PROCEEDINGS OF

The Institute of Radio Engineers

Volume 10 | OCTOBER, 1922 | Number 5

CONTENTS

CORRECTION

On page 131 of the PROCEEDINGS, April, 1922 (Volume 10, Number 2), for the final equation:

$$R_r = 32.9 \frac{(1-\gamma^2)}{\left(\frac{\lambda}{\lambda_o}\right)} [1+\gamma+\gamma^2]$$

read the following equation:

$$R_r = 32.9 \frac{(1-\gamma)^2}{\left(\frac{\lambda}{\lambda_o}\right)^2} [1+\gamma+\gamma^2]$$

GENERAL INFORMATION

PUBLISHED BY

THE INSTITUTE OF RADIO ENGINEERS, INC.

THE COLLEGE OF THE CITY OF NEW YORK

EDITED BY

ALFRED N. GOLDSMITH, Ph D.

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RECEIVING MEASUREMENTS AND ATMOSPHERIC DISTURBANCES AT THE UNITED STATES NAVAL RADIO RESEARCH LABORATORY, BUREAU OF STANDARDS, WASHINGTON, MAY AND JUNE, 1922*

By

L. W. Austin

(United States Naval Radio Laboratory, Washington, D. C.)

(Communication from the International Union for Scientific Radio Telegraphy)

The measurements for May and June have been taken in the manner already described. The tables show the increase in strength of the atmospheric disturbances with the advancing season. The signal observations in May are, in general, similar to those of the earlier months except for a decrease in the afternoon values. This decrease in afternoon strength becomes very marked in the June table, which also shows the great differences in absorption at different wave lengths. During the greater part of the month, Nauen, with a wave length of 12,500 meters, became unmeasurable at three o'clock thru the disturbances and interference, falling probably to less than 3 micro-volts per meter, while Lafayette, at a wave length of about 24,000 meters, except for three or four days, did not fall much below one-fourth of its morning value.

This afternoon absorption appears to vary with the location of the receiving station, since, according to my observations made in 1919, it was very slight at Bar Harbor, Maine, but appeared to be fairly uniform for all the coast region lying between New York and Washington, as far as our observations went. The nature of this absorption is not yet understood. It appears to be more common in the warmer regions and on damp warm afternoons. Since it is almost as important as the atmospheric disturbances in interfering with reception during several hours a day in the warm months, it is well worthy of careful investigation.

*Received by the Editor, July 19, 1922.

FIELD INTENSITY OF NAUEN AND STATIC ($\lambda = 12{,}500$ m.) IN MAY, 1922, IN MICRO-VOLTS PER METER

Date	A. M.		P. M.	
	Signal	Static	Signal	Static
1	38.5	100	....	....
2	43.0	80	....	....
3	47.0	20	21.5	100
4	30.0	60	....	150
5	26.0	100	....	....
6	....	....	11.0	100
8	34.0	80	....	150
9	47.0	60	....	....
10	56.0	80	....	....
11	34.0	50	....	....
12	38.5	150	....	....
13	30.0	100	....	....
15	34.0	60	....	....
16	30.0	500	....	....
17	38.5	240	....	....
18	....	....	34.0	100
19	21.5	306	....	....
20	....	....	....	....
23	43.0	50	....	....
25	64.0	40	....	....
26	43.0	50	....	....
27	50.0	60	....	....
29	43.0	50	....	....
31	38.5	50	....	....
Average	39.8	108.8	22.2	120

Field Intensity of Lafayette and Static ($\lambda = 23,400$ m.) in May, 1922, in Micro-volts per Meter

Date	A. M.		P. M.	
	Signal	Static	Signal	Static
1	140.0	300	85.0	400
2	110.0	150	70.0	1,000
3	110.0	60	60.0	1,000
4	80.0	200	20.0	800
5	75.0	900	....	1,000
6	25.5	400	18.0	1,200
8	....	....	17.0	260
9	110.0	200	45.0	300
10	120.0	200	15.5	5,800
11	115.0	300	3.5	1,000
12	....	....	....	1,680
13	75.0	600	....	....
15	....	500	30.0	200
16	45.0	1,500	....	....
17	31.0	982	....	....
18	....	....	15.5	200
19	45.0	840	30.0	1,500
20	15.5	1,500	....	1,000
23	75.0	60	12.5	1,265
25	200.0	80	....	....
26	100.0	100	50.0	300
27	80.0	200	25.0	80
29	115.0	150	17.0	500
31	75.0	80	15.5	200
Average	87.1	438.0	31.1	990

Field Intensity of Lafayette and Static ($\lambda = 23,400$ m.) in June, 1922, in Micro-volts per Meter

Date	A. M.		P. M.	
	Signal	Static	Signal	Static
1	50.0	200	20.0	2,000
2	65.0	80	30.0	200
3	25.0	50	35.0	250
4	65.0	60	30.0	1,000
8	75.0	80	20.0	100
9	25.0	60	35.0	1,500
12	45.0	60	25.0	600
13	45.0	300	20.0	1,000
14	50.0	100	30.0	300
15	100.0	100	20.0	650
16	65.0	300	15.0	890
17	60.0	400	....	....
19	50.0	200	....	1,650
20	30.0	400	20.0	970
21	....	....	30.0	300
22	65.0	600	20.0	560
23	40.0	200	....	690
24	95.0	300	....	....
26	....	....	15.0	560
27	60.0	400	....	680
28	75.	100	....	690
29	75.	560	....	560
30	65.0	200	....	680
Average	58.3	201.8	24.3	753.8

FIELD INTENSITY OF NAUEN AND STATIC ($\lambda = 12{,}500$ m.) IN JUNE, 1922, IN MICRO-VOLTS PER METER

Date	A. M.		P. M.	
	Signal	Static	Signal	Static
2	38.0	40	11.5	100
3	38.0	60	....	....
4	30.0	40	....	....
6	....	....	11.0	400
7	60.0	30	....	....
8	4.0	50	....	300
9	30.0	17	....	800
10	....	80	....	....
12	26.0	50	....	400
13	21.5	100	....	600
14	26.0	40	....	80
15	26.0	50	....	200
16	21.5	150	....	400
17	26.0	60	....	600
19	21.5	80	....	....
20	13.0	100	....	680
21	43.0	200	11.5	640
22	34.0	300	....	420
23	36.0	30	....	500
24	47.0	40	17.0	640
26	55.0	200	21.5	300
27	26.0	150	....	800
28	19.0	80	....	....
29	60.0	80	21.5	500
30	38.0	60	....	....
Average	33.5	86.9	15.7	464.4

SUMMARY: The daily values of the morning and afternoon strength of signals during May and June, 1922, from Nauen (12,500 meters) and Lafayette (24,000 meters), together with the strength of atmospheric disturbances are given in tabular form.

THE DYNATRON DETECTOR*

A NEW HETERODYNE RECEIVER FOR CONTINUOUS AND MODULATED WAVES

By

Albert W. Hull, E. F. Hennelly, and F. R. Elder

(General Electric Company, Schenectady, New York)

Introduction

The characteristic features of the dynatron have been fully described in a previous paper.[1] It was briefly pointed out that an oscillating pliodynatron may be used for heterodyne reception of continuous waves. This application has now been thoroly investigated, and shows unique features, which it is the purpose of this paper to describe.

The practical merits of the dynatron detector may be briefly summarized as follows: In comparison with pliotron self-heterodyne, which may be taken as the best of the systems hitherto used, the dynatron has three marked advantages:

(1) The local oscillations are not fed back into the antenna.

(2) Stray disturbances of other than signal frequency are not rectified, and hence do not produce audible disturbance.

(3) The filament may be heated by alternating current without audible disturbance.

In sensitiveness and selectivity, there is little to choose between the dynatron and self-heterodyne. In ease and stability of operation, the dynatron has considerable advantage over the self-heterodyne, and is to be compared with the separate heterodyne using two pliotrons.

Description of Dynatron

The distinguishing feature of the dynatron is its utilization of secondary or *impact* electron emission. When an electron strikes a metal plate, it generally enters it, and either attaches

*Received by the Editor, February 6, 1922. Presented before The Institute of Radio Engineers, New York, March 1, 1922.

[1]Hull, Proceedings of The Institute of Radio Engineers, 6, pages 5-35, February, 1918.

itself to one of the atoms of the plate or takes part in the electric current thru the plate and connecting wires. If the electron strikes the plate with sufficient velocity, however, its impact delivers so much energy to the electrons in the surface of the plate that some of them are able to escape. The number thus liberated increases with the velocity of impact, and may exceed the number striking, causing the current to the plate to reverse in sign. For example, if each electron that strikes the plate splashes out two, and these two escape to some other electrode, it is evident that the plate will not be gaining electrons but losing them. The electrons that are splashed off are sometimes called secondary electrons, but the name *impact electrons* is more appropriate and will be used hereafter. The plate that emits impact electrons will be called the *impact electrode*, or, when it is part of a dynatron, the *dynode*.

When the impact electrode is at a higher potential than its surroundings, so that there is no place for the impact electrons to go, they will return to the impact electrode. There is then nothing to indicate that they ever left this electrode. This is why impact emission from the plate of a pliotron amplifier is never observed. In order to utilize the impact emission there must be present another electrode at higher potential than the impact electrode, so placed that its electric field will carry off the impact electrons as fast as they are emitted. The problem of dynatron construction is to place this high potential *anode*, as it will be called, as near as possible to the impact electrode or dynode, without intercepting or deflecting the electrons so that they fail to strike the dynode. This conditioni s satisfied by a slight modification of standard pliotron construction.

The dynatron consists of a filament, grid, and surrounding cylinder, like the pliotron, but the grid is placed as close as possible to the surrounding cylinder and used as anode, the cylinder being used as dynode. (Figure 1.) The electrons from the filament fly freely thru the grid anode to the surrounding dynode cylinder, and the impact electrons they produce are attracted and carried away by the grid. It is obvious that the functions of grid and cylinder may be interchanged, that is, the grid may be used as dynode and the cylinder as anode. This is very desirable for many purposes, especially where the grid is so constructed as to intercept nearly all the primary electrons. In the tests described in this paper, however, the grid is used as anode.

The volt-ampere characteristic of the dynatron is shown in Figure 4. The negative slope of the portion AB represents the loss of impact electrons, the number of which increases with velocity of impact of the primaries, that is, with increasing dynode voltage. This *negative resistance* represents available power (at the expense, of course, of power supplied to the anode) which may be used to neutralize losses in the circuit or to produce oscillations, as described later.

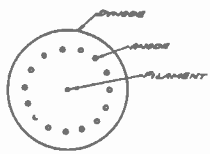


FIGURE 1 (a)—Cross-section sketch of dynatron

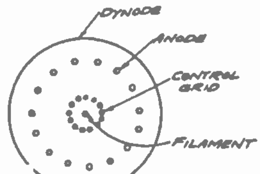


FIGURE 2 (a)—Cross-section of pliodynatron

The magnitude of the ordinates in Figure 4 depends on the magnitude of the current of primary electrons passing from the filament thru the anode to the dynode. This primary current may be controlled by a grid, close to the filament, exactly as in the pliotron. A dynatron containing such a *control grid*, in addition to the anode grid, is called a *pliodynatron* (Figure 2). Figure 5 shows the volt-ampere characteristic of the pliodynatron which was used for the oscillograms in Figures 11 and 12, at three different voltages of the control grid. The three curves are exactly similar, the scale of ordinates only being changed. It will be noted that the change of grid voltage from zero to +10 produces much greater amplification than the change from −10 to zero, in accordance with the well-known parabolic law of pliotron amplification.

Instead of using a control grid for controlling the primary emission, one may use a magnetic field, as in the magnetron. In this case the anode is preferably made of radial slats, and both anode and dynode are in the form of circular cylinders concentric with the filament, as shown in Figure 3. The control current passes thru a solenoidal coil wound on the tube, producing a magnetic field parallel to the filament. This magnetic field deflects the electrons so that fewer of them are able to get thru between the slats to the dynode. If the slats are set at a slight angle with the radius, as in Figure 3, a magnetic field in one direction will cause an increase, in the other direction a decrease, in the number of electrons that get thru, giving a nearly linear amplification. Figure 6 shows the volt-ampere characteristics, for three different

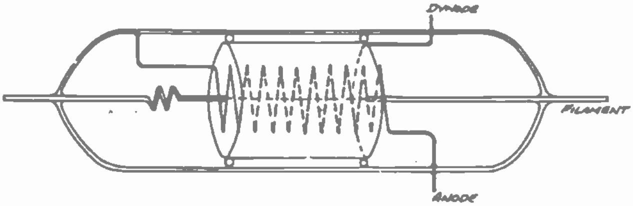


FIGURE 1 (b)—The arrangement of electrodes in the dynatron

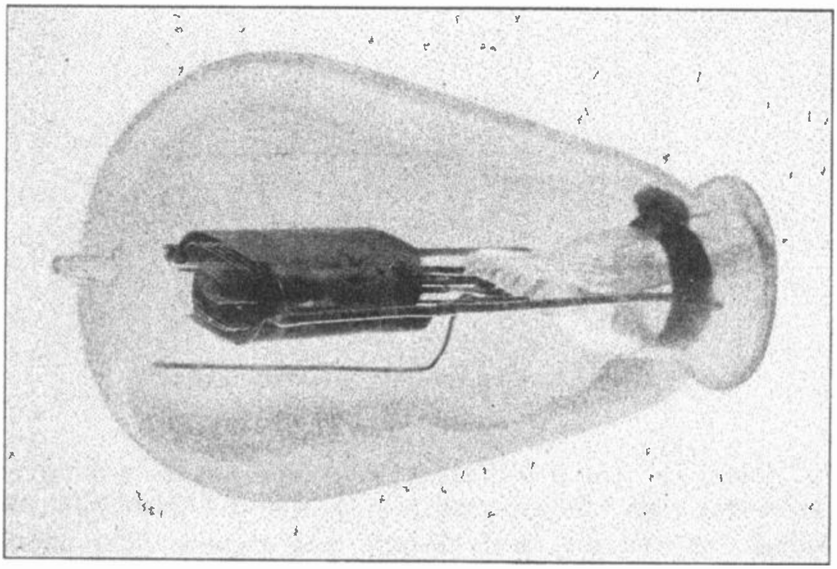


FIGURE 2 (b)—Laboratory model of pliodynatron

values of magnetic field, of the magnetron used in taking the oscillograms of Figures 13-17.

Operation

The pliodynatron detector circuit is shown in Figure 7. The signal is impressed on the grid in the usual way. The anode is connected directly to the source of high potential *E*, preferably about 110 volts, altho any potential above 40 volts is satisfactory if the tube is designed for it. The dynode is connected thru a tuned circuit L, *C*, and telephones *T* to a lower potential E_o, preferably the potential at which the dynode current is zero

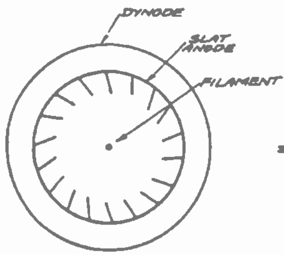


FIGURE 3 (a)—Cross-section of magnetically controlled dynatron

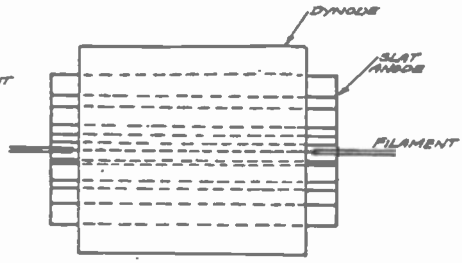


FIGURE 3 (b)—Longitudinal section of magnetically controlled dynatron

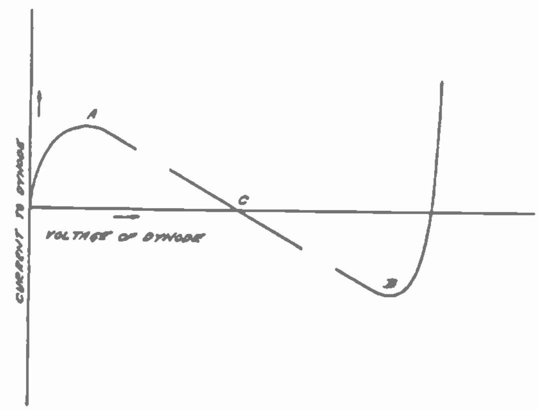


FIGURE 4—Ideal volt-ampere characteristic of dynatron

(C, Figure 4). This potential may most conveniently be obtained as shown, from a high resistance potentiometer (20,000-100,000 ohms) between the anode battery and ground. The circuit L, C is tuned to approximately the signal frequency, and generates continuous waves by virtue of the negative resistance of the dynode-anode circuit. It was shown in a previous paper,[2] that these oscillations are always maintained providing the ratio L/C of inductance to capacity is greater than the product of the negative resistance by the resistance of the tuned circuit. These radio frequency oscillations pass thru the telephones as capacity current, or may be by-passed by a small condenser in multiple with the telephones. Their rectified component passes thru the telephone windings and produces audible effects. It will be shown later that the action of the signal is to produce rectification of these local dynatron oscillations, without changing their amplitude.

[2]Previous citation, page 18.

The process of operating the detector consists simply in varying the frequency of the dynatron oscillations by means of the variable condenser C until it is approximately equal to the signal frequency, as evidenced by an audible note, and then increasing the intensity and purity of this note by tuning the input circuit in the usual way.

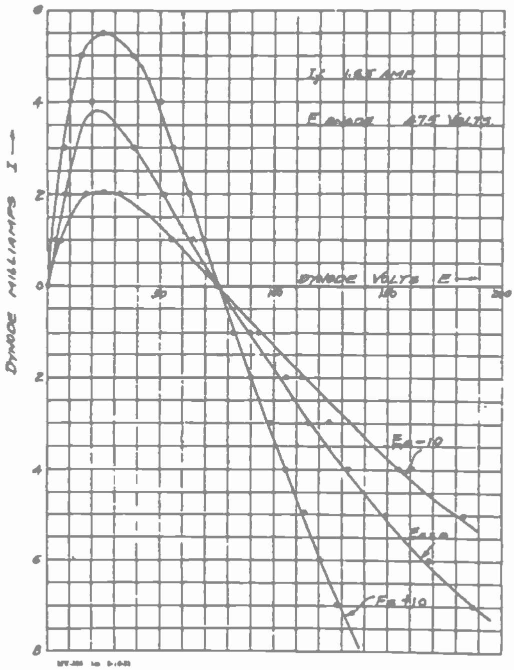


FIGURE 5—Experimental volt-ampere characteristics of pliodynatron at three different grid voltages

The detector action may be seen from Figure 8, which gives the volt-ampere characteristics of the pliodynatron at three different grid voltages (reproduced from Figure 5).

A. No Signal

The curve marked e_o represents the characteristic at normal grid voltage, with no signal impressed. If the ratio of inductance to capacity in the dynode circuit is appropriate, that is, only slightly in excess of that necessary for oscillations, the dynode oscillates symmetrically over the straight portion of this curve, between voltage limits represented by E_- and E_+. The average voltage during oscillation is E_o, and the average current per cycle, if E_o is the voltage at which the curve crosses the axis, is zero. This is the current that flows thru the telephones. Since e_o may be any grid voltage, it follows that the current thru the telephone windings is identically zero *for all grid voltages, including varying grid voltages of any frequency except signal frequency.* Hence all audio frequencies impressed on the grid, including those produced by alternating current in the filament, and all radio frequencies, except those within heterodyne range of the local oscillations, produce no effect whatever in the telephones. This fact has been carefully checked by experiment.

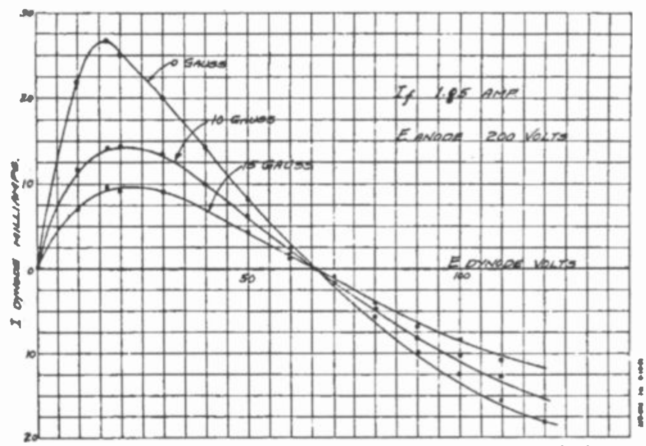


FIGURE 6—Experimental volt-ampere characteristics of magnetically controlled dynatron for three different magnetic fields

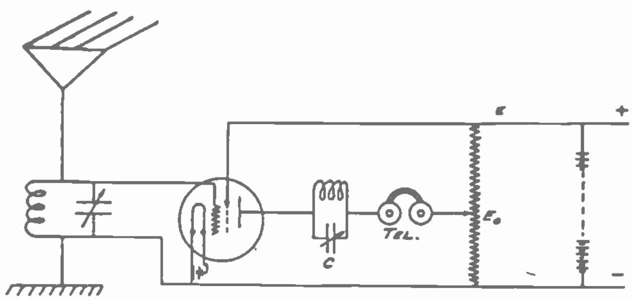


FIGURE 7—Pliodynatron detector circuit

In order to obtain complete silence from interfering signals, the three conditions mentioned above must be fulfilled: viz. (1) The volt-ampere characteristic must be straight; (2) the dynode voltage E_o must be that at which the dynode current is zero; (3) the amplitude of oscillation must be small, which requires proper ratio of inductance to capacity and resistance. In practice, however, considerable variations from these ideal conditions may be made without appreciable disturbance.

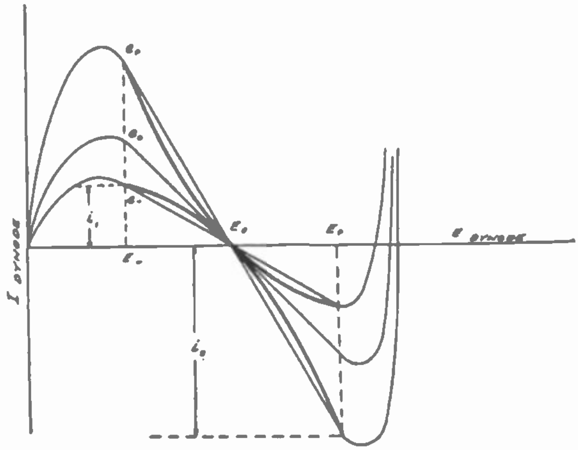


FIGURE 8—Dynamic volt-ampere characteristics of pliodynatron detector with impressed signal

B. Synchronous Signal

The effect of a signal of the same frequency as the local oscillation is unique. Referring to Figure 7, let continuous waves of the same frequency as the local oscillations be impressed on the grid, and let e_+ and e_- be the maximum and minimum values of the alternating grid voltage. The dynode volt-ampere characteristics corresponding to these grid voltages are represented in Figure 8 by the curves marked e_+ and e_-. If the phase relation is such that the grid voltage is e_- when the dynode is at its most negative value E_-, the instantaneous dynode current will be represented by the ordinate i_1. Half a cycle later the dynode voltage will be E_+, but the grid will now be at its most positive voltage e_+, and the instantaneous current will be the value i_2 on the e_+ curve. The volt-ampere characteristic corresponding to this condition of operation *with synchronous signal* will therefore be that given by the lower heavy curve. The average current is no longer zero, but has a definite negative direct current value, proportional to the amplitude of the signal. If the signal is modulated by voice frequency, the current in the telephones will be a true reproduction of the voice current. This has been checked by experiment and oscillograms.

C. Slightly Asynchronous Signal

If the frequency of the local oscillation, instead of being exactly synchronous with the signal, is slightly different, the current a any time wi e the same as in the case o exact synchronism described above; but the relative phases of grid and dynode will slowly change until the grid is at e_+ when the dynode is at E_-. The volt-ampere characteristic will then be that represented by the upper heavy curve, and the average current will be positive. These positive and negative average currents will follow each other at a frequency equal to the difference between the frequencies of signal and local oscillation, producing an audible note in the telephones. It will be shown later, both by mathematical analysis and oscillograms, that this note is a pure sine wave of frequency equal to the difference between signal and local oscillation, the only other currents present being of radio frequency, namely, the local oscillation and a frequency equal to the sum of local oscillation and signal.

In Figure 8, it is assumed that the *amplitude* of the dynode voltage oscillations is not changed by the signal. The justification for this assumption is two-fold:

(a) It is required by the theory of operation. The time average, over a signal frequency cycle, of negative conductance, and hence the average negative damping co-efficient of the dynode voltage oscillation, is constant and equal to its value when there is no signal (compare equation (13) below). This negative damping is a measure of the power fed into the oscillating circuit and hence of the amplitude of the oscillations. The stored en ergy in the local oscillating circuit, L, C, is very large, so that its change during a single cycle is inappreciable. It is, there fore, to be expected that the amplitude of the local oscillations, and consequently of the voltage at the terminals of the oscillating circuit, which is the dynode voltage, will be constant.

(b) It is shown to be true by the oscillograms, and the operating characteristics based upon it agree with the experimental tests. The oscillograms, Figures 10 to 13, show only extremely slight variations in the amplitude of the current in the oscillating circuit, even when the signal is so strong that the telephone current is completely rectified. These signals are many times stronger than those used in radio, and the slight changes in amplitude of oscillation are completely accounted for by the departure from linear amplification at these large grid amplitudes. The dynode voltage amplitude is equal to the product of this oscillating current by the impedance of the coil L.

In the case of synchronous signal the oscillograms (Figures 21 to 26) show a considerable increase in amplitude when the signal is impressed, but this is due to the unsymmetrical operating conditions, as will be discussed later.

The operation of the magnetically-controlled dynatron is identical with that of the pliodynatron, except in the method of applying the control power of the signal. In the magnetically-controlled detector the antenna coil surrounds the dynatron[3]

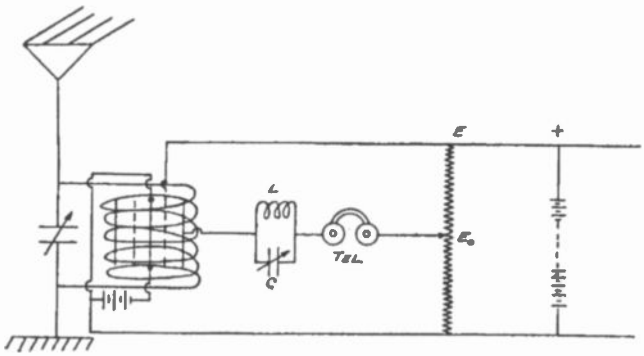


FIGURE 9—Circuit of magnetically-controlled detector

as shown in Figure 9, and the *magnetic field* of this coil is used for control; in the pliodynatron detector the *electrostatic field* between the ends of this coil is used for control. The sensitiveness of the magnetic control was slightly less than that of the electrostatic (pliodynatron) in the tests that we made, but the difference was so small as to indicate that if the tube and circuit were as well developed as those of the pliodynatron the sensitiveness of the two would be identical. The operating characteristics of the magnetically-controlled detector are identical with those of the pliodynatron, as shown by the oscillograms (See Figure 13).

Oscillographic Tests of Heterodyne Operation

In order to test the theory of detector action described above, an audio frequency model of the circuit shown in Figure 7 was constructed, in which the telephones were replaced by the vibrator of an oscillograph. Other vibrators were inserted in the oscillating circuit, the anode circuit, and the input signal circuit; so as to record the frequency, phase, and wave-shape of the dynode or telephone current, the circulating current in the tuned circuit, the anode current, and the signal voltage. The latter was a sine wave of from 5 to 10 volts amplitude, obtained by tapping off the voltage drop between two points on a fine wire resistance in the output circuit of a pliotron oscillator.

Figure 10 shows the currents in the dynode, anode, and tuned

[3]Electrostatic capacity reactions between coil and dynode were prevented by covering the tube, except for a narrow gap, with grounded tin foil or sheet copper. The control coil was either wound directlv on this foil, or wound on a form that fitted loosely over it.

circuits when no signal is impressed. The only thing to be noted is that the dynode or telephone current is symmetrical and constant with respect to its zero line. It is not a perfect sine wave, but this does nor affect the conclusions to be drawn. This film was taken with the magnetically controlled dynatron. An exactly similar film was obtained with the pliodynatron with no signal.

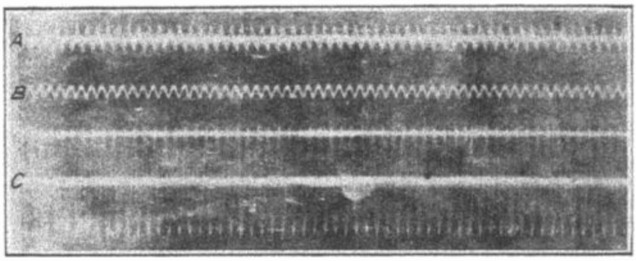


FIGURE 10—Oscillographic test of magnetically-controlled detector with no signal (frequency = 1,500 cycles)
Curve A—Current in dynode (telephone) circuit
Curve B—Current in anode circuit
Curve C—Current in tuned circuit

Figure 11 shows the effect of a signal impressed upon the grid of the pliodynatron. *The dynode high frequency current remains unchanged in amplitude, but its average value varies sinusoidally between complete rectification in the positive direction and complete rectification in the negative direction.* The anode current shows the same rectified components, in opposite phase. This was necessary, since all electrons lost by the dynode go to the anode. The current in the tuned circuit (lower curve), on the other hand, shows very little change in amplitude, which justifies the assumption of constant amplitude of dynode voltage oscillation referred to above.

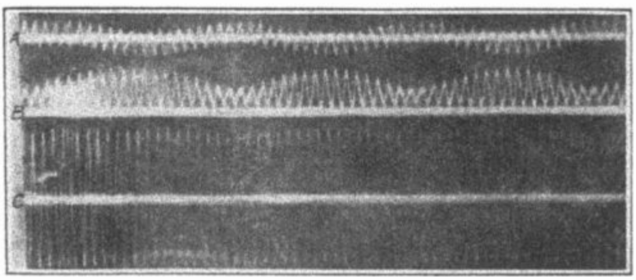


FIGURE 11—Pliodynatron detector with slightly a synchronous signal (heterodyne operation)
A—Telephone current
B—Anode current
C—Circulating (tuned circuit) current

The "snake frequency," as it may be called, is the difference between signal and local oscillation frequencies. This may be verified by measurements on Figure 12, in which the anode current is omitted and the signal voltage recorded (lower curve.) The constancy of the current amplitude in the tuned circuit, as evidence of the constancy of dynode voltage amplitude, is again to be noted.

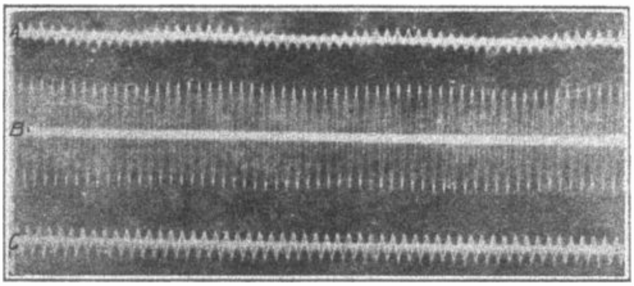


FIGURE 12—Pliodynatron detector in heterodyne operation, showing signal voltage
A—Telephone current
B—Circulating current
C—Signal voltage

Figure 13 shows the corresponding behavior for the magnetically-controlled dynatron. The upper curve is the dynode current, modulated by the signal; the middle curve is the signal *voltage*, that is, the voltage at the terminals of the control coil, which is wound around the tube (Figure 9). The controlling *magnetic field* is 90 degrees out of phase with this control voltage, as is very evident in the oscillograms of Figures 14 to 17. The lower curve is the circulating current. The characteristics are the same as for the pliodynatron.

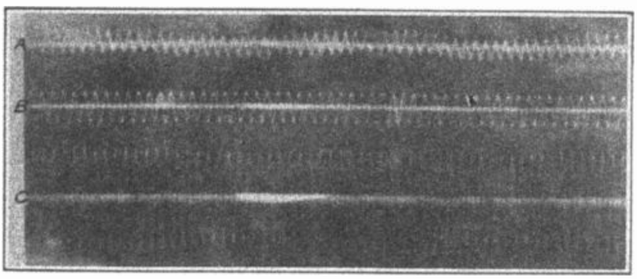


FIGURE 13—Magnetically-controlled detector in heterodyne operation
A—Telephone current
B—Signal voltage
C—Circulating current

Figures 14 to 17 show the effect of signal frequencies which are not near the frequency of the local oscillation. These films were taken with the magnetically-controlled dynatron.

Figure 14 shows the effect of a 40 cycle signal of approximately the same amplitude as the signals in Figures 11 to 13. The modulation of the amplitude of the oscillating current is clearly shown, and the dynode current is probably modulated in the same ratio; but there is no sign of rectification.

Figure 15 shows the effect of a much larger 40 cycle signal producing complete modulation of the dynode current. Here the dynode current is slightly distorted, owing to the asymmetric amplification, but even in this extreme case there is very little rectification. Figures 16 and 17 show the effect of signal fre-

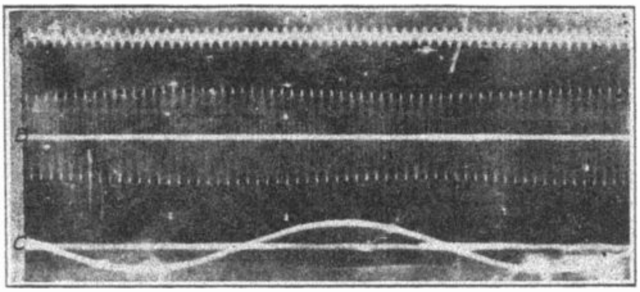


FIGURE 14—Magnetically-controlled detector oscillating at 1,500 cycles with moderate 40 cycle signal

A—Telephone current
B—Circulating current
C—40 cycle voltage

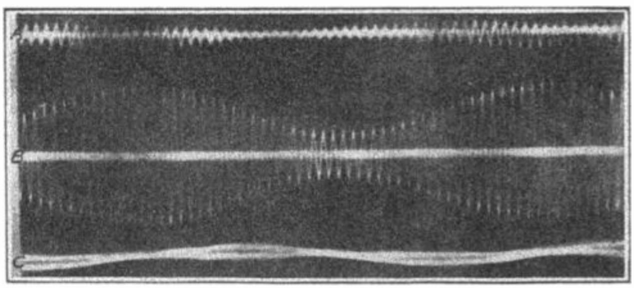


FIGURE 15—Magnetically-controlledd etector oscillating at 1,500 cycles with large 40 cycle signal

A—Telephone current
B—Circulating current
C—40 cycle voltage

quencies approximately 1/7 and 1/4 of the dynatron frequency. The dynode current is perfectly symmetrical, with no sign of rectification, in spite of complete modulation. These results, showing the absence of rectification except at synchronous frequeney, are completely in accord with the theory of operation outlined above, and with the mathematical analysis which follows.

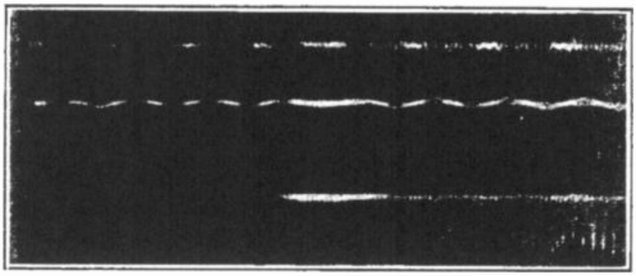


FIGURE 16—Magnetically controlled detector oscillating at approximately 1,500 cycles with 250 cycle signal

A—Telephone current
B—Signal voltage
C—Circulating current

Oscillographic tests of purely synchronous operation are shown in Figures 21 to 26, and are equally in accord with the theory. Before discussing them, however, the mathematical analysis will be briefly outlined.

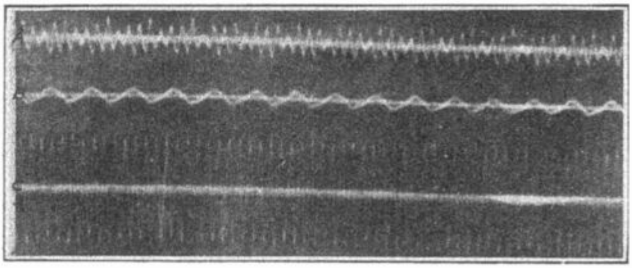


FIGURE 17—Magnetically-controlled detector oscillating at 1,500 cycles with 500 cycle signal

A—Telephone current
B—Signal voltage
C—Circulating current

MATHEMATICAL ANALYSIS

The analysis will be applied specifically to the pliodynatron. The same analysis will hold for the magnetically-controlled dynatron if magnetic field is substituted for grid voltage.

The circuit is shown again in Figure 18. In order to simplify the analysis, the effect of the inductance of the telephones will be neglected for the present. The results will then be applicable to oscillographic tests, and to audible operation with low

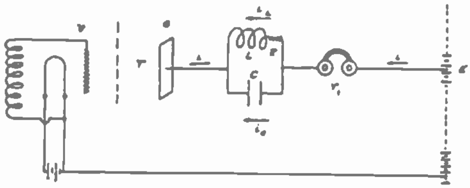


FIGURE 18—Circuit of pliodynatron detector

impedance telephones, or with a pure resistance output circuit working into a pliotron amplifier. In practice it is found that the inductance of the telephones has no noticeable effect on the operation until its ratio to distributed capacity approaches the value sufficient for sustained oscillations. Resistance in the dynode circuit has no effect except to reduce by that amount the negative resistance of the circuit, as has been shown in the previous paper already mentioned. This resistance will be included in the general solution (Equation 8.)

The operation is completely determined by the constants of the circuit and the two electrical characteristics of the pliodynatron, namely:

(a) NEGATIVE RESISTANCE

This is defined empirically from the volt-ampere characteristic as the slope of the volt-ampere curve (Figure 19):

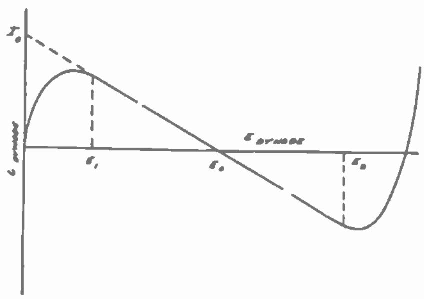


FIGURE 19—Volt-ampere characteristic of pliodynatron at constant grid voltage

$$\bar{r} = \frac{\partial e}{\partial i} = -\frac{E_o}{I_o};\ E_1 < e < E_2 \qquad (1)$$

The positive current from the dynode thru the vacuum is then given by the experimental characteristic (Figure 19) as

$$i = I_o - \frac{e}{r} \tag{2}$$

where r is the positive numerical value of the negative resistance.

(b) Grid Control Constant

This is defined in the pliotron as

$$g = \frac{\partial\, i_p}{\partial\, e_g}$$

where e_g is the control grid voltage and i_p the anode current at constant voltage. In the case of the pliodynatron only a small part of the current represented by i_p strikes the anode. The rest, a nearly constant fraction, passes thru the anode to the dynode, and is represented by the current of primary electrons, I_o, which strikes the dynode. Hence the grid control characteristic of the *primary* dynode current will be similar (Figure 20) to that of the pliotron, and the corresponding control constant may be defined as

$$g^{d}{}_{ynode} = \frac{\partial\, I_o}{\partial\, V_g} \tag{3}$$

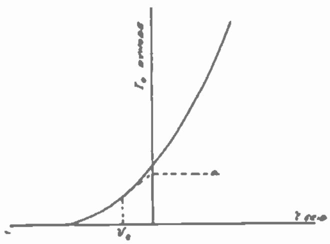


Figure 20—Grid-control characteristic of pliodynatron

The primary dynode current may then be represented, for small amplitudes of grid voltage, by the linear equation

$$I_o = a + g\, v \tag{4}$$

This equation holds with sufficient accuracy for all operating conditions, since the anode voltage is always constant and the signal amplitude small.

A. Dynode Oscillations with no Signal

It will be convenient to derive first the equation for the dynode oscillation when no signal is impressed.

Referring to Figure 18, the electrical behavior of the circuit is defined by Equation 2 above, together with the three following equations:

$$R\,i_L + L\frac{d}{dt}(i_L) = E - i\,r_1 - e \tag{5}$$

$$i_c = C\frac{d}{dt}(E - i\,r_1 - e) \tag{6}$$

$$i = i_L + i_c \tag{7}$$

Elimination of i i_L and i_c between these four equations gives, as the differential equation of the circuit when no signal is impressed:

$$\frac{d^2 e}{dt^2} + \left[\frac{R}{L} - \frac{1}{C(r-r_1)}\right]\frac{de}{dt} + \frac{1}{LC}\left(1 - \frac{R}{r-r_1}\right)e + \frac{I_o(R+r_1) - E}{\frac{LC(1-r_1)}{r}} = 0 \quad . \; . \; . \tag{8}$$

The solution of which if $\left[\frac{R}{L} - \frac{1}{(r-r_1)C}\right]^2 - \frac{4}{LC} < 0$, is

$$E_o + \frac{E - E_o}{1 - \frac{R+r_1}{r}} + A\varepsilon^{-\frac{1}{2}\left[\frac{R}{L} - \frac{1}{(r-r_1)C}\right]t} \sin\left[\sqrt{\frac{1}{LC} - \left(\frac{R}{2L} + \frac{1}{2C[r-r_1]}\right)^2}\,t - \alpha\right] \quad ($$

where A and α are arbitrary constants representing amplitude and phase, respectively, at the time $t=0$; and E is the impressed dynode voltage, which need not be the voltage E_o at which the dynode characteristic crosses the axis (Figure 19).

This solution represents a sine wave oscillation of dynode voltage, which will die out, continue unchanged, or increase indefinitely in amplitude according as the damping factor, which is the multiplier of t in the exponential term, is positive, zero, or negative. The case under consideration is that in which the damping is slightly negative, so that continuous stable oscillations are maintained. It is evident that the amplitude cannot increase indefinitely, since the constant value r of the negative resistance which Equation 9 assumes is limited to the range of voltage represented by E_1 . . . . E_2 (Figure 19). Beyond the range E_1 . . . . E_2 the resistance is positive, so that oscillations of greater amplitude are strongly damped.

The frequency of oscillation is slightly lowered by the negative resistance, as represented by the parenthesis $\left(\frac{R}{2L} + \frac{1}{2(r-r_1)C}\right)^2$ under the radical. This is the same as the damping coefficient, with the sign between the terms changed. When the damping coefficient is small, which is the condition for best operation, these two terms must be approximately equal. Hence the frequency correction due to resistance is approximately 4 times that of the freely oscillating tuned circuit. This is of the order of one part in one million in radio frequency circuits.

B. Dynode Oscillation with Signal Impressed on Grid

The effect of varying grid voltage is to vary the value of I_o (Figure 19) according to Equation 4, and hence the negative resistance (Equation 1). Substituting the values of r and I_o from Equations 1 and 4, Equation 2 becomes

$$i=(a+gv)\left(1-\frac{e}{E_o}\right) \tag{10}$$

Let the signal voltage impressed on the grid be a sine wave of frequency ω and amplitude V, so that

$$v=v_o+V\sin\omega t \tag{11}$$

Equation 2 then becomes

$$i=[a+gv_o+gV\sin\omega t]\left[1-\frac{e}{E_o}\right] \tag{12}$$

Combining Equation 12 with Equation 5, 6, and 7, and elim inating i, i_c, and i_L, one easily obtains the differential equation for dynode voltage when a signal is impressed on the grid:

$$\begin{aligned}\frac{d^2e}{dt^2}&+\left[\frac{R}{L}-\frac{1}{CE_o}(A+gv_o+gV\sin\omega t)\right]\frac{de}{dt}\\&+\frac{1}{LC}\left[1-\frac{R}{E_o}(a+gv_o+gV\sin\omega t)-\frac{LgV\omega}{E_o}\cos\omega t\right]e\\&+\frac{R}{LC}(a+gv_o+gV\sin\omega t)+g\frac{V\omega\cos\omega t}{C}-\frac{E}{LC}=0\end{aligned} \tag{13}$$

This equation is probably soluble. An exact solution is unnecessary for the present purpose, however, as a simple approxi mation will cover the operating conditions under discussion.

Comparing Equation 13 with Equation 8, remembering[4] that $r_o-\dfrac{E_o}{a+gv_o}$, it is seen that the co-efficient of e, which represents the damping factor, is the same in both cases except for the term $g\,V\sin\omega t$. This term, which represents the amplitude of the signal voltage, is always small compared with $(a+g\,v_0)$ and *its average value is zero*. A signal whose frequency, ω, is low compared with the frequency of the dynatron oscillations will therefore produce small periodic variations, of frequency ω,

[4]In order to shorten Equation 13, the telephone resistance r has been taken as zero. Its effect is purely additive, as in Equation 8, and all the above conclusions hold if it is included.

in the amplitude of the dynode voltage oscillations, and hence of the current in the tuned circuit, which is proportional to dynode voltage. This is shown for a very large signal (compared with radio signals) in Figure 14, and for enormous signals in Figures 15 to 17. For signal frequencies equal to or higher than the dynatron frequency, however, the inertia of the tuned circuit prevents its response to these rapid variations in damping, and only the average value of the damping factor is effective. This is the same as when there is no signal. Hence these radio frequeney signals will have no effect upon the amplitude of dynode voltage oscillation. The same argument applies to the average value of dynode voltage, represented by the constant term in Equation 9.

As regards the effect of the coefficient in the third term of Equation 13 on the frequency,[3] it is easily verified that for normal signals the variable part of this coefficient is of the same relative magnitude as $\frac{R}{r-r_1}$. in Equation 8, which has been shown to affect the period by only about one part in one million. For very strong signals, the last member in the parenthesis of this coefficient becomes appreciable, and produces a "synchronizing effect" between the signal and the local oscillating circuit, which is of importance in "zero beat" reception. The correctness of these conclusions both as to frequency and synchronizing effect can be verified from the oscillograms shown in Figures 21 to 26.

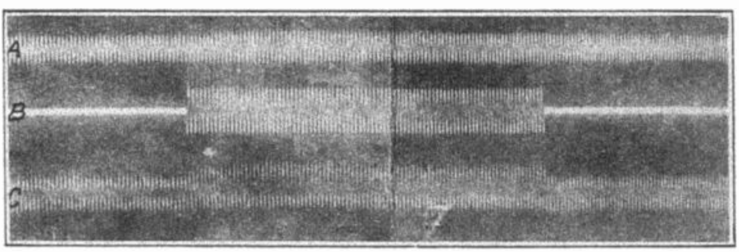


FIGURE 21—Oscillographic test of pliodynatron detector with synchronous signal (zero beats)
A—Circulating current
B—Signal voltage
C—Telephone current

Equation 9 may therefore be taken as a sufficiently accurate solution of (13) under the conditions of operation.

C. DYNODE CURRENT WITH SIGNAL IMPRESSED ON GRID

The dynode current is obtained by substituting the value

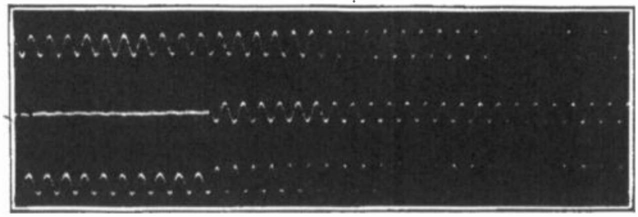


FIGURE 22—Pliodynatron detector with synchronous signal. Signal applied in phase
A—Circulating current
B—Signal voltage
C—Telephone current

of e from Equation 9 in Equation 12. Writing A' and $\frac{\omega_o}{2\pi}$ for the amplitude and frequency of the oscillation in Equation 9, namely,

$$e = E_o + \frac{E - E_o}{1 - \frac{R + r_1}{r}} + A' \sin(\omega_o t - \alpha) \quad \cdot \cdot \cdot \quad (g')$$

and writing r_0 for the normal negative resistance (with no signal), one thus obtains, from (9′) and 12).

$$\overbrace{= \frac{E - E_o}{r_o\left(1 - \frac{R + r_1}{r_o}\right)}}^{\text{Direct Current Component}} + \overbrace{\frac{A'}{r_o} \sin(\omega t - \alpha)}^{\text{Dynatron Frequency}}$$

$$+ \overbrace{\frac{E - E_o}{E_o\left(1 - \frac{R + r_1}{r_o}\right)} g V \sin \omega t}^{\text{Signal Frequency}} + \overbrace{\frac{A' g V}{2 E_o} \cos([\omega - \omega_o] t - \alpha)}^{\text{Difference Frequency}}$$

$$\overbrace{- \frac{A' g V}{2 E_o} \cos([\omega + \omega_o] t - \alpha)}^{\text{Double Frequency}} \qquad (14)$$

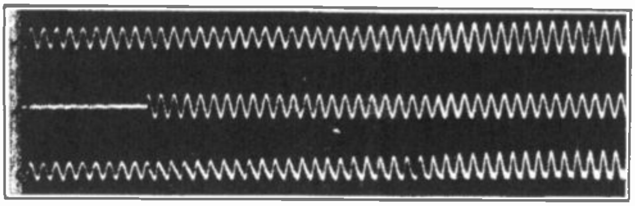


FIGURE 23—Pliodynatron detector with synchronous signal. Signal applied 180° out of phase
A—Circulating current
B—Signal voltage
C—Telephone current

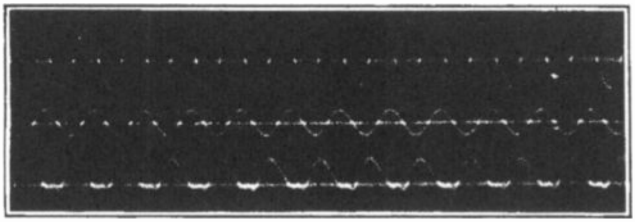


FIGURE 24—Pliodynatron detector in synchronous operation. Signal frequency lower than normal pliodynatron frequency

A—Circulating current
B—Signal voltage
C—Telephone current

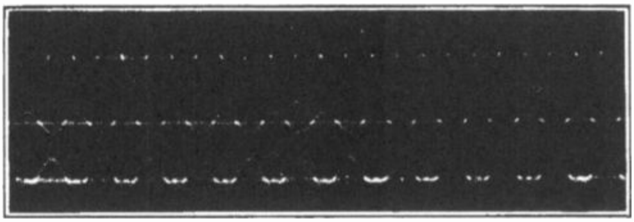


FIGURE 25—Pliodynatron detector in synchronous operation. Signal exactly in synchronism

A—Circulating current
B—Signal voltage
C—Telephone current

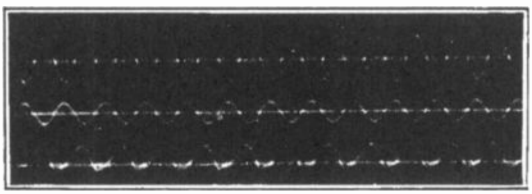


FIGURE 26—Pliodynatron detector in synchronous operation. Signal frequency higher than normal pliodynatron frequency

A—Circulating current
B—Signal voltage
C—Telephone current

It is seen that the dynode current contains, in general, a constant direct current component, the dynatron oscillation, a signal frequency component, a difference frequency or heterodyne component, and a component which is the sum of signal and local oscillation. If the impressed dynode voltage is the voltage E_o at which the dynode characteristic crosses the axis (Figure 19), both the constant term and the signal frequency component are zero. The dynode current, which is the current thru the telephones, contains then only the dynatron oscillation and the sum and difference of the dynatron and signal frequencies. There is no rectification and no direct current component. Hence, unless the difference frequency is audible, there will be no audible sound in the telephones, even when the signal is modulated or interrupted. This conclusion holds, of course, only for signal amplitudes for which the assumption of linear grid amplification is justified.

In the case of exact synchronism between local oscillation and signal the difference frequency disappears, or rather becomes a direct current component the amplitude of which is propor tional to that of the signal. Besides this direct current component there remains, if $E=E_o$, only the dynatron sine wave and its second harmonic. This is clearly shown in the following oscillograms. (Figures 21 to 26.)

Oscillographic Tests of Synchronous Operation

As a further test of the theory of operation outlined above, oscillographic tests were made with a pliodynatron oscillating at low frequency, with a synchronous signal, obtained from a pliotron oscillator, impressed on the grid. In order to obtain large dynode currents for the oscillograph, the anode voltage was raised so high that the dynode characteristic was considerably distorted, being much longer on the negative side (below the axis) than on the positive, and considerably curved. The result was that it was not practical to operate at dynode voltage E_0, so that the average dynode current was not zero; and the amplitude of the current in the oscillating circuit increased considerably when the signal was applied, owing to the curvature of the characteristic. In spite of this the dynode current curves show very well the characteristic features predicted by Equation 14, namely, the direct current component and the second harmonic. In each film the upper and lower curves are the currents in the tuned circuit and dynode (telephone) circuit, respectively, and the middle curve the signal voltage. The signal amplitude was from 5 to 10 volts. This is, of course, large compared with normal signals.

Figure 21 shows a typical signal dash. The dynode current is increased slightly in amplitude and raised bodily, producing a direct current component of current in the telephones.

Figure 22 is a lower frequency, and shows clearly the sine wave form of the dynode current before the signal is impressed, and the second harmonic after the signal is impressed. In this case, the signal and dynode happened to be in exactly the same phase at the instant when the signal was impressed on the grid, so that only 3 or 4 cycles were required to build up the rectification and the new wave form. In general, from 10 to 20 cycles are required, and the number might be much larger in the case of a very weak signal.

Figure 23 shows a case in which the signal was just 180° out of phase with the dynode current at the instant it was applied. The distorting effect of the signal is evident at once, but it is only after 30 cycles that the dynatron oscillation is finally pulled into phase and shows its characteristic features.

Figures 24, 25, and 26 show the effect of a slight difference in frequency between the signal and the local oscillation. The synchronizing force of the signal is sufficient to hold the dynatron in synchronism, but a marked distortion of the dynode current is produced. In Figure 24 the signal frequency is slightly lower than that of the dynatron; in Figure 25 the two are exactly equal, and in Figure 26 the signal is higher. The second harmonic is clearly shown in Figure 25.

Remarks on Operation

Operation of the pliodynatron as heterodyne receiver has given the following results:

(a) The detector action is reliable and reproducible.

(b) The sensitiveness is the same as that of separate heterodyne or normal self-heterodyne. By adjusting the ratio of inductance and capacity of the telephones, so as to utilize negative resistance amplification in the telephones, the sensitiveness can be increased to the limit obtainable by regeneration in self-heterodyne.

(c) The operation is the same whether the filament is heated with alternating or direct current, and no noise due to the alternating current can be detected when the dynode voltage is properly chosen.

(d) No signal can be detected by a circuit coupled to the oscillating circuit of the pliodynatron, even after rectification by a pliotron detector. This proves the absence of signal frequency in the oscillating circuit, that is, the absence of modulation of the amplitude of local oscillations by the signal, thus confirming the conclusions of the mathematical analysis and the oscillograms.

(e) The stray ratio is superior to that of ordinary rectifying heterodyne detectors.

Research Laboratory,
General Electric Company,
Schenectady, New York,
February 3, 1922.

SUMMARY: The dynatron is a tube having a filament, an anode, and a supplementary anode called a "dynode." The tube operation is based on the secondary emission of electrons from the anode. Its use as an oscillating detector for heterodyne reception is described, and the mathematical theory of its operation is given. The theory is verified experimentally with oscillographic illustrations.

APPLICATION TO RADIO OF WIRE TRANSMISSION ENGINEERING*

By

LLOYD ESPENSCHIED

(DEPARTMENT OF DEVELOPMENT AND RESEARCH, AMERICAN TELEGRAPH AND TELEPHONE COMPANY, NEW YORK)

One of the most interesting aspects of the development of radio during the last few years, and particularly of radio telephony, is the obvious convergence of its technique with that of wire transmission. It is, of course, the advent into both of these arts of that remarkable device, the electron tube, which is responsible for the close technical relations which now exist between them.

This community of interest, however, altho thus greatly stimulated by a device of such range of utility as to find important applications in both arts, is not due primarily to anv device *per se*, but rather to the fact that both type of systems are subject fundamentally, as communication systems, to the same gen eral requirements and design considerations concerning their intelligence-carrying capabilities. These underlying communi cation requirements lead to similar considerations in both types of systems as to the efficiency and fidelity with which the transmission of intelligence is effected and give rise to a transmission background, as it were, which is common to both arts.

The engineering handling of the transmission problems, which arise from these fundamental communication requirements, has been quite highly developed in the older of the two arts—wire transmission—in connection with telephone repeaters and carrier telephone and telegraph systems. It should be, therefore, interesting and profitable to apply some of the transmission technique thus developed in the wire art to several of the more important radio problems. In so doing, we obtain rather new viewpoints of radio transmission and a useful correlation of it with the better established wire methods. It is hoped, therefore, that the picture which is presented of radio and wire transmission, treated from a common standpoint, may con-

*Presented before THE INSTITUTE OF RADIO ENGINEERS, New York January 23, 1922. Received bv the Editor, April 17, 1922.

tribute to a better appreciation of both arts by radio and wire engineers alike and may make clear the underlying transmission principles which are common to them.

Principal among the problems of electric communication is the one of delivering at the receiving end the required volume of signal with the necessary freedom from interference. The delivering of the required volume is a matter of controlling the transmission losses and the amplification of the system; while the obviating of interference is, of course, concerned with the reduction of the ratio of the interfering to the signaling energy.

Transmission Losses

In considering these factors we will take up first the primary one of the losses which are suffered by the carrier waves as they are propagated thru the transmission medium. In both wire and radio transmission, of course, the actual propagation of the electromagnetic wave energy occurs in the "ether," the difference being that in the wire case, the waves are bound to a guiding path, whereas in the radio case they are transmitted freely in all directions and bound merely to the earth's surface. This difference in the mechanism of transmission gives rise to an important difference in the transmission losses occurring in the two cases. In order to assist in visualizing the two cases they are indicated diagrammatically in Figure 1.

WIRE AND RADIO TRANSMISSION SYSTEMS

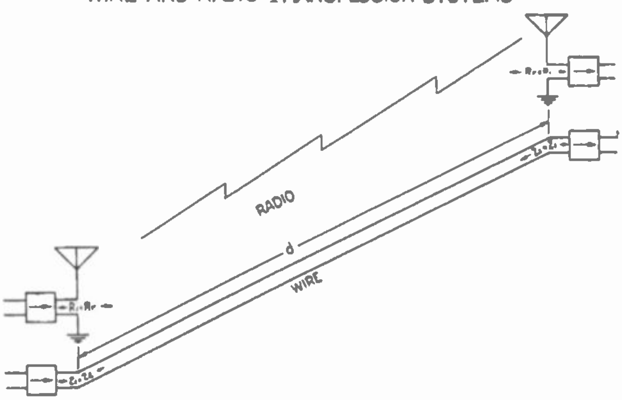


Figure 1

Referring first to the wire case, the law in accordance with which the current and voltage strength decrease as the transmission wave travels along the wire, is the familiar one of atten uation.

$$I_1 \varepsilon^{-al} = I_2 \qquad (1)$$
$$E_1 \varepsilon^{-al} = E_2$$

which simply expresses the fact, that as the wave proceeds along the wire, the losses in the resistance of the conductor and in the insulation, extract for each mile a certain definite proportion of the voltage and current which arrives at that point. After traveling (l) miles the original current I_1 is attenuated down to a value $I_1 \varepsilon^{-al}$ which represents the received current I_2. This is the same general law of damping as applies to the dying down of the voltage and current in an oscillation circuit, except that here the damping is with respect to distance along the line rather than time. We are assuming, of course, that the circuit is so terminated as to avoid reflection effects at the terminals—a condition readily met, by making the terminal impedance equal to the characteristic line impedance. This is indicated in the figure by the designations, Z (internal) equals Z (line). A similar relation is taken for the radio case. The "line" impedance is here the antenna radiation resistance while the "terminal" impedance is the resistance internal to the antenna and the apparatus, assuming resonance; thus R (internal) equals R (radiation).

We know that in radio there are two distinct causes of the transmission loss: (1) that due to the spreading out of the waves, which is characteristic of non-guided wave transmission; and (2) that due to absorption in the air and earth's surface, which extracts a definite percentage loss for each mile of the radio circuit and which conforms, therefore, to an exponential law similar in its general nature to that of wire attenuation.

This transmission law, as expressed by the familiar Austin-Cohen formula, is given in the appendix.[1] In order to express the radio transmission loss in some general manner which will be comparable to the expression of wire transmission loss, these conditions have been taken for the radio case:

(1) That we will use as the measure of the transmission loss the ratio of the square root of the power radiated from the sending antenna to the square root of the power delivered in the receiving antenna. This is, of course, the same criterion as is used for wire transmission.

[1] Measurements on ship-shore transmission made since the above was written, indicate that the Austin-Cohen law holds quite well for frequencies as high as about 1,000,000 cycles.

(2) The radiation resistance of the two antennas, sending and receiving, are made equal, analogous to the equality of line impedance at the two ends of the wire system.

(3) Also the internal antenna resistance, which corresponds to the terminating impedance in the wire case, is made equal to the radiation resistance for both ends. This is the condition of maximum power transfer between the "line" and the terminal.

These assumptions set up the two cases, radio and wire, on a comparable basis and facilitate a comparison of them. They are favorable to radio in that *they do not take account of practical limitations which obtain in antennas.* The radio curves below should be read, therefore, as giving the minimum possible losses for daylight transmission over water.

These curves show the manner in which the transmission loss varies with distance, for various frequencies, for both radio and wire. The ordinates are plotted in terms of the logarithm of the ratio of the sent to the received currents, or voltages, in circuits of equal impedances. In so doing we are plotting the losses on the straight attenuation basis upon which they are usually plotted in the wire art; that is, the ordinates represent the exponent (αl) of the wire attenuation law, and may be directly interpreted in terms of miles of standard cable[2] by multiplying by 21. approximately. The advantage of dealing with the exponent rather than the current ratio itself is the very considerable one which is characteristic of logarithms, namely, that when thus expressed the individual losses and gains thruout a system may be summed up algebraically, and the over-all transmission equivalent of the system thus readily determined.

It should be noted that the transmission loss given in the radio curves is that obtaining between the point at which power is delivered to the ether at the sending end and that at which it is delivered to the dissipative load of the receiving antenna circuit. In Figure 1 these points are represented by R_r at the transmitter and R_i at the receiver. If at the sending end, we start with the power developed within the generator, meaning

[2] For the mile of standard cable the attenuation α (at 800 cycles) equals 0.109. Therefore the equation for current ratio, in terms of miles of standard cable, becomes

$$\frac{I_1}{I_2} = \varepsilon^{\alpha l} = \varepsilon^{0.109 \times l}$$

from which

$$l = \frac{1}{0.109} \log_\varepsilon \frac{I_1}{I_2} = 21.13 \log_{10} \frac{I_1}{I_2}$$

in R_t instead of R_r, then the power ratio is simply doubled, for the conditions assumed, and the attenuation is 0.15 or about 3 miles greater than given in the curves. The curves can be used *for obtaining the loss in any practical case* simply by taking the minimum loss as given by the curves and adding thereto the additional loss obtaining in the actual antenna.

Referring now to Figure 2—the transmission losses in the two cases are given for distances up to 200 miles (320 km.). The straight lines represent the wire losses, the bending-over curves the radio losses. Of the radio curves, the dash lines give the spreading-out losses alone, while the full lines give the total losses, including absorption.

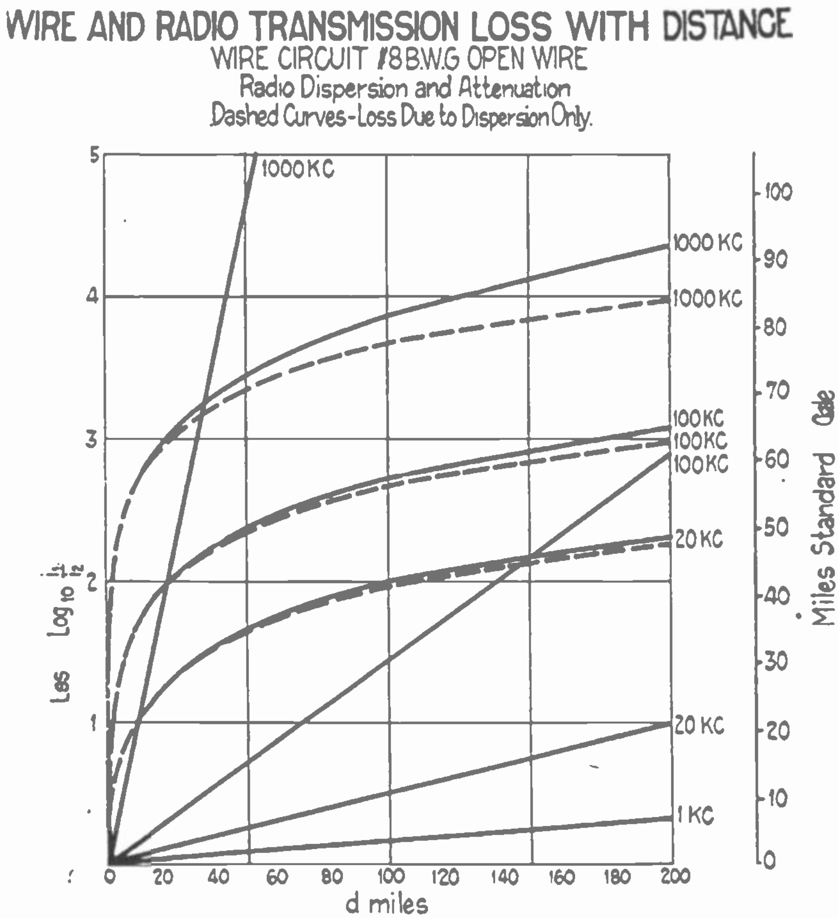


FIGURE 2

The first thing one observes is the difference in the nature of the two sets of curves—the wire losses being represented by straight lines, because of their exponential law and the fact that it is the logarithm or the exponent itself which is being plotted, while the radio curves jump up rapidly at first and then straighten out, in accordance with the "inverse-with-distance" law.

The second thing one notes is the fact that as a result of the large initial (or "jump off") loss, the radio values run on the whole higher than do the wire for the more usable wire frequencies and very much greater than the wire losses at telephone frequencies 1 k. c.).[3] For the wire case the number 8 Birmingham wire gauge open wire circuit is taken.[4] This is the standard long distance telephone circuit of the United States. The constants are given in the appendix.

A third characteristic which one notes in the radio curves is that the losses are greater for the higher frequencies or, conversely, lower for the lower frequencies. This is because the efficiency of the antenna has been kept constant for all frequencies. In practice the transmission losses at the lower frequencies are higher than here indicated, because of limitations in antenna heights.

Were we to take the ideal condition *as regards the transmission medium itself*, where for wires there is no conductor or dielectric loss, and for radio there is, likewise, no earth or air absorption loss, we would note: (1) that, for wires, there would be no attenuation whatever, the curve following along the X axis; (2) for radio, there would remain the loss due to dispersion, inherent in the unguided method of transmission, the magnitude of which loss is, ot course, very substantial. The dash-line radio curves show the radio losses without attenuation, the full-line curves with attenuation.

Considering the actual condition, where there is dissipative loss in the transmitting medium, we find that for moderate distances, up to 200 miles (320 km.), as plotted in Figure 2, the wire losses are in general less, and at telephone frequencies very much less, than the radio losses. The low wire attenuation at telephone frequencies is, of course, in keeping with experience and accounts for the economical terminal apparatus which is employed in telephone practice. Likewise the relatively high losses for radio accounts for the large amplification at either the sending or receiving end or both, which experience has proven to be necessary. This brings in an interesting side-light, namely, that altho in radio the transmission medium is provided by

[3] 1 k. c. is 1 kilocycle per second or 1,000 cycles per second.
[4] Diameter of number 8 Birmingham wire gauge wire = 0.165 in. = 0.42 cm.

nature, the effective *use* of this medium is not as economical as might be expected, because it requires considerable equipment, amplifiers at both ends for overcoming the large attenuations, selective means for dividing-up frequency range and thereby "multiplexing" the ether, and antennas for getting into the medium and out again.

For the higher frequencies, the wire attenuations increase relatively more rapidly than the radio, thus limiting the frequency range which can be employed on wires without likewise running into large amplification requirements. For example, the loss at 100,000 cycles for a distance of 200 miles (320 km.), is about as great over wires as the minimum loss which it is theoretically possible to obtain over radio.

Referring now to the attenuations for longer distances, as given in Figure 3, it is of interest to note that for distances of

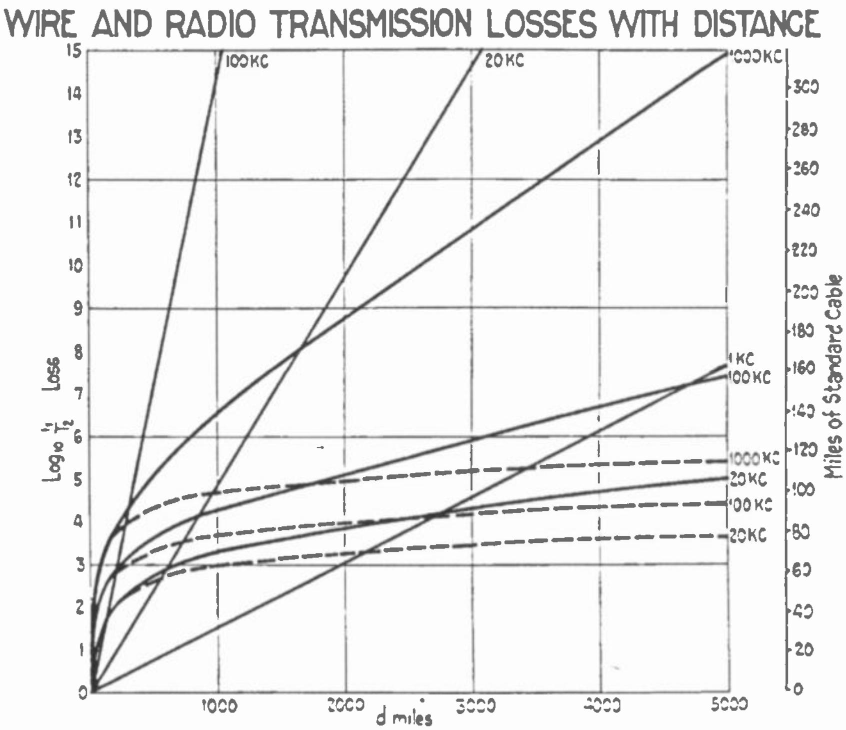


FIGURE 3

the order of 2,000 to 3,000 miles (3,200 to 4,800 km.), the lower radio frequency curves cross the 1,000 cycle wire curve, meaning that for these distances it is possible for radio transmission to be as efficient as straight telephone transmission. The wires present to carrier frequencies for these long distances losses which are generally greater than prevail for radio.

These attenuation relations cannot be directly converted into an economic comparison, however, for the economies depend not upon the attenuation itself, but upon, among other factors, the cost involved in *overcoming* the losses by means of amplification; and this cost in turn depends largely upon the extent to which the amplification can be applied at weak powers, as by the frequent application of telephone repeaters. By applying repeaters every few hundred miles in the wire case, the attenuation is prevented from piling up and the amplification is handled at relatively weak and therefore economical power levels. This brings us to the point of requiring that the attenuation values given above be considered in reference to the amplification and power required to overcome them and yield the necessary volume of transmission at the receiving end over and above interference.

Interference and Its Effect Upon the Transmitting Power Required

In both the radio and wire cases there is always present in the transmission medium a certain amount of stray wave energy which tends to interfere with the proper reception of the message-carrying waves. It is necessary that the communication waves arrive at the receiving end of the system with such power as to be large compared with the interfering waves—by a factor determined by the type and grade of communication involved. Inasmuch as the stray energy always has some finite value, this requirement of freedom from interference will determine, in the radio case as well as in some types of wire transmission, the minimum wave power required at the receiving end of the transmission system.

In the wire system the minimum power requirement may be expressed directly in terms of a power, or—as it is usually—a current, in the receiving apparatus, the "transfer" between the line and the receiver being a constant and efficient relation. In radio the power delivered out of the ether or "line" into the receiving antenna is so largely a function of the antenna dimensions that it is necessary to express the necessary received power in terms of the received field, usually simply as a field intensity—and not as a certain current in the terminal apparatus. Of course the transmission loss occasioned by getting from the "ether" into the receiving circuit does not affect the interference relation but merely the absolute amount of terminal amplification required. On the other hand the transmission loss occasioned by getting into the "ether" at the sending end affects the interference relation vitally, as we shall see.

Transmission Levels

This necessity of having to keep the power of the received waves above the interference level may be visualized by reference to Figure 4.

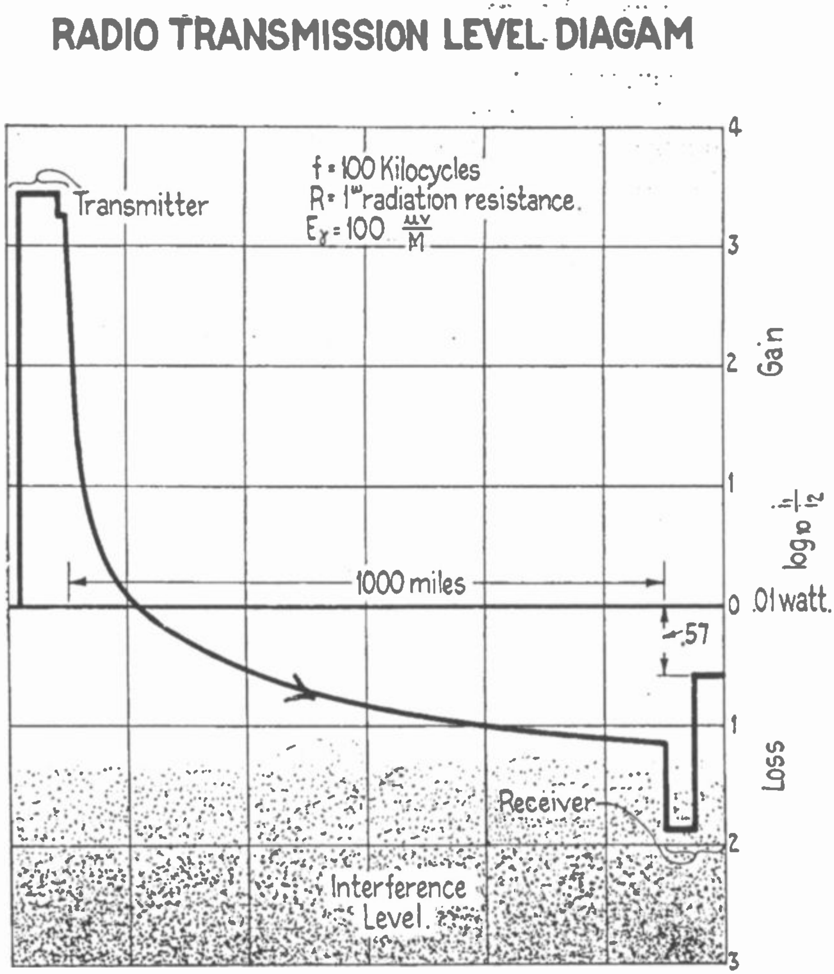


Figure 4

Here we have what in wire practice is called a "transmission level" diagram. Such a diagram is useful in showing what goes on in the system from the power and interference standpoints. The vertical scale is plotted in terms of the transmission level expressed as the logarithm of the current or field intensity ratios, and the horizontal scale represents progression along the system. For illustration purposes, the presence of interference is indicated at the bottom of the transmission-level scale by the shading.

Tracing thru the diagram we proceed as follows:

The point of "zero" level is taken roughly as that corresponding to the power delivered into a telephone circuit by a certain telephone transmitter when spoken into by the average talker, and is here taken to be 0.01 watt. As the voice currents are amplified to power proportions in the transmitting station, at the left, the transmission level is greatly increased, as illustrated by the vertical jump in the curve. The amplified voice currents are assumed to be converted by modulation into high frequency currents at this high power level and put into the antenna. The high frequency loss in the antenna system is indicated by the perpendicular jog in the curve. The drooping-off curve then commences, starting with a point which represents the power usefully applied to the ether in accordance with the expression I^2R, where I is the antenna current and R is the radiation resistance. The level curve falls off in accordance with the transmission loss curves previously discussed, as it extends across the transmitting medium to the receiving station. It will not do to permit the transmission level to fall as low as that of the interference, so I have shown that the transmission reaches the receiving station before dropping down very far into the interference level. At the receiving point a further transmission loss occurs in getting into the receiving antenna circuit, shown by the drop in the curve, but this loss obtains for the interference as well as the desired signals and does not affect the interference ratio. The terminal amplification brings the level up to that required for suitable audition and the difference between where this level leaves off and the original zero level, measures the overall transmission equivalent of the circuit, shown in this case as about $log_{10}\frac{I_1}{I_2}=0.57$, or about 12 miles of standard cable. This corresponds to a current ratio of about 4, a value ample for good "talk." Of course, in a one-way circuit the terminal amplification can be raised to any value desired. In a two-way circuit, however, a limit in the terminal amplification is imposed by interference between the two transmissions, as will be understood subsequently.

We may make the following useful observations from this curve:

1. The net transmission equivalent represents the difference between the over-all loss and the over-all gain.
2. The over-all gain is divided between the transmitting and receiving ends. We should like to throw as much of this amplification as possible to the receiving end because of the economy with which amplification can be provided at low powers.
3. The extent to which we can do this, however, is distinctly limited by the fact that the transmission level obtaining at the receiving end in the transmission medium must be held above a certain amount in order to overcome interference.
4. It is, then, the absolute intensity of the interference which determines the receiving power level required, and in turn this together with the attenuation back to the transmitting station which determines the transmitting power and, in turn, the amplification required.

Thus the two transmission features most fundamentally important in the radio communication system are: (1) the interference level, and (2) the transmission loss thru the medium. These once given, the other engineering considerations follow naturally. There are analagously fundamental factors in wire communication systems. In the latter case, however, the art has advanced to a point where means of controlling the interference level are available, so that the ratio of interference to transmitted power may be made small by decreasing the former rather than increasing the latter.

Minimum Transmission Levels Obtaining in Practice

The working values which should be assumed for the ratio between the transmission level of the received signals and the interference, depend upon the type of communication involved, whether it is telephone or telegraph, for example, and upon the grade of service to be given. There is a wide difference between the transmission level which will enable telephonic signals to be barely discerned by an expert ear and that which is required for a public service communication system which must provide sufficient operating margin to enable the average person to converse with ease and certainty under all ordinary conditions.

Under favorable static conditions, the transmission level can be permitted to fall to extraordinarily low values. When this condition is accompanied by a substantial reduction in the effective attenuation, which sometimes occurs at short wave lengths, especially at night, apparently due to the effective absence of ether absorption, then it becomes possible to "get thru" over relatively long distances with powers diminutive as compared with those required for giving a regular service. With these exceptional transmission conditions we are, of course, familiar. They are exemplified by the long distances reached at night by the amateurs, as across the Atlantic, and by the hearing of the normally 30-mile (48 km.) Catalina Island system in Australian waters. The transmission curves of Figure 3 account for these unusual long distance transmissions if we assume that the attenuation due to absorption is eliminated on these occasions by some natural cause. Thus, at 3,000 miles (4,800 km.) the curves for 1,000 kilocycles (300 meters), for example, show that were the absorption eliminated, the transmission equivalent would be improved by the difference between about 10.8 and 5.2 for $log_{10}\frac{I_1}{I_2}$ or 5.6, an improvement equivalent to a little over 100 miles of standard cable. The remaining or purely spreading-out loss of about 5 units, or 100 miles of standard cable, is then taken care of by the sending and receiving amplification.

Interference may occur in either or both of two ways—by the interference level rising to a point comparable with the normal transmission level at the receiving end of the ether circuit, or by the transmission level of the waves themselves dropping so low, due to excessive atmospheric absorption, as to fall below that of the atmospheric disturbances. For reliable transmission it is necessary, therefore, to deliver normally at the receiving end, a wave intensity sufficient to allow for the fluctuations which occur in atmospheric absorption and in the intensity level of the atmospherics. The importance of working to transmission level standards which give an adequate operating margin against interference, for the types of service required, will be appreciated from the foregoing. The following values of minimum transmission levels will be of value to know:

(a) For carrier wire telephone transmission at frequencies in the tens of thousands of cycles, the limiting interference may be our old friend "static," or interference from high frequency transients in power systems. Unless the lines are especially well transposed for these frequencies, the interference requires that the transmission level be kept above a minimum value of the order of $log_{10}\frac{I_1}{I_2}=1.2$ (about 25 miles of standard cable below zero level).

(b) While for radio telephone transmission the available data are as yet very meagre, we have obtained a few order-of-magnitude figures which should be of interest. For the Catalina Island radiophone system, for example, the minimum field intensity is estimated at roughly 1,000 microvolts per meter. The circuit is sometimes quite noisy during the summer months, altho not prohibitively so. In our ship to-shore radio telephone experiments along the Atlantic coast, we have on occasions worked with lower field intensities, as low as 100 microvolts per meter. The latter figure, however, gives a grade of service far below wire standards.

(c) The best data on the minimum permissible transmission level for radio telegraphy are those obtained from the experience in trans-Atlantic telegraph operation. The figures prevailing for present trans-oceanic radio-telegraph operation are understood to lie in the order of 10 to 100 microvolts per meter, depending upon individual cases and the time of the year.

The Net Transmission Equivalent

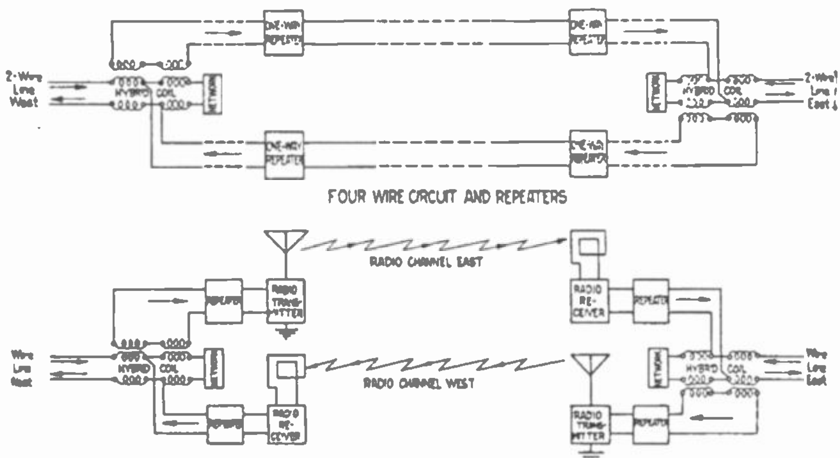
The net over-all transmission equivalent of the system is measured by the ratio of the transmitted to the received signaling power, and is shown in Figure 4 as the difference between the transmission levels at the two ends. This relatively small loss represents the difference between two large values, the transmission loss and the transmission gain thruout the system. Relatively small changes in either the attenuation or amplification may, therefore, cause large changes in the net equivalent of the circuit, thus tending to give rise to instability in the transmission performance of the circuit.

This problem of fluctuation becomes very serious with the use of very high frequencies, whether transmitted by wires or by radio. Were we to attempt to employ, for example, a million cycles for wire carrier transmission over considerable distances, as has been proposed, not only would the losses be very large, but they would be unstable, changing with weather conditions, so that the maintenance of a constant volume of transmission, would become extremely difficult. Similarly in radio transmission the fluctuations in the ether attenuation, particularly at short wave lengths where over long distances we experience the well-known "swinging" or fading effects, render the maintenance of a satisfactory volume of transmission a difficult problem. As noted, above these fluctuations, particularly as between day and night transmission with very high frequencies, may be enormous.

It is of value to the radio engineer to have some idea of the over-all circuit transmission equivalents which are necessary for satisfactory telephone communication. In the wire telephone art, the maximum equivalent between subscribers is ordinarily taken as about 30 miles of standard cable, or $log_{10} \frac{I_1}{I_2} = 1.4$. Under quiet conditions, considerably larger transmission equivalents can be talked over. The long distance toll lines themselves are usually designed for transmission equivalents of 0.5 to 0.75 or 10 to 15 miles of standard cable. These figures will serve as a general guide for the transmission equivalents which radio telephone circuits should provide. Where a radio circuit forms a link in a direct wire circuit as, for example, in the case of Catalina Island, it is desirable to work the radio link as close to a zero equivalent as possible, that is, to give out at the receiving end a volume nearly equal to that fed in at the transmitting end.

Two-Way Operation

When the two one-way radio channels are merged at their two ends into a regular telephone circuit for connection to the wire network, as illustrated in Figure 5, then there is a limit in the transmission equivalent which can be given over the radio part of the circuit.



TWO WAY RADIO CIRCUIT

FIGURE 5

This limit will be appreciated by reference to Figure 5. It is imposed by the tendency of the two one-way channels to form a round-trip circuit by "feeding-back" from one to the other via the voice frequency connecting circuit. If the total amplification around the circuit, including the voice-frequency line, exceeds the total losses in the circuit, "singing" will result. Were no line balance provided at the voice frequency terminals, then it would be impossible to operate the circuit at a zero equivalent. By setting up a balancing circuit at each end in the manner illustrated, a transmission loss is, in effect, inserted between the sending and receiving sides of the voice circuit, which tends to prevent this sing-around action. Actually, there is a limitation in the degree of balance which can be realized between the telephone line and the balancing network, especially if the telephone line is to be switched at a nearby central office, and this factor, together with the margin of safety which is required between the operating condition and the singing condition, prevents the radio channels from being operated much better than the zero equivalent. This whole matter of realizing in practice an adequate transmission equivalent, will be appreciated to be an especially difficult problem in the case of marine radio telephony, where the connection is switched from one vessel to another at varying distances.

It should be noted further, with reference to two-way operation, that the difficulty of effecting simultaneous sending and receiving at a station arises primarily from the large attenuation which must be overcome and the resulting large ratio between the energies transmitted into and received from the ether. The receiver must be prevented from being overloaded by the home transmitter and this, in general, requires that there be provided between the high frequency side of the transmitter and that of the receiver, a transmission loss comparable in size to that obtaining over the radio circuit itself. This "separating" transmission loss is ordinarily provided (a) by frequency-selecting circuits (tuned circuits and filters), the sending and receiving transmissions being placed on different frequencies, (b) by balance, as when using the blind spot of a loop-antenna receiver, and (c) by spatial separation between sending and receiving points, where the large step-off loss is used to advantage.

Transmission Levels on Combination Wire and Radio Telephone Systems

It will be of interest to trace thru the approximate transmis-

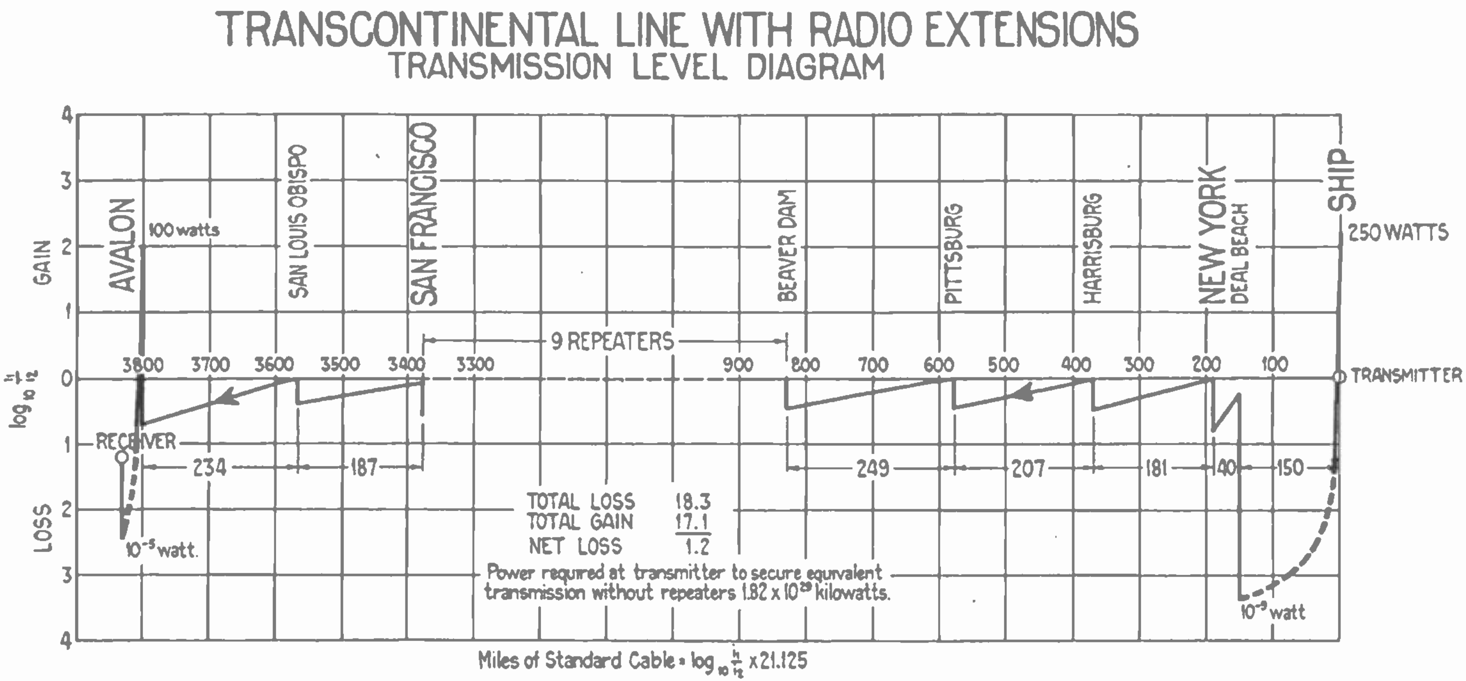

TRANSCONTINENTAL LINE WITH RADIO EXTENSIONS
TRANSMISSION LEVEL DIAGRAM
GAIN
LOSS
$\log_{10}\frac{I_1}{I_2}$
AVALON
100 watts
SAN LOUIS OBISPO
SAN FRANCISCO
BEAVER DAM
PITTSBURG
HARRISBURG
NEW YORK
DEAL BEACH
SHIP
250 WATTS
TRANSMITTER
RECEIVER
9 REPEATERS
3800
3700
3600
3500
3400
3300
900
800
700
600
500
400
300
200
100
234
187
249
207
181
40
150
10^{-5} watt.
10^{-9} watt
TOTAL LOSS 18.3
TOTAL GAIN 17.1
NET LOSS 1.2
Power required at transmitter to secure equivalent transmission without repeaters 1.82×10^{29} kilowatts.
Miles of Standard Cable = $\log_{10}\frac{I_1}{I_2} \times 21.125$

Fig re 6

sion levels which obtain for a radio system linked up with a long repeatered land line system.

In Figure 6, there is taken a rather striking example of this case in the trans-continental telephone line as connected up to radio extensions at its termini—to Catalina Island on the Pacific and to a vessel at sea on the Atlantic. The tranmission illustrated is that occurring from east to west. The voice currents start out from the vessel at zero level, are amplified to a relatively high level and upon being transmitted to the shore 150 miles (240 km.) away, drop to a very low level. At the shore radio station they are boosted up, at New York amplified again, and put upon the trans-continental circuit. Regularly at about 300 miles (480 km.) the telephone repeaters pull back the transmission level to about its original value. In the radio link at the western end the currents are again amplified to a high level at the transmitting station, drop down to a very low level at the receiver and are brought back to a level at which they can be heard. Actually in the receiving telephone the transmission is about $log_{10}\frac{I_1}{I_2}=1.2$ below zero level, or roughly 25 miles of standard cable "down." The total loss and the total gain in the circuit is enormous, as is shown by the figures given in the diagram. This is a rather striking illustration of the extent to which amplification properly distributed and maintained can be used to overcome attenuations enormous in the aggregate. Just to give a better idea of what these values of attenuation and amplification mean, I would note that were it necessary to supply at the transmitting end all of the amplification required for delivering this volume of transmission to the receiving end thru the combination circuit, the kilowatts required would be measured by a twenty-nine place figure, an amount of power unavailable in the world. The importance of correctly distributing the amplification along the system is well illustrated by this figure by comparing it with the signaling power actually represented in the system, which sums up to something less than 1 kilowatt. The difference is simply a question of the transmission level at which the amplification is worked.

Figure 7 gives a view of the interior of one of these radio telephone stations—that of the American Telephone and Telegraph Company and Western Electric Company—located at Deal Beach, New Jersey. In the foreground is the switchboard for enabling the operator to control the radio-wire circuit at the connecting point. In the background are the transmitter units—

four of them. These, together with the four antennas with which the station is equipped, "multiplex" the ether, in effect, and permit four channels to be established to as many distant stations. It is intended that three of these be telephone talking channels and the fourth a signaling or a reserve talking channel. The

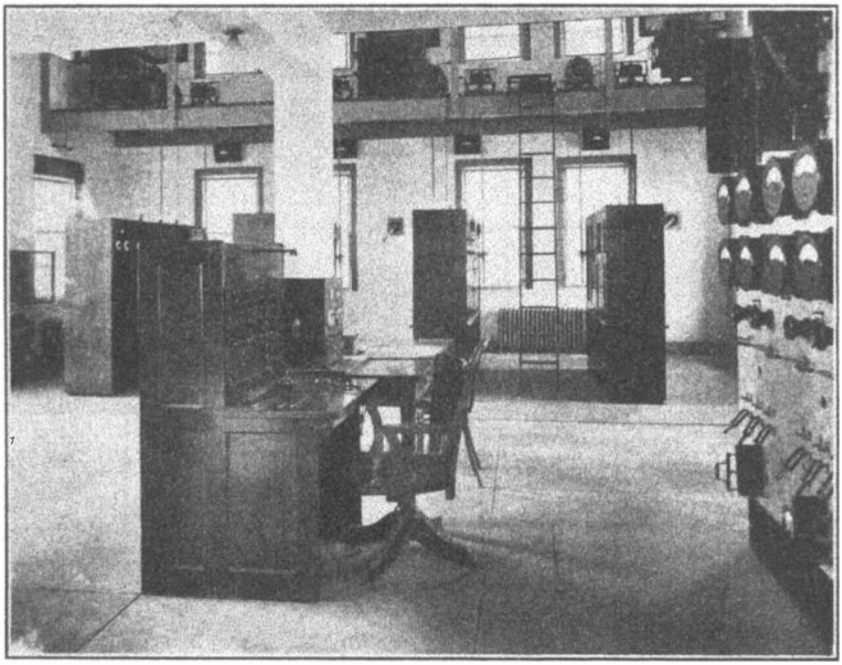


FIGURE 7

receiving station is located at another point. It is not desired to describe this station in any detail, but merely to illustrate it as an example of a radio repeating station functioning to connect the wire system with ships at sea and capable of effecting simultaneously three different connections. It is hoped that this ship-to-shore development may be itself the subject of an INSTITUTE paper.

INTERMEDIATE REPEATERS

The trans-continental line with radio extensions as shown in Figure 6 is a good illustration of the use of intermediate repeaters generally. Two types of repeaters are represented, the straight wire telephone repeaters and the shore radio stations which are in effect huge repeaters relaying between the land line and the radio circuits.

Because of the moderate attenuation obtaining in the wire transmission system, we can work with fairly long repeater spacing, about 300 miles in this case, and with moderate amounts of power and yet keep the transmission levels at the receiving end relatively much higher than is usual in radio systems. Due to the large attenuations obtaining over the radio extensions,

WIRE AND RADIO REPEATER SYSTEMS
TRANSMISSION LEVEL DIAGRAM
Wire Transmission f=1 K.C.
Radio Transmission f=1000 K.C.

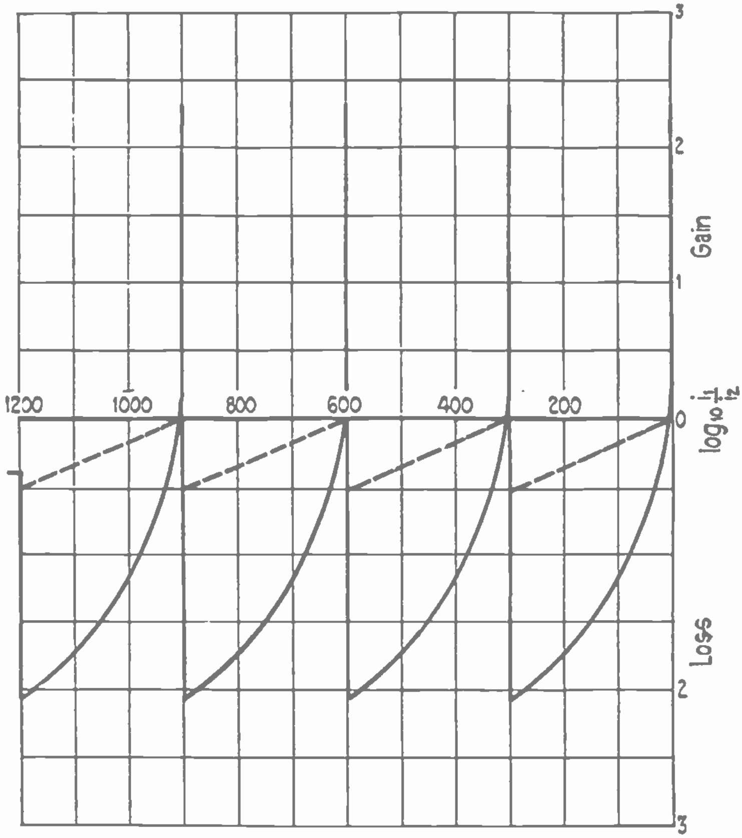


FIGURE 8

the radio repeaters must put out a high transmission level, making them costly, and even with this relatively high output, the level drops to very low values at the receiving end. This falling off occurs largely in the get-away loss at the transmitting end of the radio circuit, as the diagram indicates.

Figure 8 depicts an all-radio system provided with intermediate repeaters and compares for illustration purposes the transmission levels obtaining therein with those for a wire system. The solid lines are for radio and the dash for wire. It will be seen that the radio system courses thru wide transmission level variations as compared to the ordinary wire system, due to the large attenuations obtaining and particularly to the large step-off loss near the sending station

The figure illustrates the same spacing for both radio and wire repeating and gives a measure of the difference in amplification required in the two cases. Altho in the radio repeaters the level can be permitted to drop to low values, nevertheless a large part of the total amplification has to be supplied at relatively high power levels, and it is this fact, together with the antenna structures required at each point to "get into" the ether transmission medium anew, that militates against the economics of radio repeaters as compared with straight-away radio transmission. The tendency will be to "stretch out" the straight-away transmission due to the fact that for the longer distances the transmission loss increases relatively slowly. While we may look for some important uses of radio repeaters in special cases, we should not, in general, expect them to be as important to the radio art as are wire repeaters in wire operation.

Transmission of Side Band Without Carrier

In dealing with the subject of power levels in radio transmission, it is important to recognize that a modulated radio telephone wave consists of two components, one, the carrier frequency itself and the other, the so-called side bands, which are the actual modulated components. This resolution of the modulated carrier into two or, rather, three components, the carrier and two side bands, has been given mathematically a number of times and need not be repeated. It is physically analogous to the resolution of the uni-directional current of a microphone transmitter into direct current and alternating current components, the direct current corresponding to the carrier and the alternating current to the modulated components.

Now, the important thing about this matter of side bands and the unmodulated carrier component, with reference to transmission considerations, is this:—that it is the side bands alone, and not the carrier, which convey the actual intelligence. The function of the carrier comes in merely at the receiving end, in the detector, as a means for translating the side band from radio frequency back to audio frequency.

This will be made clearer by reference to Figure 9. At the bottom of the figure is shown schematically a one-way radio system. Above it is depicted the voice-frequency band, showing the manner in which it is shifted by modulation up to the carrier frequency range, and at the receiving end, by detection, back to the voice-frequency range. The voice-frequency band, as it comes out of the ordinary telephone transmitter, is shown at

RADIO TRANSMISSION WITH CARRIER AND SIDE BAND SUPPRESSION

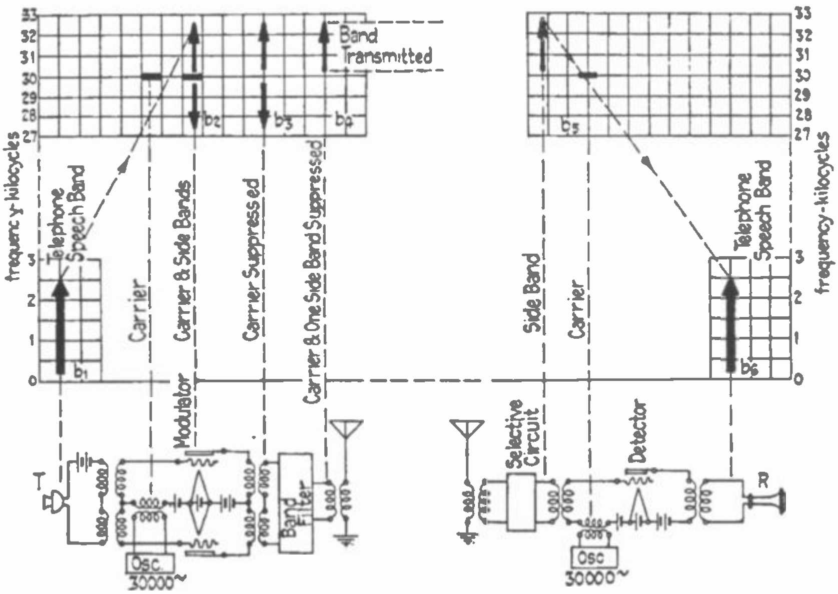


FIGURE 9

(b_1) at its normal telephone-frequency position. Upon modulation with the carrier the reference point of the voice-frequency band is shifted from zero frequency (direct current) up to the carrier frequency as shown at b_2 where the two side bands appear. The effect of modulation is, therefore, simply to shift the band of signaling frequencies upward in the frequency range and refer it in a double relation to the carrier frequency.

Located between the upper and lower side bands in the figure, there is indicated the unmodulated component of the carrier. The fact that this component is unnecessary so far as the actual intelligence-carrying energy is concerned, is proven by the fact that it need not be transmitted to the receiver. The carrier may be suppressed as shown at b_3. A means for doing this is the Carson balanced tube modulator circuit illustrated below in the figure.

A reduction of the total band can be effected by filtering out one of them as shown at b_4. The remaining single band is the simplest component of the modulation process with which intelligence can be transmitted to the distant end. Upon arriving at the receiving end at b_5, this side band is fed into the detector along with a carrier of the same frequency as that employed at the sending end; these two components demodulate one another, with the result that the side band is shifted down to its original audio-frequency position in the scale, as indicated at b_6.

Actually this general method of transmission, involving both carrier suppression and side band elimination, is being employed in wire carrier systems in the Bell Telephone Plant.[5] I am briefly explaining it here because it represents a valuable improvement in wire transmission which should have important applications in radio.

From the standpoint of transmission levels its application is in showing that the real intelligence-carrying component of a radio wave is the side band and not the carrier itself. In considering transmission levels accurately care should be taken, therefore, to deal in terms of the level or wave-intensity of the side band component and not the carrier. It is because of this that as nearly complete modulation as possible is desired at the transmitting station.

It follows that the power resident in the carrier is a pure waste in so far as overcoming interference is concerned. An important power saving can be effected in the transmitting station by providing some such means as is illustrated whereby the carrier power is held back in the circuit. The two side bands together can never be greater in current and voltage value than the carrier, and each side band alone cannot be greater than half the carrier. The power of the carrier is therefore always at least four times the power of one side band or twice that of both together. Thus by "holding-back" the carrier at the transmitter we can transmit with but one-third the power ordinarily required. Actually the power saving is much greater than this because of the necessity of normally working with larger ratios between carrier and side band in order to accommodate the peaks of the telephone waves and thereby preserve the quality of transmission. The power saving is, of course, especially important in long distance work.

[5] "Carrier Current Telephony and Telegraphy," by Colpitts and Blackwell, "Journal of the American Institute of Electrical Engineers," February 16-18, 1921.

The suppression of one of the two side bands halves the frequency band required for transmission and would double the message-carrying capacity of the ether were no frequency range required to space the channels apart. This advantage of the present method is likewise of special importance in long-distance-long-wave transmission.

Quality of Transmission

We have spoken above of factors concerned with the volume of transmission and only incidentally of that other requisite of transmission, namely, good quality. Without going into this matter in much detail, it will be well to make note of the several factors involved in obtaining good quality, as follows:

1. It is important that a substantial *band* of voice frequencies be transmitted. Of course distorted talk can be transmitted on a relatively narrow band, but commercial transmission has been found to require a single side band width of the order of 2,000 or more cycles, the band width increasing with the quality desired, up to about 5,000 cycles.

2. It is necessary that the distortion which is due to non-linearity of transmission with respect to amplitude, be avoided. This is equivalent to saying that there should not be permitted to take place self-modulation between the components of the side band, nor the too close cutting-off of the peaks of the telephone waves due to saturation effects.

3. The transmission must be kept free from interfering noises. The ratio between interfering noise current and voice currents of the order of 0.1 is regarded as large in wire practice. While this amount of interfering current will not prohibit service, it does seriously impair the effectiveness of transmission and annoys the listener. In radio the ratio of static noise to signal strength is very often much greater than this value. As the radio art progresses, it will be necessary to work toward standards more nearly in keeping with those which have been found necessary for wire service.

The writer wishes to express his indebtedness to the following of his associates for helpful suggestions and assistance: Messrs. J. R. Carson, Ralph Bown, and D. K. Martin.

SUMMARY: The transmission losses in wire and radio communication are compared for various frequencies and distances of transmission. The effect of natural and signal interference at the receiving ends of such systems is described, and conclusions drawn as to the necessary minimum transmission levels to be maintained in wire and radio circuits. The limitations on two-way operation resulting from "singing" of the entire system are considered in each case, and for combination wire and radio circuits as well. The comparative usefulness of repeater stations for wire and radio used is derived. Several means of effective wire and radio telephony are described, notably transmission on one side band only. The transmission requirements in numerical form of a radio telephone system are then given.

APPENDIX:—The curves of Figures 2 and 3 are based upon the following equations and data:

The radio curves are based on the familiar Austin-Cohen formula:

$$I = \frac{7.8\times 10^{-10}\, h_r\, h_s\, f\, I_s}{Rd}\, \varepsilon^{-4.4\times 10^{-6} d\sqrt{f}} \qquad (1)$$

where I = amperes

R = ohms

h = meters

f = cycles

d = miles

Taking equal antenna heights at two ends $h_s = h_r$.

As regards antenna resistance we assume symmetry as between the two ends, and that the external (radiation) resistance of the antenna equals the resistance within the antenna (which resistance would be internal apparatus resistance in the case of a perfect antenna). This makes R_r radiation resistance $= R_i$ ohmic resistance; and R of (1) becomes $= R_r + R_i$

$$\text{where } R_r = 17.8\times 10^{-13}\, h^2 f^2 \qquad (2)$$

Expressing in terms of current ratio and substituting values of R, (1) becomes,

$$\frac{I_s}{I_r} = 45.5\times 10^{-6} f\, d\, \varepsilon^{4.4\times 10^{-6} d\sqrt{f}} \qquad (3)$$

In order to plot this equation on the same basis as we usually plot wire attenuation, the logarithm of the ratio is used, thus:

$$log_{10}\frac{I_s}{I_r} = log_{10}\, 45.5\times 10^{-6} f\, d + \frac{4.4\times 10^{-6}}{2.303}\, d\sqrt{f} \qquad (4)$$

which is the equation of the curves plotted.

The ratio of the currents in the two antennas is in this case a true measure of the transmission because they are in circuits of equal impedances, by the assumption of antenna symmetry.

Data for the Wire Curves

$$\alpha = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}}$$

For number 8 Birmingham wire gauge open wire; dry weather. the constants per mile are:

diameter = 165 mils = 4.19 mm.
wire spacing = 12 inches = 30.5 cm.
40 poles per mile.

$$L = 3{,}370\ \mu\ \text{h.}$$

$$C = 9{,}140\ \mu\mu\ \text{f.}$$

Frequency, Kilocycles

	1.	20.	100.	1,000 units
$R=$	4.14	10.0	21.5	65.7 ohms per loop mile
$G=$	0.55	10.0	*50.0	*500.0 μ mhos per loop mile
$\alpha=$	0.003488	0.0112	0.03289	0.2059

*Estimated

DISCUSSION*

John R. Carson (by letter): Mr. Espenschied in his interesting paper has performed a distinct and important service to the radio profession in translating into the terms of radio transmission the engineering principles which have been so successfully applied to telephonic transmission and which, indeed have raised the latter art to the level of a science. It is my belief, however, that there are certain limitations inherent in radio transmission which prevent a complete correlation of the engineering principles governing the two arts. A failure clearly to understand this fact may lead to incorrect deductions from the paper under discussion.

In this connection I particularly wish to discuss briefly the bases of comparison of the formulas and curves (Figures 2 and 3) of wire and radio transmission. This comparison is, I believe, very likely to be misinterpreted in a way unduly favorable to radio unless the underlying assumptions are clearly understood and their limitations recognized.

In wire transmission it is almost universal practice to fit the terminal impedances to the line; that is, to make the terminal impedances, as seen from the line, equal to the line impedance. As is well known, this condition eliminates reflection at the terminals and assures maximum energy transmission. In comparing wire and radio transmission Mr. Espenschied has assumed a corresponding condition for the case of radio—namely, that the *ohmic* or *dissipative resistance* of the antenna (which includes the equivalent load or terminal resistance in series with the antenna) is equal to the *radiation resistance.* This is the theoretical condition for maximum energy transfer and might seem like a fair and reasonable basis for comparison. The point I wish to emphasize, however, is, that in the case of radio transmission, this condition is only realizable over a relatively narrow range of frequencies, and that deductions drawn regarding variation of efficiency with frequency are quite deceptive except within narrow limits.

The reason, in brief, why the conditions assumed in wire transmission lead to deceptive and even impossible conclusions when applied to radio transmission, is the relatively low value of the radiation resistance of an antenna even at very high frequencies, and its rapid decrease with decreasing frequencies. For example, consider an antenna with an effective height of

*Received by the Editor, July 10, 1922.

100 meters; its radiation resistance for representative frequencies is given in the following table:

f	Radiation Resistance
10^6	178 ohms
10^5	1.78 ohms
10^4	0.0178 ohms
10^3	0.000178 ohms

Now the conditions assumed are that the dissipative resistance of the antenna itself plus the equivalent resistance of the generator (or receiver) in series with the antenna is equal to the radiation resistance. Inspection of the table above shows that this condition is not realizable at frequencies less than 100 kilocycles. On the other hand, at frequencies above 1,000 kilocycles (10^6 cycles/second), the antenna height begins to approach the wave length and the radiation resistance ceases to increase in accordance with the square law. Furthermore, at high frequencies, the equivalent dissipative resistance of the antenna becomes quite large and increases rapidly with the frequency.

A brief analysis of the transmission formulas will serve to emphasize the foregoing considerations. The current in the transmitting or sending antenna is

$$I_s = \frac{E}{R_\omega + R_r}$$

where E is the equivalent emf. of the generator in series with the antenna; R_ω is the dissipative resistance of the antenna, including the equivalent resistance of the generator in series therewith; and R_r is the radiation resistance which may be taken as $k\,(h\omega)^2$ k being a numerical factor depending on the units employed. The energy radiated is

$$W_r = \frac{E^2 R_r}{(R_r + R_\omega)^2}$$

the total energy developed is

$$W_T = \frac{E^2}{R_r + R_\omega}$$

and the transmitting efficiency is

$$\eta = \frac{R_r}{R_r + R_\omega}$$

Now the condition for maximum radiated energy is clearly $R_w = R_r$; introducing this condition we have:

$$W_r = \frac{E^2}{4R_r} = \frac{E^2}{4k(\omega h)^2}$$

$$W_T = \frac{E^2}{2R_r} = \frac{E^2}{2k(\omega h)^2}$$

$$\eta = 1/2$$

We are thus led to the conclusion that for a given generator, *the maximum energy radiated is inversely proportional to the square of the frequency and the square of the antenna height.*

Now in accordance with this condition let us suppose that we make the product (ωh) smaller and smaller, ultimately the radiation resistance would become so small that the condition $R_w = R_r$ can no longer be maintained and we should actually get

$$W_r = \frac{E^2 R_r}{R_w^{\ 2}} = \frac{E^2 k(\omega h)^2}{R_w^{\ 2}}$$

$$W_T = \frac{E^2}{(R_r + R_w)} = \frac{E^2}{R_w}$$

$$\eta = \frac{k(\omega h)^2}{R_w}$$

That is, we should actually find that the factor $(\omega h)^2$ appears in an entirely different way than it does in the ideal case where $R_w = R_r$.

The same considerations apply to the phenomena of reception. Consider an antenna of height h, dissipative and radiation resistances R_w and R_r, in a radiation field in which the electric intensity parallel to the antenna is E. Assuming perfect tuning, the current I_k in the antenna is

$$I_k = \frac{Eh}{R_w + R_r}$$

and the energy absorbed is

$$I_k^{\ 2} R_w = \frac{R_w E^2 h^2}{(R_w + R_r)^2}$$

This is a maximum when $R_w = R_r$, in which case

$$I_k^{\ 2} R_w = \frac{E^2 h^2}{4\, R_w} = \frac{E^2}{4k} \frac{1}{\omega^2}$$

In other words, the maximum energy absorbable by the antenna is independent of the height and inversely proportional to the frequency

Here again the conclusions are quite deceptive when pushed to the limit. For if the frequency is reduced sufficiently, R_r becomes relatively negligible and ultimately

$$I_k^2 R_\omega = \frac{E^2 h^2}{R_\omega}$$

This formula shows that the energy absorbed under these conditions is proportional to the square of the antenna height but independent of the frequency.

In a similar manner the formula given in Mr. Espenschied's paper for the ratio of the sending and receiving currents is based on assumptions which are realizable only within narrow limits.

I do not wish to be misunderstood as failing to recognize the practical usefulness of the curves and formulas under discussion. As Mr. Espenschied says, they can be used for obtaining the loss in any practical case simply by taking the minimum (or theoretical loss) as given by the curves and adding thereto the additional loss obtaining in the actual antenna. I do wish, however, to emphasize the danger of drawing incorrect inferences regarding variation of efficiency with respect to frequency and height of antenna from the curves alone.

June 14, 1922.

A METHOD FOR TESTING AND RATING ELECTRON TUBE GENERATORS*

By

L. M. Hull

(Associate Physicist, Bureau of Standards, Washington, D. C.)

When used as a generator of alternating currents, an electron tube is associated with a mono-periodic electrical system and with a source of direct-current power. These are fundamental. Departures from this ideal state due to extraneous capacities, distributed circuit constants, and the like, are frequently encountered in practice. But like most ostensibly basic theory, the calculations employed in the present discussion will treat only the predominant effects.

Over two years ago a scheme was outlined by the writer for calculating approximately the power output from a given triode, in terms of the constants of a highly periodic output circuit, and of the "oscillation characteristic" of the valve, as determined by static method.[1] It was assumed at the outset that in the steady-state oscillation process the alternating plate and grid voltages are sinusoidal, opposite in phase, and maintained in a constant ratio of amplitudes, for a given coupling ratio. Graphical analysis of a given oscillation characteristic taken on this basis, tho tedious, yields results which are in striking quantitative agreement with experimental data obtained on oscillating systems, provided that certain circuital conditions are complied with, which are not seriously restrictive when dealing with radio-frequency circuits.

The brief theory in the paper referred to above was descriptive, rather than mathematically elegant. Similar conceptions of tube operation have since been presented in more complete mathematical form by Appleton and van der Pol[2] and others.

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* Received by the Editor, June 6, 1922.

[1] "Determination of the Output Characteristics of Electron Tube Generators." Scientific Paper of the Bureau of Standards, Number 355, December 1, 1919.

[2] Appleton and van der Pol, "Phil. Mag.," 42, 201-220, August, 1921. See also van der Pol, "Radio Review," 1, 701-710, and 754-762, November and December, 1920.

For practical applications the existing theory is not satisfactory or final. In a triode generator the primary phenomena are almost invariably complicated by secondary and thermal emission from grid and plate, non-uniform filament temperature, and other irregularities. As a consequence of the impossibility of obtaining even an approximate solution of the current-flow problem within the triode, the only useful volt-ampere valve characteristics are those experimentally determined. Computations from these characteristics are necessarily graphical. It is quite impossible, moreover, to obtain the data for even static volt-ampere characteristics with most commercial valves, on account of the excessive heating of the electrodes. For these reasons it seems desirable that certain coefficients of electron tube generators be defined, which serve to determine their behavior in radio-frequency circuits; also that methods of measurement of these coefficients for any triode be devised, these methods being analogous to those employed in measuring the coefficients of amplification and detection. An audio-frequency bridge arrangement is being used by the writer for this purpose, with a degree of success which seems to merit a detailed description.

Preliminary to a discussion of practical results, we shall consider briefly the basic theory of the conversion of direct-current power into alternating-current power by an electron tube.

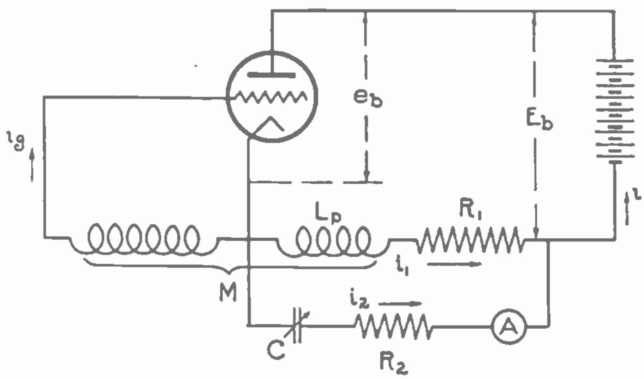

FIGURE 1—Semi-Direct-Coupled Generating Circuit

Take as a typical case the familiar semi-direct-coupled circuit shown in Figure 1. Forgetting for the moment just how and why the instantaneous volt-ampere relation happens to be as it

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E. Takagishi, "Electrician," 86, 346-348 and 374-375, March 25 and April 1, 1921; "Sci. Abs.," B, Number 642, June, 1921.

E. V. Appleton, "Radio Review," 2, 419-424, August, 1921.

is, for the triode, it appears that the fundamental features of this system are as follows: a parallel periodic circuit connected in series with a source of direct-current power of voltage E_b and with a variable resistance, T, of such a nature that its volt-ampere characteristic has a slope which is negative for at least a portion of the operative range. The symbols refer to Figure 2.

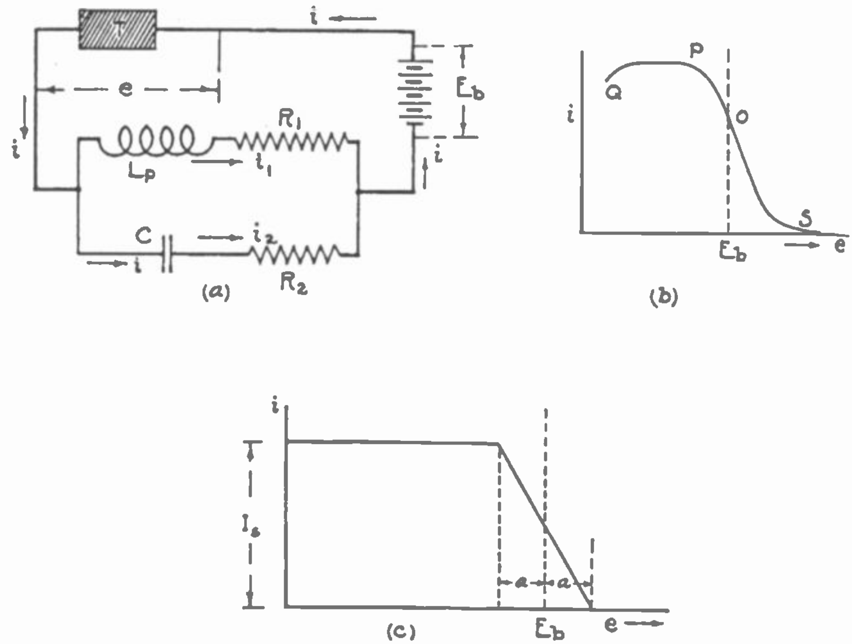


FIGURE 2

A solution for the vibrations of such a system has been the purpose of much diligent investigation since Barkhausen's first work on the theory of the singing arc. Dr. van der Pol[3] develops the function $i = f(e)$ as a power series about the point $e = E_b$, and obtains an approximate solution by a method similar to that suggested by Lord Rayleigh ("Theory of Sound") for small vibrations under a variable restoring force. It seems probable, however, that such a method tends, by the nature of its approximations, to conceal factors which are most conspicuous in fixing the stable amplitudes. In determining power output, the region PQ is usually more significant than the region PS of the characteristic. The terms in a series expansion about the point O which describe the shape of the characteristic at the point Q are just those terms which must be lopped off to fit into the

[3] "Radio Review," previous citation.

differential equation, if the latter is to be manageable. It is possible, however, by a slightly roundabout process, to arrive at an approximate solution which checks with a goodly variety of experimental results.

This procedure is based upon the assumption of small dissipative resistance in the output circuit, compared with the reactance of either branch, and upon the fact that the fundamental angular frequency of the steady-state oscillation is approximately $\frac{1}{\sqrt{LC}}$. Altho the frequency at which this steady oscillation occurs cannot be deduced precisely by mathematical analysis, it is not difficult, by theoretical reasoning, to set sufficiently narrow limits upon the range of frequencies over which this oscillation can occur. Suppose the function $i=f(e)$ is specified, empirically or otherwise. A possible form for $f(e)$ is that shown in Figure 2 (b). This oscillation characteristic can be represented to any desired degree of accuracy by a discontinuous curve composed of rectilinear segments. The simplest approximation to it is the symmetrical curve of three segments shown in Figure 2 (c). Then at any instant of time when the total voltage e across the impedance T lies between the limits:

$$-a<e<+a$$

or when

$$e<-a$$

$$e>+a$$

the motion of the system can be described by a linear differential equation with constant coefficients. The voltage e can be expressed as a function of time, in terms of the circuit constants and of the constant slope of the characteristic, for periods or epochs of time during which it is passing over any one of the rectilinear segments of the characteristic. Each of the alegebraic equations so obtained contains two unknown constants. The values of e at the end points of the segments are known, however, and the problem is narrowed down to a determination of the *time derivative* of e at one end point of each segment. By making use of the criterion for steady oscillation, namely, that at some convenient end point, say $e=+a$, the time derivative of e shall always be the same, each time the oscillation reaches this particular displacement, we obtain n transcendental equations in n unknowns, n being the total number of segments of the characteristic. The unknowns are the end-point values of the time derivative of e.

In the case of three segments, disposed about the operating point as shown in Figure 2, the solution of these steady-state equations can be readily obtained. The process of solution, while tedious, is perfectly straightforward and involves nothing but algebraic manipulations. The resulting equation for the fundamental frequency of oscillation exhibits the following interesting fact: The steady-state oscillation which takes place on characteristic composed of three rectilinear segments of the type shown in Figure 2 (c) cannot have a fundamental frequency which exceeds the transient frequency of the circuit isolated from T, or which is less than the transient frequency of the circuit when it is shunted by a constant resistance equal to the absolute value of the minimum resistance assumed by the variable impedanee in the region $-a$ to $+a$.

Of course the curve of Figure 2 is not a practical oscillation characteristic. When carried thru in detail, however, the analysis sketched in the foregoing paragraphs shows that any alteration in the slope of a section of the characteristic which is small compared with the region a a will influence the stable frequency to an extent which is small, even in comparison with the difference between the limiting values of frequency specified above. Thus while it may be impossible to calculate exactly the stable frequency in the case of a continuously curved characteristic, we can, at least, always determine the degree of approximation involved in assuming that this frequency is $\frac{1}{\sqrt{LC}}$, by measuring the slope of the oscillation characteristic at the point where this slope is greatest. If the value of $\frac{\partial i}{\partial e}$ at this point is h, then it is known that the stable frequency, ω_f can not be less than

$$\omega_1 = \sqrt{\frac{1}{LC} - \frac{1}{4}\left(\frac{R}{L} + \frac{h}{C}\right)^2} \tag{1}$$

nor greater than

$$\omega_2 = \sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}} \tag{2}$$

In these equations R is the total dissipative resistance of the output circuit.

In most cases, to a high degree of accuracy, $\omega_2{}^2 = \frac{1}{LC}$; that is, the difference between $\omega_2{}^2$ and $\frac{1}{LC}$ cannot be detected experimentally, whereas the difference between ω_1 and ω_2 is sometimes measurable. The writer has measured the mean slope of the upright portion of the oscillation characteristic of a number of tungsten-filament valves varying from 5 watts to 250 watts power capacity. These measurements were made on a number of oscillation characteristics for each valve, taken to correspond to various operating conditions in the radio-frequency circuits. All of these mean slopes lay in the comparatively narrow range of values: 10^{-4} to 3×10^{-3} inverse ohms. These values of h when substituted in equation (1) with values of L and C appropriate to sharply tuned radio-frequency circuits varying in natural wave length from 600 to 4,000 meters, gave values for the frequency which always lay between ω_1 and ω_2. The outstanding fact is, however, that with high-frequency output circuits having a free decrement per semi-period anywhere between 0.01 and 0.25, the computed difference between ω_f and ω_2 never exceeds one per cent., even with the highest values of h, and generally is of the order of one part in a thousand. This is contrary to common experience, where the quantities L and C are measured separately, and then determinations are made with a wave-meter of the fundamental frequency of oscillations maintained in the circuit by a subsequently attached triode. Absolute measurements of L and C can seldom be made with an accuracy greater than one per cent. Measurements of $(\omega_f-\omega_1)$ can, however, be made and repeated consistently with a total variation not exceeding five parts in ten thousand, by the use of an auxiliary driving circuit and a wave-meter with a sensitive current measuring instrument. The oscillatory circuit is connected to the triode in condition for use, except that the filament is not lighted. A thermo-galvanometer is loosely coupled to the output circuit and the auxiliary generator is tuned to resonance with this circuit. The wave-meter is tuned to resonance with this generator, the generator turned off, and the test triode allowed to maintain oscillations in the original circuit. The wave-meter is retuned to resonance with the fundamental of these oscillations, and the difference between the two wave-meter settings gives $(\omega_f-\omega_2)$. In this way the corrections due to electrode capacities, coil capacities, etc., are avoided, and surprisingly small values for $(\omega_f-\omega_2)$ are found. Measurements made in a given high-frequency circuit at wave lengths of from 1,500 to 4,000 meters, using consecutively two 5-watt triodes, a 50-watt tungsten-filament triode, and a 50-watt oxide-filament triode yielded absolute values for the quotient $\frac{\omega_f-\omega_2}{\omega_f}$, which under no conditions exceeded 0.6 per cent. It was greatest when using the oxide-filament tube, owing to the excessive filament emission.

It is not worth while to relate in detail the results of these experiments, since the whole question is one of what might be termed second-order importance in determining the net power output. The outstanding fact is that with an output circuit in which the effect of resistance on the free oscillation frequency is negligible, the difference between the free oscillation and the fundamental of the forced oscillation is less than one per cent of the former, with most existing types of triodes. The use of a hard, high-voltage triode tends to decrease this difference; the use of a soft tube or of a high-current tube tends to increase it. These considerations do not apply to generating circuits in which the series resistance of any branch is comparable with the reactance of that branch.

In what follows it will be assumed that the fundamental frequeney of the forced oscillation is $\frac{1}{\sqrt{LC}}$. This neglects the correction due to electrode capacities, but, as was remarked at the beginning, the present simple theory concerns only monoperiodic systems.

Reverting to Figure 2, if we eliminate all reference to the volt-ampere characteristic of the impedance T, and express i_1 and i_2 in terms of i, we have:

$$\begin{aligned} \frac{d^2i_1}{dt^2}+\frac{R}{L}\frac{di_1}{dt}+\frac{1}{LC}i_1&=\frac{R_2}{L}\frac{di}{dt}+\frac{1}{LC}i \\ \frac{d^2i_2}{dt^2}+\frac{R}{L}\frac{di_2}{dt}+\frac{1}{LC}i_2&=\frac{d^2i}{dt^2}+\frac{R_1}{L}\frac{di}{dt} \end{aligned} \qquad (3)$$

Having established the fact that the quantity $\frac{1}{\sqrt{LC}}$ may be taken as the angular frequency of the fundamental forced oscillation, a method for approximating the fundamental power output at once appears.

Suppose that $e=E_b+E \sin \omega t$ (4)

Then from the characteristic, Figure 2 (c),

$$i=A_o(E)+A_1(E)\sin\omega t+\sum_{n=1}^{\infty}[A_n(E)\sin\omega n t + B_n(E)\cos\omega n t] \qquad (5)$$

The coefficients are all functions of E, for a given characteristic. The fundamental is in phase with $E \sin \omega t$, that is, the coefficient $B_1(E)=0$, because $\int_0^{2\pi} f(E \sin x)\sin x\,dx=0$ for any form of the function f. If we substitute for i

in equation (3) and solve separately the resulting equations for i_1 and i_2, we have expressions for the fundamentals and all the harmonics of i_1 and i_2 in terms of the functions $A_1(E)$, $A_n(E)$ and $B_n(E)$.

On restricting ourselves further to output circuits in which the current distribution is determined practically by the reactances, the theory is limited to low-resistance audio-frequency systems and to a fairly wide range of radio-frequency systems, but the resulting expressions for the output currents are simplified to:

Fundamental:

$$
\begin{aligned}
{}_1 i_1 &= -\frac{A}{R\,C\,\omega}\cos\omega t \\
{}_1 i_2 &= \frac{A_1\,\omega\,L}{R}\cos\omega t
\end{aligned} \tag{6}
$$

Harmonics:

$$
\begin{aligned}
{}_n i_1 &= \frac{\beta}{n}\sqrt{(A_n{}^2+B_n{}^2)\,(1+\xi^2)}\,\sin\left\{n\,\omega\,t+\tan^{-1}\left(\frac{B_n+\xi\,A_n}{A_n-\xi\,B_n}\right)\right\} \\
{}_n i_2 &= -\beta\,n\sqrt{(A_n{}^2+B_n{}^2)\,(1+\xi^2)}\,\sin\left\{n\,\omega\,t+\tan^{-1}\left(\frac{B_n-\gamma\,A_n}{A_n+\gamma\,B_n}\right)\right\}
\end{aligned} \tag{7}
$$

Here $$\beta=\frac{n}{1-n^2}$$

$$\xi=\omega\,C(nR_2-\beta R) \text{ and } \gamma=\omega\,C\left(\frac{R_1}{n}+\beta\,R\right)$$

It is assumed explicitly that $\frac{1}{n^2-1}$ is very large in comparison with $C\,\omega$, and that $(1-n^2)$ is very large in comparison with $R_2\,RC^2\,\omega^2$.

From these values for ${}_n i_1$ and ${}_n i_2$ a trigonometric series in $\omega\,t$ is obtained for e, and then a new series for i is developed in terms of this series for e, in place of the single term, equation (4). In order to estimate the relative magnitude of the above quantities, assume that γ is negligible compared with one.

If $\quad R_1=R_2=0.01\,L\,\omega$, then $\gamma=0.01\left[\frac{1}{n}+2\,\beta\right]$

Then $$ {}_n i_2 = -\beta n\,\sqrt{A_n{}^2+B_n{}^2}\,\sin\left(n\,\omega\,t+\tan^{-1}\frac{B_n}{A_n}\right) \tag{8}$$

and the alternating part of e is:

$$e=\frac{L}{RC}A_1\sin\omega\,t+\sum-\beta\sqrt{\frac{L}{C}}(A_n{}^2+B_n{}^2)\,\sin\left\{n\,\omega\,t-\tan^{-1}\frac{A_n}{B_n}\right\} \tag{9}$$

as compared with the quantity $E\,\sin\,\omega\,t$ with which we started.

It is pleasantly apparent from equation (9) that this process is very rapidly convergent. This fact is a direct result of the assumption that $\omega^2 = \frac{1}{LC}$ and R_1 and R_2 are small compared with $L\omega$. In fact, if the circuit be so arranged that $\sqrt{\frac{L}{C}}$ is small compared with its square, the process need be carried no further, provided only that none of the harmonic amplitudes $\sqrt{A_n^2(E)+B_n^2(E)}$ exceed $A_1(E)$ for any value of the argument E. There are no definite limits. The approximation grows better, for a given value of E, the smaller the resistance relative to L ω and the greater the rapidity with which the amplitudes of harmonics in i converge with increasing values of n. This latter depends upon the form of the volt-ampere characteristic. It is fairly apparent, without more detailed computation, that the difference is negligible between the coefficients of the i series, when $i = f(E \sin x)$ and the coefficients when $i = f(E \sin x + \sum a_n \sin nx)$ where no value of a_n exceeds one per cent of E and where the series of coefficients $\sum a_n$ converges. The physical reason for this is, of course, that the parallel impedance of a periodic circuit is high for its resonant frequency and relatively low for all harmonics of this resonant frequency. Thus the "voltage drop" across the output circuit is high for the fundamental of i and low for all the current harmonics, even tho they are as large as the fundamental in amplitude.

Having established the fact, which is familiar to most experimenters, that under certain restrictions on the ratio of resistance to reactance in the output circuit, the alternating voltage e, and consequently the currents i_1 and i_2, are approximately sinusoidal, we have the following simple relations concerning the useful or fundamental power output and input:

$$\text{Power generated} = \frac{A_1^2(E)\,L}{R\,C} \qquad (10)$$

$$\text{Power generated} = i_1^2 R = i_2^2 R$$

$$\text{Power supplied} = E_b\, A_o(E) \qquad (11)$$

The latter expression for the power drawn from the battery E_b is the only positive term resulting from the multiplication of $(E_b + E \sin \omega t)$ by the trigonometric series for i, equation (5). The coefficients of the alternating terms are all negative if the characteristic is "falling," as shown in Figure 2(b), and indicate power furnished by the impedance T.

Having found the conditions existing during the steady state oscillation of a general system as set forth in Figure 2(b), these considerations can be applied with only slight modifications to the tube-driven system of Figure 1. As is well known, a starting transient, and hence an ultimate steady state is possible for a wide range of conditions of the volt-ampere characteristic of the triode, provided that the mutual inductance M is of the right sense to make the total derivative $\frac{d\,i}{d\,e_p}$ negative at the starting point. If the circuit and valve conditions permit a starting transient with inverse damping, a steady state will ultimately be reached. Then during the steady state the instantaneous relation between e_p, the plate voltage, and i, the plate current, must have a negative slope for part of the operating range, and be of the same nature, qualitatively, as the characteristic of Figure 2(c). With oscillations of this sort, the current i_1 has a phase angle of 90° lagging with respect to the alternating plate voltage, equations (6) and (9), and the alternating grid voltage is 90° behind this current. This results in a constantly maintained phase relation of 180° between the alternating plate voltage and alternating grid voltage, provided that the quadrature component of voltage due to the flow of grid current thru the coil L_g is negligible compared with the voltage induced in this coil. The amplitude of the alternating plate voltage is $L_p\,\omega\,i_1$ and of the alternating grid voltage, $M\,\omega\,i_1$, the latter being sinusoidal to the same degree that the plate voltage is sinusoidal. Hence, to a first approximation, the alternating component of plate voltage, in addition to being opposite in phase to the grid voltage, bears a constant ratio thereto, for a given value of M and L regardless of the values of R and C. This may be called the plate-grid ratio, and is thus dependent upon the circuit, and not upon the tube. This fact enables the determination of the oscillation characteristic for a given triode, as a function of the alternating plate voltage, or of the grid voltage. Such characteristics, taken by static methods, must necessarily be made separately for each value of the plate-grid ratio, corresponding to each value of the ratio $\frac{\mathrm{L}}{M}$ to be used in oscillatory circuits.

The flow of current in the grid circuit abstracts useful power from the output circuit because the fundamental of grid current is in phase with the alternating grid voltage, in contrast with the fact that the fundamental of plate current opposes the alternating plate voltage. The grid power is thus a loss, and is important, in that it limits the alternating-current power which any triode will supply. Hence the oscillation characteristic of the grid current and that of the plate current are, in practical cases, of almost equal importance. In Figure 3(a) are shown the oscillation characteristics of a typical, tungsten-filament, 5-watt, hard triode, operating at 300 volts plate voltage; the

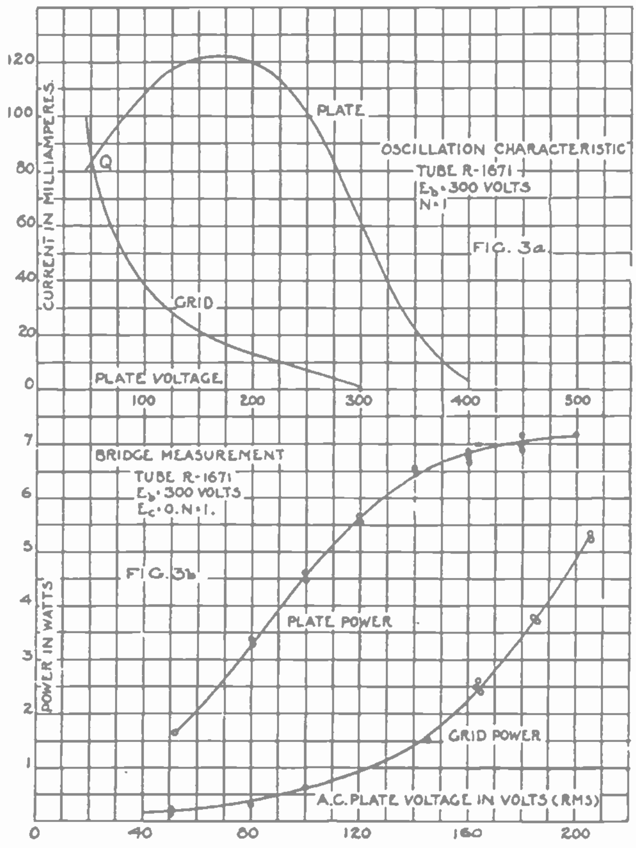


FIGURE 3

characteristic is shown for the value one of the plate-grid ratio, because that is the value which was used in subsequent measurements. The currents are plotted against the total plate voltage, e_p. The alternating plate voltage is merely $e_p - E_b$. These characteristics show how the grid current limits the useful power in two ways: first, by direct abstraction of power thru the grid coupling; second, by reducing the saturation current to the plate at the point Q, which increases the harmonics of plate current—chiefly the third—at the expense of the fundamental. If it were not for the flow of grid current the fundamental of plate current would approach the constant value $\frac{2\, I_s}{\pi}$ as the alternating voltages are increased, this value being the fundamental amplitude of a perfectly rectangular current wave, or $\frac{1}{\pi}\int_{\pi}^{2\pi} I_s \sin x\, dx$. This fundamental would remain constant until the alternating plate voltage approached the peak E_b, beyond which the fundamental of plate current would speedily disappear. Thus the maximum of power at the fundamental which can possibly be obtained from a triode having filament emission I_s operating at a steady plate voltage E_b, is $E_b \frac{I_s}{\pi}$

Two quantities which are important factors or characteristics of a power tube are the fundamental grid resistance, R_g and the fundamental plate resistance, R_p. These may be defined as the quotients, respectively, of a sinusoidal impressed plate voltage by the fundamental plate current and a sinusoidal impressed grid voltage by the fundamental grid current; it is understood in the definition that these voltages are impressed simultaneously, bearing a definite ratio to each other, and being exactly opposite in phase. Thus for a given value of the plate-grid ratio, R_p and R_g are both functions of the amplitude of the alternating plate voltage. Suppose that R_p and R_g are known, empirically, as functions of E for a given value of $N = \frac{L_p}{M}$. Then for each value of E the average useful power output is:

$$P_o = \frac{E^2}{2}\left(\frac{1}{R_p} - \frac{1}{N^2 R_g}\right) \tag{12}$$

This equation, which involves only the triode characteristics, is related to the constants of an oscillatory output circuit thru the equation:

$$\frac{L}{R\, C} = \frac{E^2}{2\, P_o} \tag{13}$$

Thus if R_p and R_g are known graphically as a function of E, then substituting the same values of the parameter E in equation (13), we obtain an approximate relation between the power output and the parallel impedance, $\frac{L}{R\, C}$, of the output circuit, which should hold for any oscillatory circuit, in which the grid coupling reactance is in the ratio $\frac{1}{N}$ to the plate coupling reactance.

It should be noted here that while the fundamental amplitude of plate current, $A_1 = \frac{E}{R_p}$, the fundamental plate resistance R_p is always considerably less than the quantity $\frac{L}{RC}$. In fact, when $\frac{L}{RC}$ is large, involving large values of E, $\frac{L}{RC}$ may be ten times as great as R_p and the triode will still oscillate. This is due to the presence of the dissipative grid resistance R_g.

With electron tubes of power capacity not exceeding 50 watts, the audio-frequency bridge shown in Figure 4 furnishes a con-

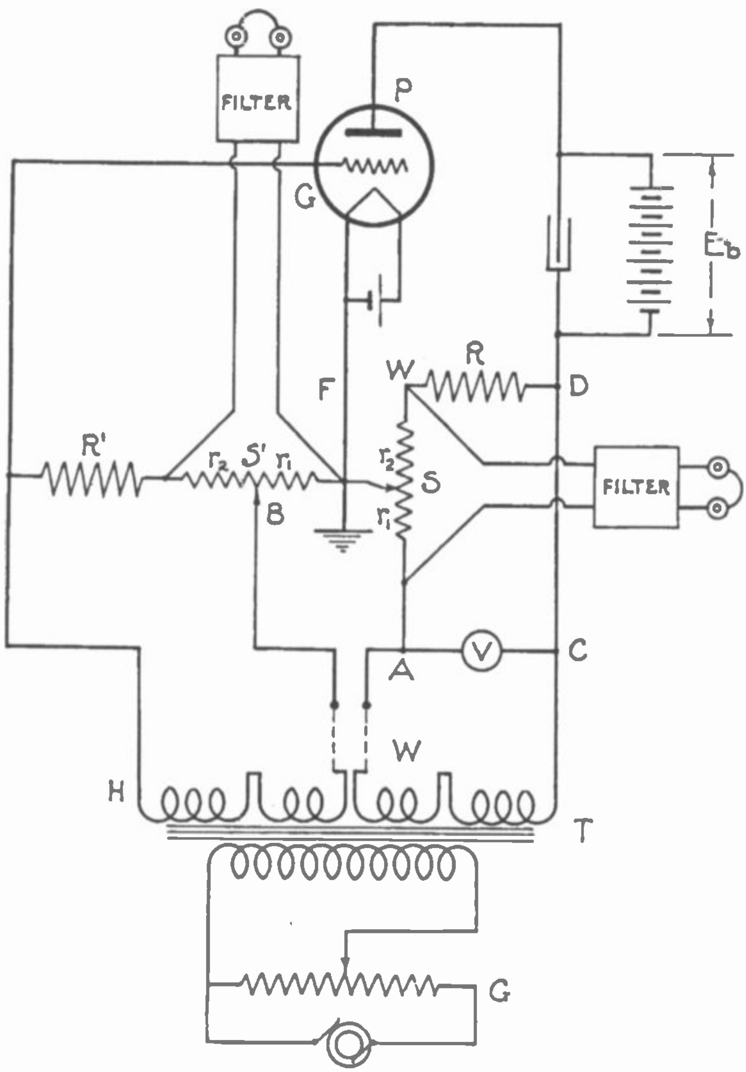


FIGURE 4—Alternating-Current Bridge

venient method for measuring R_p and R_g for various values of E and N. A practical question intrudes itself at this point: If the power output or i^2R in a radio-frequency generating circuit depends only upon the effective resistance $\frac{L}{RC}$, and upon the plate-grid ratio, why not test a triode in a carefully designed generating circuit, of which the constants are known, obtaining P_o directly as a function of $\frac{L}{RC}$ instead of measuring R_p and R_g separately. The answer is, that the results of an audio-frequency measurement, applied to one radio-frequency circuit after another, agree, in general, more closely with measured values of the oscillating power output than do measurements on one radio-frequency circuit, when applied to circuits of a different type, or to circuits of the same type having different constants. The whole foundation upon which these computations are based is onlv an approximation of the true conditions. Using measurements on an oscillating triode, the computations are based upon the readings of a thermal ammeter, a circuit resistance which usually varies rapidly with the frequency, and upon the assumption of negligible extraneous capacities and negligible variations in the plate-grid ratio. The latter assumption is a departure from experimental facts only if the plate-grid ratio is decreased considerably below its optimum value, or if the alternating voltages are increased excessively above the values which yield maximum power output. But it is preferable from a practical standpoint to test a tube in a system which stimulates the conditions of an *ideal* generating system. Then the departure from these conditions may be in one direction or the other, when the triode is used as a generator, in different circuits. But at any rate the errors are not augmented by similar departures in the conditions of test.

Referring to Figure 4, the transformer T introduces into the plate and grid circuits of the triode alternating voltages having a constant ratio (which can be regulated by the switching arrangement W), and an absolute value which can be varied by the voltage divider G. The frequency is any convenient audible frequency, possibly 500 cycles. Under the influence of these voltages, currents flow in the plate circuit $A\ F\ P\ D\ C$ and in the grid circuit $B\ F\ G\ H$, which are distorted by the characteristic plate and grid harmonics. The transformer T has, however, low-resistance windings, designed to furnish alternating-current power perhaps ten times as great as the maximum power supplied in the triode. As a consequence, the 180-degree phase relation between the voltage impressed in the grid circuit and that impressed in the plate circuit is maintained at all times. In addition, these voltages e_p and e_g can be made very closely sinusoidal if the transformer is properly designed. When used with 15-watt valves, ordinary commercial 150-watt instrument transformers have been found to operate satisfactorily. On investigation with an audio-frequency oscillograph the plate and grid voltages are found to be as nearly opposite in phase as can be detected visually; also the oscillograms of either voltage, taken with full power supplied to the tube, can be superimposed upon oscillograms of the open-circuit voltages with perfect coincidence, as nearly as can be detected by the eye.

In the plate circuit, R is a resistance of 10,000 or 20,000 ohms, S is a 6-ohm slide wire, and V is an alternating current voltmeter. The filter employed with the telephones may be dispensed with entirely, for rough measurements. A two-stage audio-frequency amplifier, in which both coupling transformers are tuned to 500 cycles, eliminates the harmonics in the telephone receivers to such an extent that settings can be made on the null point of the fundamental, with a total variation (from one setting to another) of 2 per cent in the region of maximum output and 4 per cent at excessively high voltage. This is about the nearest precision measurements that are possible of achievement with electron tubes, anyway. If a wave filter is used, its "cut-off frequency" should be slightly above 500 cycles.

If E is the sinusoidal alternating voltage of frequency induced in the plate circuit, then a current I, of frequency ω and of amplitude $\frac{E}{R+r_1+r_2}$, flows around the circuit $C\ D\ W\ A$.

Thru the circuit $A\ F\ P\ D\ C$ currents of frequency ω, $2\ \omega$, $3\ \omega$, and so on, will flow, and the fundamental of amplitude A_1 has a phase of 180 degrees with respect to the current I, in the branch r_1, since the fundamental of plate current is in phase with the grid voltage and opposes the plate voltage. Hence, if the reactance of the slide-wire is small compared with the input impedance of the filter, we have as the condition for the disappearance of the fundamental note in the telephones:

$$Ir_2 = A_1 - Ir_1$$

whence, if $R_p - \frac{E}{A_1}$, then:

$$R_p = \frac{r_1 R}{r_1 + r_2} \tag{17}$$

In this equation the slide-wire resistance (r_1+r_2) is neglected in comparison with R.

In the grid circuit, remembering that the fundamental of grid current is in phase with the grid voltage, we have, as the condition for the disappearance of the fundamental note in the telephones

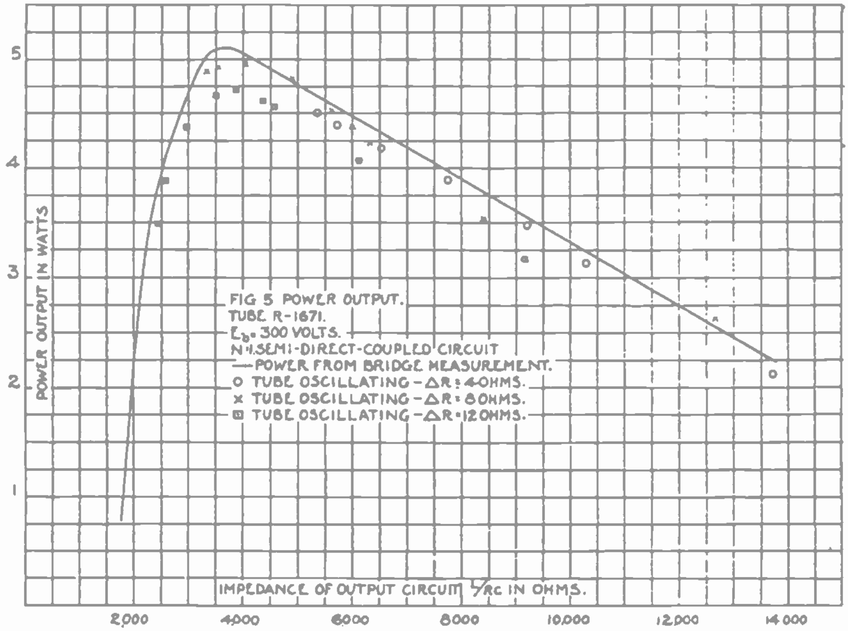
$$R_g=\frac{r_1}{r_2}R \tag{18}$$

By means of a suitable double-throw switch, the same slide-wire and filter can be used for both plate and grid measurements.

Accompanying Figure 3(a), which shows the static "oscillation characteristic" of a 5-watt valve, for $N=1$, is Figure 3(b), showing the fundamental as power dissipated in the grid circuit of this triode, for the same value of N. The data for Figure 3(b) were obtained from measurements of R_p and R_g on the bridge shown diagrammatically in Figure 4, for different values of voltage supplied from the transformer. A complete set of measurements of this type, for a given value of N, can be made in less than a half hour's time. The separate plotted points on Figure 3(b) are the results obtained by two different operators of the bridge, taken at different times on the same value. Settings of the power bridge can be repeated with approximately the same degree of precision as can be attained in measuring the amplification factor and output resistance of an amplifier tube. Figure 3(b) shows how rapidly the grid power increases as the alternating plate voltage approaches an amplitude of 250 volts sufficient to bring it to the point Q of the characteristic. At $E=$ 280 volts (root-mean-square value 200 volts) the fundamental of plate current has not yet started to decrease, and yet the grid power loss is almost sufficient to reduce the net power output to zero. The grid current is the controlling factor in small triodes.

Figure 5 shows a comparison of the results of the bridge measurements, or test, with the actual power output in a radio-frequency circuit. The heavy line shows the power output as a function of $\frac{L}{RC}$ in the oscillatory circuit, as predicted from the bridge measurement. To express the computed power as a function of $\frac{L}{RC}$ in any output circuit, equation (13) was used. The separate points on Figure 6 shows measured values of i^2R with the triode feeding an oscillatory circuit of the type shown in Figure 1. These are plotted against the impedance $\frac{L}{RC}$

as measured on the output circuit *isolated from the tube.* No attempt was made to correct C for the triode capacities, and the



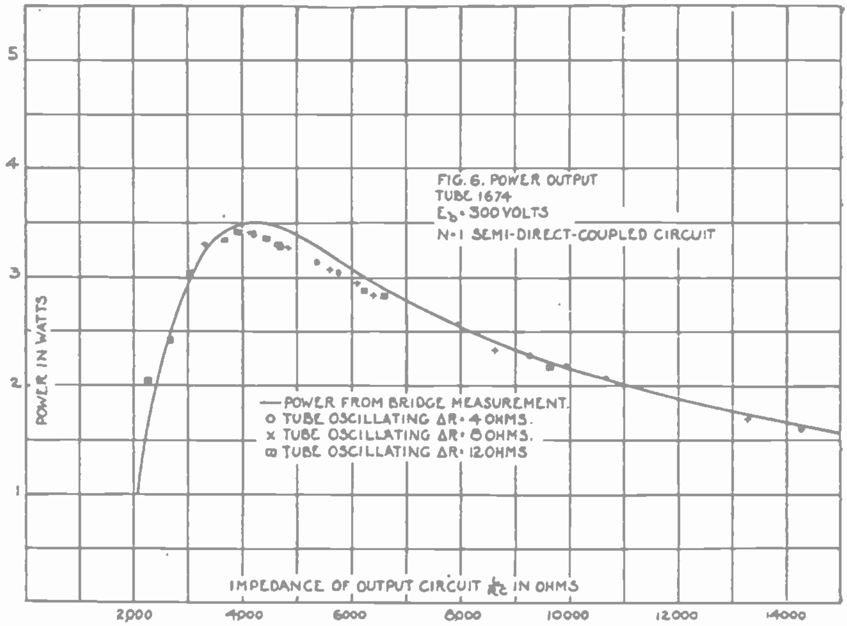


FIGURE 6

measurements of R were made by the resistance-variation method on the L C circuit with nothing else connected to it. If a method of this sort is to be successful in practice it must be applicable with some degree of certainty to computations on antenna circuits and the like, the constants of which may be known only when they are isolated. It would be practically impossible to select a triode to operate in a given circuit, by a test of this sort, if the circuit constants had to be corrected and altered with each change in the tube or in the coupling reactances.

The absolute values of the constants of the radio-frequency circuits were as follows:

$$L = 498 \text{ microhenrys (low-frequency measurement)}$$
$$C = 0.0015 \text{ to } 0.015 \text{ microfarad}$$

Measurements were made for three values of inserted resistance, $R=4$ ohms, and 12 ohms. The coil, condenser and ammeter resistance added about two ohms, this being variable, of course, with the frequency.

In Figure 6 is shown for a different triode the tested power output, and actual power output as measured in the same oscillatory circuit as was used previously. The oscillation characteristic for this second triode was considerably different from that of the first one. The agreement of the predicted results with those obtained in the generating circuit is slightly better than in the previous case.

Enough experimental evidence has been obtained to indicate that this method of bridge measurement is a sufficiently close approximation, and can be performed with such a degree of facility as to make it worth while from a practical standpoint. As a matter of research it would be interesting to investigate the departures from the ideal state encountered in circuits operating at short wave lengths or with high values of R.

The writer has recently applied the bridge measurements of power output described above to radio-frequency generating circuits of two other types: the direct-coupled (Hartley) circuit, and the symmetrical iductively coupled (Meissner) circuit. The steady-state analysis of both of these systems shows that under the previously mentioned restriction (resistance of any branch of the output circuit small compared with its reactance) the voltage reaction of the output circuit upon the tube is sinusoidal over a wide operating range, and the grid and plate voltages are opposite in phase and constant in their ratio to each other. Thus the fundamental resistances of the tube as measured on the power bridge are adequate for the calculation of the alternating power output when the tube is associated with either of these circuits.

It is found, moreover, that in the steady state of oscillation of these circuits, as well as the semi-direct-coupled circuit, the fundamental resistance of the tube enter the expression $\frac{E^2}{P_o}$ only in *second-order terms*, E being the plate-voltage amplitude and P_o the power dissipated in the output circuit. In both cases this ratio involves only the constants of the output circuit, to first-order terms, and constitutes in effect the real (pure resistance) load in the plate circuit of the oscillating tube, containing a first-order correction for the effect of the grid current. Thus the bridge measurement for a given tube, giving power output as a function of the plate voltage, E for a given coupling ratio, can be applied as before to the calculation of the power output as a function of the constants of the output circuit by use of equations (12) and (13). For the Hartley circuit we have, to first-order terms

$$\frac{E^2}{2P_o} = \left(\frac{N}{N+I}\right)^2 \frac{L_p+L_g+2M}{RC}$$

$$N = \frac{L_p}{L_g}$$

and for the Meissner circuit:

$$\frac{E^2}{2P_o} = \frac{M_p{}^2}{LRC}$$

wherein R and C are the lumped inductance and capacity of the output circuit, L_p and L_g the inductance of the plate and grid coils, M the mutual inductance between the coils, L the self-inductance of the output circuit, and M_p is the mutual inductance between the plate coil and the output circuit in the Meissner arrangement.

Experiments performed with a number of 5-watt tubes have shown excellent agreement between the power output as computed from the fundamental resistance and as measured in Hartley and Meissner generating circuits having values of resiatance of from two to twelve ohms and over a range of wave lengths of from one thousand to five thousand meters. Note-Noteworthy departures of the radio-frequency i^2R in the output circuit from the computed values are encountered in the Meissner circuit only if the mutual coupling reactance be so reduced that the load departs widely from a pure resistance and the true oscil lation characteristic becomes a loop instead of a single line. Even than the discrepancy in power may not be serious because the useful power output depends upon the *area* under the current-voltage curve, and a partial compensation occurs when the load becomes reactive. The same effect is encountered, of course, when the plate and output circuits are brought into an inductance coil thru separate taps in the Hartley circuit, or when a loosely-coupled tuned-grid arrangement is used for semi-direct coupling. In general the efficiency of conversion is decreased and the proportion of harmonics is increased when the power factor of the load is reduced appreciably below unity by the introduction of leakage reactance or of separate self-inductive reactance into the plate circuit of the tube.

The writer was given much careful and painstaking assistance in the experimental work described in this paper by H. A. Snow, of the Bureau of Standards. He is also indebted to Professor G. W. Pierce and Professor E. L. Chaffee, of Harvard University, where the experimental work was begun, for valuable suggestions.

Radio Laboratory,
Bureau of Standards,
Washington, D. C.

June 14, 1922.

SUMMARY: Two new coefficients of the three-electrode tube are defined, which are significant in determining its behavior and merit as a generator of alternating currents in a highly periodic output circuit or antenna. These coefficients are based upon a simple theory of the operation of the tube as a generator and are adequate for a calculation of the power output in circuits with any type of coupling, provided certain circuital requirements are complied with which are consistent with the usual requirements of good design in radio-frequency circuits. A bridge method for measuring these coefficients is described, and experimental data upon a particular generating circuit are presented, illustrating their use in calculating the high-frequency power output.

MATHEMATICAL TREATMENT OF RECTIFICATION PHENOMENA*

By

D. C. Prince

(General Electric Company, Schenectady, New York)

If an attempt is made to present an exact mathematical treatment of kenotron rectifier phenomena, taking account of all the factors entering in—such as transformer reactance, varying tube resistance, and so on, the expressions become so involved as to be practically almost useless. However, a very good perspective of the whole problem can be obtained by analysis of the voltage waves produced by various rectifier connections. Such waves are actually obtained at no load and in somewhat modified forms with various types of loads.

The Fourier series analysis of a half wave rectifier is:

$$e_r = E\left[\frac{1}{\pi} + \frac{1}{2}\sin\theta - \frac{2}{3\pi}\cos 2\theta - \quad \frac{2}{\pi(m^2-1)}\cos m\theta\right]$$

where e_r is the instantaneous voltage.

E is the maximum voltage.

$e = E \sin\theta$ is the voltage before rectification.

m = any even integer.

For rectifiers utilizing two or more tubes, that is, full wave two, three, and six or more phases, the direct current component is given by the formula

$$a_o = \frac{p}{\pi}\sin\frac{\pi}{p} \text{ where}$$

"p" is the number of tube phases counting single phase full wave as two, and quarter phase as four, and so on. For three and six phase, "p" is three or six, respectivelv.

Table I shows the direct current components of the most used simple connections.

TABLE I—*Average Direct Current Potentials*

*Received by the Editor, April 10, 1922.

		A	B
Single phase, half wave	$1/\pi$	0.318	0.450
Single phase, full wave	$2/\pi$	0.637	0.900
Two phase....................	$\frac{2\sqrt{2}}{\pi}$	0.900	1.275
Three phase..................	$\frac{3\sqrt{3}}{2\pi}$	0.827	1.170
Six phase......................	$3/\pi$	0.955	1.350
Infinite phases................	1	1.00	1.41

The values in column "A" are referred to the maximum voltage. Those in column "B" are referred to the nominal or mean effective voltage. From these values the tube drop must be subtracted. If there is a storage condenser across the rectifier terminals, the voltages under light load will varv from those for the phases used to the value for infinite phases at no load.

Omitting the storage condenser, the ripple voltage for any number of tube phases greater than one can be obtained from the following formula which is the result of a Fourier series analysis:

$$a_n = \frac{2\,a_o}{(n^2-1)} \quad \text{where}$$

a_n is the coefficient of the n^{th} harmonic as referred to the original alternating current frequency.

$n = Kp$ where "p" is the number of tube phases and "K" is any integer. The number of tube phases is taken as the number of anodes as in the determination of the direct current component.

Table II gives the application of this formula to the most usual rectifier combinations.

TABLE II—*Ripple Composition of Rectifier Combinations*

Coefficients of	$\cos\theta$	$\cos 2\theta$	$\cos 3\theta$	$\cos m\theta$
Single phase. (Full Wave)	$-4/3\pi$	$4/15\pi$	$4/35\pi$	$\frac{4}{\pi(4m^2-1)}$
Two phase...	$\frac{4\sqrt{2}}{15\pi}$	$4\sqrt{2}/63\pi$	$4\sqrt{2}/143\pi$	$4\sqrt{2}/\pi(16m^2-1)$
Three phase..	$3\sqrt{3}/8\pi$	$3\sqrt{3}/35\pi$	$3\sqrt{3}/80$	$3\sqrt{3}/\pi(9m^2-1)$
Six phase....	$6/35\pi$	$6/143\pi$	$6/323\pi$	$6/\pi(36m^2-1)$

For simplicity, one θ is taken as the lowest ripple frequency present in each case which is equal to the alternating current frequency multiplied by the number of tubes.

There are several other combinations which can be made up by taking two or more of the foregoing simple combinations and connecting them in series or parallel thru an interphase transformer. The resultant voltages are found by addition in the former case, while in the latter they are the average of the individual combinations. The usual combinations under this head are as follows:

Two single phase full wave rectifiers may be fed from a two-phase system and connected in series, or, thru an interphase transformer, in parallel. The average direct current voltage will be $4/\pi$ or $2/\pi$, respectively. The ripples will be the same in percentage as for an ordinary two-phase rectifier.

Three single-phase full-wave rectifiers may be fed from a three-phase system and connected in series, or, thru an interphase transformer. The voltages are $6/\pi$ and $2/\pi$, respectively, while the ripples are of the same percentages as with an ordinary six-phase rectifier.

Two three-phase rectifiers may be fed from three-phase lines in such a way as to be displaced 180° in phase and connected in series or parallel thru an interphase transformer. The voltage will be $3\sqrt{3}/\pi$ or $3\sqrt{3}/2\pi$ and the ripples will be of the same percentages as with ordinary six phase.

These special combinations have the advantage that each tube is utilized for a greater proportion of the time than with the simple connection of the same total number of phases. This results not only in both greater output and efficiency for each individual tube, but also for the high voltage transformer windings. For instance in a single phase rectifier each tube may operate up to 180°, three phase up to 120°, two phase up to 90°, six phase up to 60°. If the tube is limited in emission, the current output per tube goes up directly as the angle during which current flows. If it is limited by heating, the current output increases as the two-fifths power of this angle.

Transformer secondary capacity is always reduced when used with the usual rectifier connections. If we assume that the output current is constant, the current in any secondary transformer winding is a square wave. Suppose a 10,000-volt 10-kilowatt transformer to be used as one leg of any of the simple rectifier arrangements. The current, voltage and power outputs are shown in Table III.

TABLE III—*Transformer Outputs with Various Rectifier Connections Based on Secondary Heating*

	Amperes	Volts Average	Fraction of a Cycle	Power	Per Cent
Non-rectifying.......	1 (r.m.s.)	10,000 (r.m.s.)		10,000	100
Single phase	1.41	9,000	0.5	6,370	64
Three phase.	1.73	11,700	0.333	6,750	68
Two phase..	2.00	12,750	0.25	6,375	64
Six phase...	2.45	13,500	0.167	5,510	55

In considering any transformer rectifier connection, it is necessary to remember that the ordinary transformer is not intended to carry direct current. A comparatively small amount of unbalanced direct current in the windings will saturate the core iron and cause heavy exciting current to be drawn, thus reducing power factor and efficiency and overheating the transformer. In a single phase half-wave rectifier, this effect is most pronounced so that this arrangement is only useful for small powers. Full-wave single phase, two phase, and six phase are normally balanced with good transformer design, so that this phenomenon requires no concern. Three-phase rectifiers have direct current components in each leg for which there must be no return magnetic circuit, if the best operation is to be secured. To meet this requirement the transformer must be core type rather than shell, and only one end of the core may be permitted to make contact with the iron of the case if the latter has any appreciable cross-section.

Before leaving the subject of transformers, it is well to consider the character of the wave which will be drawn from the alternating current supply lines. Table IV shows the harmonics which are represented by a square block of current of unit height drawn for different fractions of a cycle. Such blocks would be drawn by a rectifying transformer of low reactance feeding direct current into a highly inductive circuit. From this table the harmonics drawn by a polyphase rectifier can be made up. For instance, with a three-phase rectifier, current is drawn for one-third of the time. The second harmonic is equal to one-half of the fundamental. With the fundamentals displaced

120° the second harmonics are displaced 240°, which is equivalent to opposite phase rotation. A second harmonic current, equal to one-half the fundamental and having opposite phase rotation, is therefore drawn from the line by a three-phase rectifier. With a six-phase rectifier, the second harmonic is nearly 87 per cent. of the fundamental, but is opposite in phase in the two legs fed from one primary phase so that no second harmonic is drawn from the line, and so on.

TABLE IV—*Harmonic Composition of Square Waves*

Fraction of Cycle	1/2	1/3	1/4	1/6
Direct Current Component	0.5	0.333	0.25	0.167
Fundamental	0.637	0.552	0.450	0.318
2nd Harmonic	0	0.276	0.318	0.276
3rd "	0.212	0	0.150	0.212
4th "	0	0.138	0	0.138
5th "	0.127	0.110	0.090	0.064
6th "	0	0	0.106	.0
7th "	0.091	0.079	0.064	0.045
8th "	0	0.069	0	0.069
9th "	0.071	0	0.050	0.071
10th	0	0.055	0.064	0.055
11th	0.058	0.050	0.041	0.029
12th "	0	0	0	0

The foregoing discussion is ideal to the extent that the transformer reactance has been neglected in all cases. The only generality that can be drawn concerning the effect of transformer reactance is that it tends to damp out higher harmonics more and more. Tube drop also has been neglected. This is a small item for high voltage work and does not affect the form of the conclusions.

In calculating the effect of smoothing networks, the harmonic analysis presents a simple method of attack which is quite exact for the case where some current is always flowing from the rectifier into the load and network. If the current is completely interrupted by the supply of all the current from a storage condenser for part of the cycle, the equations become discontinuous and the harmonic analysis no longer applies exactly. If current always flows thru one or more tubes, the impedance of the network to any ripple frequency can be ascertained and also the cur rent division between network and load. It is possible to tell at a glance which harmonics appear with considerable amplitude and which need not be guarded against. It often happens that a simple aperiodic filter will care for all ripples except one and that one can then be eliminated by the use of a tuned trap, or equivalent means.

In any case a thoro appreciation of the fundamentals underlying the rectification phenomena is very useful in analyzing and interpreting the significance of experimental investigations.

SUMMARY: By Fourier series analysis of the no-load voltage waves of multiphase tube rectifiers, the direct current component, the ripple voltage of the various harmonic frequencies, the available transformer output, and the harmonic composition of the supply current are obtained and tabulated for various cases.

The effect of any unbalanced direct current component in the secondary circuit on transformer operation is described, and practical conclusions as to design are drawn. The effect of smoothing networks is considered, as well as the efficiency of utilization of the tubes in various multiphase combinations

DISCUSSION ON "RADIO TELEGRAPHY"*

By

G. Marconi

Stuart Ballantine (by letter): Dr. Marconi's paper touches briefly upon the interesting phenomena of the propagation of electric waves over the earth's surface, especially viewed from the important geographical position of the antipole of a terrestrial source. Of special interest to me was his recorded opinion (evidently based upon directional observations made in the neighborhood of the antipole) that waves from certain American stations prefer to travel three-quarters of the earth's circumference rather than by the direct route of the great circle. Now this is a very surprising thing, and even marshalling all possible effects, coast-line diffraction, Heaviside layer reflection—and if you please, the relativity deflection due to the earth's east-west rotation—is hard to account for. In the absence of a fuller statement of the experimental conditions I do not see how it is possible to infer this from observations made at one or more closely located points, unless these points are located on the sea and the direction-finding apparatus itself is installed on a wooden raft. For experience with the radio direction-finder indicates with a high degree of probability that most of the permament deviations in direction, at least in the case of an apparatus located on land, are due to diffraction and scattering by local objects, coast lines, and so on, and that these disturbances of the approximately linear propagation do not persist at distances from their foci, which are great compared with the wave length.[1] I presume that the opinion in question is not based upon those transient shifts, studied extensively by Taylor, Kinsley, Pickard, and Loughlin in this country and by Eckersley in England, which have probably no local terrestrial genesis and are for that reason more likely to cause the large deviations which have been frequently observed. The idea that the wave consistently takes any other route than that of the great circle certainly violates

* Received by the Editor, July 18, 1922.

[1] Compare Ballantine; "The Radio Compass," page 10, and following, "Year Book of Wireless Telegraphy and Telephony," London, 1921.

our physical intuition, and should be fortified by experimentally mapping the direction of the Poynting-vector along the entire course of propagation.

The whole matter is one of great scientific fascination, and the important practical economies made possible by the antipodal rise in signal strength impart a great interest to the theoretical problem. The spreading of waves around a spherical earth has been discussed mathematically in the important papers of Rayleigh, Macdonald, Poincaré, Nicholson, Whittaker, and others, but with considerable disagreement of results, and in most cases the discussions have failed to embody such practical features as the finite conductivity of the earth and the possible influence of the Heaviside layer. The effect of the finite earth conductivity has been considered in but one case, in a paper by Rybcynski, and if we accept the experimental evidence lately submitted by Eckersley as to the existence of a super-atmospheric conducting stratum, the approximate influence of this may be treated by a method of Macdonald's[2] for the oscillations in simply-connected spaces. I have carried this calculation thru, assuming perfect earth conductivity, and have obtained formulas which with the complete experimentally determined attenuation law may help to fix the possible height of this reflecting layer. But in view of the great difficulty of the problem, augmented by the impossibility of taking into the mathematical equations all the practical deviations from the idealized problem, I think that practical men will welcome the experimental data promised by Dr. Marconi with considerable satisfaction. It would be especially interesting to know if Fresnel nodal-lines are discovered, and whether the vibrations conspire sufficiently to give an antipodal rise of the order of the theoretical calculations.

[2] Compare H. M. Macdonald; "Electric Waves," page 49 and following, Cambridge, 1902.

DIGEST OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY*

Issued June 27, 1922–August 22, 1922

By

John B. Brady

(Patent Lawyer, OuRay Building, Washington, D. C.)

1,420,629—J. H. Hammond, Jr., filed March 29, 1918, issued June 27, 1922.

Teledynamic Orientation System, wherein a transmitting station is arranged to vary the frequency of radiated impulses from time to time in proportion to the space thru which the controller at the transmitter is moved and a receiving apparatus having a controller arranged to move over a variable path in response to transmitted impulse for local control of circuits in accordance with transmitted impulses.

1,420,824—H. P. Donle, filed May 31, 1921, issued June 27, 1922.

Rectifier, comprising a tube containing two cold electrodes within a vacuum. The electrodes are shaped and arranged so that negative particles tending to leave an electrode or move away from its vicinity will move in a line substantially at right angles to the electrode surface giving a predominating one-way transmission and consequently the quality of rectification.

1,421,015—M. C. Latour, Paris, France, filed April 18, 1916, issued June 27, 1922.

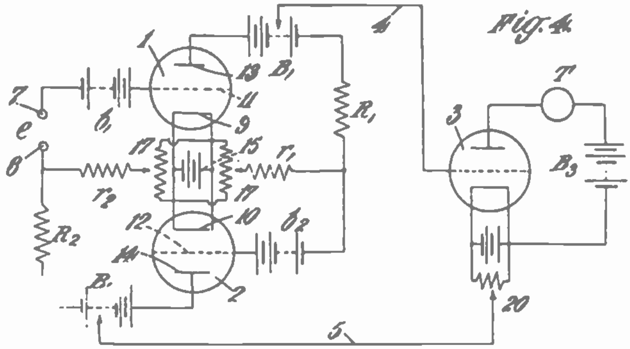
A Telephone having a diafram provided with slits extending in a direction from pole to pole in the diafram for reducing eddy current losses.

1,421,447—Le Roy Kelsay, New York, N. Y., assignor to Western Electric Company. Filed February 8, 1921, issued July 11, 1922.

Switching Key.

*Received by the Editor, September 6th, 1922.

1,422,013—L. B. Turner, filed February 6, 1920, issued July 4 1922.



NUMBER 1,422,013—Thermionic Apparatus Applicable for Radio Telegraphy and Other Purposes

THERMIONIC APPARATUS APPLICABLE FOR RADIO TELEGRAPHY AND OTHER PURPOSES, having a circuit arrangement of two electron tubes. The circuit comprises connections between the anode and filament and grid and filament of each tube, a battery for supplying current to each of the circuits, and a common branch connection containing high resistance in the anode and grid circuits of both tubes whereby the anode of one tube and the grid of the second tube in said apparatus are connected, and the anode of said second tube and the grid of the first-mentioned tube are connected, to form a closed circuit cascade system. Increased amplification by resistance retroactions is claimed for this circuit.

1,422,312—F. S. Smith, filed February 12, 1922, issued July 11 1922. Assigned to The Products Protection Corporation.

ELECTRIC CONDENSER, constructed in tubular form containing as a dielectric a gas such as nitrogen, carbon dioxide, or air permanently sealed. An extended round rod forms one armature of the condenser and a concentric tube the opposite armature thereof.

1,422,429—C. D. Filikins, Schenectady, New York, assignor to General Electric Company. Filed April 21, 1919, issued July 11, 1922.

CONNECTION FOR ELECTRICAL APPARATUS, particularly designed for bringing out connections from the inner rotatable coil of a variometer or variocoupler.

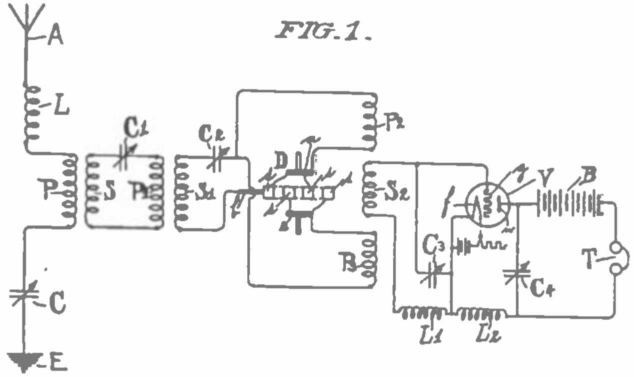
1,422,837—I. B. Crandall, Nahant, Massachusetts, assignor to Western Electric Company. Filed October 13, 1917, issued July 18, 1922.

CONNECTING TRANSMITTERS TO VACUUM TUBE AMPLIFIERS.

1,422,882—H. W. Nichols, filed September 1, 1915, issued July 18, 1922. Assigned by mesne assignments to Western Electric Company, Incorporated.

RADIO TRANSMISSION. This invention relates to a modulation circuit for a transmitter. The modulator tube is inserted as a shunt upon an impedance in the antenna which is comparable with that of the tube so that only one-half or a small portion of the antenna power is thereby carried by the modulator tube.

1,423,345—D. G. McCaa, filed May 20, 1921, issued July 18, 1922. Assigned one-half to Cornelius D. Ehret, of Philadelphia, Pennsylvania.



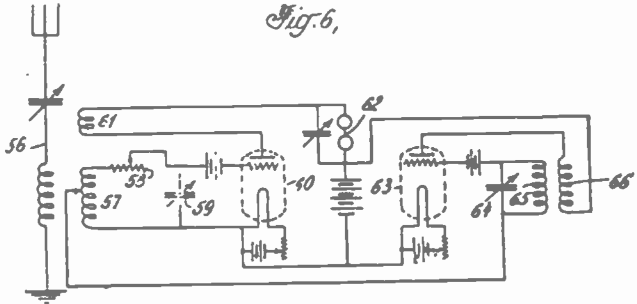
NUMBER 1,423,345—Signaling Method and Apparatus

SIGNALING METHOD AND APPARATUS for receiving radio signals. A local oscillator is employed for generating oscillations of a frequency equal to the frequency of the received signal oscillations. A rotating commutator is used periodically to reverse the coaction of the received and locally produced oscillations at an audible frequency and the resultant impressed upon a detector and telephone receivers. The production of beats is avoided.

1,424,065—Edwin H. Armstrong, filed June 27, 1921, issued July 25, 1922.

SIGNALING SYSTEM. This is the patent on the Armstrong super-regenerative circuit. Twelve diagrams are shown for connecting electron tube circuits in different ways for securing super-regeneration. The electron tube is connected for regenerati n and the circuit adjusted to its maximum sensitiveness. Then the relation between the amount of feedback action and t damping is periodically varied over a certain minimum range a a state of equilibrium produced in which super-regeneratio r obtained.

Fig. 6,



NUMBER 1,424,065—Signaling System

1,324,091—C. R. Fountain, filed January 21, 1919, issued July 25, 1922.

RADIO ELECTRON OSCILLATOR. A cylindrical construction of tube is shown in this patent intended for high power operation.

1,424,141—L. F. Fuller, filed June 16, 1919, issued July 25, 1922. Assigned to Augustus Taylor, of San Francisco, California.

ELECTRICAL OSCILLATION GENERATOR. The patent relates to an anode tip for an arc transmitter. The tip comprises a U-shaped hollow metallic tube thru which cooling water may circulate while the arc discharges from the outer surface of the tube.

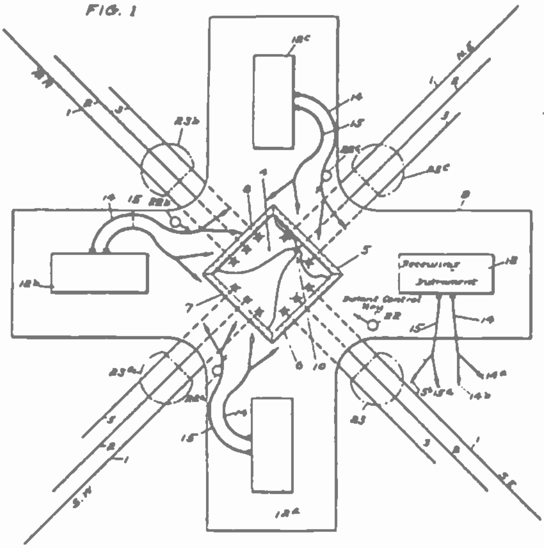
1,424,294—S. G. Frost, filed May 11, 1922, issued August 1, 1922.

SIGNALING UNIT FOR RADIO TELEGRAPHY, comprising a magneto electric generator for furnishing direct current for operation of a radio transmitter. An automatic switch is operated by the rotation of the armature shaft for closing the transmitter circuit in a desired sequence.

1,424,365—E. H. Loftin and H. H. Lyon, filed April 5, 1920, issued August 1, 1922.

RADIO SIGNALING. This patent relates to an arrangement of antenna in a low horizontal system. A plurality of low hori zontal wires of different lengths are extended in different directions and connected with a selector board whereby a receiver may be connected to any number of the wires to choose the components or combinations of components which give the desired optimum antenna length and direction for reception.

FIG. 1



NUMBER 1,424,365—RADIO SIGNALING

1,424,515—John Parkin, Jr., filed April 25, 1921, issued August 1, 1922.

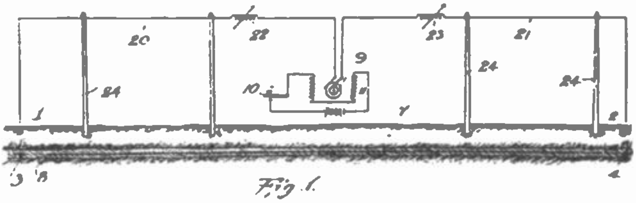
RHEOSTAT having a rotatable graduated dial and a variable resistance element circularly disposed around the back of the dial for contact with a stationary terminal.

1,424,641—J. H. Hammond, Jr., filed December 23, 1918, issued August 1, 1922.

MARINE TRAILER FOR RADIANT-ENERGY RECEIVING SYSTEMS. This patent shows a form of trailer adapted to be automatically detached from a projected body such as a torpedo for extending an antenna conductor out behind the torpedo for radio control of mechanism within the torpedo.

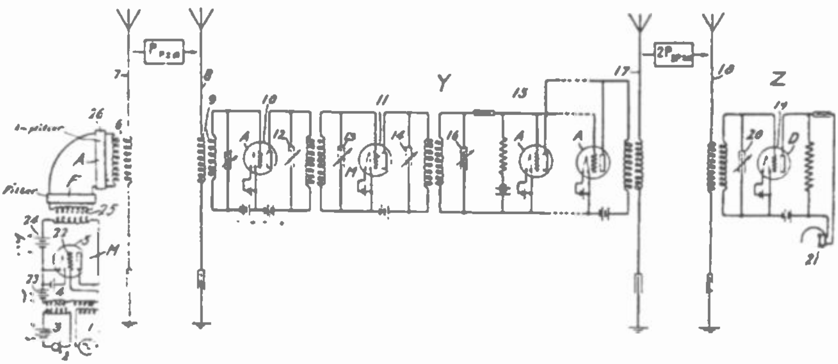
1,424,805—L. De Forest, filed June 16, 1917, issued August 8, 1922.

SUBTERRANEAN SIGNALING SYSTEM, in which overhead conductors 20 and 21 are extended out on pole lines and dropped vertically to plates 3 and 4 buried in the earth in a conductive strata below the surface of the earth. The radio apparatus is connected with the overhead system.



NUMBER 1,424,805—Subterranean Signaling System

1,424,866—P. I. Wold, filed February 17, 1917, issued August 8, 1922. Assigned to Western Electric Company, Incorporated.



NUMBER 1,424,866—Method and Means for Relaying Modulated Carrier Waves

METHOD AND MEANS FOR RELAYING MODULATED CARRIER WAVES, wherein a transmitting station is arranged to control a repeating station having a receiving circuit connected to modulate a transmitter which transmitter retransmits the signal energy to a more distant receiver. The patent is directed to a circuit arrangement for eliminating the tendency of the system to "sing."

1,425,154—R. A. Weagant, filed December 13, 1917, issued August 8, 1922. Assigned by mesne assignments to Radio Corporation of America.

APPARATUS FOR RECEIVING RADIO SIGNALS, including a screened loop antenna shielded from static disturbances. An inverted trough screen covers the loop above and an adjustable circuit provided for connecting the screen to earth.

Fig. 1.

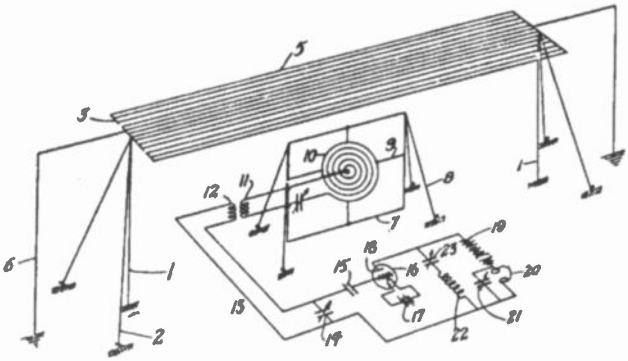
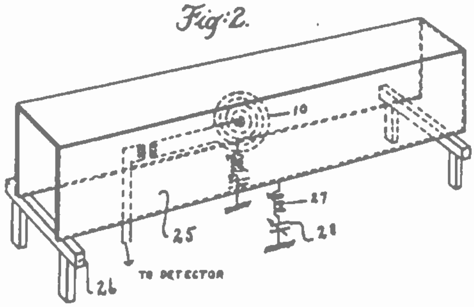


Fig. 2.



NUMBER 1,425,154—Apparatus for Receiving Radio Signals

1,425,522—J. H. Hammond, Jr., filed June 7, 1917, issued August 15, 1922.

SYSTEM FOR SOUND TRANSMISSION, wherein a receiver circuit is used with separate radio controls for generating and trans mitting submarine sound signals of different frequencies in accordance with radio signals transmitted from a distant station.

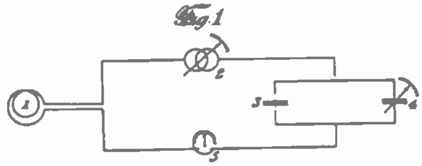
1,425,523—J. H. Hammond, Jr., filed June 22, 1917, issued August 15, 1922.

TRANSMISSION SYSTEM FOR RADIANT ENERGY, wherein a central radio transmitting station is arranged to have its circuits closed by line wire from any one of a plurality of different stations with an indicator at each of the central stations connected by the same wires which form the radio control circuit to advise the operator whether or not the radio transmitter is properly functioning.

1,425,912—R. T. Staples, filed August 17, 1920, issued August 15, 1922. Assigned to Western Electric Company, Incorporated.

Condenser and the Method of Making Same. The condenser unit of this invention is impregnated with paraffin having a solidifying temperature of 120° F. (48.9° C.) and heated to a temperature of approximately 110° F. (43.3°C.) and maintained at such temperature for a period of not less than three hours.

1,426,132—F. N. Waterman, filed March 23, 1917, issued August 15, 1922. Assigned by mesne assignments to Radio Corporation of America.



Number 1,426,132—Electrical Measuring Instrument

Electrical Measuring Instrument for the direct reading of decrement of a circuit. The instrument comprises a pick-up inductance coil 1, a variometer 2 shielded from inductance 1, a fixed condenser 3, and variable condenser 4 and a current-squared meter 5. The instrument is designed after Bjerknes' formula relating to the decrements of two loosely coupled circuits as follows:

$$\delta_1 + \delta_2 = 2\pi\left(1 - \frac{n_r}{n_1}\right)\sqrt{\frac{I_1^2}{I_r^2 - I_1^2}}$$

where:

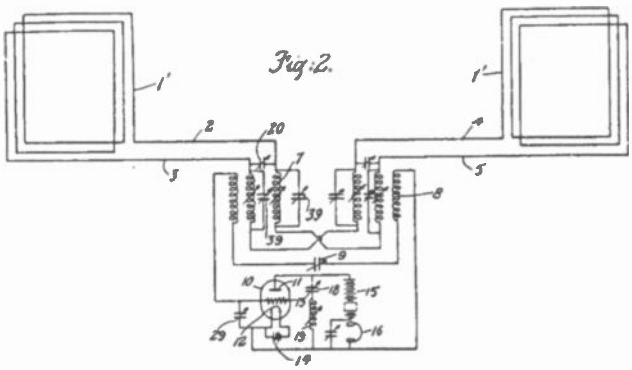
δ_1 = decrement of circuit under test per cycle.
δ_2 = decrement of circuit of instrument per cycle.
n_r = frequency at resonance.
n_1 = frequency off resonance.
I_r^2 = current squared at resonance.
I_1^2 = current squared off resonance.

In operation the condenser 4 is set to its zero or minimum value.

The coil is placed in inductive relation with the exciting circuit, but the variable inductance 2 screened from such action. For this purpose the coil 1 is preferably separate from the rest of the apparatus, but connected thereto by a flexible cable of sf-u ficient length so that, while coil 1 is in inductive relation, the variable inductance, is sufficiently distant to escape any appreciable influence. The meter circuit is then adjusted to resonance by the variometer 2 and the maximum value I_r^2, of the square of the current noted. The variometer is again adjusted, by decreasing the inductance 2, until the current squared is one-half its resonance value. The variometer is left in this last position.

It will be seen that these operations have been conducted with a fixed value of capacity and that the operations necessary to determine the value of the radical in the equation of Bjerknes have been performed. Hence, if the capacity be now changed so as to alter the wave length of the circuit to the other side of the resonant value until I_r^2 is again equal to one-half I_r^2, for the new value of capacity, C_2, there will be but one value of the equation for $\delta_1+\delta_2$. Hence, if the value of condenser 3 is known, and 4 is a calibrated condenser, a scale for $\delta_1+\delta_2$ can at once be calculated, calibration by comparison being unnecessary. Such a scale, graduated in terms of decrement, is directly attached to condenser 4 and when the adjustment for I has been made by means of 4, the sum $\delta_1+\delta_2$ can be read directly from this scale. The instrument is thus direct reading.

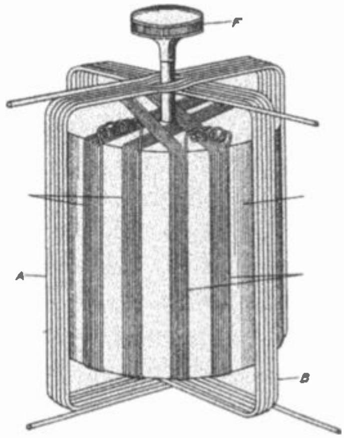
1,426,133—R. A. Weagant, filed December 12, 1917, issued August 15, 1922. Assigned by mesne assignments to Radio Corporation of America.



NUMBER 1,426,133—Apparatus for Preventing Static Interference in Radio Signaling

APPARATUS FOR PREVENTING STATIC INTERFERENCE IN RADIO SIGNALING, comprising a pair of loop collectors mounted vertically and spaced horizontally a fraction of a wave length. A receiving current is provided whereby impulses or electromotive forces set up in the collectors simultaneously, as by static disturbances, are made to oppose each other while oscillations set up by signal waves, being out of phase, act cumulatively on the detector.

1,426,137—G. N. Wright, filed April 16, 1921, issued August 15, 1922. Assigned to Radio Corporation of America.



NUMBER 1,426,137—Radio Goniometer

RADIO GONIOMETER constructed with antenna coils *A* and *B* and search coils *C* and *D*, each consisting of two windings at an acute angle to the other set.

1,426,337—E. A. Sperry, filed July 9, 1917, issued August 15, 1922.

SIGNALING APPARATUS FOR DETECTING SUBMARINES, comprising a system of submarine nets in which a submerged submarine might become entangled with the succeeding carrying away of the net whereupon buoys, normally submerged and anchored, are released and rise to the surface. The buoy contains a complete radio transmitting equipment. The rising of the buoy automatically starts the transmitter, which functions to radiate a particular set of signals enabling a radio direction-finder aboard ships in vicinity to determine the position of the submarine for pursuit.

1,426,465—E. G. Danielson, San Francisco, California, assignor to Elmer T. Cunningham, San Francisco, California. Filed February 14, 1921, issued August 22, 1922.

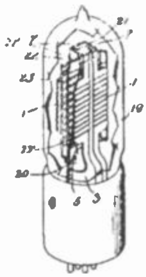
VARIABLE RESISTANCE, comprising a molded mixture of a finely divided conducting material and a synthetic gum which hardens in molding. Contacts are provided along the resistance element whereby its value is made variable by steps.

1,426,516—L. Steinberger, Brooklyn, N. Y., filed April 29, 1918, issued August 22, 1922.

INSULATOR.

1,426,734—W. F. Hendry, filed April 30, 1919, issued August 22, 1922. Assigned to Western Electric Company, Incorporated.

Fig. 14.



NUMBER 1,426,734—Method of Manufacturing Audions

METHOD OF MANUFACTURING AUDIONS, which avoids the use of glass arbor or stem for supporting the electrodes. The plates of this tube are cut from a blank with a collar portion arranged to embrace the glass support in the tube. A metallic box-like connection for the top of the plates carries insulated material forming a support for the other electrodes.

1,426,743—F. J. Kachni and W. L. Kaehni, Cleveland, Ohio, filed June 9, 1922, issued August 22, 1922.

ELECTROMAGNETIC SOUND-PRODUCING DEVICE for operation with a radio receiving apparatus. An arrangement of magnets is provided whereby a vertical armature is vibrated transversely in accordance with the telephonic signal energy in the plate circuit of the last tube. A depression in the end of the armature receives a phonograph stylus imparting mechanical vibrations to the diafram thereof and reproducing the signals through the usual phonograph horn.

1,426,751—P. K. McCall, West Orange, N. J., and H. R. Menefee, New York, N. Y., assignors to Western Electric Company, Incorporated. Filed August 13, 1919, issued August 22, 1922.

THERMIONIC REGULATOR for controlling the potential of an air fan generator for aircraft radio apparatus. An electron tube circuit is provided for regulating the potential. The electron tube is mounted within the stream line casing of the generator.

1,426,754—R. C. Mathes, New York, N. Y., assignor to Western Electric Company. Incorporated. Filed October 23, 1916 issued August 22, 1922.

CIRCUITS FOR ELECTRON DISCHARGE DEVICES.

1,426,755—R. C. Mathes, New York, New York, and S. Read, East Orange, New Jersey, assignors to Western Electric Company, Incorporated. Filed May 9, 1919, issued August 22, 1922.

VACUUM-TUBE CIRCUITS AND METHODS OF OPERATING THEM.

1 426,801—W. Wilson, filed March 21, 1917, issued August 22, 1922. Assigned to Western Electric Company, Incorporated.

REPEATER FOR UNDULATORY CURRENTS, constructed to increase the amplifying power of a tube and prevent the formation between the anode and cathode of an arc or glowing discharge characteristic of positive ionization. An apertured screen is mounted in the space between the cathode and anode and charged positively, increasing the thermionic discharge of thee athode.

1,426,807—H. D. Arnold and J. P. Minton, East Orange, New Jersey, assignors to Western Electric Company. Filed November 12, 1917, issued August 22, 1922.

METHOD AND SYSTEM FOR TESTING TRANSMITTERS OR RECEIVERS.

1.426,788—L. Steinberger, Brooklyn, N. Y., filed November 26, 1918, issued August 22, 1922.

INSULATOR STRAIN MEMBER.

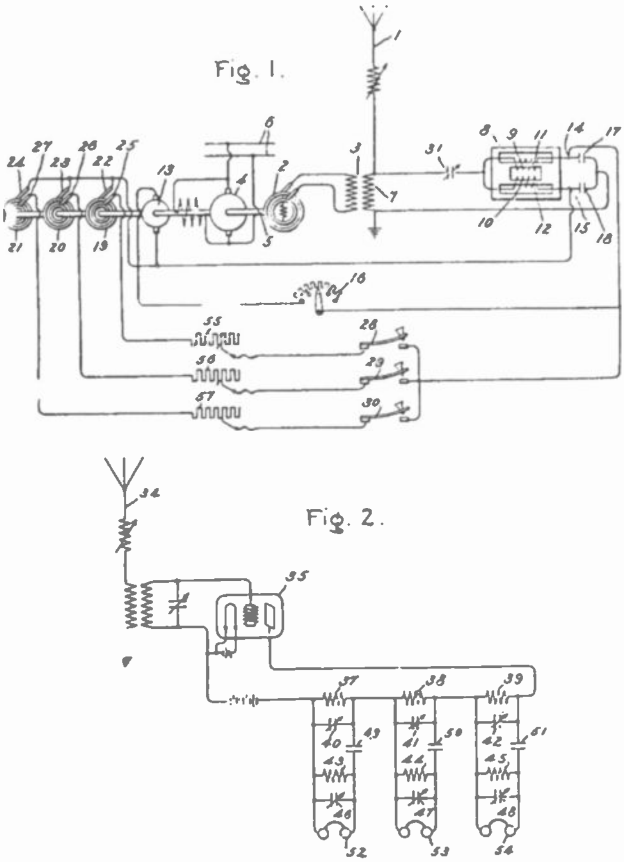
1,426,789—L. Steinberger, Brooklyn, N. Y., filed January 28, 1922, issued August 22, 1922.

INSULATOR.

1,426,826—H. C. Egterton, filed July 19, 1917, issued August 22, 1922. Assigned to Western Electric Company, Incorporated.

Electron Discharge Device Circuits, arranged for the elimination of excessive positive charges on the grids with inherent "blue haze." A leakage path to the filament is provided for the excess positive charges on the grid either by an auxiliary electrode within the tube or by a two-electrode tube connected externally across the grids and filament electrodes of the tube.

1,426,944—E. F. W. Alexanderson, filed September 4, 1917, issued August 22, 1922. Assigned to General Electric Company.



Number 1,426,944—Radio Signaling System

Radio Signaling System for multiplex transmission and reception on the same wave length. A radio frequency alternator is employed at the transmitter with the antenna current controlled by a magnetic amplifier. A plurality of different audio frequency generators are connected thru separate key circuits with the control winding of the magnetic amplifier. In transmitting simultaneously by the separate key circuits the amplitude pulsations produced in the fundamental wave will be the resultant or combination of the amplitude pulsations produced by the several different frequencies employed. At the receiver the antenna is tuned to the fundamental frequency. A secondary receiving circuit is employed with branch circuits, each tuned to the particular frequency of the amplitude pulsations used in the corresponding key circuit at the transmitter.

Partial List of Registered Radio Trademarks

In response to several requests more data is given on those marks published in the list appearing in the August, 1922, issue. A trademark registration remains in force for a period of twenty years with privilege of renewal for similar periods upon the expiration of those periods, by payment of fees prescribed by law.

113,751—"Lenzite," for electromagnatic wave detectors or other electrical devices for receiving or conveying sound waves. Lenzite Crystal Corporation of Pasadena, California. Registered October 31, 1916.

129,698—"Radio Amateur News" for magazines published monthly. Experimenter Publishing Company of New York. Registered March 9, 1920.

140,120—"Radio News" for publications issued monthly. Experimenter Publishing Company of New York. Registered March 8, 1921.

134,313—"Radiophone" for radio apparatus. De Forest Radio Telephone and Telegraph Company of New York. Registered August 24, 1920.

134,314—"Radiofone" for radio apparatus. De Forest Radio Telephone and Tehegraph Company of New York. Registered August 24, 1920.

150,948—"Radiotron" for electron tubes and valves. Radio Corporation of America. Registered January 10, 1922.

134,881—Diagrammatic arrangement of circles with design within the inner circle for telephones and parts thereof. American Thermophone Company, Boston, Massachusetts. Registered September 21, 1920.

152,961—"RADIOLA" for radio apparatus. Radio Corporation of America. Registered March 7, 1922.

153,172—"RADIO MFG. CO." (Mark arranged within a triangular ornamentation) for radio apparatus. Joseph D. R. Freed, of New York. Registered March 14, 1922.

153,238—Representation of Magnavox Loud Speaker with lion head protruding from horn, for loud speaking telephones and other amplifying reproducers. The Magnavox Company, Oakland, California. Registered March 14, 1922.

155,802—"U V 200" for vacuum tubes and valves. Radio Corporation of America. Registered June 6, 1922.

155,803—"U V 201" for vacuum tubes and valves. Radio Corporation of America. Registered June 6, 1922.

155,804—"U V 202" for vacuum tubes and valves. Radio Corporation of America. Registered June 6, 1922.

155,805—"U V 203" for vacuum tubes and valves. Radio Corporation of America. Registered June 6, 1922.

156,671—"VOCALOUD" for loud speakers. John Firth and Company of New York. Registered July 11, 1922.

156,630—"SUPERPHONE" for telephones. G. Boissonnault Company. Registered July 11, 1922.

156,693—"ABC STANDARDIZED RADIO" for radio apparatus. Jewett Mfg. Corporation. Registered July 11, 1922.

156,781—"AERIOLA JUNIOR" for radio apparatus. Westinghouse Electric and Mfg. Co. Registered July 11, 1922.

The following trademarks have been published in the Official Gazette of the Patent Office and unless a notice of opposition is filed within thirty days after the publication, the marks will be registered in the regular course of business. A notice of opposition is simply an opportunity for others legally to contest the registration who may believe that the trademark sought to be registered is substantially identical with a trademark previously used upon goods of the same descriptive qualities; or that a similar registration has been previously made for that same mark by another; or that the mark so nearly resembles such a trademark or known trademark owned or used by another as to be in conflict therewith and void; or that the business of another might be unjustly injured by the registration of such a trademark with its inherent legal force.

It, therefore, behooves members vitally interested in the business of radio to watch the publications of trademarks to eliminate legal difficulties. The numbers given in the following list are serial numbers of pending applications:

159,777—"AIR O PHONE" for radio apparatus. Air-o-phone Corporation, New York, N. Y., claims use since February 9, 1922. Published May 9, 1922.

161,242—"E" arranged with ornamental design for radio apparatus. Everett Electric Corporation, New York, N. Y., claims use since March 21, 1922. Published July 18, 1922.

162,616—"EDELMAN RAYTAINOR" for radio apparatus. Philip E. Edelman, of New York, claims use since August 1, 1921. Published July 18, 1922.

162,669—"R. I. Co." Letters arranged in the form of filament, grid and plate electrodes of an electron tube. Brent Daniel Radio Instrument Company, Incorporated, of Washington, D. C., claims use since September 1, 1920. Published July 18, 1922.

163,422—"RADIO BROADCASTING NEWS" for publications. Westinghouse Electric and Mfg. Co., East Pittsburgh, Pennsylvania, claims use since December 29, 1921. Published August 1, 1922.

163,455—"ETHEROLA" in ornamental design for radio receiving apparatus. Ralph A. McKinney, Pittsburgh, Pennsylvania, claims use since February 1, 1922. Published August 15, 1922.

164,117—"LISSENIN" for radio apparatus. Louis L. Rappaport-Lissenin Radio Parts Co., Ridgefield Park, New Jersey, claims use since January 20, 1922. Published August 15 1922.

164,249—"RADIO WORLD" for publications. Roland B. Hennessy-Radio World Co., New York, N. Y., claims use since April 1, 1922. Published August 15, 1922.

164,438—"SOCOSTAT" for electron tube mounting. J. Roy Hunt, Inc., Long Island City and Astoria, N. Y., claims use since April 15, 1922. Published August 15, 1922.

164,697—"SUPERADIO" for radio transmitting and receiving sets. De Witt-La France Co., Cambridge, Massachusetts, claims use since May 10, 1922. Published August 15, 1922.

164,800—"CHAMPION" for radio telephone jacks and plugs, and radio bulb sockets. Champion Radio Co., Detroit, Michigan, claims use since April 6, 1922. Published August 15, 1922.

164,832—"MINATRON" for vacuum tubes and valves. Radio Corporation of America claims use since May 27, 1922. Published August 15, 1922.

164,833—"RADIOLETTE" for vacuum tubes and valves. Radio Corporation of America claims use since May 27, 1922. Published August 15, 1922.

165,195—"RADIO BUILDER" for magazines. Raymond Kanofsky Radio Builder Publishing Co., New York, N. Y., claims use since June 6, 1922. Published August 15, 1922.

165,321—"VOCAROLA" for loud speakers. Westinghouse Electric and Mfg. Co., East Pittsburgh, Pennsylvania, claims use since December 1, 1921. Published August 15, 1922.

165,657—"RADIOSTAT" for compression resistor carbon pile rheostats. Allen-Bradley Co., Milwaukee, Wisconsin, claims use since April 1, 1922. Published August 15, 1922.