent in the final stage plate supply, and the plate dissipation is reduced by a factor of three or four, with a resulting economy in the cooling system and an improvement in tube life.

The absence of any such requisites as the complicated driving stages of the Chireix system or the large audio equipment involved in high level modulation gives the new circuit an important advantage over other high efficiency systems in cost of apparatus and simplicity of design.

To accomplish high level modulation with the same total tube complement as required with linear radio-frequency amplifiers it is necessary to subject the smaller number of radio-frequency tubes to higher instantaneous plate voltages, with consequent likelihood of shortening of tube life and operating difficulties due to flashovers, particularly in case of overmodulation. In the proposed system, as in any low level system, the plate voltage is held constant at a value consistent with safe operation, and overmodulation is of no particular significance as far as any harmful effect on the tubes is concerned.

The new amplifier has been found to operate in complete accord with theoretical predictions, and is believed to offer a most logical and practical solution to the problem of efficient operation of high power transmitters.

Acknowledgment

The author wishes to express his appreciation of the encouragement and opportunity that have been given him to carry out this work, and the valuable assistance he has received from his associates of Bell Telephone Laboratories, Inc., in setting up and testing the high efficiency circuit.

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A MODERN TWO-WAY RADIO SYSTEM*

By
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Summary—*The general problems and limitations encountered in two-way police communication are discussed. A receiver circuit especially adapted to the adverse conditions imposed upon this class of equipment is described in detail and performance curves are shown. Several transmitters are described and their salient features pointed out. A new type of ultra-high frequency antenna system is described and its general theory of operation explained.*

Introduction

THE acceptance of radio communication as an essential part of a municipal police system has opened up a new field for the radio engineer. It has presented to him a set of conditions and requirements not heretofore encountered in radio communication. Give a radio engineer an efficient antenna, a skilled operator, and power on the proper frequency, and his communication range far exceeds that required for municipal police service. The conditions presented to him by municipal police radio are quite different; low antennas on a moving vehicle, in many cases unskilled operators, and limited power supply. Modern municipal police radio systems are being put on the higher frequencies with their more or less line-of-sight transmission. The frequencies now assigned by the Federal Communications Commission for experimental use by stations of this class lie between thirty and forty-two megacycles. The choice of this frequency range to localize the communications has made possible two-way communication. Today, two-way duplex radio communications from a moving automobile in a properly equipped city with an efficient maintenance organization approaches the simplicity and reliability of a telephone call from the corner drug store. In fact, a call put through from a car to a wire extension of a police telephone system can often not be distinguished from a call from a regular telephone station.

General Considerations

Police radio systems in common use today are classified as either "one-way" or "two-way." In a one-way system, headquarters can talk to the cars, but the cars are not equipped with transmitters for reply-

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ing to headquarters. In a two-way system, the cars are equipped with transmitters and can reply to headquarters. Two-way systems are divided into two types, duplex and simplex. When using the duplex system, transmission and reception may be carried on simultaneously as is the case during an ordinary telephone conversation. The simplex system does not permit this feature; i.e., it is not possible to receive when the transmitter is being used, nor is it possible to transmit while receiving a signal.

In the development of police radio communication apparatus, it was necessary to solicit the co-operation of police departments and to consider carefully the merits of these systems with the view of determining which appeared to be the more suitable. After numerous comparative tests, it became obvious that the duplex method offered several major performance features over those which it was possible to obtain with other systems. A number of these outstanding points are listed below.

1. It is the normal method of direct verbal or telephone conversation.
2. It allows messages to be transmitted, acknowledged, and discussed more quickly.
3. It is simpler to operate.
4. It allows the use of telephone extensions on the radio circuit.
5. It allows headquarters to supervise all car-to-car communications.
6. It increases the car-to-car communication range since headquarters acts as a relay station.

Of course, two frequencies are required for a duplex system and only time will tell whether the operational advantages listed above outweigh the additional channel and equipment required.

A block diagram of a complete duplex system is shown in Fig. 1. The car equipment is all controlled from a remote control unit which is mounted on the dashboard of the car. A handset is mounted on top of this unit and it is only necessary to remove this handset to put the car transmitter on the air. The removal of this handset automatically connects the earphone in the handset across the loud speaker. In the event that the operator wishes to listen to this earphone alone, a switch is provided on the loud speaker for turning it off and substituting a resistance load. The remote control unit also contains a calling tone button, a receiver volume control knob, a receiver vernier tuning knob, and two pilot lights, one for the transmitter and one for the receiver. As far as the operator is concerned, this unit is the only one which requires his attention. This remote control unit also has mounted on it a switch for changing the equipment from duplex operation for talking with headquarters to simplex operation for talking between cars. A push button is mounted on the handset for controlling the transmitter plate voltage and shorting the receiver output when simplex operation is used.

Although a good antenna can usually be constructed on police headquarters, the antennas on the car must of necessity be low and, therefore, relatively inefficient. The transmitting antenna usually takes the form of a vertical rod or flexible "fish pole" as long as practical. There has also been developed a bumper antenna for use on the

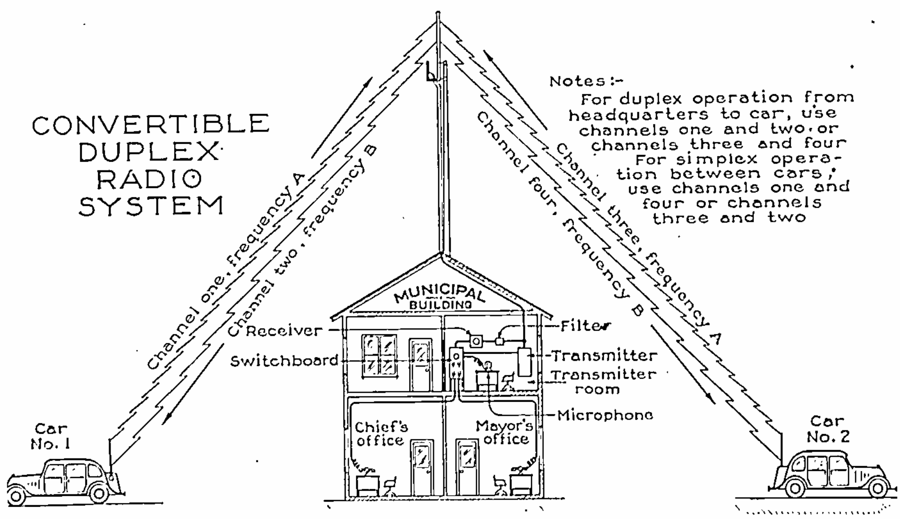

Fig. 1—Municipal police radio installation with duplex operation from headquarters to each car and simplex operation from car to car through the headquarters transmitter.

transmitter when the radio installation must be inconspicuous. This antenna looks exactly like the normal rear bumper and is insulated and fed at one end. Its efficiency as a radiator, is of course, not as great as that of a vertical radiator. The receiving antenna may be the usual roof antenna although, for greatest signal pickup, the transmitting antenna, when this is of the vertical type, should be used. This antenna, when properly connected to the car receiver, usually gives from two to five times the pickup that the roof antenna gives. It is a bit more complicated to use the same antenna for both transmitting and receiving, however. In a two-frequency duplex system, a very sharp filter must be used in the antenna lead to the receiver and in a single frequency simplex system, the antenna must be switched back and forth between transmitter and receiver. The filter used in the two-frequency duplex system must be very selective since the signal on the antenna from the car transmitter may be several million times stronger than the signal from headquarters to which the receiver is tuned. A filter has been developed for this purpose making use of a section of transmission line. The attenuation characteristic of this filter is such that it is possible to obtain satisfactory receiver operation on the same antenna which is being used to radiate the signals from a fifteen-watt transmitter only four per cent removed in frequency.

A car receiver used in this class of service is of the fixed tuned variety and is required to stand by on a given frequency waiting for calls. The time of receiving signals runs from approximately fifty per cent of the time on a busy day in a large city to approximately five per cent of the time in a small municipality giving half-hour routine calls. The receiver, then, must be on frequency when the call comes and for most efficient operation should be silent between calls. These two points are considered in more detail later in this paper.

The car transmitter for this class of service must be ready for instant operation and yet not draw any power from the car storage battery in the stand-by condition. Therefore, filament type tubes are more suitable than the indirect heater type with their longer warming-up period.

The Receivers

A police radio receiver must necessarily meet service demands that are unusually severe. The receiver must not only be of sturdy mechanical construction to withstand the vibration and shocks encountered in mobile service, but, also, should have an electrical design which incorporates features for meeting the adverse reception conditions that are frequently encountered.

Let us consider, for example, a patrol car that is operated in one of the larger cities. In such cases, the car may be operating in the midst of heavy traffic, surrounded by high buildings containing much steel work and, of course, in close proximity to various types of electrical systems, each of which acts as a potential source of interference to radio reception. To function satisfactorily in an area of this nature, the radio receiver should possess electrical characteristics which provide for (1) maximum rejection of electrical interference; (2) exceedingly rapid and effective automatic volume control, and (3) excellent sensitivity. Even in the small municipalities where reception conditions are, in general, better than those encountered in the larger metropolitan areas, it will contribute substantially to the over-all performance of the system if the car receivers possess the above electrical features.

For quite some time, radio engineers have known that both the superheterodyne and superregenerative principles of reception possess very desirable electrical characteristics. However, neither principle in itself appeared to afford all that was desirable in a receiver. In the development of a receiver for police service, a circuit was needed which would provide for the electrical performance mentioned above, and at the same time give stable and reliable operation. With these points as an objective, a receiver was developed which, in effect, incorporates the most desirable features of both the superheterodyne and superregenerative performance. In brief, the combination of a superregenerative second detector with the standard principle of superheterodyne action has made possible the production of a police radio receiver unit which possesses great sensitivity, almost entirely excludes interference from automobile ignition systems and other similar electrical devices and with an automatic volume control that is remarkably fast.

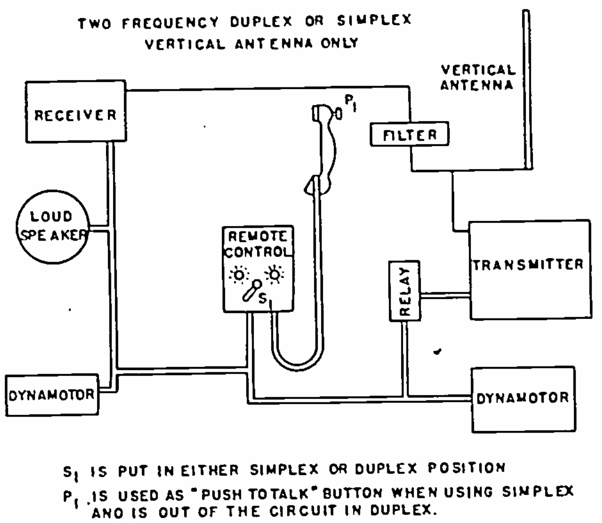

Fig. 2—Block diagram of patrol car installation with duplex operation from car to headquarters and simplex operation from car to car through the headquarters transmitter.

The advantages to be gained through the use of a receiver of this type will, it is believed, become more apparent through the points brought out in the following sections:

Noise Suppression

The interference to police radio reception which is occasioned by the operation of electrical machinery, appliances, electric trolley systems, and automobiles is a matter of great importance and, therefore, merits very careful consideration. It has been found that receivers which are quite satisfactory for reception in rural localities are practically unusable in metropolitan areas where many electrical devices are in use. Of the various forms of electrical interference that may be encountered from time to time perhaps that originating from automobile ignition systems is the most troublesome. Fig. 3 shows two oscillograms in which the effect on reception due to near-by automobile ignition is pictured. The lower oscillogram shows the audio output of an ordinary superheterodyne police receiver, while the upper oscillogram shows the audio output of a similar receiver using a combination of superheterodyne and superregenerative circuits. The vertical

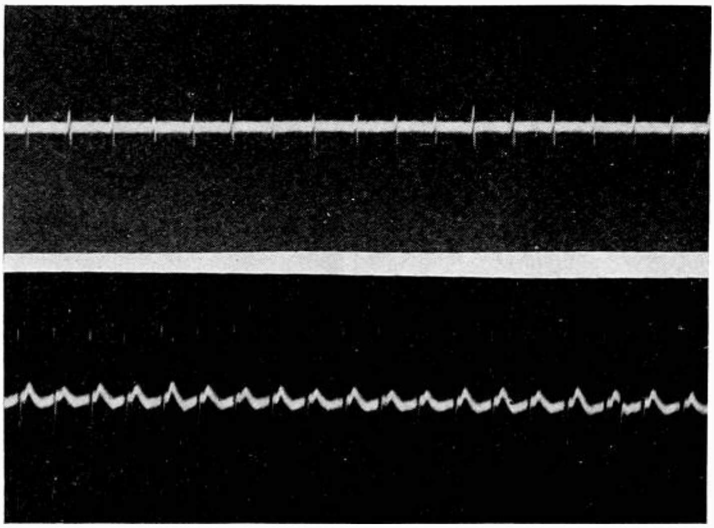

Fig. 3—Oscillograms showing the quenching effect on noise of the combined superheterodyne-superregenerative receiver.

lines which appear at regular intervals indicate the magnitude of interference originating from an automobile ignition system. A comparison of these two oscillograms very strikingly shows the quenching effect on this interference that is produced by the use of a superregenerative second detector.

Automatic Volume Control

At frequencies available for use by municipal police organizations, the signal strength of the transmitter will vary in intensity as the distance between the transmitter and receiver is increased and, also, further variations may be produced due to the presence of structures which absorb or reflect the signal. It follows, therefore, that a mobile receiver is called upon to meet field strength variations which quite frequently are found to undergo a 100:1 ratio within a linear distance of two or three feet. It is at once obvious that an extremely rapid automatic volume control is necessary in order to maintain the signal at the proper level whenever such a condition is encountered. Due to the inherent nature of a superregenerative circuit to produce automatic volume control, such a receiver as is here described will deliver a loud speaker output that is substantially constant under extreme conditions of antenna input variation.

Sensitivity

It is a well-established fact that a superregenerative circuit is capable of producing unusually high sensitivity. By combining the

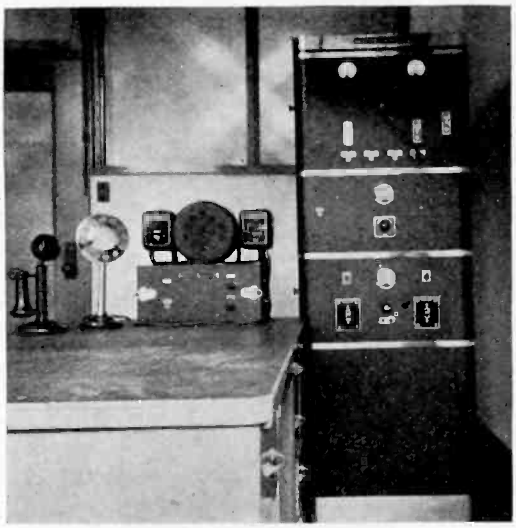

Fig. 4—75/150-watt 30- to 42-megacycle transmitter and superheterodyne-superregenerative receiver using two antennas installed in Hamilton, Ohio.

superregenerative feature with a superheterodyne, which in itself has good sensitivity, the resultant sensitivity of the arrangement is ample for meeting unusually strenuous service demands.

The combined principles that are utilized in the circuit of this receiver are such as to eliminate effectively the difficulties that have sometimes been encountered in the past in connection with superregenerative equipment. It has been possible to provide the salient electrical features outlined herein and, at the same time, maintain good audio quality that is substantially free from hiss, features that are not found in ordinary receivers employing the usual method of superregeneration.

The ability of the superregenerative detector to discriminate against ignition interference depends upon the relative duration of the oscillatory condition, and for best performance, the ratio of carrier frequency to quench frequency should be kept at 300 or greater.

In consideration of this fact, there are two paths open to the receiver development engineer; either to make a straight superregenerative receiver with its two disadvantages of poor selectivity and reradiation or to combine the superregenerative principle with the superheterodyne principle to get the advantages of both.

There are two methods of combining these two principles in a receiver. One way is first to convert to an intermediate frequency in

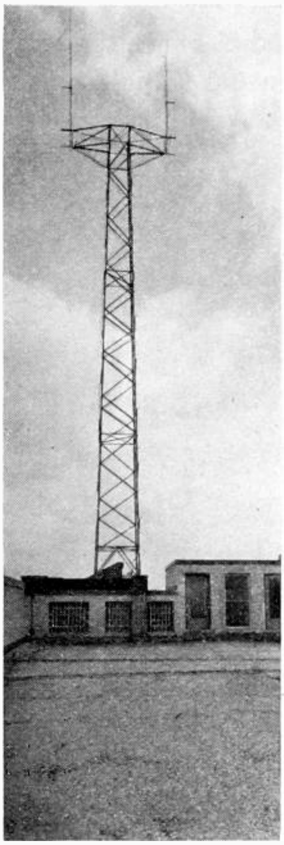

Fig. 5—The transmitting and receiving antennas installed on one tower in Hamilton, Ohio.

the neighborhood of 3000 kilocycles and, after getting selectivity at this frequency, convert again to a relatively high frequency, say 25,000 kilocycles, and then design the superregenerative detector to operate at this frequency. In this way the superregenerative detector operates at a frequency where good noise discrimination can be obtained and yet the selective circuits in the intermediate-frequency amplifier are tuned to a low enough frequency so that excellent selectivity can be obtained. Another method of combining the two principles into a single receiver is to convert to an intermediate frequency which is approximately 300 times the quench frequency. For example, if a quench frequency of thirty kilocycles is used, design the superheterodyne converter to give an intermediate frequency of 9000 kilocycles and then design the superregenerative detector to work directly at the output of the intermediate-frequency amplifier. This method makes a much simpler receiver for police use and at the same time gives ample selectivity and noise discrimination.

Although it is possible to adjust the superregenerative detector in the laboratory so that the hiss is not objectionable, it is impossible to hold this adjustment over long periods of time in police service under conditions of varying supply voltage, temperature, and humidity. Also, the best adjustment for sensitivity and noise elimination does give some hiss. Of course, this hiss disappears when a carrier is present at the receiver but, in police service, the receiver is standing by at maximum sensitivity a great deal of the time waiting for a signal to appear. It is, then, very desirable to incorporate in this receiver, as in any police receiver, a carrier-off noise elimination circuit. In this receiver, the superregenerative detector is adjusted well into the hiss area and then a carrier-off noise suppressor circuit is used to cut off this hiss when the headquarters transmitter is off the air. Of course, the presence of the headquarters carrier which opens this noise suppressor circuit also causes the superregenerative detector to stop hissing and so, in a properly operating system, the hiss is seldom audible within the reliable service area.

The adjustment of this noise suppressor circuit can be varied in the service shop and is set for the conditions to be encountered. A receiver which is to work in areas of high signal strength and extreme noise would be adjusted to a less sensitive condition than one to operate in rural areas of weak signal strength and low noise.

The phenomenon used to obtain a carrier-off noise suppressor is that the plate current of the superregenerative detector decreases with the presence of a signal on its grid.

A schematic diagram of the superregenerative circuit and the noise suppressor is shown in Fig. 6. It can be seen from this schematic that the direct voltage between point A and ground will increase with the presence of a signal on the grid of the superregenerative tube since its plate current will decrease. N_1 is a neon glow tube which will have a constant voltage drop across it when it is glowing regardless of the current through the tube. All of the voltage change at point A is then transferred to resistor R_{01}. This voltage change is used in such a way and is of sufficient amplitude to cause the high-mu audio amplifier tube V_5 to be cut off in the no-signal condition and to work at normal bias when a signal of two microvolts or more is present at the input of the receiver. The small copper oxide rectifier, RTX_1, is connected as shown to prevent the bias on V_5 from becoming more positive than

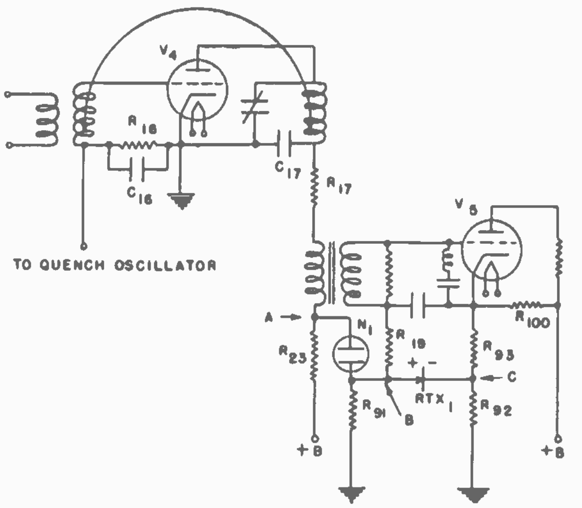

Fig. 6—Schematic diagram of superregenerative circuit and carrier-off noise suppression circuit used in model 4SH2A1 receivers.

its normal operating bias when high signal levels are present at the receiver input terminals.

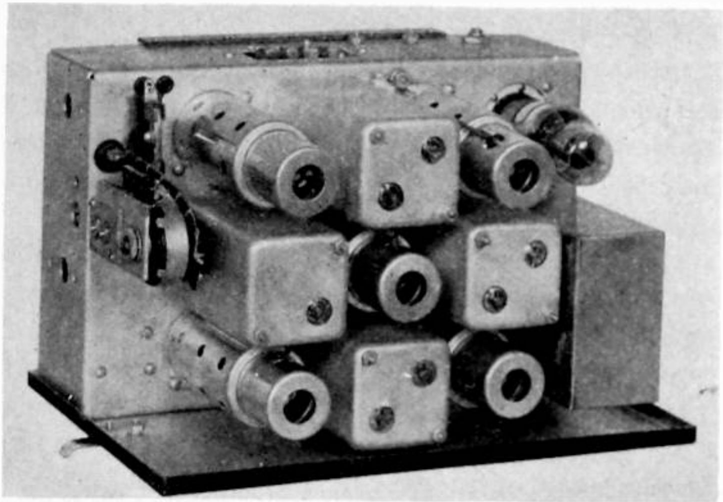

Fig. 7—Superheterodyne-superregenerative car and station house receiver chassis model 4SH2A1.

In order to take care of variations in signal strength in excess of that which the superregenerative detector can take care of alone, the conventional type of automatic volume control is used also. This makes the output response substantially flat over wide variations of signal strength. Fig. 8 shows the automatic volume control curve of the 4SH2A1 receiver.

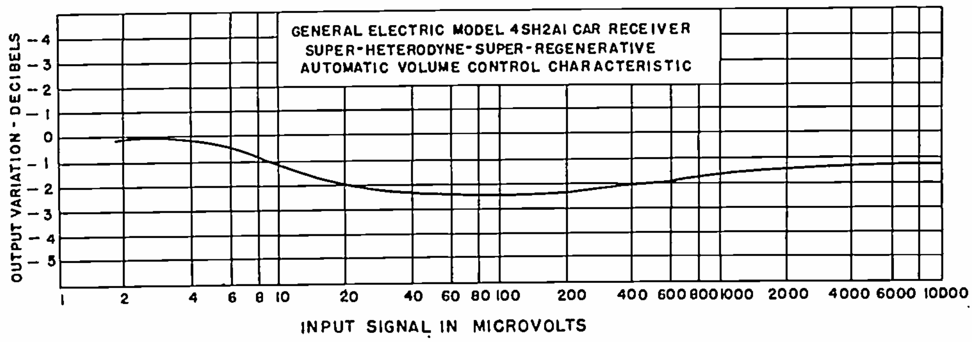

Fig. 8

The police car receivers here described employ a dynamic type loud speaker capable of giving excellent reproduction of the voice frequencies. The over-all fidelity of the receiver is shown in Fig. 9.

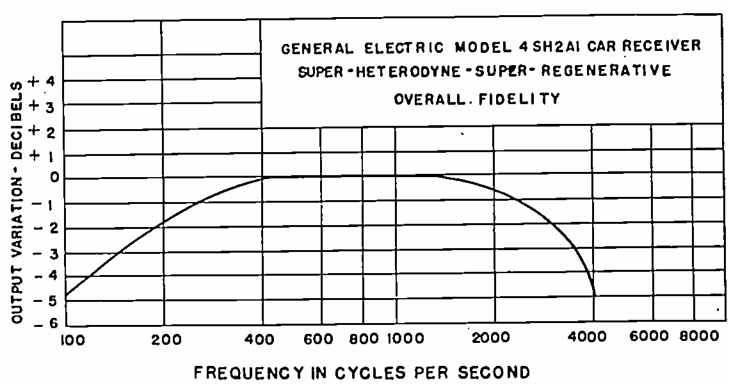

Fig. 9

Another important feature in police radio equipment is that of making the equipment so that it can be quickly repaired or replaced. These receivers are held in their cases by snap catches and the same receiver chassis will fit into either a car or headquarters installation. This simplifies the stocking of spares and makes it unnecessary for the serviceman to become familiar with more than one chassis.

The Car Transmitter

The car transmitter for municipal police service also has imposed upon it a unique set of adverse conditions. It may be required to operate at a temperature of minus forty degrees centigrade or of plus fifty degrees centigrade. It is required to do this under conditions of extreme shock and vibration day after day without attention. It must usually be placed at the rear of the car in the trunk or under the rear deck to be near the antenna. This is the point of greatest throw when

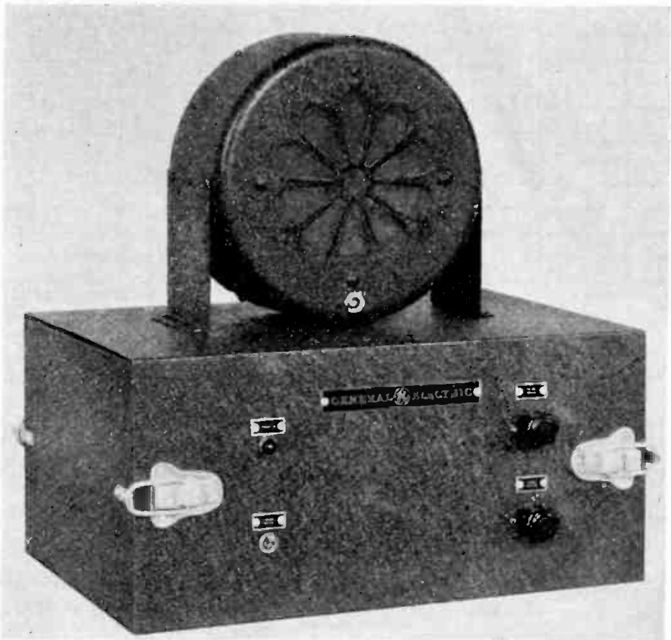

Fig. 10—Superheterodyne-superregenerative station house receiver model 4SH3A1 for municipal police use, 30 to 42 megacycles.

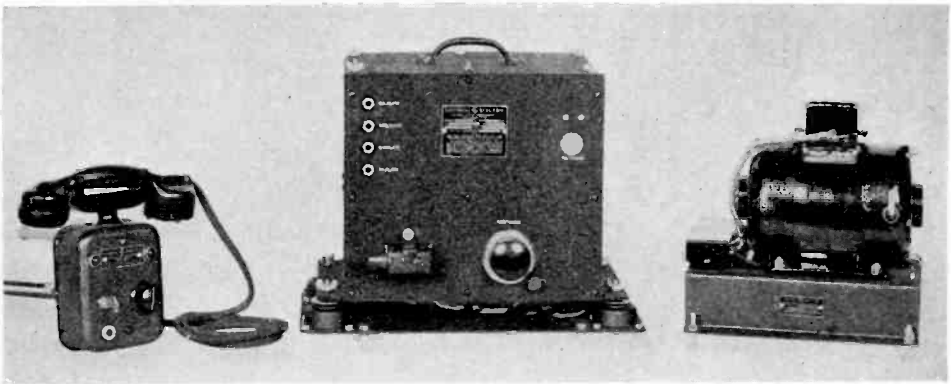

Fig. 11—Control unit, 15-watt 30- to 42-megacycle car transmitter model 4G1A1 and six-volt dynamotor unit used in municipal police patrol car installation.

riding over rough roads. The transmitter should take no continuous drain from the storage battery and yet be ready for service in a time not exceeding two or three seconds after turning it on. The transmitter is thus not given any warming-up period and must be on frequency every time it is turned on. Its frequency stability is of prime importance since the operator at headquarters does not have time to tune his receiver for every call nor does he know when a call is about to come in. In order not to require a special storage battery, the car transmitter must be as efficient as possible in order to insure maximum power output with minimum power input. Frequency stability, ruggedness, and efficiency cannot be overstressed in this class of equipment.

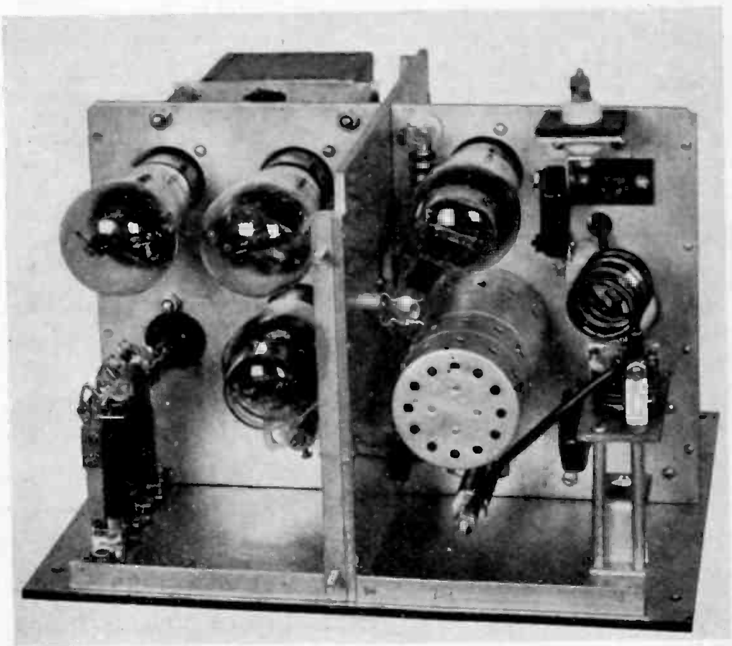

Fig. 12—The 15-watt car transmitter model 4G1A1 removed from its case.

The transmitter which has been developed to operate under these conditions makes use of five tubes. A type 47 tube is used to amplify the voltage from a double button carbon microphone. This 47 tube drives two 46's in class B push-pull. These 46's modulate a 2A3 power amplifier which is excited by a 2A3 master oscillator. This tube complement will give fifteen watts output 100 per cent modulated. It uses a specially designed master oscillator with very high frequency stability. This power output is made possible by the proper choice of high quality component parts and low loss connections throughout.

Only low loss insulation is used in the radio-frequency circuits of this transmitter and both the master oscillator and power amplifier coils are made of silver-plated copper tubing to insure low loss and good contact at the loading tap on the power amplifier coil. All tank connections in the master oscillator and power amplifier circuits are made with heavy cadmium-plated copper strips.

The transmitter is of extremely rugged construction as shown in Fig. 12. It is laid out in the form of a cross with the speech amplifier and modulator tubes in one section, the iron-core devices in the second section, the master oscillator parts in the third section, and the power amplifier parts in the fourth section where they are accessible for connection to the antenna.

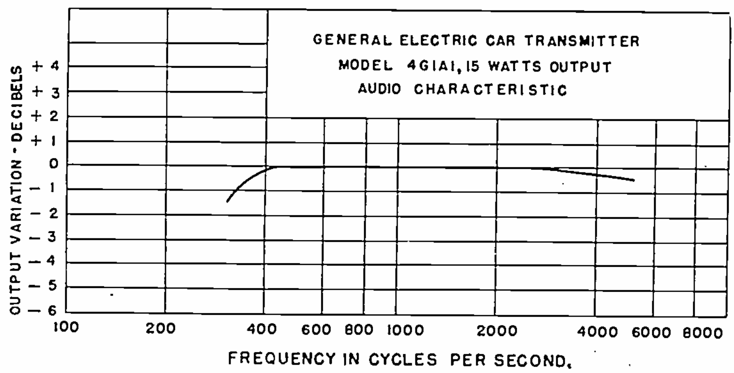

Fig. 13

The transmitter has four test jacks on its panel for checking its adjustment during operation. These jacks permit checking the plate current of the speech amplifier, the modulator, the master oscillator, and the power amplifier. Thus, it is possible to check rapidly the operation of the tubes in the transmitter and to adjust neutralization and loading without disturbing the installation of the equipment in any way. By talking into the microphone or pressing the tone button, the modulation can be checked. These jacks are indispensable in servicing or adjusting the transmitter.

The frequency characteristic of this transmitter is shown in Fig. 13. Full advantage is taken of this characteristic by the use of a double button carbon microphone which is mounted in a handset.

The model 4MA1A1 frequency monitor shown in Fig. 14 has been designed for checking and setting the frequency of the car transmitter.

This frequency monitor uses a low temperature coefficient crystal and the crystal oscillator coil is an integral part of the crystal plug-in unit, thus making it extremely easy to change frequency.

The Headquarters Transmitter

The power output of the transmitter at headquarters, of course, depends upon the service area to be covered. It is not the purpose of this discussion to set up coverage limits versus transmitter power output. This problem has been covered in The National Electrical Manufacturers Association Publication No. 35–27 dated August, 1935. This is largely a function of local conditions, both natural and man-made, of the antenna height and of receiver sensitivity.

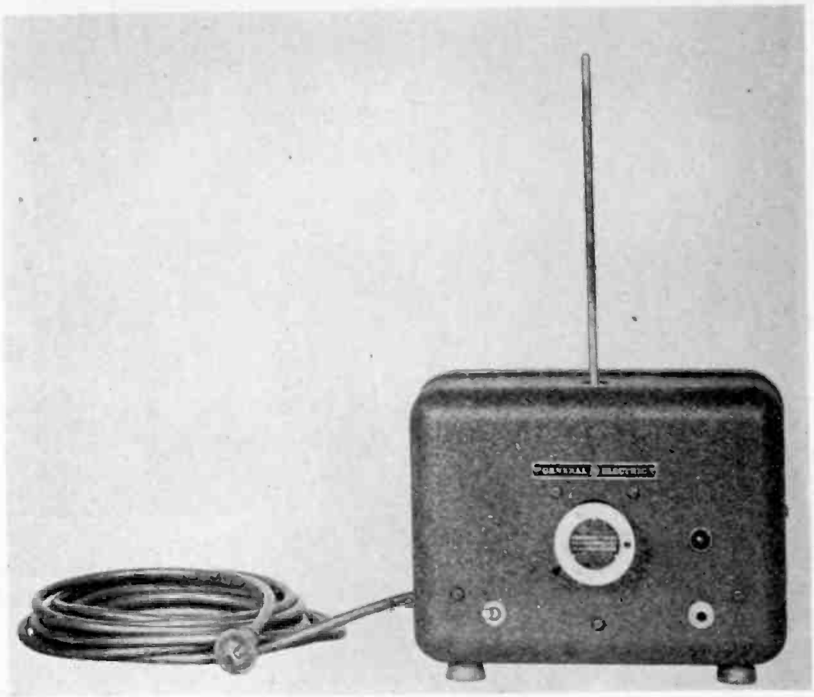

Fig. 14—Crystal controlled frequency monitor, 30 to 42 megacycles, for checking frequency settings.

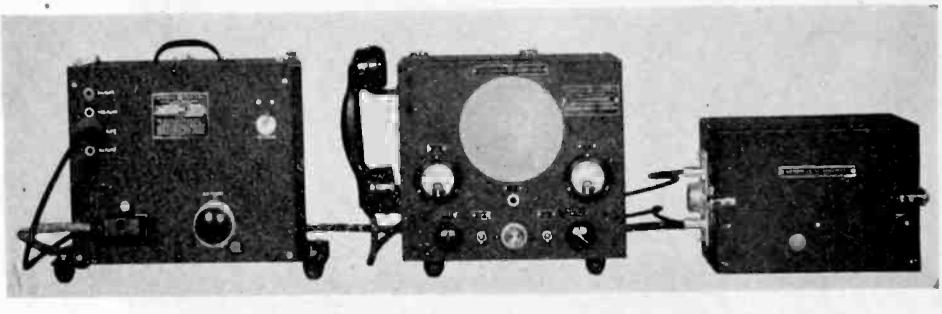

Fig. 15—General Electric 30- to 42-megacycle two-way radio communication equipment for police station house.

The present tendency is for the municipal police assignments to be in the thirty to forty-two megacycle band and, therefore, there has been developed a line of transmitters for use in this band, ranging in output power from fifteen watts to 1.5 kilowatts.

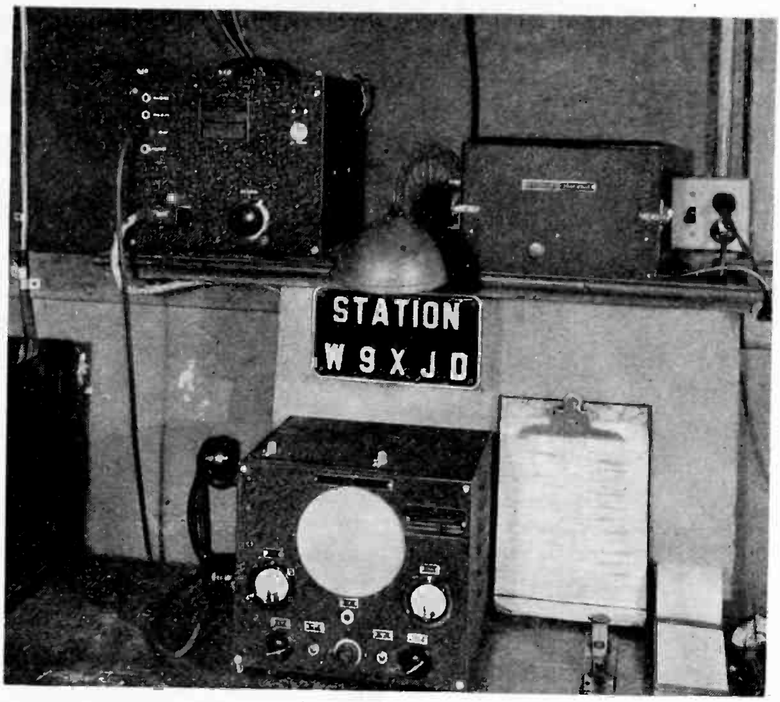

Fig. 16—Fifteen-watt 30- to 42-megacycle transmitter and superheterodyne-superregenerative receiver working from a common power supply and on a single antenna installed in Whiting, Indiana.

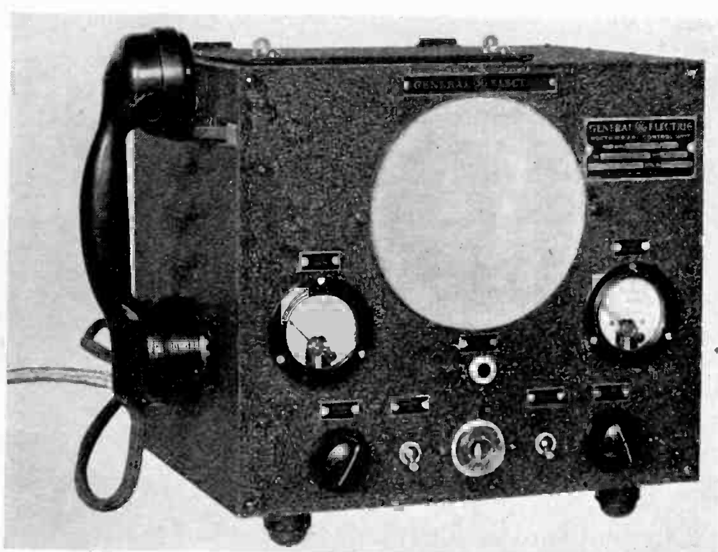

Fig. 17—Rectifier and control unit model 4MR1A1 for 15-watt 30- to 42-megacycle transmitter and receiver used in two-way police radiotelephone communication system.

The fifteen-watt headquarters transmitter is identical in every respect both mechanically and electrically with the fifteen-watt car transmitter. When operated as a headquarters transmitter, it is, of course, alternating-current operated using the same rectifier power supply for both transmitter and receiver. This single power supply and interchangeability of transmitter chassis and receiver chassis with those in the car installation greatly simplifies the problem of servicing

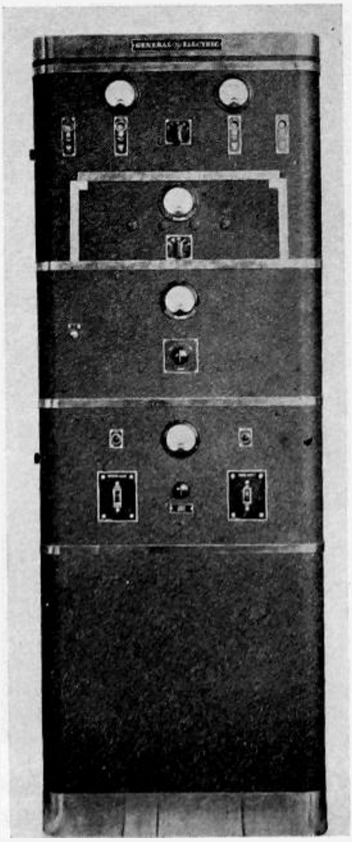

Fig. 18—75/150-watt 30- to 42-megacycle radiotelephone transmitter model 4G2C1 for municipal police use.

and stocking of spares. Such a headquarters equipment is shown in Fig. 15.

The 75/150-watt transmitter for 110- or 220-volt sixty-cycle operation has many unique features. It is designed for either local or remote control and for operation with either an open-wire or a concentric tube type of transmission line. The unit is of sectional design with rounded corners, black crackle finish, and stainless steel trim. The meters and controls are logically and conveniently arranged. This transmitter is shown in Figs. 18 and 19.

The rectifier unit contains two complete rectifiers; one to furnish voltage to all of the high voltage vacuum tubes and the other to provide for the low voltage tubes. Both of these rectifiers are of the mercury-vapor type. The high voltage rectifier tubes are protected by a time delay relay which prevents voltage from being applied to them before the filaments have reached their correct operating temperature.

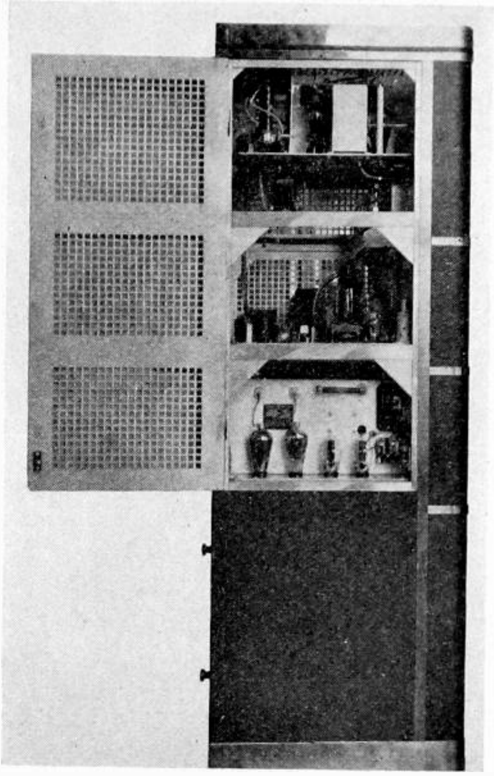

Fig. 19—75/150-watt 30- to 42-megacycle radiotelephone transmitter model 4G2C1 with side shields removed.

The transmitter is crystal controlled using a special low temperature-frequency coefficient quartz crystal which is practically unaffected by ordinary room temperature changes. This crystal operates in the band five to seven megacycles and the output of the transmitter in the band thirty to forty-two megacycles is within plus or minus 0.02 per cent of the assigned frequency for all normal temperatures encountered in a police headquarters installation.

The crystal oscillator is a type 802 pentode driving an 802 frequency tripler which in turn drives an 802 frequency doubler. Another 802 is used as the first intermediate power amplifier which drives the second intermediate power amplifier consisting of two 800's in push-pull.

This second intermediate power amplifier drives the power amplifier which consists of two 800's in push-pull in the seventy-five watt equipment and four 800's in push-pull parallel in the 150-watt equipment. A meter is located on the panel of the power amplifier compartment which can be used for monitoring the power amplifier plate and grid currents and the intermediate power amplifier plate current by means of a rotary selector switch. Operating personnel is positively protected from high voltage by means of door interlocks. A large storage compartment free from live wires is provided for the storage of spare tubes, etc.

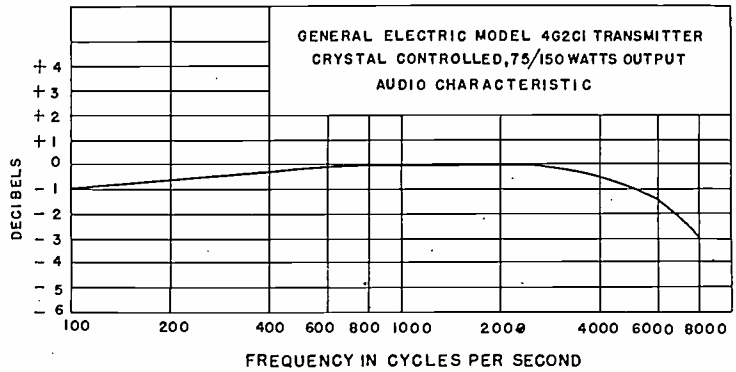

Fig. 20

This transmitter is capable of 100 per cent modulation either by voice or by a 1000-cycle calling tone. When modulating by voice, the output of the double button carbon microphone is amplified by a type 57 tube, the output of which drives a type 56. When modulating by tone, this type 56 tube is driven by another type 56 tube in an audio oscillator circuit. The output of the 56 drives two 2A3's operating push-pull class AB. These in turn drive two FP-146's working class B which fully modulate the radio-frequency power amplifier with a total harmonic distortion not exceeding nine per cent. The frequency characteristic of this transmitter is shown in Fig. 20.

Ultra-High-Frequency Antenna System

An end-fed vertical half-wave radiator has several advantages over other types of antennas for municipal police use. Among them may be mentioned the symmetrical horizontal field pattern and the amenability to installation on top of a tower, building, water tower, or other structure. In order to utilize a concentric transmission line, which inherently has a low surge impedance, to excite the half-wave radiator from the end, a quarter-wave impedance matching section is resorted to. To obtain the utmost efficiency of energy transfer, a method of connection has been developed which reduces the standing waves on the transmission line to a negligible value.

Fig. 21 shows the arrangements of the antenna system. The quarter-wave impedance matching section is short-circuited at the bottom and the grounded outer transmission line conductor connects to the short-circuiting bar. In as much as this point is at practically zero radio-frequency potential, the antenna system may be here connected to a

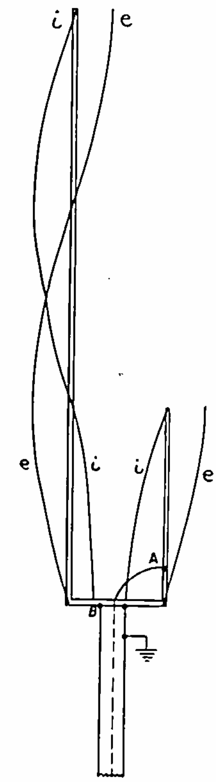

Fig. 21

grounded supporting structure, thus affording an easy method of mounting, and at the same time obtaining lightning protection.

It will be noted that the half-wave radiator is an extension of one of the quarter-wave conductors so that the antenna load is not reflected symmetrically to the quarter-wave transformer. By connecting the transmission line inner conductor to a correctly chosen point, A, and the outer conductor to point B, and by *properly dimensioning* the entire system, the quarter-wave section can be balanced, and the transmission line terminated in its surge impedance.

The concentric transmission lines are of usual construction utilizing isolantite beads to maintain the inner conductor coaxial. It is interesting to note that the three-eighth-inch line, which would have a surge impedance of seventy-two ohms if no beads were present (optimum ratio of diameters for minimum losses), actually has a surge impedance of forty-nine ohms due to the increased capacity per unit length. Thus, when transmitting an unmodulated carrier of fifteen watts, the radio-frequency voltage existing on the line is only 27.2 volts. The advantages of this antenna system may be briefly summarized as follows:

1. Properly matched throughout, thus obtaining maximum energy transfer to the antenna.

2. No feeder wires, guy wires, or other structures directly in the field of the antenna to alter its radiation pattern.

3. Entire antenna system at direct-current ground potential thus affording lightning protection.

4. Unaffected by the elements as are two-wire line antenna systems.

5. Adaptability of installation.

Concentric Band Elimination Filter

To obtain two-frequency duplex operation for mobile units, a vertical rod attached to the rear of the car has usually been used for transmitting while the horizontal antenna in the roof of the car has been used for receiving. It was soon recognized that this arrangement was not ideal because of the vertical polarization of the transmitted wave from the fixed station. This together with the advent of the steel-topped car, spurred on the development of a band elimination filter which would permit duplex operation with a single rod antenna.

Those familiar with high-frequency filter design are cognizant of the difficulties encountered and the practical impossibility of using lumped constants to obtain a transducer with sharply defined filter characteristics at ultra-high frequencies. This, of course, is due in part to the high resistance of an inductance and the low inductive reactance obtainable because of stray capacities, and results in a low circuit Q. This, in turn, results in poor filter characteristics, lack of selectivity, and attenuation in the pass bands comparable to that in the attenuated bands.

The concentric band elimination filter exhibits a high Q, a sharply defined impedance versus frequency characteristic with a high maximum impedance at the elimination frequency, and a pass characteristic at frequencies closely adjacent to that to be attenuated. Figs. 22 and 23 show the two fundamental arrangements. Observe that in Fig.

22, the free end is open-circuited, while in Fig. 23 the free end is short-circuited.

The concentric transmission line shown in Fig. 22 has an electrical length equal to one wave length at the frequency to be attenuated. The input is at the left end and the output is tapped off either at a point one-quarter wave length from the free end or at a point three-quarters wave length from the free end. In general, the open-ended type may be any number of half wave lengths long at the frequency to be attenuated and the output tap may be made at any point which is an odd number of quarter-wave lengths from the free end.

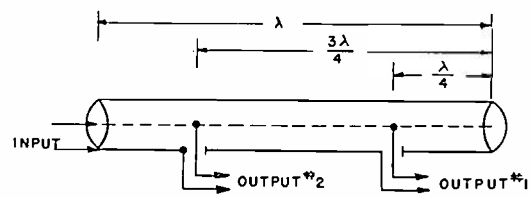

Fig. 22

The line shown in Fig. 23 is of the closed-end type; that is, at the free end the inner conductor is short-circuited to the outer conductor. In this type, the electrical length may be any number of odd quarter-wave lengths, three or greater, and the output tap may be made at any point which is any number of half-wave lengths from the short-circuited end.

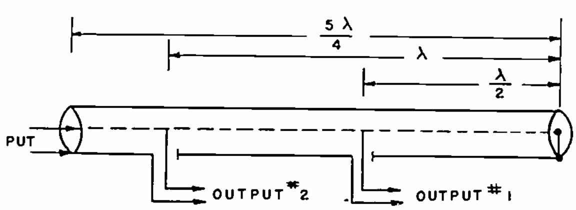

Fig. 23

When the outer transmission line conductor is grounded and the filter excited by applying a voltage whose frequency is f_t, a standing wave will exist on the inner conductor. For purposes of discussion and explanation of the band elimination characteristic, we need consider only the inner conductor and the standing waves of voltage thereon.

Fig. 24 shows three standing voltage waves in an open-ended line whose electrical length is one wave length at frequency f_t. Frequency f_h is higher than f_t and f_L is lower than f_t. It is apparent, from Fig. 24 and the general theory of standing waves on wires, that at the quarter-wave and at the three-quarter wave points practically no voltage exists at frequency f_t. However, at frequency f_h or f_L, a voltage does exist at the output points. At the quarter-wave point, the ratio of output to input voltage at frequency f_h is e_7/e_3, at frequency f_L is e_5/e_2.

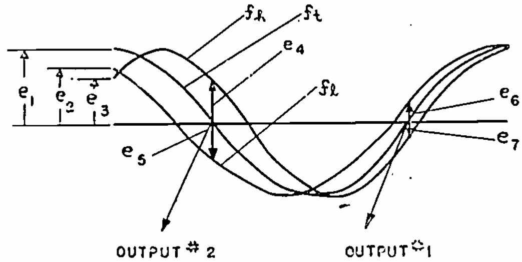

Fig. 24

The analysis of the closed-end type is similar to that of the open-end type, the difference being that at the free end of the closed type, the voltage is always zero instead of a maximum as in the open-end types. Fig. 25 shows the standing wave voltage plot for the closed-end type five-fourths wave length long.

It will be noted that a multiplicity of possible output points exists in each type of filter dependent upon the electrical length of the filter. At each of these output points, the voltage to be attenuated, at frequency f_t, is the same practically zero value. However, at frequencies other than f_t, the ratio of output to input voltage will, in general, be different at each output point.

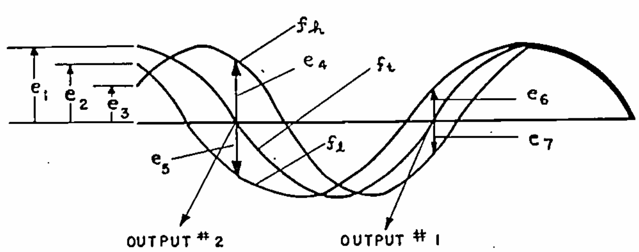

Fig. 25

Incidentally, the use of the filter is not limited to mobile units. A number of headquarters installations have been made wherein a filter and single antenna system have performed very satisfactorily in two-frequency duplex operation.

The operational details of the filter may be summarized as follows:

1. The filter is connected between the transmitting antenna and the receiver and operates in the thirty- to forty-two-megacycle band.

2. It is designed to attenuate the transmitted frequency.

3. Satisfactory operation of the receiver in conjunction with a fifteen-watt transmitter at frequencies to within four per cent of the transmitted frequency can be achieved.

short-wave range. It includes descriptions of a signal generator for the 20–200-centimeter wave-length range, thermocouples for the measurement of power, very small diodes for the measurement of voltage, and a discussion of their uses and limitations. Because of the great utility of voltage measurement in the study of oscillators and amplifiers, particular emphasis has been placed on this phase of the work.

A brief examination of the differences between circuits and circuit elements at the long wave lengths and at the ultra-short wave lengths will aid in an understanding of the motives which underly the design of the devices to be described later in the paper. At the longer wave lengths the physical dimensions of the circuits and devices used are small compared with the wave length. This permits the use of lumped circuits, such as coils for inductances and condensers for capacities, the connecting leads making almost negligible contributions to the constants of the circuits. The radiation resistance of such circuits is negligible, so it is not necessary to shield for radiation fields. Further, since the physical length of any branch of a circuit is small compared to the wave length, the current is uniform through the branch. Finally in electronic devices such as vacuum tubes the time of transit of electrons between electrodes is negligible compared to the period of the alternating components of the applied voltages.

This is not the situation at the ultra-short waves. The use of lumped circuits is not convenient since their small physical size renders them difficult to construct and use; and when they are used, the connecting leads contribute a large part, in some cases the major part, to the resulting circuit constants. Because the physical dimensions are comparable with the wave length in this wave-length range, the current distribution is constant throughout a branch of a circuit only if the branch is very short, and the radiation resistance may be an appreciable part of the total circuit resistance. The ohmic resistance itself can differ greatly from the direct-current value because of skin effect. Conventional vacuum tubes also become subject to various disturbing effects. Aside from the reductions in electrode impedances due to the interelectrode capacities at ultra-high frequencies, there are other effects. These are electron loading—the absorption of energy by the electrons from the input electrode, and a slight shift in the interelectrode capacities with frequency due to the high-frequency behavior of the electron space charge,[1,2] and in some cases a complete failure of the

[1] W. R. Ferris, "Input resistance of vacuum tubes as ultra-high-frequency amplifiers," Proc. I.R.E., vol. 24, p. 82; January, (1936).

[2] D. O. North, "Analysis of the effects of space charge on grid impedance," Proc. I.R.E., vol. 24, p. 108; January, (1936). (This paper has a good bibliography of the subject.)

tube to respond to the signal. The last of these effects will be described later in connection with diode voltmeters.

Circuits

For tuned circuits, resonant transmission lines have been used exclusively in this work. In some cases transmission lines have been used as inductances and capacities. Such circuits are convenient in size and can be made very good electrically. Two types of transmission lines have been used; the parallel wire, or Lecher type, and the concentric cylinder type. Of the two, the second has been preferred because it is self-shielding and has very low radiation resistance.[3] Except in the cases where the lines have been loaded at the receiving end, the sending end impedance has been high compared with the impedance across which the line was connected. In general, it can be said that transmission lines can be made to perform as well at short wave lengths as lumped tuned circuits at long wave-lengths.

For the measurement of wave length, a Lecher system has been used exclusively.[4]

The Signal Generator

A good signal generator is essential to measurements. The requirements of such a generator are:

1. It must deliver sufficient output to drive the devices to be studied over the required wave-length range.
2. It must be stable so that neither its frequency nor output varies with time while measurements are in progress.
3. It must be thoroughly shielded so that its output appears at the output terminals and not about the laboratory in the form of stray fields.
4. Its output must be free from harmonics.

These requirements are reasonably well met in the generator now to be described.

The split-anode magnetron was selected as an oscillator,[5,6] since it affords the simplest circuit arrangement, requiring only a tuned circuit between anodes. By proper design a single circuit can be made to cover the entire wave-length range. The associated circuits consist of:

[3] Robert Whitmer, "Radiation resistance of concentric conductor transmission lines," Proc. I.R.E., vol. 21, p. 1343; September, (1933).

[4] August Hund, "Theory of determination of ultra-radio frequencies by standing waves on wires," *Sci. Pap. Bur. Stand.*, no. 491.

[5] G. R. Kilgore, "Magnetostatic oscillators for generations of ultra-short waves," Proc. I.R.E., vol. 20, p. 1741; November, (1932).

[6] E. C. S. Megaw, "An investigation of the magnetron short-wave oscillator," *Jour. I.E.E.*, vol. 72, no. 346, p. 313; April, (1933).

1. The primary oscillating circuit which is a parallel wire transmission line short-circuited at the end and adjustable in length. The spacing between wires of the line and its position are such as to make the parallel wires and the tube leads a nearly uniform transmission line.

2. The secondary circuit which is also a parallel wire transmission line of variable length. This line can also be moved as a unit so that it overlaps the primary line by an amount determined by the degree of coupling desired between primary and secondary.

3. A supply circuit for the electromagnet providing the field for the magnetron.

4. A filament supply circuit for the magnetron. Because it was desirable in the interest of the operator's safety to operate the tube with the anodes and transmission lines at ground potential, this circuit was designed to withstand voltages in excess of 3000 volts.

5. The anode supply circuit. This consists of a 2000-volt rectifier and a vacuum tube voltage regulator which holds the voltage output of the unit constant within a few per cent for changes of load from ten to 100 milliamperes and for input voltage variations of twenty per cent.[7] The regulator contributes much to the stability of the oscillator. The circuit of the regulator, which was suggested by F. H. Shepard, Jr., of this company, is shown diagramatically in Fig. 1.

The tube and lines are placed along the axis of a series of concentric brass cylinders in sliding contact, the cylinders serving as a shield. (See Fig. 2.) The first of these cylinders supports the tube; the second, which slides within the first, supports the end of the primary line. The third cylinder, which slides within the second, supports the end of the secondary line while the fourth, which in turn slides within the third, supports the output terminals of the secondary line. These cylinders are moved by suitable lead screws. It is possible to make any one of the three necessary adjustments without changing either of the other two. The entire shield may be rotated about its axis by means of a worm and gear to orient the tube properly in the magnetic field.

For wave lengths above fifty centimeters the magnetron is operated as a "negative resistance" oscillator, the frequency being determined solely by the tuned circuit.[6] The primary line is then approximately a quarter-wave length long. For wave lengths below fifty centimeters, the magnetron is operated as an electronic oscillator, the frequency being determined by the magnetic field strength, and to a small extent by the length of the primary line. The primary line is then approxi-

[7] T. H. Johnson and J. C. Street, "The use of a thermionic tetrode for voltage control," *Jour. Frank. Inst.*, vol. 214, p. 155, (1932).

mately a half-wave length long. This difference in primary length for the two modes of operation considerably reduces the required amount of variation in primary length and somewhat simplifies the design.

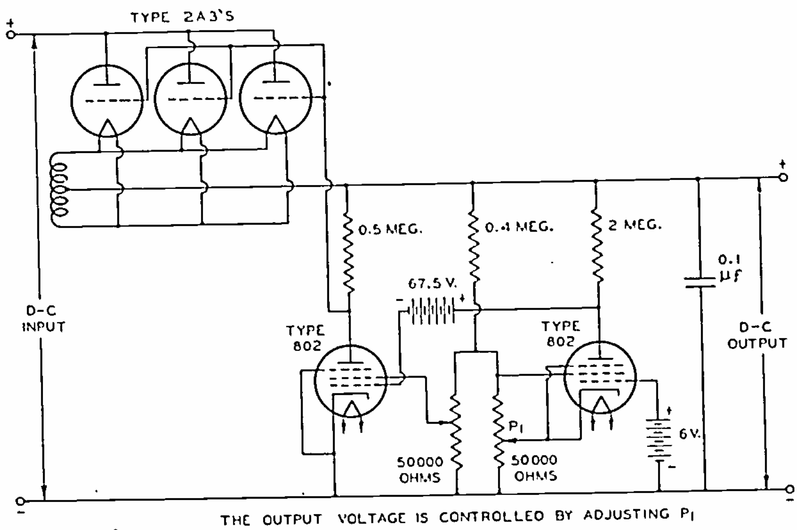

Fig. 1—Circuit of the voltage regulator.

A capacity attenuator has been built to provide known voltage ratios. This attenuator is of a capacity type and consists of two parallel, circular, condenser plates enclosed in a cylinder. The spacing between the condenser plates can be varied by means of a micrometer screw.

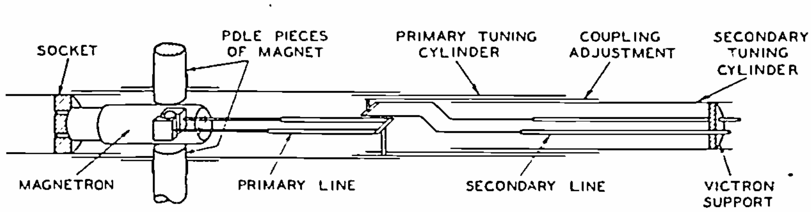

Fig. 2—Schematic diagram of the signal generator.

The radius of the cylinder and the pitch of the micrometer thread were so chosen that the voltage is changed by a factor of ten for ten turns of the micrometer head. The dimensions have been made small so that sufficient attenuation can be obtained before the spacing between condenser plates approaches a tenth of a wave length. This prevents errors due to the radiation field, which attenuates less rapidly than the induction field. The attenuator has been calibrated at a wave length of 112 centimeters and the attenuation has been found to agree very well with the calculated attenuation.

Fig. 3 shows the tube end of the oscillator and the electromagnet. This end of the oscillator is enclosed by an interlocked cage to protect the operator from high voltage. Fig. 4 shows the other end of the oscillator with the attenuator in place.

The Measurement of Power

To measure power, vacuum thermocouples have been built. These thermocouples cover the power range from one-tenth milliwatt to fifty

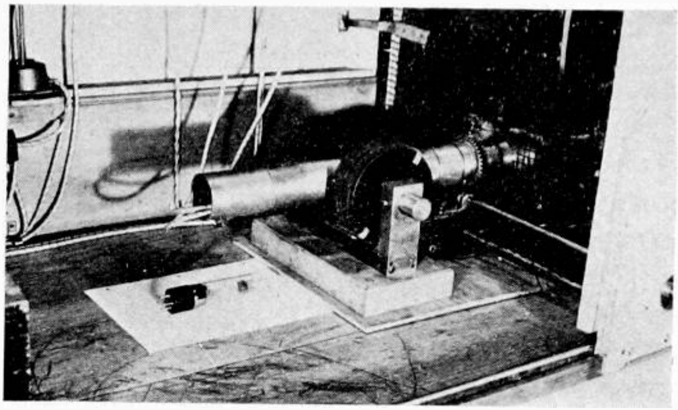

Fig. 3—The tube end of the signal generator. The tube shown is a magnetron of the type used in the generator.

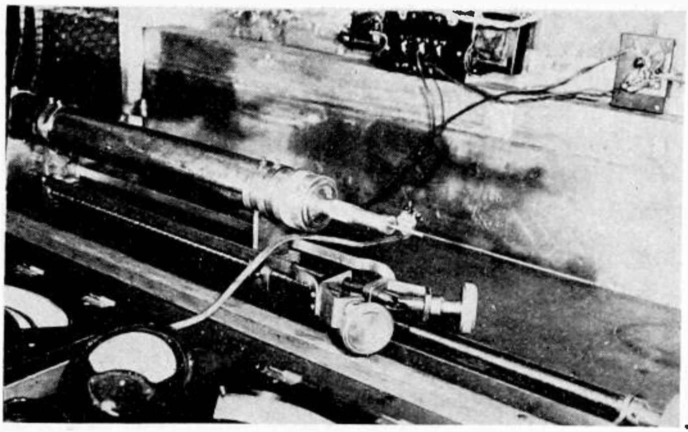

Fig. 4—The output end of the signal generator with the attenuator in place.

watts. The requirements of a satisfactory thermocouple are:

1. It should have adequate sensitivity.

2. The heater should be short and straight so that the current will be uniform throughout the length of the heater, permitting the use of a low-frequency calibration.

3. Its resistance should be large compared to the circuit resistance. If this is not the case, the thermocouple will not read the total power

but only that part of the power actually dissipated in the heater.

4. The junction and its leads should be arranged so that there is no inductive coupling between the heater and junction circuits.

5. The heater leads should be short and the mount free from beads and spacers, to make the effective parallel susceptance of the heater circuit a minimum. If these requirements are not satisfied, it may not be possible to get power into the heater.

The first requirement presents some difficulty in the low power range; none in the high power range. The second and third can be met in the low power range where the power to be dissipated is small. However, in the high power range it becomes impossible to meet both the second and third requirements. If the heater is made short, it must necessarily be of small diameter to keep the resistance up to a reasonable value, and as a result it will not stand the temperature rise. This difficulty has been partially overcome by the use of an indirectly heated couple. The heater is enclosed in a metal shield of good thermal conductivity; the thermal junction is attached to the shield. Then, whatever the distribution of temperature on the heater, the couple will read the total power dissipated within the shield.

The following descriptions of thermocouples as built will serve to illustrate the compromises made in design:

1. Laboratory thermocouple R-446.

This thermocouple has a straight heater of 0.3-mil carbon wire, 8 millimeters long; a couple of 0.5-mil gold-palladium wire and 0.5-mil iron wire, the junction being fastened to the heater with Aquadag. The tungsten leads supporting the heater and couple are held rigid by a small Nonex bead, a hard glass being used to reduce high-frequency dielectric losses. The couple wires are placed in a plane normal to the heater wire to minimize the inductive coupling between the heater and the junction. The couple leads are brought out of the "acorn" type bulb close together and well separated from the heater leads to reduce coupling. The heater resistance is about 1500 ohms. This resistance is not constant, nor is that of any thermocouple to be described, because of the temperature coefficient of the wire. This thermocouple covers the power range from 0.1 to twenty milliwatts.

2. Laboratory thermocouple R-253.

This thermocouple has a straight tungsten heater,[8] 8 millimeters long, 0.7 mil in diameter; and an iron-nickel couple, the couple wires

[8] The use of the tungsten heater was suggested by P. D. Zottu of this laboratory.

being 0.7 mil in diameter. In this couple it was found possible to dispense with the glass bead, the leads being supported solely by the radial seal of the "acorn" type bulb. The heater leads and heater wire lie on a straight line, an ideal fulfillment of the first and fifth requirements. The couple is arranged in an acute V, the leads being brought out close spaced at ninety degrees to the heater to reduce inductive coupling between the heater and the couple. This thermocouple is particularly useful for low power measurements. It has a long useful working range and, because of the tungsten heater, is able to stand

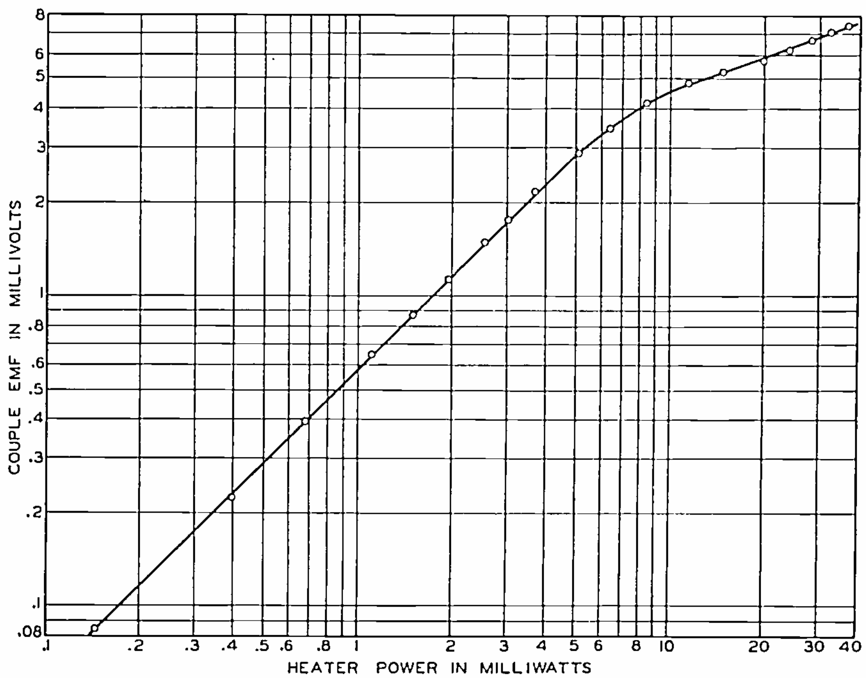

Fig. 5—Direct-current calibration of laboratory vacuum thermocouple R-253-1. The couple electromotive force was read on a meter of 10 ohms resistance.

large accidental overloads. The heater resistance is about seven ohms. Fig. 5 shows the direct-current calibration of this thermocouple.

3. Laboratory thermocouple R-307.

This thermocouple is of the indirectly heated type. The heater is a spiral of two-mil tungsten wire, 1.5 centimeters long, sixteen mils in diameter; enclosed by a nickel cylinder, three millimeters in diameter. This cylinder is supported by two leads, one of ten-mil nickel, the other of ten-mil iron. These leads serve as the junction. The heater and the cylinder are supported solely by the lead seals, thus eliminating the dielectric losses and the shunting capacities which would otherwise occur in the supporting beads and spacers. The heater leads and couple leads are arranged on the corners of a square, the heater leads being on one diagonal and the couple leads on the other to reduce inductive coupling. This thermocouple measures power up to four watts.

4. Laboratory thermocouple R-347.

The heater of this thermocouple is the filament of a seventy-five-watt incandescent lamp mounted in the form of a V, the point of the V being supported by a tungsten spring. To this spring is attached a nickel-tungsten couple, its close spaced leads being brought out of the bulb at the end opposite the heater leads. This thermocouple is not particularly accurate at wave lengths below 100 centimeters, the current not being uniform throughout the length of the filament. At a wave length of fifty centimeters the nonuniformity is quite apparent when the filament is brought to red heat. This thermocouple measures power up to fifty watts.

5. Laboratory thermocouple R-356.

This thermocouple is also of the indirectly heated type. The heater consists of the filament of a seventy-five-watt lamp enclosed by a carbonized copper cylinder. To the cylinder is attached an iron-nickel junction which serves to measure the temperature of the copper cylinder. The direct-current calibration of this thermocouple is shown in Fig. 6.

Fig. 7 is a photograph of the thermocouples described above. All of these thermocouples have been put to practical use. The low power thermocouples have been used to measure the power output of amplifiers and oscillators, to serve as tuning indicators on wavemeters, and to perform as indicators for measuring field strengths. They have proved satisfactory in all these applications. The high power thermocouples have proved less adaptable to general use. As has been pointed out, the thermocouples having the junction in direct contact with the heater cannot be used to measure total output at the shorter wave lengths. However, since they have a rapid response, they are useful in making circuit adjustments and have been used for this purpose. The indirectly heated thermocouples are inherently slow in response. This makes them undesirable for making circuit adjustments and other purposes where a quick succession of reading must be taken. Their major utility has been in making measurements of power where more than usual accuracy was necessary.

Despite the efforts made to minimize the coupling between the heater and the couple of these thermocouples, it has been found advisable to tune the couple leads with a transmission line to prevent circulating high-frequency currents in the couple circuit.

In a few instances, where it has been desirable to measure power over a small range, the power has been dissipated in a short, straight tungsten filament and the thermionic emission of the filament used as a measure of the power input.

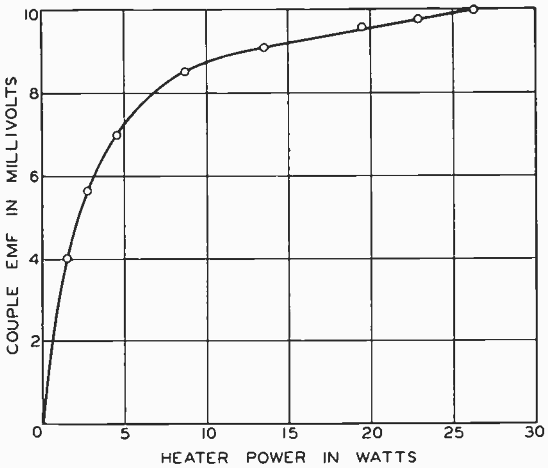

Fig. 6—Direct-current calibration of laboratory thermocouple R-356. The couple electromotive force was read on a meter of 1000 ohms resistance.

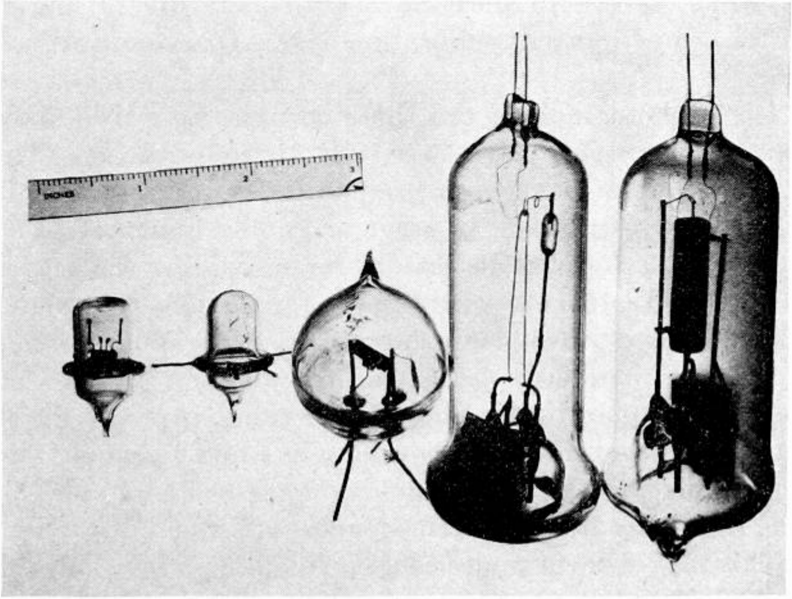

Fig. 7—Laboratory thermocouples. From left to right are R-446, R-253, R-307, R-347, and R-356.

The Measurement of Voltage

The attributes of a good voltmeter are:

1. It has adequate sensitivity.

2. It does not perturb the circuit to which it is connected; i.e., its input impedance is very high.

3. Its calibration is not a function of frequency.

Vacuum tube voltmeters have these attributes for longer wave lengths, so it seemed natural to use them as a starting point at the ultra-short waves. Reflex voltmeters and square-law voltmeters of the conventional type using RCA-955 acorn tubes were tried. It was soon found that these had input impedances of the order of 10^4 ohms at a wave length of about 100 centimeters which did not satisfy the second requirement.

A few small diodes were then built having anode diameters of twelve mils and filament diameters of 2.6 miles. These were used as rectifiers to charge a string electrometer or, alternatively, a condenser, the potential across the condenser being measured by placing a microammeter in series with a resistance of the order of a megohm across the condenser. These voltmeters were found to have very small loading effects on the circuits to which they were attached, their resistances being greater than 10^5 ohms at a wave length of one meter.

With this requirement fulfilled, sources of error in the calibration of the voltmeter were considered. The possibilities were:

1. An increase of the voltage between cathode and anode of the voltmeter due to series resonance between the interelectrode capacity and inductance of the leads. This would increase the alternating voltage between electrodes of the tube to

$$\frac{1}{1-\left(\dfrac{\text{resonant wave length}}{\text{operating wave length}}\right)^2}$$

times the voltage across the terminals of the voltmeter. This formula assumes that the operating wave length is sufficiently remote from the resonant wave length to render the resistance of the circuit negligible compared to the total reactance.

2. The time of transit of electrons between electrodes of the diode may be comparable with the period of the signal, in which case the electrometer or condenser will not charge to the peak value of the applied alternating voltage.[9,10]

[9] C. L. Fortescue, "Thermionic voltmeters for use at very high frequencies," *Jour. I.E.E.* (London), vol. 77, no. 456, p. 429; September, (1935), points out the possibility of these errors. He sets up the equations for the transit-time error, and calculates the error for a few points, but gives no experimental data. It is believed that the approximate formula given in this paper is considerably more useful.

[10] Since this paper was written a paper "Voltage measurement at very high

Both of these effects were observed when two diode voltmeters were compared at frequencies of sixty cycles and 2.38×10^8 cycles (126 centimeters wave length). The first diode had an anode diameter of twelve mils and a filament diameter of 2.6 mils; the second had an anode diameter of twenty-five mils and a filament diameter of 2.6 mils. It was found that the two voltmeters read the same at sixty cycles while at a wave length of 126 centimeters the twenty-five-mil anode voltmeter read about thirteen per cent less than the twelve-mil anode voltmeter.

The following theory was then developed to account for this discrepancy:

(*a*) *General considerations.*

Consider a diode and condenser connected across a source of constant high-frequency voltage. When the voltage is applied, the diode will pass current on the positive peaks of voltage and charge the condenser. As the charge on the condenser increases, the peak positive swing across the diode electrodes decreases. If the time of transit of the electrons were zero, the condenser would charge up to the peak value of the applied alternating voltage; i.e., until the electrons at the diode cathode felt no accelerating force at any part of the alternating-current cycle. In reality, the transit time is not zero so the condenser will charge only to such a voltage that the electrons just fail to reach the anode before their velocity becomes zero because of the retarding field on the inverse half cycle. Since the condenser does not charge to the peak of the alternating voltage when the transit time is appreciable, the voltmeter will read low. If the applied voltage is less than that necessary to bring electrons to the anode, the voltmeter will not respond. This effect will be designated "premature cutoff."

(*b*) *Premature cutoff in a parallel plane diode voltmeter.*

Consider a diode consisting of parallel planes, infinite in extent, and separated a distance d centimeters. Let the first plane be the cathode, emitting electrons at zero velocity. Suppose this diode to be connected in series with a condenser across a source of alternating voltage $V \sin(\omega t+\alpha)$, ω being 2π times the frequency and α a phase angle which will be fixed later, and that the condenser has charged to a potential such that the diode no longer passes current at any part

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frequencies," by E. C. S. Megaw, has appeared in the *Wireless Engineer*, Part I, vol. 13, no. 149, p. 65; February, (1936); Part II, vol. 13, no. 150, p. 135; March, (1936); Part II (concluded), vol. 13, no. 151, p. 201; April, (1936). Megaw has studied the errors considered in this paper and has arrived at the same approximate transit time error formula that is obtained here. The experimental constants in the formulas differ because of differences in diode structures.

of the cycle. Let this potential difference across the condenser, which is less than V due to the premature cutoff effect, be $V \sin \alpha$. This defines the phase angle ω. Hence, the potential across the diode at any time t is

$$V[\sin (\omega t + \alpha) - \sin \alpha]. \tag{1}$$

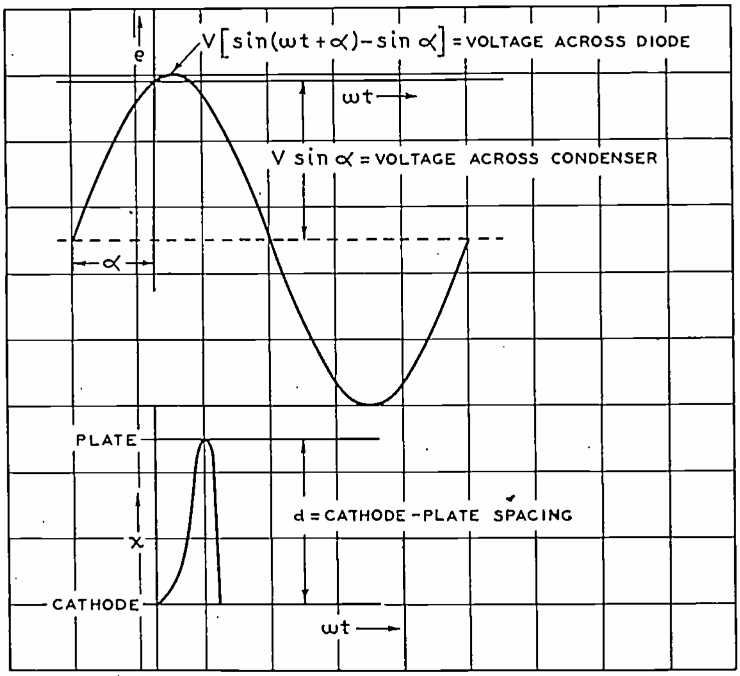

Fig. 8—Top curve: Voltage relations in a diode voltmeter at cutoff. Bottom curve: Displacement from the cathode of an electron as a function of ωt.

(See Fig. 8.) The equation of motion of an electron is then,

$$\ddot{x} = \frac{-eV}{md} [\sin (\omega t + \alpha) = \sin \alpha] \tag{2}$$

in which

$\ddot{x}$ = acceleration of the electron,

$\dfrac{e}{m}$ = specific charge of an electron.

Integrating once gives the velocity

$$\dot{x} = \frac{-eV}{m\omega d} [- \cos (\omega t + \alpha) - \omega t \sin \alpha + \cos \alpha] \tag{3}$$

the integration constant having been determined from the condition that $\dot{x}=0$, when $t=0$. Integrating again gives the distance from the cathode which the electron has traveled in time t

$$x = + \frac{eV}{m\omega^2 d}\left\{\sin(\omega t + \alpha) + \frac{(\omega t)^2}{2}\sin\alpha = \omega t\cos\alpha - \sin\alpha\right\} \tag{4}$$

the integration constant having been determined from the condition that $x=0$, when $t=0$. Because the velocity is to be zero when the electron just reaches the anode, we have the following simultaneous equations from which t is to be eliminated and $\sin\alpha$ determined:

$$0 = \frac{eV}{m\omega^2 d}\left[-\cos(\omega t + \alpha) - \omega t\sin\alpha + \cos\alpha\right] \tag{5}$$

$$d = + \frac{eV}{m\omega^2 d}\left[\sin(\omega t + \alpha) + \frac{(\omega t)^2}{2}\sin\alpha - \omega t\cos\alpha - \sin\alpha\right]. \tag{6}$$

When $\cos(\omega t+\alpha)$ and $\sin(\omega t+\alpha)$ are expanded, these equations become

$$0 = \frac{eV}{m\omega^2 d}\left\{[\cos\omega t - 1]\cos\alpha - [\sin\omega t - \omega t]\sin\alpha\right\} \tag{7}$$

$$d = \frac{eV}{m\omega^2 d}\left\{\left[\cos\omega t - 1 + \frac{(\omega t)^2}{2}\right]\sin\alpha + [\sin\omega t - \omega t]\cos\alpha\right\}. \tag{8}$$

If it is assumed that ωt is small and α is near $\pi/2$ when cutoff occurs—and such will be the case for large applied voltages—(7) becomes approximately

$$\omega t \approx \frac{3}{\tan\alpha} \equiv -3\tan\phi \tag{9}$$

and (8), on substitution for ωt,

$$d \approx -\frac{9}{8}\,\frac{eV}{\omega^2 md}\cdot\phi^4. \tag{10}$$

Approximations (9) and (10) have been checked against (4) for the case of the trajectory having $\alpha=75$ degrees. The values calculated from (9) and (10) checked those from (4) within two per cent. Then,

$$1 - \sin\alpha = 1 - \cos\phi \approx \frac{\phi^2}{2} = \frac{1}{2}\sqrt{-\frac{8}{9}\,\frac{m}{e}\,\frac{\omega^2 d^2}{V}}. \tag{11}$$

Now as V is the amplitude of signal and $V\sin\alpha$ is the voltage when cutoff occurs, the ratio of voltage error to applied peak voltage is

$$\frac{\Delta V}{V} = \frac{1}{2}\sqrt{-\frac{8}{9}\frac{m}{e}\frac{\omega^2 d^2}{V}}, \tag{12}$$

or, in practical units,

$$\frac{\Delta V}{V} = \frac{2110d}{\lambda\sqrt{V}} \tag{13}$$

where ΔV and V are expressed in peak volts and d and the wave length λ are expressed in centimeters.

The time of transit of an electron between electrodes of a diode can be expressed in the following way.[1]

$$\tau = \frac{kd}{5.95 \times 10^7\sqrt{V}} \text{ sec.} \tag{14}$$

where d is the spacing between cathode and anode; V, the applied direct voltage; and k is a parameter depending on the geometry of the diode.

The error formula can then be written

$$\frac{\Delta V}{V} = \gamma \cdot \frac{kd}{\lambda\sqrt{V}} = \eta \cdot \frac{\tau}{T}, \tag{15}$$

where,

k = Ferris' parameter k,

γ = a constant,

η = a dimensionless constant,

τ = the time of transit,

T = the period of the signal.

For the parallel plane case without space charge, $k=2$. Hence, in this case,

$$\gamma = \frac{2110}{k} = \frac{2110}{2} = 1055 \text{ expressed in (peak volts)}^{1/2}. \tag{16}$$

That the diodes used as voltmeters are free from space charge seems reasonable when it is recalled that the electrons are confined to the cathode surface by a large retarding field during most of the cycle.

(c) *Premature cutoff in cylindrical diodes.*

It has not been possible to solve the equations of motion for the cylindrical diode and from it to obtain the transit-time error formula;

[1] W. R. Ferris, *loc. cit.* In this paper is given a plot of k against the ratio of anode diameter to filament diameter for both the space-charge and no space-charge cases.

but from the form of the error formula in the parallel plane case, the error in the cylindrical diode voltmeter would naturally be written as a function of τ/T

$$\frac{\Delta V}{V} = f\left(\frac{\tau}{T}\right) = f\left(\frac{kd}{\lambda\sqrt{V}}\right) \tag{17}$$

which, when expanded in a Maclaurin series, becomes

$$\frac{\Delta V}{V} = \frac{kd}{\lambda\sqrt{V}}\cdot f'\left(\frac{kd}{\lambda\sqrt{V}}\right)_0 + \frac{1}{2}\left(\frac{kd}{\lambda\sqrt{V}}\right)^2\cdot f''\left(\frac{kd}{\lambda\sqrt{V}}\right)_0 + \cdots . \tag{18}$$

If the error is small, the error formula becomes

$$\frac{\Delta V}{V} = \gamma'\cdot\frac{kd}{\lambda\sqrt{V}}, \tag{19}$$

γ' being a constant which is to be determined experimentally.

(*d*) *The discrepancy between the readings of two voltmeters suffering series resonance and transit time errors.*

Consider two voltmeters one and two suffering resonance and transit-time errors, their terminals connected to the same alternating voltage E. Then, the first will read,

$$R_1 = A_1EF_1\left(1 - \frac{\beta_1}{\lambda\sqrt{F_1E}}\right) \tag{20}$$

and the second,

$$R_2 = A_2EF_2\left(1 - \frac{\beta_2}{\lambda\sqrt{F_2E}}\right) \tag{21}$$

in which,

$$\beta_1 = \gamma'kd_1, \tag{22}$$

$$\beta_2 = \gamma'kd_2, \tag{23}$$

$$F_1 = \left[1 - \left(\frac{\lambda_{r1}}{\lambda}\right)^2\right]^{-1}, \tag{24}$$

$$F_2 = \left[1 - \left(\frac{\lambda_{r2}}{\lambda}\right)^2\right]^{-1}, \tag{25}$$

λ_{r1} = series resonant wave length of the first voltmeter,

λ_{r2} = series resonant wave length of the second voltmeter.

Because neither voltmeter reads correctly, the true applied voltage E is unknown. The best, then, that can be done is to express the discrepancy R_2-R_1 in terms of the reading of one of the voltmeters, say the first. Solving the first equation for E gives,

$$E = \frac{1}{F_1}\left\{\frac{R_1}{A_1} + \frac{\beta_1{}^2}{2\lambda^2} + \frac{\beta_1}{\lambda}\sqrt{\frac{R_1}{A_1}}\sqrt{1 + \frac{\beta_1{}^2}{4\lambda^2}\frac{A_1}{R_1}}\right\}. \tag{26}$$

Substituting this in the equation of the second voltmeter, subtracting R_1, and neglecting terms of the second order in β_1/λ give

$$R_2 - R_1 = \left(\frac{A_2}{A_1}\frac{F_2}{F_1} - 1\right)R_1 + \frac{A_2}{\sqrt{A_1}}\frac{F_2}{F_1}\frac{\beta_1 - \beta_2\sqrt{\dfrac{F_1}{F_2}}}{\lambda}\sqrt{R_1}. \tag{27}$$

The neglect of higher order terms in β_1/λ invalidates the formula for small voltages, but makes it much more manageable over its valid range.

If the equation be written

$$R_2 - R_1 = C_1R_1 + C_2\sqrt{R_1}, \tag{28}$$

C_1 and C_2 have the values

$$C_1 = \frac{A_2}{A_1}\frac{F_2}{F_1} - 1, \tag{29}$$

$$C_2 = \frac{A_2}{\sqrt{A_1}}\frac{F_2}{F_1}\frac{\beta_1 - \beta_2\sqrt{\dfrac{F_1}{F_2}}}{\lambda}. \tag{30}$$

These can be solved to give

$$\frac{F_2}{F_1} = \frac{A_1}{A_2}(C_1 + 1), \tag{31}$$

$$\beta_2 = \beta_1\sqrt{\frac{A_1}{A_2}(C_1 + 1)} - \frac{C_2\lambda}{\sqrt{A_2(C_1 + 1)}}. \tag{32}$$

(*e*) *Critique.*

If the theory is valid within the limitations imposed in obtaining the final solution:

1. The discrepancy between readings of two voltmeters with the same applied voltage should be

$$R_2 - R_1 = C_1R_1 + C_2\sqrt{R_1}.$$

Hence, when $(R_2-R_1)/\sqrt{R_1}$ is plotted against $\sqrt{R_1}$, a straight line should be obtained. The slope of this line should be C_1, and its intercept on the $(R_2-R_1)/\sqrt{R_1}$ axis should be C_2.

2. The values of β_2 determined from C_1 and C_2 for a given pair of voltmeters at various wave lengths should be the same.

3. If a set of diodes of various spacing are calibrated against a "standard" voltmeter, the value of γ' obtained should be the same for all of them.

To check the theory and, what is more important, to determine the errors occurring in diode voltmeters a set of diodes was built.[11] See Figs. 9 and 10.

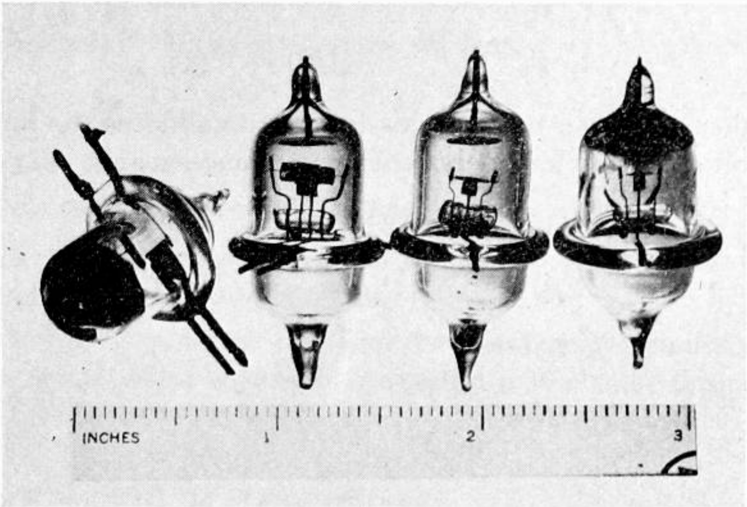

Fig. 9—Laboratory diodes. From left to right are diodes having anode diameters of 25, 80, 9, and 6.5 mils.

Laboratory No.	*Anode Diameter*	*Filament Diameter*
R-353	9 mils	2.6 mils
R-308	12	2.6
R-384	25	2.6
R-445Ac	50	2.6
R-445Ab	80	2.6
R-445Aa	120	2.6

Because of the small anode diameters, the filament alignment was not particularly good in the first three. The most that can be said about the spacings of these diodes is that they do not exceed the nominal spacings. The diode R-384 was chosen as the standard and each of the others calibrated against it. The discrepancy between voltmeters was measured with a reflex vacuum tube voltmeter, this method being more accurate than taking the difference between readings to obtain

[11] It may be of interest to note that a diode having an anode diameter of 6.5 mils has been built.

the discrepancy, which is small compared to the readings. The circuit connections are shown in Fig. 11, while Fig. 12 shows the arrangement of the voltmeters on the oscillator. Calibrations were made at sixty

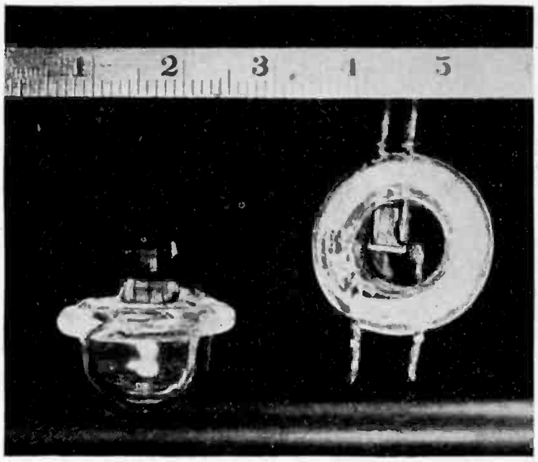

Fig. 10—The structure of the 9-mil and 25-mil diodes. Note the short, heavy leads used to minimize lead inductance in the 25-mil diode.

cycles, at a wave length of 112 centimeters, and at a wave length of eighty-four centimeters. The sixty-cycle calibrations served to determine A_1 and A_2, the proportionality constants in (20) and (21).

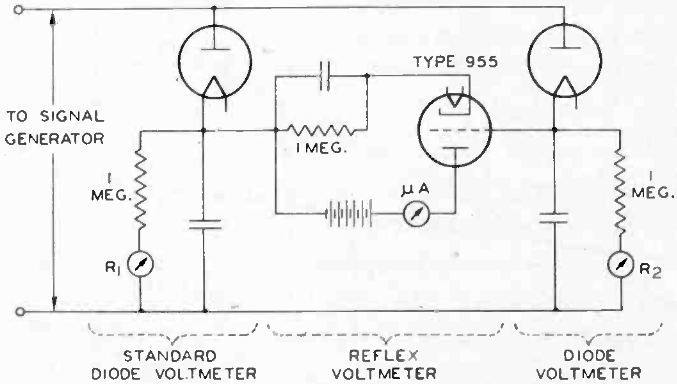

Fig. 11—Schematic diagram of the circuit for studying diode-voltmeter errors.

In Fig. 13 is shown the plot of $(R_2-R_1)/\sqrt{R_1}$ against $\sqrt{R_1}$ for the case of R-445Ac. Inspection shows that the plots are reasonably linear down to about $\sqrt{R_1}=4$. This is about the value of $\sqrt{R_1}$ at which the approximations of the theory break down. The plots for the other tubes are similar. In Fig. 14 is shown a plot of β_2 against kd for all the diodes, the circled points being the 112-centimeter wave-length points, the others the eighty-four-centimeter points. These data are rather scattered, but are believed to be within the present experi-

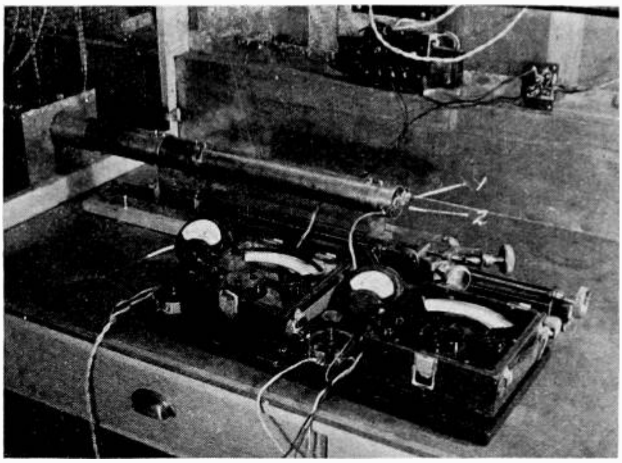

Fig. 12—A pair of diode voltmeters (1 and 2) in position for test. Note the short connections from output to the diode anodes and from the diode filaments to ground.

mental accuracy. The straight line shown has a slope of 562, so $\beta_2 = \gamma' kd \approx 562 \cdot kd$ centimeters (peak volts).$^{1/2}$ This value of γ' for the

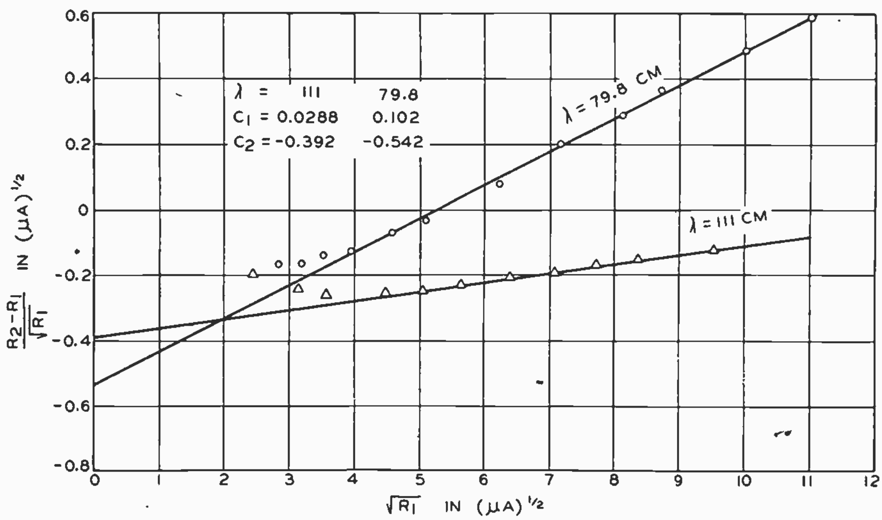

Fig. 13—Determination of the constants C_1 and C_2 using laboratory diode R-445 Ac.

cylindrical tubes is approximately one half of the γ for the ideal parallel plane tube of the theory.

The determination of the resonant frequencies of the diodes from the constant C_1 has not given accurate results. This inaccuracy is inherent in the method, as can be seen by examination of (31) from which the resonant frequencies are to be determined. This presumes, of course, that C_1 has been determined at several values of λ. C_1 is usually small compared to unity, of the order of 0.01. Hence the formula is much more sensitive to errors in the determination of A_1 and A_2 than it is to changes in C_1. The measurements indicate that the resonant frequencies of all the diodes are of the order of twenty-five centimeters.

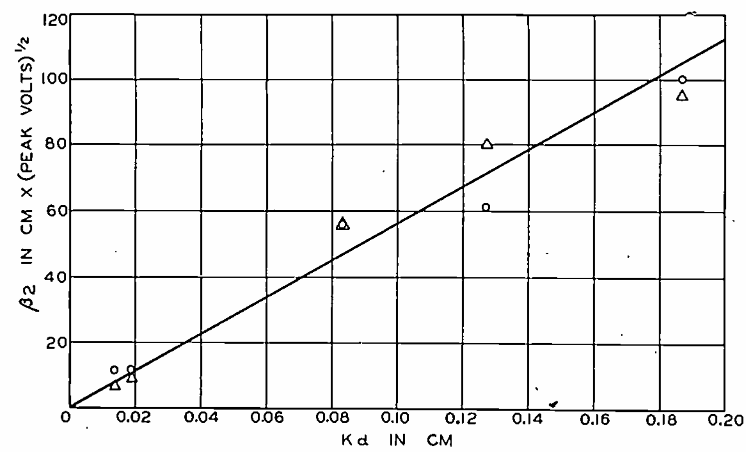

Fig. 14—Determination of γ'.

In Fig. 15 are shown the calibration curves of diode R-353, which has an anode diameter of nine mils and a filament diameter of 2.6 mils, for wave lengths of 100 centimeters and fifty centimeters. The sixty-cycle calibration and the calibrations corrected only for premature cutoff are also shown, to illustrate the magnitude of the transit-time error. It will be seen that the series resonance effect causes the greater of the errors, about six per cent at $\lambda = 100$ centimeters, about thirty per cent at $\lambda = 50$ centimeters.

The RCA-955 with grid and plate tied together has been studied to determine its applicability as a peak voltmeter. This study showed that the premature cutoff error of such a voltmeter is

$$\Delta V \approx \frac{30\sqrt{V}}{\lambda}$$

as compared with

$$\Delta V \approx \frac{7.5\sqrt{V}}{\lambda}$$

for the nine-mil diode R-353. The series resonant wave length is about forty centimeters compared with about twenty-five centimeters for R-353. Hence, the resonance error of the RCA-955 is approximately sixteen per cent at a wave length of one meter, and about sixty-four per cent at a wave length of fifty centimeters. Since the resonance error is much the larger of the two, the conclusion as to the utility of the RCA-955 voltmeter can be based on a consideration of this error. The

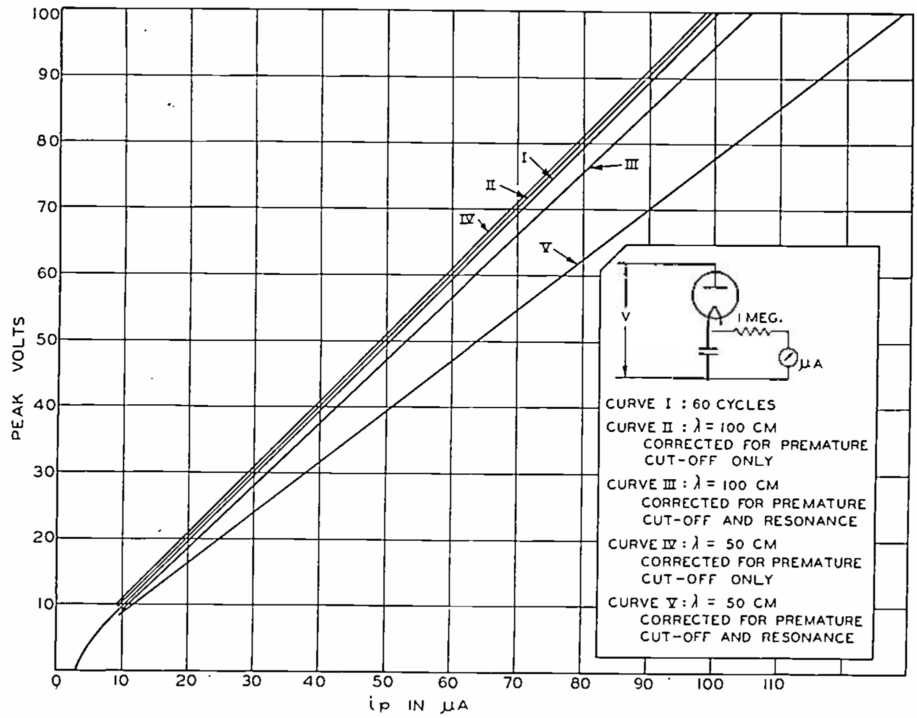

Fig. 15—Calibration of diode R-353 as a peak voltmeter.

conclusion is the RCA-955 is useful as a voltmeter above sixty-centimeter wave length, but for precision measurements below 150 centimeters and for general measurements below sixty centimeters, the small diodes are necessary.

Conclusions

Power can be measured with reasonably good accuracy throughout the power range 0.1 milliwatt to fifty watts. The thermocouples for power measurements below 100 milliwatts are very satisfactory. The indirectly heated thermocouples for powers exceeding 100 milliwatts are too slow in response to be entirely satisfactory for general use, but when time is not a factor these thermocouples give reliable results.

Peak voltage can be measured with accuracy and with very little loading by the voltmeter. Unfortunately, the calibrations of the voltmeters are not independent of frequency. However, by proper design of the diodes the discrepancy between high- and low-frequency calibrations can be made less than ten per cent for wave lengths above eighty centimeters. If greater accuracy is desired above eighty centimeters or if the range is to be extended below eighty centimeters, the calibrations can be corrected for resonance errors by measuring the series resonant wave lengths of the diodes and the premature cutoff errors, by measuring the effective cathode-anode spacings.

The writer feels that the above methods and technique for the measurement of power and voltage are adequate for most present needs, and that the problems now requiring attention are:

1. *The measurement of current.*

Thermocouples seem most promising for the measurement of small currents. However, before thermocouples can be used for this purpose an accurate knowledge of their heater reactance and resistance is necessary. For large currents the measurement of voltage drop across a resistor seems feasible. This again involves a knowledge of impedance, the high-frequency impedance of the resistor.

2. *The measurement of impedance.*

Measurements of impedance have been made by placing the unknown impedance across a transmission line loosely coupled to an oscillator and measuring the length of the line for resonance and the sharpness of resonance. This method seems capable of considerable accuracy and may be a satisfactory solution to the problem.

Acknowledgment

The writer wishes to express his appreciation of many helpful discussions of the problem with Mr. B. J. Thompson, and with his own co-workers, Dr. A. V. Haeff, Mr. G. R. Kilgore, and Mr. P. D. Zottu. To Mrs. Edna McDowell, whose manual skill and knowledge of tube mounting technique have made many new tubes possible, must go the credit for having constructed the diodes and low power thermocouples.

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THE LIMITATIONS OF RESISTANCE COUPLED AMPLIFICATION*

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Summary—*It is demonstrated that the frequency response curve of a resistance coupled amplifier is identical to that of an idealized tuned amplifier. Certain theoretical limitations on the stage gain and the band of frequencies amplified are derived, and the practical limitations are estimated. The use of a pliodynatron as a resistance coupled amplifier in order to overcome certain of the limitations is suggested. Curves and formulas are presented which will aid in the design of amplifiers by Luck's method. Analogous formulas for tuned amplifiers are derived.*

SOME time ago, D.G.C. Luck published a method[1] for predicting the performance of resistance coupled amplifiers. This method has been used by the author for designing amplifiers with highly specialized characteristics on several occasions, and has been found to yield excellent results. During the course of these investigations, it appeared that resistance coupled amplifiers were subject to certain limitations. These limitations are inherent in Luck's formulas, but their nature is not at all obvious when the formulas are in the form given. Furthermore, the curves in Luck's paper, and in fact all curves drawn from his data, seemed to show a resemblance to the selectivity curves of tuned amplifiers. Accordingly, it was decided to investigate the limitations of resistance coupled amplifiers and the relationship between resistance coupled and tuned amplification, using Luck's equation as a starting point.

Luck gives the following expressions:

$$A_0/A = 1 + j(f/f_0 - f_0/f)/w \tag{1}$$

$$A_0 = g_m R_s/(1 + C_g/C) \tag{2}$$

$$w = (1 + C_g/C)\sqrt{\frac{R_g C}{R_s(C_p + C_g + C_p C_g/C)} \cdot \frac{R_1}{R_s}} \tag{3}$$

$$1/f_0 = 2\pi\sqrt{R_1 R_g (C_p C + C_g C + C_p C_g)} \tag{4}$$

where,

A = (vector) voltage amplification at any frequency

* Decimal classification: R363.6. Original manuscript received by the Institute, January 6, 1936. *Note:* The views expressed herein are the private ones of the writer and are not to be construed as reflecting the views of the Navy Department.

[1] D. G. C. Luck, "A simplified general method for resistance-capacity coupled amplifier design," PROC. I.R.E., vol. 20, pp. 1401–1424; August, (1932).

A_0 = maximum voltage amplification
f_0 = frequency of maximum amplification
R = plate resistance of amplifier tube
g_m = transconductance of amplifier tube
R_p = plate load resistance
R_g = resistance of grid resistor (of following stage)
C = capacity of coupling condenser
C_p = total capacity shunting R_p
C_g = total capacity shunting R_g
$1/Rs = 1/R + 1/R_p + 1/R_g$
$1/R_1 = 1/R + 1/R_p$
$\mu = g_m R$

The assumptions involved in the derivation of (1), (2), (3), and (4) are:
Harmonic distortion in the tube negligible.
Regeneration negligible.
$(C_p - C_g)/C$ negligible in comparison with unity.

Luck shows that A_0, f_0, and w are fundamental constants of the amplifier from which its performance may be calculated, and gives curves in terms of these constants which make calculation rapid once these quantities are found. Consider now the tuned transformer circuit of Fig. 1. The potential induced into the secondary is $j\omega M i_p$. If the

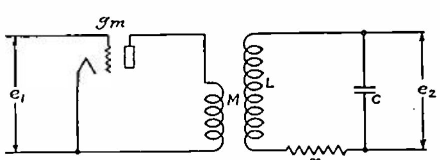

Fig. 1

primary impedance is small with respect to the plate resistance of the tube, $i_p = e_1 g_m$. Then

$$i = \frac{j\omega M g_m e_1}{r + j(\omega L - 1/\omega C)}$$

$$e_2 = i/j\omega C; \qquad A = \frac{e_2}{e_1} = \frac{g_m}{r + j(\omega L - 1/\omega C)} \cdot \frac{M}{C}$$

$$A_0 = \frac{M g_m}{Cr} \tag{5}$$

$$\frac{A_0}{A} = \frac{r + j(\omega L - 1/\omega C)}{r} = 1 + j\left(\omega L/r - \frac{1}{\omega C r}\right)$$

$$\frac{A_0}{A} = 1 + jQ\left(\frac{f}{f_0} - \frac{f_0}{f}\right) \tag{6}$$

where,

$$Q = \frac{1}{r}\sqrt{\frac{L}{C}} \tag{7}$$

$$f_0 = \frac{1}{2\pi\sqrt{LC}}. \tag{8}$$

Equation (6) is identical to (1) if $Q = 1/w$. From this it follows that the law which governs the way in which amplification varies with frequency is the same for both resistance coupled and tuned amplifiers.

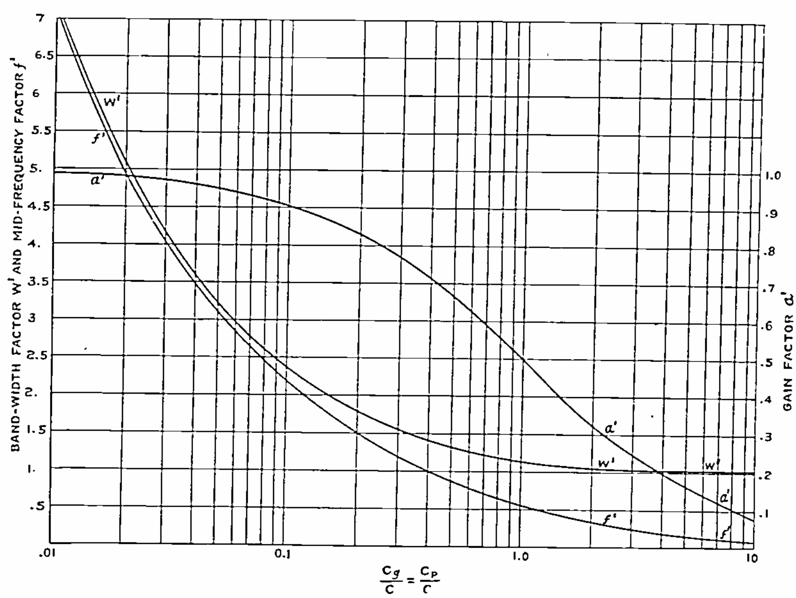

Fig. 2—Effect of large capacity shunts. Gain factor a', band-width factor W', and mid-frequency factor f' plotted as a function of C_g/C where $C_p = C_g$. In this range C_g and C_p must be nearly equal if the actual performance is to agree with the calculated performance.

The apparently great difference in the performance of the two types is therefore a difference in degree rather than a difference in kind. The difference in degree is of course very great. It will be shown later that the equivalent Q of a resistance coupled amplifier cannot exceed one half, while the Q of a tuned amplifier is generally greater than ten. The equivalent Q of the usual resistance coupled audio amplifier is roughly three orders less than the Q of a normal tuned radio-frequency amplifier, and the difference is much greater yet if extreme cases are considered.

Another consequence of the equivalence of (6) and (1) is that the performance of tuned amplifiers can be estimated from Luck's curves. To verify this, selectivity curves for some typical tuned radio-frequency amplifiers were calculated from Luck's curves and from Abac

No. 22 in R. T. Beatty's "Radio Data Charts." The results agreed as closely as the curves could be read.

As a corollary of this, it follows that the performance of tuned transformer amplifiers can be estimated from Luck's curves. The approximation involved is one that is frequently used;[2] and will often be satisfactory where high impedance tubes such as the conventional radio-frequency pentodes are used. Also, the reason for the symmetry of Luck's curves and their resemblance to resonance curves becomes obvious. The relation between tuned and resistance coupled amplification, and the possibility of designing tuned amplifiers by Luck's method is considered more rigorously in the Appendix.

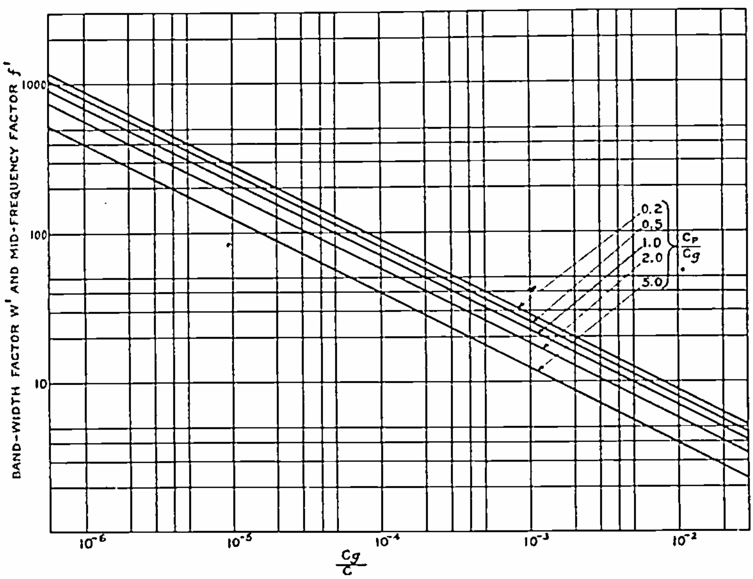

Fig. 3—Effect of small shunt capacities. Band-width factor W' = mid-frequency factor f' plotted as a function of C_g/C for five values of C_p/C_g. Gain factor a' may be regarded as unity in this range.

In order to calculate the limiting values of A_0, f_0, and w it is convenient to throw Luck's equations into a somewhat different form. Let

$$R/R_p = p, \quad R/R_g = q, \quad C_g/C = k, \quad C_p/C = h.$$

Then $R_s = R/(1+p+q)$, $R_1 = R/(1+p)$ and Luck's equations become

[2] While this assumption is inadmissible from a rigourous viewpoint, it does not lead to any serious misconception of the behaviour of a tuned amplifier. The principal error it involves is the effect of the tube plate circuit losses. The exact calculation of tuned amplifier performance is very laborious and leads to results too complicated to be useful. However, by making only the approximations usually employed for handling radio circuits; i.e., neglecting the square of the tank resistance as compared with the square of the coil reactance, and considering the grid circuit of the following tube to be a pure capacity load—the author has been able to obtain expressions exactly similar to (6), (7), and (8), except that r is replaced by $r+L/CR$. As this computation is rather long and presents no unusual features, it has not been included.

$$A_0 = \frac{\mu}{(1+k)(1+p+q)} \tag{9}$$

$$= a'a'' \text{ where } a' = 1/(1+k), \ a'' = 1/(1+p+q)$$

$$w = \frac{1+k}{\sqrt{k+h+kh}} \cdot \frac{1+p+q}{\sqrt{q(1+p)}} \tag{10}$$

$$= w'w'' \text{ where } w' = \frac{1+k}{\sqrt{k+h+kh}}, \quad w'' = \frac{1+p+q}{\sqrt{q(1+p)}}$$

$$f_0 = \frac{1}{2\pi RC}\sqrt{\frac{q(1+p)}{k+h+kh}} \tag{11}$$

$$f_0 = \frac{1}{RC} \cdot f'f'' \text{ where } f' = \frac{1}{\sqrt{k+h+kh}}, \quad f'' = \frac{1}{2\pi}\sqrt{(1+p)q}.$$

It will be noted that a', w', and f' are functions of the ratio of the capacities only, and that a'', w'', and f'' are functions of the ratios of the resistances only. For convenience in design work, these quantities are plotted in Figs. 2 to 6. As a' is substantially unity for small values of k, it has not been plotted for k less than 0.01. As the approximation condition $(C_p - C_g)/C = 1 - k \ll 1$ requires C_p and C_g to be nearly equal for large values of k, only $C_p = C_g$ is considered for k greater than 0.01.

A very useful expression is formed by multiplying (9), (10), and (11) together:

$$A_0 f_0 w = \frac{\mu}{(1+k)(1j\,p+q)} \cdot \frac{1+k}{\sqrt{k+h+kh}} \cdot \frac{1+p+q}{\sqrt{q(1+p)}}$$

$$\cdot \frac{\sqrt{q(1+p)}}{\sqrt{k+h+kh}} \cdot \frac{1}{2\pi RC}$$

$$= \frac{\mu}{2\pi RC(k+h+kh)}$$

$$A_0 f_0 w = \frac{g_m}{2\pi\left(C_p + C_g + \dfrac{C_p C_g}{C}\right)}. \tag{12}$$

This, of course, gives any one of Luck's constants as soon as the other two have been determined. Both A_0 and w, however, are subject to independent limits which must be taken into account. As (9) shows, A_0 must be less than μ; if A_0 is equal to or less than one the tube can no longer be considered an amplifier. The factor w' approaches a lower limit of unity, which is approached as k and h both become very large.

(It would also become small if h was large and k small, but this violates the original assumption that $(C_p - C_g)/C$ is to be small). It can be shown by differentiation that w'' attains a minimum value of two when $p = q - 1$. Two is therefore the lower limit for $w = w'w''$. As w increases indefinitely when k and h both decrease, there is no theoretical upper limit for w. In practice, C_p and C_g cannot be reduced much below 20×10^{-12} farads; w'' cannot be increased beyond six without seriously reducing the stage gain. Substituting these values in (10) it will be found that values of w up to about a thousand times the number of microfarads coupling capacitance can be realized.

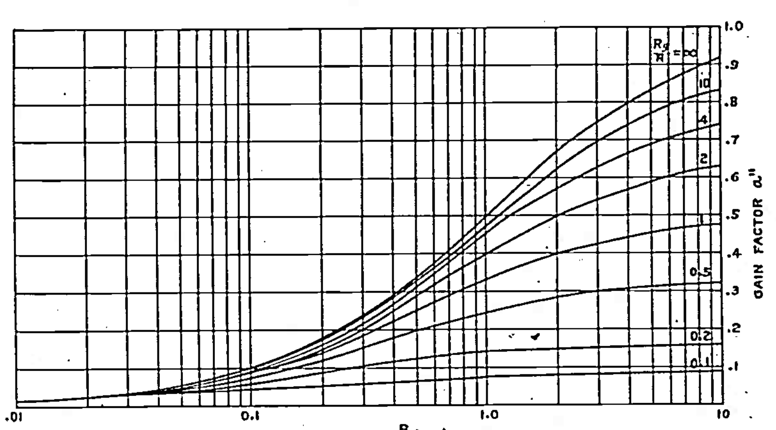

Fig. 4—Gain factor a'' as a function of $R_p/R(=1/p)$ and $R_g/R(=1/q)$.

We shall now discuss frequency limitations in the problem of obtaining maximum selectivity. Introducing the minimum values of A_0 and w in (12) gives the maximum value of the mid-frequency f_0

$$f_0(\max) = \frac{g_m}{4\pi\left(C_p + C_g + \dfrac{C_p C_g}{C}\right)}. \tag{13}$$

In this equation C cannot be chosen arbitrarily: putting $w = 2$ implies that C_p/C is large. A value of $C_p/C = 2$ gives a reasonable approach to $w = 2$; assuming C_p and C_g to be about 20×10^{-12} as before and assuming $g_m = 10^{-3}$ as a typical value, gives

$$f_0(\max) \doteq \frac{10^{-3}}{4\pi(20 + 20 + 40)10^{-12}} \doteq 10^6 \text{ cycles}$$

as the upper limit. In practice, one generally has to be content with a value of the mutual conductance of one half to one third its rated value, in order to keep the plate supply voltage to a reasonable value.

This sets the upper limit for present tubes between 300 and 500 kilocycles. With the resistance coupled stage thus adjusted for maximum selectivity, a voltage amplification of ten per stage could not be obtained at frequencies greater than a tenth of this frequency.

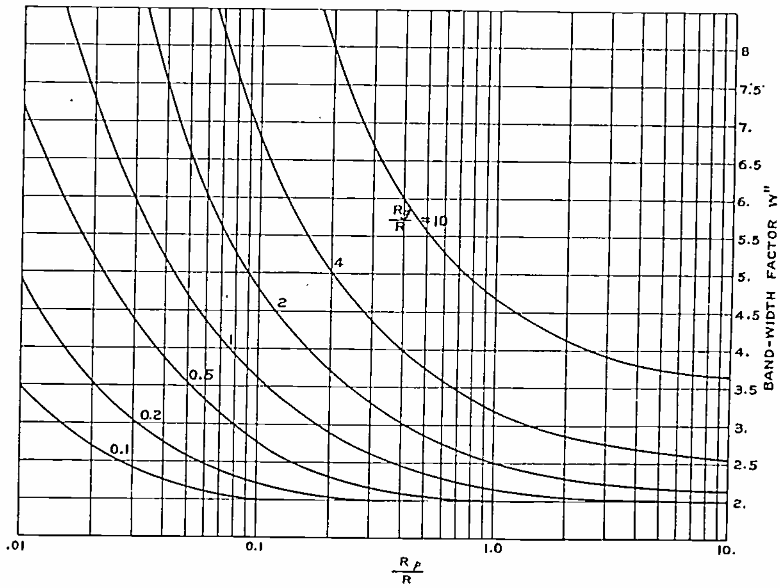

Fig. 5—Band-width factor W'' as a function of $R_p/R(=1/p)$ and $R_g/R(=1/q)$.

It can be shown[3] that if w is large compared with 2, the frequencies at which the gain is 70.7 per cent of its maximum value (amplification about three decibels down) are $f_1=f_0/w$, $f_2=f_0w$. If the pass band is defined as the region in which the gain is within 70.7 per cent of its

[3] It is stated in Luck's article that the frequencies at which the absolute value of the voltage amplification is $A_0/2$ are wf_0 and f_0/w. It can be readily seen from (1) that this gain occurs when

$$\frac{f/f_0 - f_0/f}{w} = 1$$

or,

$$f^2 - wff_0 - f_0^2 = 0, f = \frac{wf_0 \pm \sqrt{w^2f_0^2 + 4f_0^2}}{2}$$

$$= \frac{f_0}{2}(w \pm \sqrt{w^2+4})$$

if, $w^2 \gg 4$, then,

$$f = \frac{f_0}{2}\left[w \pm \left(w + \frac{2}{w} + \cdots\right)\right]$$

$$\doteq f_0w, (-)f_0/w.$$

Luck's statement, which he gives as the definition of w, is then approximately correct for large values of w. For small values of w the statement is inadmissable. Accordingly, the definition of w given later in this paper is to be preferred. The remainder of Luck's method is valid for all values of w.

maximum value, f_2 is the highest frequency in the pass band. If the reciprocal of w is negligible by comparison with w, f_2 is also approximately equal to the width of the pass band. From (12),

$$f_2(\max) = \frac{g_m}{2\pi A_0\left(C_p + C_g + \dfrac{C_pC_g}{C}\right)} \tag{14}$$

$$\doteq \frac{10^6}{A_0} \text{ for present tubes.}$$

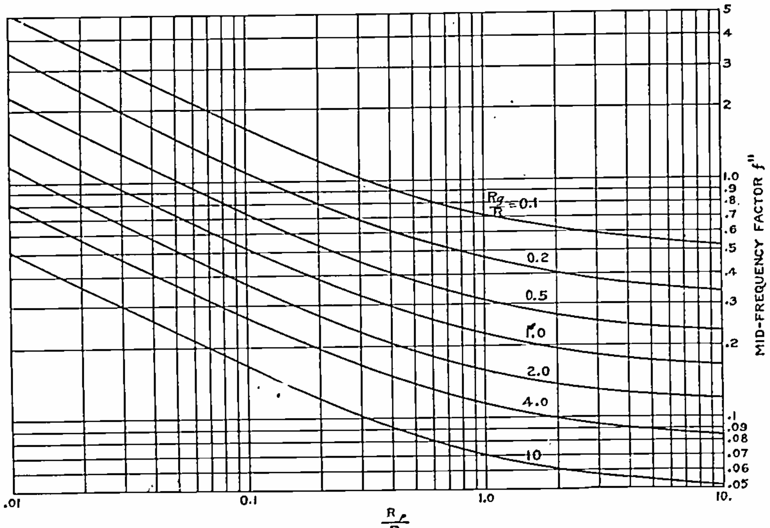

Fig. 6—Mid-frequency factor f'' as a function of $R_p/R(=1/p)$ and $R_g/R(=1/q)$.

Due to the fact (previously derived) that there is no theoretical upper limit on the value of w, there is likewise no theoretical lower limit on the value of f_0. By artificially increasing the capacity shunts C_p and C_g, a low value of f_0 can be obtained in conjunction with a low value of w. Trial computations show that without using any condenser of more than a microfarad capacitance, or any resistance larger than a quarter megohm, mid-frequencies of less than one cycle can be obtained in conjunction with broadness factors as low as $w = 2.3$. The limitations in this direction are therefore of little consequence at present. It must be noted, however, that in case very low frequencies are included in the pass band of an alternating-current operated amplifier, power-supply filter considerations are likely to introduce more severe limitations.[4]

[4] *See* L. B. Arguimbau, "An alternating-current operated resistance-coupled voltage amplifier," *General Radio Experimenter*, vol. 10, September, (1935).

It is obvious that if p and q in (9) and (10) were allowed to assume negative values, the limitations just derived could be greatly exceeded. This effect could be obtained by using a tube with a negative plate resistance, i.e., a pliodynatron,[5] in a conventional resistance coupled connection. Letting $Rp = -R/p'$, $Rg = -R/q'$, Luck's equations become

$$A_0 = \frac{\mu}{(p' + q' - 1)(1 + k)} \tag{15}$$

$$w = \frac{1 + k}{\sqrt{k + h + kh}} \cdot \frac{p' + q' + 1}{\sqrt{q'(p' - 1)}} \tag{16}$$

$$f_0 = \frac{1}{2\pi RC}\sqrt{\frac{q'(p' - 1)}{k + h + kh}}. \tag{17}$$

The theoretical limitations on A_0 and w are now completely removed, for, as $p'+q'$ approaches unity, A_0 becomes large beyond all limitations and w approaches zero. However, the instability common to negative resistance devices, and the indefiniteness of the resistance where the characteristic is curved now enter as limiting factors. It would be difficult to estimate the magnitude of the practical limits without long experience with the device. Comparing (11) and (17) it will be seen that the limiting frequency for pliodynatron tubes is slightly less than the limiting frequency for normal tubes. The resistance coupled pliodynatron has in common with regenerative tuned amplifiers the property that stage gains greater than the amplification factor of the tube may be readily obtained without a step-up transformer. The effective values of $Q(=1/w$, as previously noted), which are obtained are apt to be more like the Q factors of nonregenerative tuned amplifiers. The circuit is in many respects similar to the resistance tuned amplifier described by Cabot,[6] but it requires fewer circuit elements, only one critical balance instead of two, and the operation of the circuit seems easier to understand. The natural frequency can be most conveniently varied by a condenser, instead of a resistance as in Cabot's amplifier. If this is a three-gang condenser so designed that C_p/C and C_g/C remain constant, the selectivity is independent of frequency, as in Cabot's amplifier.

[5] The author knows of no tube now on the market which is entirely satisfactory as a pliodynatron. Published curves for the RCA 236 indicate that an average tube may be operated under conditions such that $R = -0.13$ to -0.48 megohm, $g_m = 300$–500 micromhos, $\mu = 65$–150. This tube might be used as suggested where uniformity was not a factor.

[6] Sewell Cabot, "Resistance tuning," Proc. I.R.E., vol. 22, pp. 732–737: June, (1934).

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A POTENTIOMETRIC DIRECT-CURRENT AMPLIFIER AND ITS APPLICATIONS*

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Introduction

STABLE amplification of direct currents has engaged the attention of numerous investigators, especially since the advent of the photocell for the measurement of light and associated values. The result has been the development of many conductively coupled amplifier circuits capable of a relatively high degree of stability consistent with a high gain, but unfortunately still far too variable for trouble-free industrial usage.

The system here described is basically an automatic null potentiometer, but may be classified as an amplifier because its output is proportional to the input but larger in amplitude. Null potentiometric comparison of input to output by means of a fixed resistor or resistors, and the use of a sensitive null device assures independence of circuit variables and a degree of stability far beyond that of the conductively coupled amplifier.

The principal applications divide roughly into two general groups; first, the amplification of direct voltage for the measuring, recording, etc., of direct-current quantities ordinarily too small to be applied directly; second, a unique system of regulation involving the interposition of the quantity to be regulated directly into the cycle of operation, and balancing to an electrical null point affected by this quantity, such as regulation of temperature by the use of a temperature sensitive bridge, thermocouple, and fixed reference electromotive force, etc.

Fundamental Circuits and Operation

In the double bridge circuit of Fig. 1, two phototubes are arranged as shown to determine the grid potential of the triode forming the adjustable arm of the bridge. A null galvanometer is placed to operate from the unbalance current of the bridge, and is optically connected to the phototubes so that deflection causes a differential change in conductance, preferably an increase in one with a decrease in the other. Upon the appearance of unbalance current, the galvanometer

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deflects in the proper direction to change the grid potential, and thus the plate resistance of the tube, until the bridge is balanced and the unbalance current disappears, at which point deflection ceases and the action stops. It is apparent that a change in any of the bridge arms, including the phototubes, giving rise to an unbalance, again initiates this action and the bridge automatically rebalances itself.

If a small potential e is now inserted in the galvanometer circuit at the point marked X, the galvanometer will deflect and the bridge will unbalance itself to a state where the unbalance current I passing through the resistance R creates an IR drop equal and opposite to e, and the condition of zero galvanometer current is again satisfied. Potentiometric balance is now established, and the current I, and the resistance R may be used as a measure of e. The action of the bridge is purely secondary and in no way can occasion errors within functional limits determined by its capacity to deliver the current required for potentiometric balance.

With the circuit functioning as a voltage amplifier, having the output I corresponding to the input e, the mutual conductance is the conductance of the resistance R

$$g_m = \frac{1}{R}\cdot$$

As the galvanometer is constantly in the cycle of operation, spurious deflections caused by extraneous influences instantly cause a flow of galvanometer current to return it to position. Thus the usual spring control torque may be removed and the movement allowed to be free floating, eliminating the practical difficulties of control torque application and consequent zero drift. As spring control torque is in effect replaced by the action of unbalance currents, the system has a natural period comparable to that resulting from the spring control torque in a conventional galvanometer. However, the zero position is always a true potentiometric zero, and not a current and spring torque balance, as would be the case with a spring control torque with the unavoidably variable mechanical zero. Furthermore, the galvanometer null position in this system is different for different values of input, and spring control torque would reduce the gain of the amplifier.

The circuit of Fig. 2(A) is substantially that of Fig. 1 with consideration given to practical usage. A prism is arranged to divide the light beam reflected from the mirror galvanometer in such a way that a very small deflection causes a relatively large change of output current. During a period of unbalance the galvanometer attains considerable velocity resulting in an overswing and damped oscillation, unavoidable by conventional means of damping such as shunting or winding the moving coil on a conducting frame. To provide the additional damping necessary, an absorption circuit consisting of a resistor

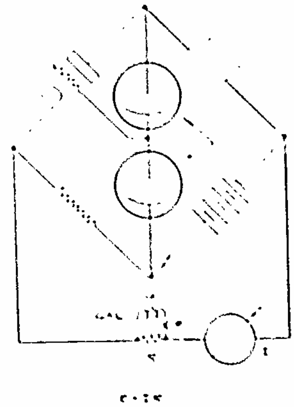

Fig. 1

and condenser connected in series between the grid and cathode is used. Proper values will critically damp the circuit after the first half cycle of overswing.

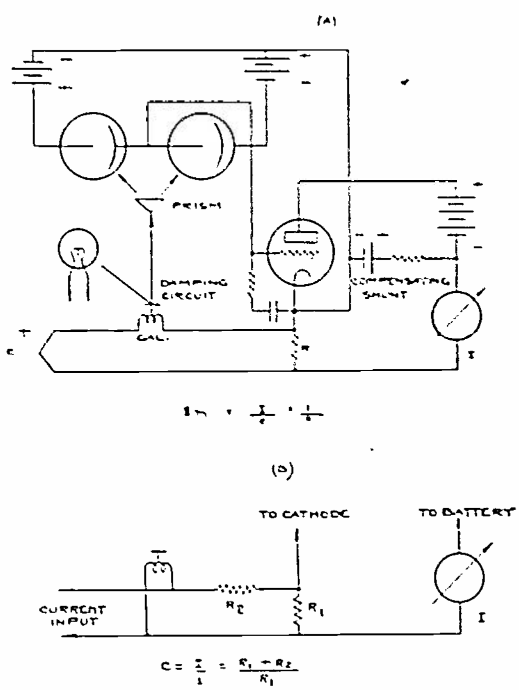

Fig. 2

The compensating shunt allows the tube to be operated over an optimum portion of its characteristic, and can allow a reversal of current polarity through the standard resistor for amplifying reversals of polarity of the input.

Adaptation of the system to current amplification is most effectively accomplished by the circuit of Fig. 2(B). A portion of the output current is attenuated into the input circuit assisting the small input current. At the point where the ratio of output to input current is equal to the attenuation factor of the resistors forming the attenuation circuit, $(R_1+R_2,)/R_1$ the galvanometer terminal electromotive force is zero, and the circuit balances to this condition. The current amplification factor of the amplifier is then equal to this attenuation factor,

$$C = \frac{R_1 + R_2}{R_1}.$$

An unusual feature of this system of amplification is that it has the ideal characteristic of infinite input impedance in the case of voltage amplification, and zero input impedance in the case of current amplification, under static conditions or at frequencies small in comparison with the period. This is the natural result of continuous potentiometric balance and advances application where this is desirable or necessary.

The Practical Instrument

As the stability of the amplifier is the stability of the potentiometer resistor or resistors, the finished instrument is operated entirely from alternating-current service power, without being affected by normal changes of line voltage or other components subject to appreciable variation. A standard type of power supply is used to replace the batteries of Fig. 2. The galvanometer and optical system are designed so that a deflection of approximately ten minutes swings the amplifier over the full range of output current. An output range of ten milliamperes was selected as sufficient for the majority of applications.

Applied as a voltage amplifier to an input circuit of negligible resistance, such as a thermocouple, and operating with a mutual conductance of five mhos, stability is well within one per cent. As the galvanometer sensitivity is constant, the stability increases as the mutual conductance is decreased, and for a mutual conductance of one mho, would be better than 0.2 per cent. Applied as a current amplifier to a circuit of relatively high resistance, such as a blocking layer photocell, and operating with a current amplification factor of 2000, the circuit exhibits about the same degree of stability. Similarly the stability increases as the amplification factor is decreased.

The speed of response to a change of input may be expressed in terms of the natural period, and with proper damping the amplifier will come to rest within one cycle irrespective of the amount or rapidity of change. Operating with a mutual conductance of five mhos, or with an amplification factor of 2000, as before stated as examples of stability, the period is roughly 0.4 second. For other values the period theoretically varies from this figure as the square root of the mutual conductance or the amplification factor, but actually has a limit at about 0.03 second, at which point it becomes necessary to overdamp the amplifier to avoid self-sustained oscillation.

Continuous Regulation

The circuit of Fig. 3(A) shows the amplifier as arranged for the amplification of voltage as heretofore described, with the addition of a

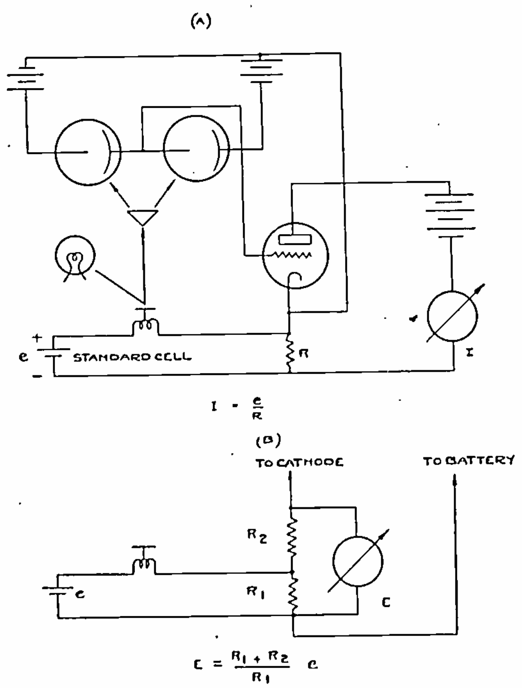

Fig. 3

standard cell or other source of constant electromotive force connected to the input terminals. As the output current is held in a fixed proportion to a constant electromotive force, it in turn is held constant regardless of the output circuit resistance, within functional limits. Similarly, by means of a voltage ratio potentiometer circuit, Fig. 3(B), the output may be made to take the form of a constant voltage that is independent of the current drawn. Thus, this circuit provides true zero slope regulation of direct voltage or current.

In a like manner any physical quantity that can be both interpreted and changed by electrical means can be regulated. As an elementary example, the circuit of Fig. 4(A) illustrates the control of temperature.

A thermocouple and heater are connected as shown so the output provides the current for the heater, and the thermocouple circuit contains a reference electromotive force equal and opposite to the electromotive force of the thermocouple at the temperature to which it is desired to regulate. To establish potentiometric balance, the heater current levels out at the value necessary to maintain the temperature at the regulation point, and changes as necessary to hold this point for varying thermal loads.

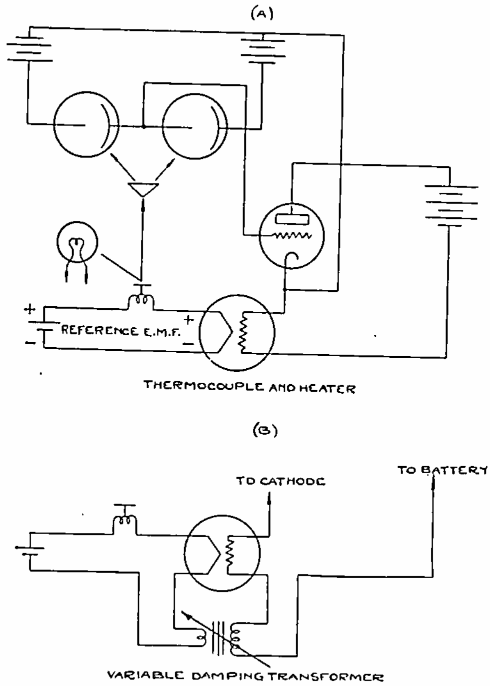

Fig. 4

Obviously the galvanometer in such a system would be extremely fast in comparison with the slow logarithmic characteristic of any such heater and thermocouple arrangement, and would result in serious overshooting or actual continued hunting. To be practical the circuit must be damped far beyond the capacity of the absorption circuit as shown in Fig. 2, and such damping preferably should be logarithmic in character. This can be done by the use of a mutual inductance or transformer, connected as shown in Fig. 4(B) to provide degenerative coupling of the output to the input circuit. This causes an electromotive force to be induced in the galvanometer circuit proportional to the rate of change of the heater current, with the result that the system now balances to the point where the rate of change is proportional to the difference between the thermocouple and reference electromotive forces, and the heater current approaches the final value logarithmically with a speed dependent upon the degree of coupling. By properly matching the damping factor to the thermal characteristics of the device being regulated, precise regulation under static conditions, and a maximum speed under varying thermal load, is attained.

Practical application of such a circuit now simply requires that a sufficient amount of power be brought under control to satisfy the requirements of the device being regulated. Instead of using the primary output current directly as an energy supply, as in Fig. 4, it may be

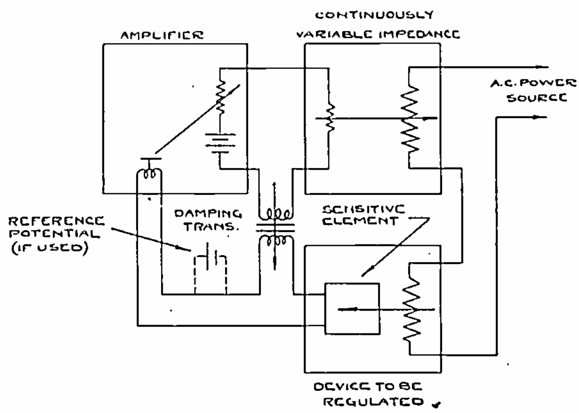

Fig. 5

used to control the value of a continuously adjustable impedance capable in turn of controlling a suitable amount of power from a separate source, as is shown schematically in Fig. 5. Theoretically, such a power control should be linear in characteristic to allow a constant damping factor to apply equally well at all power levels. However, in practice the damping factor usually is not critical and, consequently, curvature is not objectionable. This latitude allows a great number of devices and combinations to be used for this purpose.

In applications where alternating-current power can be used, saturable reactors, grid controlled rectifiers of either the high vacuum or gaseous discharge type, or various combinations of both, are usually the most economical. Where direct-current power is necessary, a battery of parallel connected power tubes, either controlling directly or through the field of a generator, appears to be most satisfactory. By these methods, power in any quantity can be applied for the regulation of large furnaces, etc.

For the precise regulation of direct current or voltage in power quantity, the direct-current methods outlined may be used, and the voltage or current held in a strict relation to a standard cell electromotive force and the potentiometer resistors.

The application of the primary direct-current amplifier to continuous regulation as shown constitutes a system of control to a predetermined level, capable of a much higher degree of precision than the conventional step methods such as by the use of relays, especially in applications usually considered difficult. It is unique in so far as it combines continuous regulation with a zero slope characteristic, and operates the null galvanometer as a free floating movement resulting in the highest sensitivity obtainable from such a device.

Practical Results of Continuous Regulation

Several installations and experimental setups have shown excellent results under conditions where the usual methods would be either entirely impracticable or unsatisfactory. One installation has been in operation for some time regulating the voltage of a direct-current line, over a load range of from zero to three amperes. The regulation attained is well within one part in 10,000. Upon instantaneously applied loads, the response is sufficiently rapid to render the momentary unbalance scarcely discernible on the galvanometer of a Leeds and Northrup Type 7 potentiometer.

The sensitivity of the system applied to the regulation of temperature is usually greater than other factors warrant. The limit usually is imposed by the degree of thermal contact between the body being regulated and the temperature sensitive element. Uninsulated water and oil baths have been held to within 0.005 degree centigrade using a temperature sensitive bridge as the interpreting element, while a well-insulated air bath showed as much as 0.05 degree centigrade departure using a similar bridge. The sensitive elements were held to a constant temperature in both cases; the departure of bath temperature in the latter case being the fault of the bath rather than the regulator.

An example of doubtful utility, but well illustrating the rapidity of response where the quantity being regulated is subject to rapid changes, is the regulation of an arc light to a steady level of illumination using a photocell as the sensitive element. The characteristic flicker was reduced to an unnoticeable value, and the cell output was held substantially constant while the arc current fluctuated to maintain the system in balance.

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BOOK REVIEWS

The National Physical Laboratory Report for the Year 1935. Published by His Majesty's Stationery Office for the Department of Scientific and Industrial Research, 1936. Paper cover, 249 pages. Price 12s.

This publication first presents a brief report of the Executive Committee to the General Board in which outstanding work of the various departments is mentioned. More detail of the work is found in the reports following for the departments, which are physics, electricity, radio, metrology, engineering, metallurgy, aerodynamics, and the William Froude Laboratory. A list of the Laboratory's publications and publications of members of the Laboratory for the years 1934 and 1935 is given.

The radio engineer may first turn to the report of the Radio Department, a section of twenty-seven pages. This report is divided into four sections describing work under the general headings of propagation of waves, direction finding, atmospherics, radio apparatus and materials. Under the first heading work is described on experimental studies of the ionosphere, measurement of the angle of incidence and polarization of downcoming waves, production and utilization of ultra-short waves including studies of electronic and magnetron oscillations. Sources of error in radio direction finding, visual cathode-ray direction finder, and the investigation of an airways marker beacon are given in the second part. The portion on atmospherics includes description of apparatus for the study of wave form, directional observations and their origin and nature. The last section describes work on the compensation of inductance coils for temperature variation, the design of a very stable variable air condenser, tests upon a constant frequency oscillator, variation of interelectrode capacity of thermionic tubes, electrical properties of the earth's surface, measurement of intensity of radio-frequency fields, and the design of a receiver for reception of time signals.

The radio engineer will also find the reports of work on other radio subjects in the report of the Electricity Department. The section on frequency standards includes work upon quartz ring oscillators, standard tuning fork, audio-frequency oscillator, international comparisons of frequency standards, and standard frequency transmissions. Radio measurements include attenuation, current, and power factor. A new method and apparatus are described for power factor measurement on insulating materials from ten kilocycles to 100 megacycles The method is based upon the sharpness of resonance. There are many other portions of the report dealing with electrical standards and measurements.

Persons interested in research on sound and its measurement will find considerable of interest in the section of the report of the Physics Department on sound.

†E. L. HALL

Radio Receiving and Television Tubes—Third Edition, by Moyer and Wostrel. Published by McGraw-Hill Book Company, Inc., New York. 626 pages and 9 page Index. Price $4.00.

†National Bureau of Standards, Washington, D. C.

In the opinion of this reviewer, the third edition is an improvement over the earlier ones, both as to material and arrangement.

Some 335 pages are devoted to tubes and an explanation of their behavior in simple circuits, 209 pages to the use of tubes in radio and industrial circuits, and 66 pages to fundamental electrical relations.

The book is directed particularly to students and to those wishing a fairly comprehensive general knowledge of the subject. It is also usable in the service field and by vacuum tube engineers as an elementary reference book.

The illustrations are plentiful and, in general, of good quality. Further editorial work would probably have improved the phraseology and accuracy in a few places.

*B. E. SHACKELFORD

An Hour a Day with Rider on D-C Voltage Distribution in Radio Receivers, by John F. Rider, 1936, 96 pages, published by the author, New York. Price 60¢.

This book opens with definitions of molecules, atoms, and electrons. It continues with the conventions of polarity and units, Ohm's law, temperature coefficient and power rating of resistors.

The body of the book comprises the application of Ohm's Law to various direct-current circuits found in radio receivers. The presentation is made by reducing typical receiver circuits to their equivalent resistor networks to which Ohm's law is applied. Grid bias methods, plate circuit drop, and power supply voltage dividers are covered. The circuits considered are all basic and the explanations for the most part are complete and lucid.

In brief the book is a textbook on Ohm's law as applied to radio receiver circuits all of which are standard and fundamental. The easy style and full explanations recommend the book for the use of service men.

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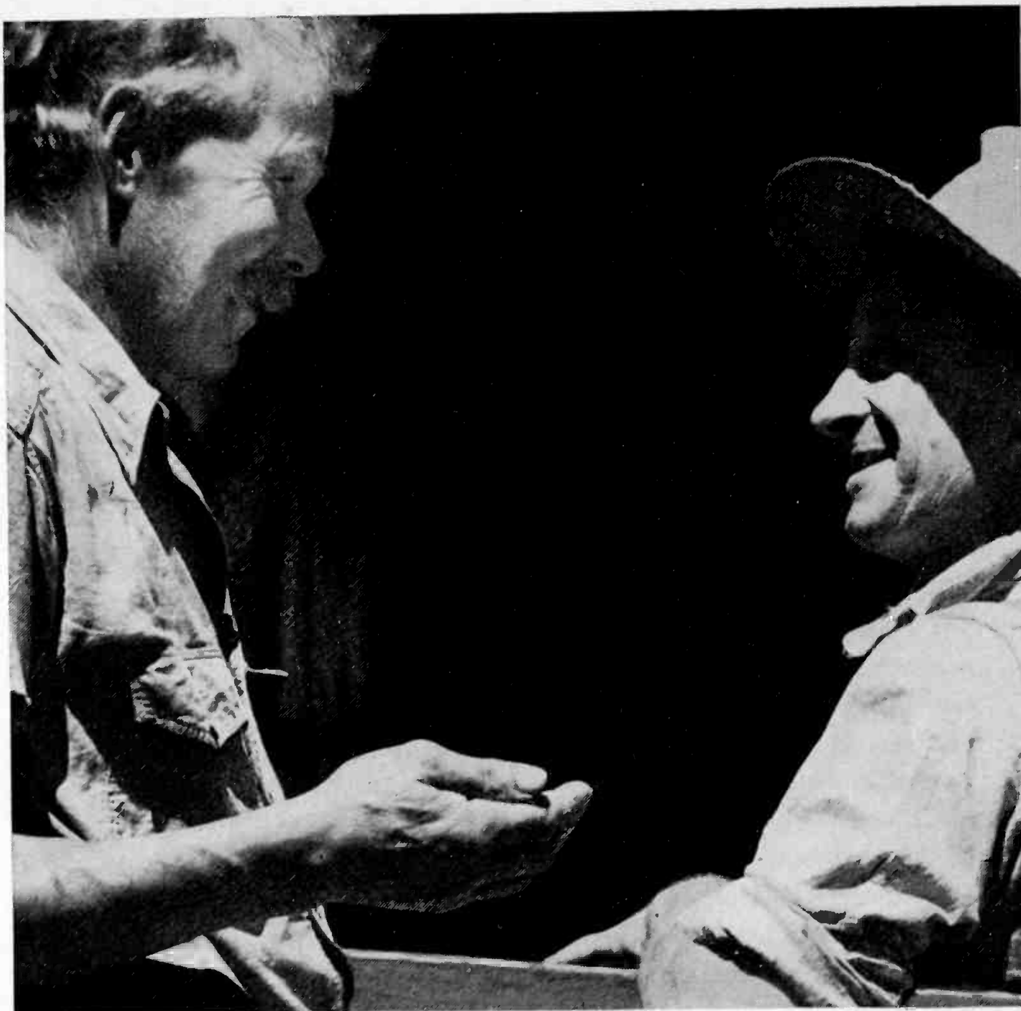
Curtis, W. F.: Born November 15, 1905. Received A.B. degree, Colorado College, 1928. Westinghouse Engineering School, August, 1928. Junior physicist, Naval Research Laboratory, 1928 to date. Associate member, Institute of Radio Engineers, 1929.

Doherty, W. H.: Born August 21, 1907, at Cambridge, Massachusetts. Received S.B. degree in electric communication engineering, Harvard University, 1927; S.M. degree in engineering, 1928. Technical employee, long lines department, American Telephone and Telegraph Company, 1928; research associate, radio section, National Bureau of Standards, 1928–1929; member, technical staff, Bell Telephone Laboratories, 1929 to date. Associate member, Institute of Radio Engineers, 1929.

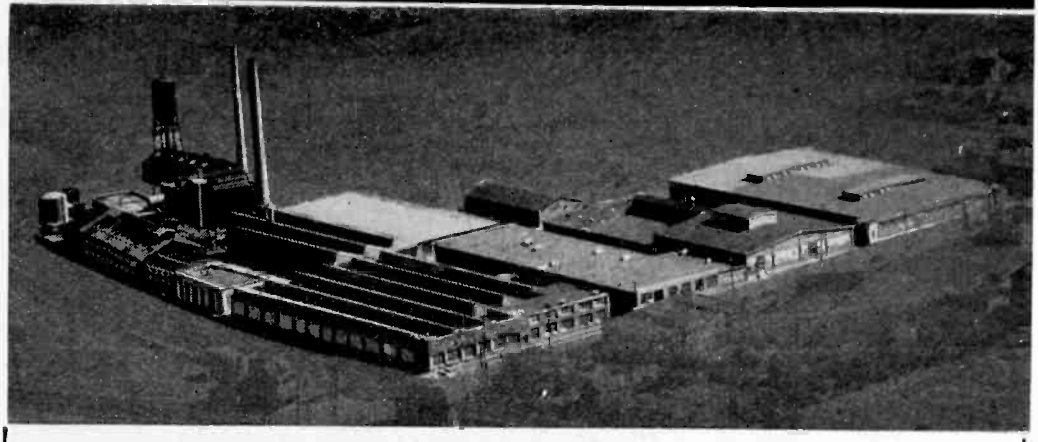
Gilbert, R. W.: Born July 21, 1909, at Brooklyn, New York. Diesel power plant design, Anglo-Chilean Nitrate Corporation, 1927. Lehigh University, 1928–1930. Gas power plant construction, Bethlehem Steel Corporation, 1931; marine engineering, turboelectric drives, Grace Line, 1932; engaged in development of industrial electronic applications, engine indicators, carrier wave telemetering, etc., private laboratory, 1933; engineering department, Weston Electrical Instrument Corporation, 1934 to date. Nonmember, Institute of Radio Engineers.

Leeds, Laurance M.: See Proceedings for June, 1936.

Nergaard, Leon S.: Born September 2, 1905, at Battle Lake, Minnesota. Received B.S. degree in electrical engineering, University of Minnesota, 1927. Research laboratory and vacuum tube engineering department, General Electric Company, 1927–1930. Received M.S. degree in electrical engineering, Union College, 1930. Teaching assistant, department of physics, University of Minnesota, 1930–1933. Research and development laboratory, RCA Radiotron Division, RCA Manufacturing Company, 1933 to date. Received Ph.D. degree, University of Minnesota, 1935. Member, Sigma XI and American Physical Society. Associate member, Institute of Radio Engineers, 1929.



ANOTHER MILESTONE OF PROGRESS



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To the engineering fraternity we extend a most cordial invitation to stop by and inspect the new factory.

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The Institute of Radio Engineers

Incorporated

330 West 42nd Street, New York, N.Y.

APPLICATION FOR ASSOCIATE MEMBERSHIP

(Application forms for other grades of membership are obtainable from the Institute)

To the Board of Directors

Gentlemen:

I hereby make application for Associate membership in the Institute of Radio Engineers on the basis of my training and professional experience given herewith, and refer to the members named below who are personally familiar with my work.

I certify that the statements made in the record of my training and professional experience are correct, and agree if elected, that I will be governed by the constitution of the Institute as long as I continue a member. Furthermore I agree to promote the objects of the Institute so far as shall be in my power, and if my membership shall be discontinued will return my membership badge.

(Sign with pen)

(Address for mail)

(Date) (City and State)

Sponsors:

(Signature of references not required here)

Mr. .... Mr. ....

Address .... Address ....

City and State .... City and State ....

Mr. ....

Address ....

City and State ....

The following extracts from the Constitution govern applications for admission to the Institute in the Associate grade:

ARTICLE II—MEMBERSHIP

Sec. 1: The membership of the Institute shall consist of: * * * (c) Associates, who shall be entitled to all the rights and privileges of the Institute except the right to hold any elective office specified in Article V. * * *

Sec. 4: An Associate shall be not less than twenty-one years of age and shall be a person who is interested in and connected with the study or application of radio science or the radio arts.

ARTICLE III—ADMISSION AND EXPULSIONS

Sec. 2: * * * Applicants shall give references to members of the Institute as follows: * * * for the grade of Associate, to three Fellows, Members, or Associates; * * * Each application for admission * * * shall embody a full record of the general technical education of the applicant and of his professional career.

ARTICLE IV—ENTRANCE FEE AND DUES

Sec. 1: * * * Entrance fee for the Associate grade of membership is $3.00 and annual dues are $6.00.

ENTRANCE FEE SHOULD ACCOMPANY APPLICATION

(Typewriting preferred in filling in this form) No. ....

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

Name ....
(Give full name, last name first)

Present Occupation ....
(Title and name of concern)

Business Address ....

Permanent Home Address ....

Place of Birth ....Date of Birth ....Age ......

Education ....

Degree ....
(college) (Date received)

TRAINING AND PROFESSIONAL EXPERIENCE

DATES	

Record may be continued on other sheets of this size if space is insufficient.

Receipt Acknowledged ....Elected ....Deferred............

Grade ....Advised of Election ....This Record Filed...........

When writing to advertisers mention of the Proceedings *will be mutually helpful.*

Alphabetical Index to Advertisements

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EMPLOYMENT PAGE

WANTED: Assistant Radio Engineer for civilian employment in the United States Military Service. College degree in electrical engineering necessary and must qualify under civil service within six months of appointment. Salary $2,600 per annum. Box No. 161, care of The Institute of Radio Engineers.

CHIEF ENGINEER wanted by Midwest manufacturer, familiar with operation, design, and manufacture of broadcast and other transmitters, with special reference to R.F. components, antennas and transmission lines. Also SALES and ADVERTISING MANAGER, familiar with transmitters (especially amateur and broadcast) but not necessarily an engineer. State qualifications in detail, including training, experience, and salary desired. Box No. 162, care of The Institute of Radio Engineers.

ENGINEERING POSITIONS

Advertisements on this page are accepted from Institute members who desire new positions and organizations which are interested in employing Institute members. A nominal charge of $2.00 is made for each insertion. Information on the preparation of material and closing dates for its acceptance will be mailed on request.

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