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of
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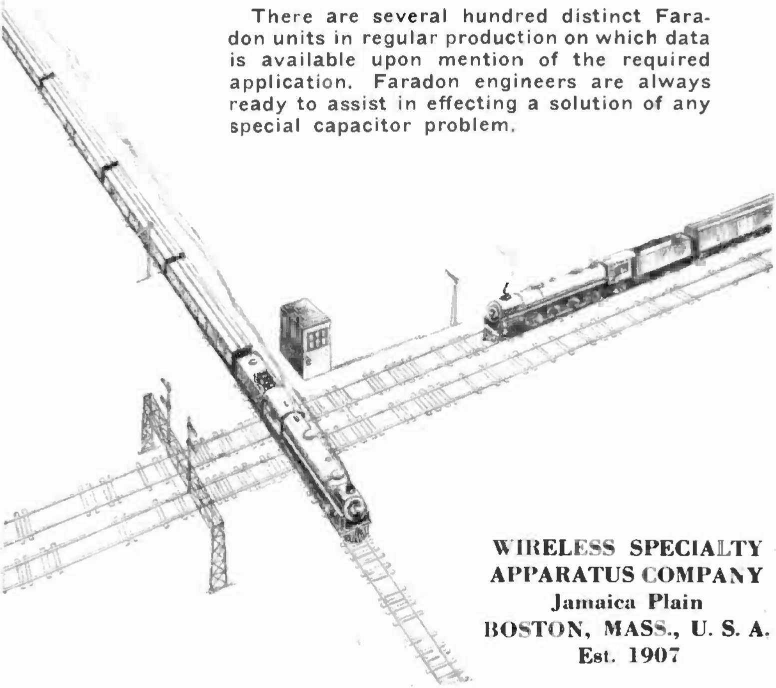

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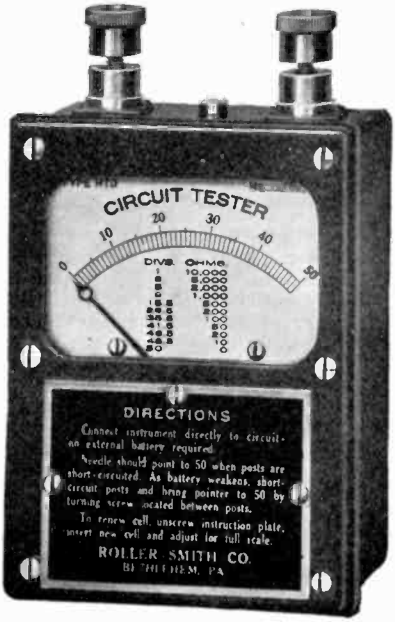
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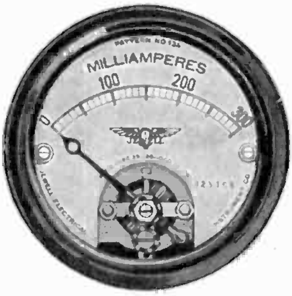
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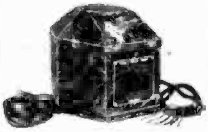

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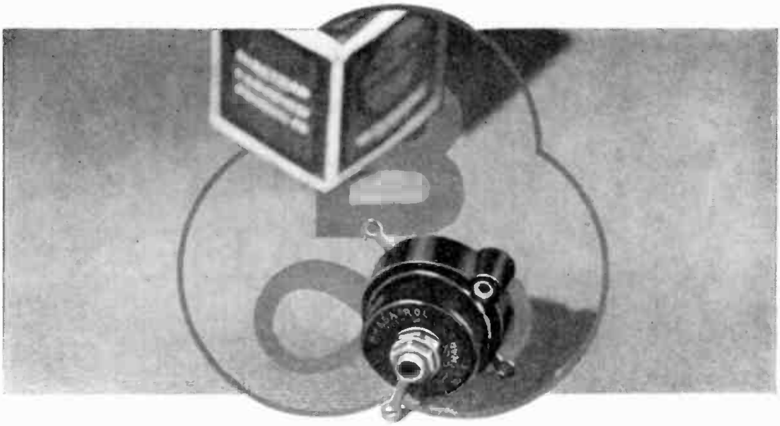

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Filament—5 Volts—.5 Amperes

Plate Voltage	90	135	180	Volts
Negative Grid Bias	16½	27	40½	Volts
Plate Current	10	16	20	Milliamperes
Plate Resistance (A.C.)	2500	2200	2000	Ohms
Mutual Conductance	1200	1360	1500	Micromhos
Voltage Amplification Factor	3.0	3.0	3.0	
Max. Undistorted Output	130	330	700	Milliwatts

R. F. & A. F. AMPLIFIER RADIOTRON UX-226

Filament (A. C.) 1.5 Volts—1.05 Amperes

Plate Voltage	90	135	135	180	Volts
Negative Grid Bias	6	12	9	13½	Volts
Plate Current	3.7	3	6	7.5	Milliamperes
Plate Resistance (A.C.)	9400	10,000	7400	7000	Ohms
Mutual Conductance	875	820	1100	1170	Micromhos
Voltage Amplification Factor	8.2	8.2	8.2	8.2	
Max. Undistorted Output	20	60	70	120	Milliwatts

DETECTOR RADIOTRON UY-227

Heater (A. C.) 2.5 Volts—1.75 Amperes

Plate Voltage	45	90	Volts
Grid Leak	2-9	¼-1	Megohms
Plate Current	2	7	Milliamperes
Plate Resistance (A.C.)	10,000	8000	Ohms
Mutual Conductance	800	1000	Micromhos
Voltage Amplification Factor	8	8	

FULL WAVE RECTIFIER RADIOTRON UX-280

A.C. Filament Voltage	5.0	Volts
A.C. Filament Current	2.0	Amperes
A.C. Plate Voltage (Max. per plate)	300	Volts
D.C. Output Current (Maximum)	125	Milliamperes
Effective D.C. Output Voltage of typical Rectifier Circuit at full output current as applied to Filter	260	Volts

HALF WAVE RECTIFIER RADIOTRON UX-281

A.C. Filament Voltage	7.5	Volts
A.C. Filament Current	1.25	Amperes
A.C. Plate Voltage (Max. per plate)	750	Volts
D.C. Output Current (Maximum)	110	Milliamperes
Effective D.C. Output Voltage of typical Rectifier Circuit at full output current as applied to Filter	620	Volts

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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 15 AUGUST, 1927 Number 8

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INSTITUTE ACTIVITIES

PERSONAL MENTION

Dr. J. H. Dellinger, Junior Past President of the Institute, recently sailed for Europe where he will be traveling for three months in the interest of the Department of Commerce, particularly to study radio aids to navigation in other countries.

Mr. L. E. Whittemore, member of the Board of Direction of the Institute, has been appointed Secretary of the American Delegation to the International Radio Telegraph Conference to be held in Washington, D. C., in October. Mr. Whittemore is now in Washington where he will remain until the conclusion of the Conference. His address is American Delegation, International Radio Telegraph Conference, United Chamber of Commerce Building, 1615 H Street, N. W., Washington.

During Mr. Whittemore's absence from New York, Professor L. A. Hazeltine is serving as Acting Chairman of the Institute's Committee on Standardization.

News of the Sections

ATLANTA SECTION

The first formal meeting of the newly organized Atlanta Section was held in the Atlanta Athletic Club on the evening of July 6th. Major Walter Van Nostrand presided.

The major portion of the meeting was devoted to permanent organization work. The officers of the Section were elected as follows: Chairman, Major W. Van Nostrand; Vice-Chairman, H. L. Wills; Secretary-Treasurer, George Llewellyn; and Assistant Secretary-Treasurer, John M. Keith. The Chairmen of the Committees are: Meetings and Papers, H. L. Wills; Publicity, Lambdin Kay; Membership, C. F. Daugherty.

Following the business session, the members present informally discussed the effect on propagation of waves under varying power using Hertzian antennas.

At a meeting of the Atlanta section held on August 3rd, Major W. Van Nostrand presiding, Ray F. Lovelee, of the Federal Radio Corporation, presented a paper entitled "Broadcast Receiver and Power Supply Designed."

A general discussion followed the presentation of this paper.

There were twenty members of the Institute present at the meeting and at the business meeting and dinner preceding.

The next meeting of the Atlanta section will be held on September 7th in the Atlanta Athletic Club.

PHILADELPHIA SECTION

A meeting of the Philadelphia Section was held on June 24, 1927 in the Franklin Institute. J. C. Van Horn was the presiding officer.

A paper was presented by C. Francis Jenkins on "Radio Vision." Prof. Lloyd M. Knoll, John Arnold and others participated in the discussion which followed.

Due to his removal to New York City, the resignation of David P. Gullette as Secretary of the Section was accepted. John C. Mevius was elected to succeed Mr. Gullette.

The attendance at this meeting was over one hundred.

The next meeting of the Philadelphia Section will be held on September 23, 1927 at which time Prof. Charles M. Weyl will present a paper on "Reproduction of Sound."

ROCHESTER SECTION

On May 20, 1927 at a meeting of the Rochester Section held in the Sagamore Hotel, Benjamin Olney of the Stromberg Carlson Telephone Manufacturing Company delivered a paper entitled "Acoustics as Related to Radio Engineering."

Following the technical portion of the meeting, the following officers were elected for the next Section-year by unanimous vote: Chairman, H. J. Klumb; Vice-Chairman, A. E. Soderhon; Secretary-Treasurer, F. Reynolds. Mr. J. F. Hitchcock was elected Chairman of the Board of Direction of the Section. The following are Chairmen of the standing Committees: Meetings and Papers, V. M. Graham; Constitution, A. E. Roeser; Membership, A. L. Schoen; Rooms, A. E. Soderhon.

SEATTLE SECTION

A meeting of the Seattle Section was held on May 11, 1927 in the Club Room of the Telephone Building, Seattle.

T. M. Libby was the presiding officer. Two papers were read. The first, by Charles L. Reynolds, was on "Our Patent Laws," covering in detail the history and development of our present laws together with an account of some of the requirements for valid patents.

The second paper was presented by E. S. Smith on "Cathode Ray Oscillographs." This paper described the development, construction and use of both hot and cold cathode ray oscillgraphs. The paper was illustrated with slides of equipment, setups and the resulting oscillograms.

At a meeting of the Seattle Section on June 11, 1927, two papers were presented. Charles Peterson delivered a paper on "Loud Speaking Receiver Design," describing the basic design and manufacturing problems of loud speaking receivers.

The second paper, by J. A. Burleigh and O. C. Smith was on the subject of "Sounds We Can and Cannot Hear," which included a striking demonstration of the frequency-response characteristics of the human ear.

The next meeting of the Seattle Section will be held early in September.

I. R. E. Notes

NOMINATING COMMITTEE

For the purpose of recommending to the Board of Direction a program of possible candidates for the offices of President, Vice-President and two Directors of the Institute a nominating committee has been appointed by President Bown. The membership of this committee is as follows: J. V. L. Hogan, Chairman; L. A. Hazeltine and Melville Eastham. The Committee desires to receive suggestions from the Institute membership.

As authorized in the amendment to the Institute's Constitution dated January 1921, "Nomination by Petition shall be made by letter, addressed to the Board of Direction, setting forth the name of the proposed candidate and the office for which it is desired he be nominated. For acceptance, a letter of Petition must reach the Board of Direction on or before October 15th of any year, and shall be signed by at least thirty-five Fellows, Members or Associates."

CHANGES OF ADDRESS

It is very important that the office of the Institute be notified promptly of any change in address of any members of the Institute. Such notification should contain the following information: the former address, the new mailing address and the new business address (if the mailing address is a residential one.)

The Institute cannot be held responsible for misdirected mail if it is not notified at least two weeks prior to the publication date of the PROCEEDINGS.

MEMBER AND FELLOW DIPLOMAS

Members and Fellows of the Institute desiring a diploma of membership are requested to advise the office of the Institute the exact reading of the name they desire to have placed on the diploma.

Diplomas for Members or Fellows can be secured free of charge upon application to the Institute.

CONFERENCE OF ENGINEERING SOCIETIES SECRETARIES

At the third Conference of Secretaries of Engineering Societies held on June 16th and 17th in Cleveland, Ohio, R. A. Carlton, a member of the Cleveland Section of the Institute, represented the Institute. Much information of value was secured at this Conference.

Committee Work

I. R. E. SUBCOMMITTEE ON POWER SUPPLY

At a meeting of the Institute's Subcommittee on Power Supply of the Committee on Standardization, held at Institute Headquarters on July 13, 1927, the following committee members were present: W. E. Holland, Chairman; L. Baker, C. T. Burke, L. W. Chubb, R. A. Klock, E. J. Moore, G. E. Rodwin, G. W. Vinal.

This was the first meeting of the Committee and, despite the summer heat, the Committee remained in session all day and made good progress on the following subjects:

a. Terms and definitions applying to radio socket-power units.

b. Standard method of ripple measurement.

c. Method of measuring inductance of filter choke coils.

d. Method of measuring capacitance and power factor of filter condensers.

Items a, b, and c have been discussed previously in meetings of the Socket Power Committee of the National Electric Manufacturers' Radio Division and have been referred recently to the Institute's Power Supply Subcommittee by the N. E. M. A. Committee as being work that should more properly be undertaken by the Institute.

The Radio Manufacturers' Association Engineering Division is also looking to the Institute's Power Supply Subcommittee to work out standards on the above subjects.

The Subcommittee plans to meet again in September. It is hoped that it will then be able to make final recommendations to the Committee on Standardization on all four of the above subjects.

Sectional Committee on Radio, A. E. S. C.

COMMITTEE ON VACUUM TUBES

A meeting of this Committee was held at the office of the Institute of Radio Engineers. The meeting considered the standardization of tube bases, the preparation of information for the American representative on the Advisory Committee on Tube Base Standardization at the International Electrotechnical Commission meeting in Italy in September, and filament voltage tolerances.

Two types of UX base and tube terminology were adopted for recommendation.

The Committees on Transmitting and Receiving Sets and Installations, Component Parts and Wiring, Electro-Acoustic Devices and Power Supply and Outside Plant have reported no further activities during the month of May.

OBITUARY

With deepest regret the Institute of Radio Engineers announces the death of

Cornelius Johannes de Groot

Dr. de Groot was born in den Helder, Holland in 1883, and was in succession a student at the Technical College, Delft, Holland and the Technical College at Karlsruhe, Germany. For many years he was a Fellow of the Institute of Radio Engineers.

He was one of the pioneers both in the long wave and short wave field of radio communication, was responsible for the installation of the highly modern radio stations in the Dutch East Indies (which communicate directly with Holland and other parts of the world), and has made many fundamental studies of atmospheric disturbances and transmission phenomena which he has published in brilliant papers in the "Proceedings of the Institute of Radio Engineers", and elsewhere. His careful observations and ingenious deductions and conclusions from them have been repeatedly verified by later investigations.

He will unquestionably be remembered as a pioneer worker and inspired investigator, one who brought the benefits of radio communication to the distant parts of the earth, and a most likeable gentleman.

A NEW FREQUENCY TRANSFORMER OR FREQUENCY CHANGER

BY
ISAAC KOGA

Many devices have been published to multiply the frequency of a given alternating current, but there is no static means to obtain 1/2, 1/3 etc. of the given frequency. This paper explains how such frequencies can be obtained by means of a three-electrode vacuum tube. The same scheme can also be used as a frequency multiplier, and even to get frequencies such as 3/2, 2/3, 4/3, 3/4 etc. of a given frequency.

The process is quite simple; to get for example half the frequency of a given alternating current of frequency f_o, construct any valve oscillation generator having natural frequency of about $\frac{1}{2} f_o$ and excite its grid or plate circuit with proper voltage of frequency f_o, the resulting frequency will be just $\frac{1}{2} f_o$. The process of getting any other frequencies having simple ratios to the given frequency is quite similar, so that I shall describe only a few cases in order to give a general idea of the method.

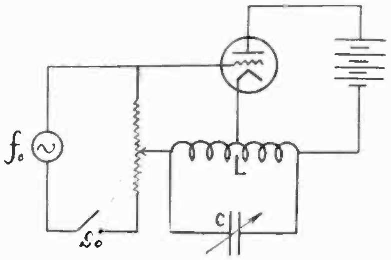

Figure 1

Take, for example, the Hartley circuit (Figure 1), the grid of which can be excited by an external source by means of a potentiometer or a transformer. Without external exci-

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Received by the Editor June 8, 1927.

tation the frequency f of the oscillatory current is given by the natural frequency f_n of the circuit.

$$f = f_n = \frac{1}{2\pi \sqrt{CL}}$$

Let now an exciting voltage of f_c ($\doteqdot 2 f_n$) be gradually increased. At first, at very low excitation, the frequency of the oscillatory current is independent upon the exciting frequency f_o and is maintained at the value f_n, but as the exciting voltage increases, the frequency of the system begins to approach ½ f_o until it suddenly coincides with ½f_o,

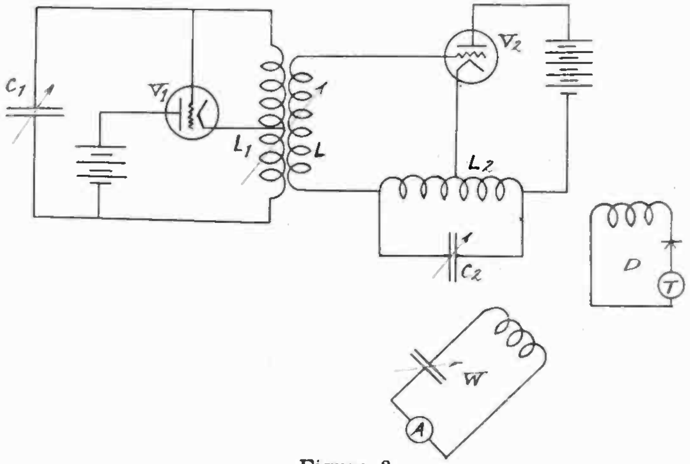

Figure 2

thereby causing sudden depression of plate current at the same time. The necessary exciting voltage may be lower, the smaller the difference of ½ f_o and f_n. Next, keeping the exciting voltage constant, vary the natural frequency of the circuit in one direction, the oscillator frequency f falls suddenly in step with ½ f_o and again suddenly out of step at a certain value of ½ $f_o \sim f_n$. This "attraction" of frequencies is already treated by E. V. Appeleton* in the case of two nearly equal frequencies, and the similar curves as in Figure 9 in his paper were also taken for our cases.

The oscillatory current thus produced in the oscillation circuit contains only a small amount of second harmonic of

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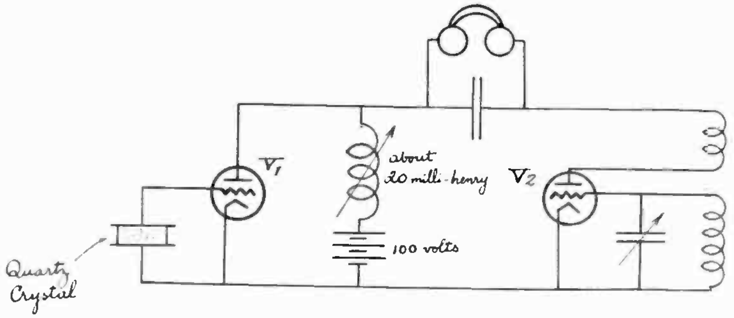
*"Automatic Synchronization of Triode Oscillators", Cambridge Phil. Soc., Proc. 21 pp. 231-248, Nov., 1922.

$\frac{1}{2} f_0$, i. e. the component of f_0 itself, and the wave form seems to be sinusoidal on inspection.

If an excessive excitation, the whole system will be evidently forced to produce a current of frequency f_0; and if the difference $\frac{1}{2} f_0 \sim f_n$ is very large, the system will take another frequency, automatically selecting the nearest to f_n which is yet in a simple ratio to f_0.

In Figure 2, V_1 is used as an exciter, V_2 as an oscillator which is controlled by V_1, both being Radiotrons UX-201-A. The frequency meter W is loosely coupled to the oscillation circuit $L_2 C_2$ only. Adjusting C_2 slowly to and fro, a beat note is heard in the crystal detector circuit D which starts and stops suddenly without giving a tone of low pitch. This shows that the oscillator V_2 suddenly falls into step and out of step with the frequency of V_1 in some simple numerical ratio. The frequency meter W gives two frequencies for which a beat is heard; one the frequency f_0 of V_1 itself and the other a frequency f slightly different from the natural frequency f_n of V_2. (This change of frequency f was already noted). But when the beat note stops, only one frequency is observed and practically no trace of f_0 in the corresponding setting of the frequency meter, which shows that there is now scarcely any component of f_0 in the oscillator V_2. One of my experimental results is shown in the following table, the third column gives the natural frequency f_n of V_2 corresponding to the both limits just before the beat notes are heard under excitation. All numbers are given in kilocycles per second.



V_1 V_2 — Radiotron UX-201-A

Figure 3

Frequency of V_1	Frequency of V_2	Natural Frequency of V_2
544 x 2	544	542 - 551
538 x 3	538	535 - 544
436 x 4	436	432 - 438
445 x 5	445	443 - 448
373 x 6	373	372.5 - 374.5
321 x 7	321	320.5 - 321.5

As a practical application of this apparatus we tried to obtain a constant radio frequency oscillation of long wavelength using a quartz crystal controlled oscillator. In Figure 3, V_2 is a triode controlled by a crystal oscillator V_1. The fundamental frequency of the crystal oscillator is about 70 kilocycles per second and V_2 could be controlled easily down to one-tenth of this frequency i. e. about 7 kilocycles per second. If we use this process in a series of steps, we will be able to get still lower frequencies.

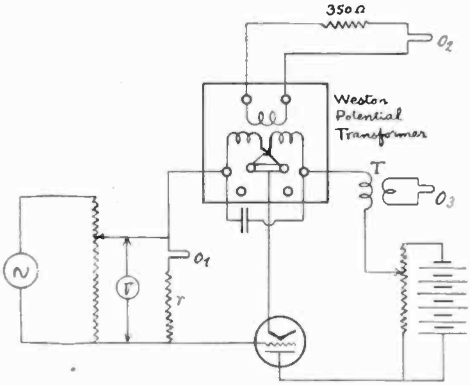


Figure 4

The wave form of the current in the various parts of the circuit and the manner of falling into step were observed by means of a Siemens electromagnetic oscillograph. Audio frequency alternating current was employed for this experiment, a potential transformer (Weston Potential Transformer Model 311, ratios 440 and 220 volts to 110 volts) was used as an inductance in the oscillatory circuit, the exciting voltage being secured from an inductor type alternating current generator designed for 500 cycles at full speed. The general connection arrangement is shown in Fig. 4. A resistance of about 350 ohms is inserted in the output circuit as a load. O_1 O_2 O_3 are oscillograph vibrators, T is a current transformer, and r indicates a series resistance. Figures 5-13 are picked out from about sixty oscillograms which were secured.

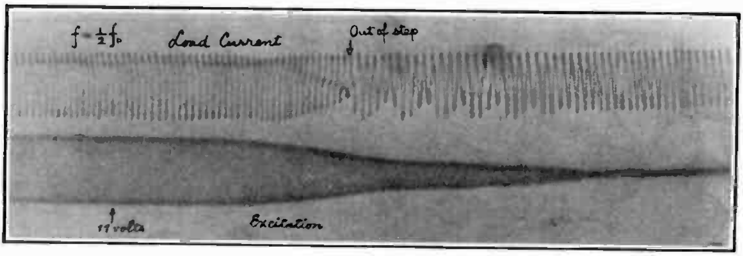


Figure 5

Figure 5 shows the oscillograms of the exciting voltage and load current. The excitation is decreased from 11 volts to zero. "Synchronization" (which is the term I shall employ for the sake of simplicity for the condition when the two oscillators fall into step with a definite frequency ratio) fails to occur below about 7 volts. With no excitation the frequency of the oscillator f is f_n

$$f = f_n \div \frac{2}{5} f_o ,$$

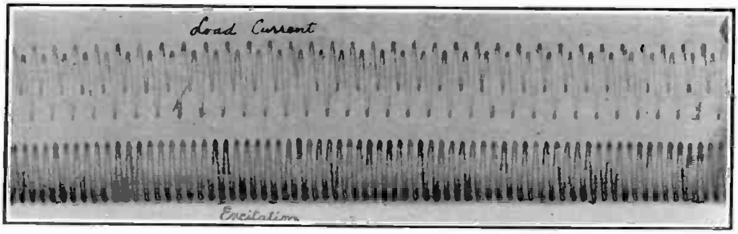


Figure 6

but over 7 volts excitation, we have

$$f = \frac{1}{2} f_o ,$$

and over 11 volts the load current contains considerable amount of exciting frequency component; and still further increase in exciting voltage gives only f_0, and the component of frequency $\frac{1}{2} f_0$ disappears.

Figures 6 and 7 cover the case when the excitations are slightly above and below 11 volts respectively. Figure 8 shows the case when the possibility of free oscillation is altered by decreasing the plate voltage, the excitation being kept at about four volts. Alternating current of 50 cycles is also shown as a timing wave. With certain plate voltage, the oscillatory current cannot be maintained at a fractional value of the given frequency, and the system acts only as a poor amplifier. From all governing considerations, it is probably certain that this type of frequency control cannot be carried out unless the oscillatory circuit satisfies the condition of self-oscillation.

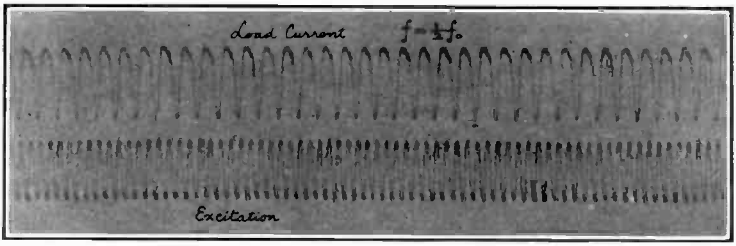


Figure 7

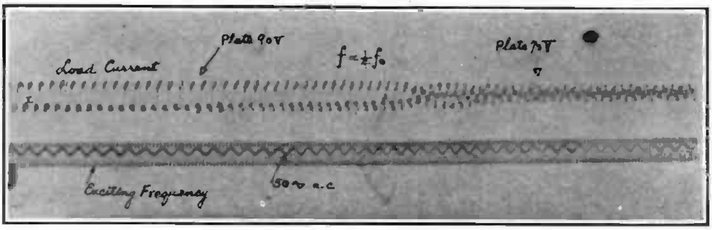


Figure 8

Sometimes when

$$\frac{1}{2} f_0 - f_n > f_n - \frac{1}{3} f_0 ,$$

the circuit is sometimes controlled to a frequency of $1/3 f_0$ at certain exciting voltage, and then to $\frac{1}{2} f_o$ at a little higher voltage. Figure 9 is another example when the oscillator is forced into the relation

$$f = \frac{1}{4} f_o .$$

The excitation is begun from 14 volts, and lowered to zero.

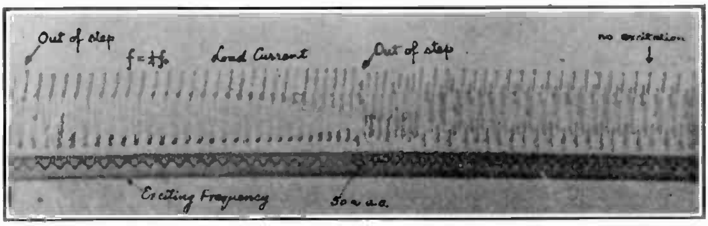


Figure 9

The load current is somewhat complicated but the general characteristics are nearly the same as those of Figure 5 as a whole. Figure 10 shows the plate current and load current when

$$f = \frac{1}{6} f_o .$$

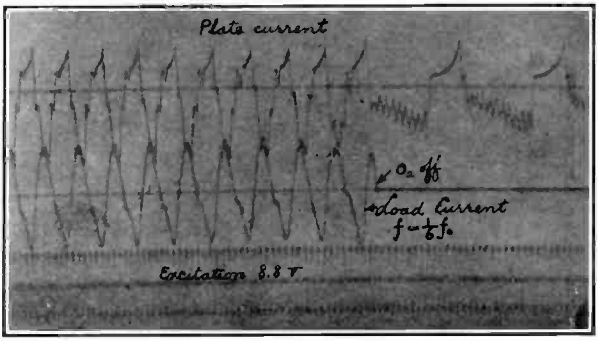


Figure 10

The load being cut off midway, the natural frequency of the oscillator decreases with the increase of inductance, the plate current is distinctly seen to contain the component of exciting frequency. Evidently the synchronization fails to occur under no-load because the oscillation becomes too strong for the same controlling voltage. Similar oscillograms were taken up to $f = \frac{1}{8} f_o$ which are omitted here.

On the other hand when the frequency of excitation f_o is lower than the natural frequency f_n of the oscillatory circuit, the amplitude of load current is generally very small compared with that in the foregoing cases. That the plate current has the exciting frequency as its fundamental frequency is again quite a different circumstance from that in the foregoing cases. In Figure 11 the grid excitation is decreased from 15 volts to zero. It is seen that while the load current is depressed markedly when controlled, it becomes of considerable amplitude when the excitation is removed. The frequency relations are:

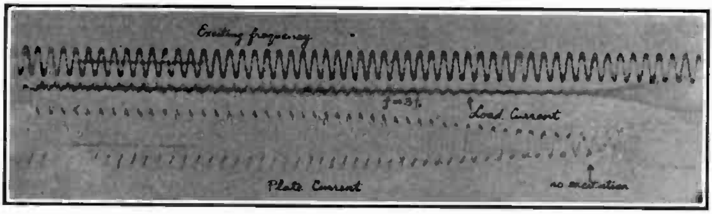


Figure 11

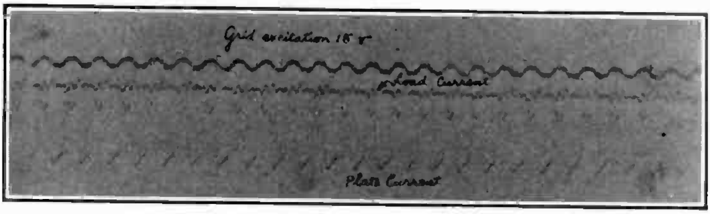


Figure 12

$$f = 3 f_o \text{ while } f_n \div \frac{7}{2} f_o$$

Figures 12 and 13 show the cases when the circuit is controlled so that

$$f = 4 f_o \text{ and } f = 9 f_o$$

These phenomena of synchronization can be explained as due to the "attraction" of two nearly equal frequencies, and to the non-linear characteristic of a vacuum tube. Suppose that the exciting frequency f_0 and the frequency f of the oscillator can be expressed as:

$$f_0 = \frac{n}{m} F, \quad f = F + \epsilon.$$

where F is a fixed value,

m and n are any integers, not including zero,

ϵ is a very small quantity, (positive or negative), compared to F.

As the grid voltage-plate current characteristic is not linear, if the exciting voltage of $\frac{n}{m} F$ is superposed, the current in the oscillatory circuit will be composed of several frequencies, among which the following three frequencies will be predominant:

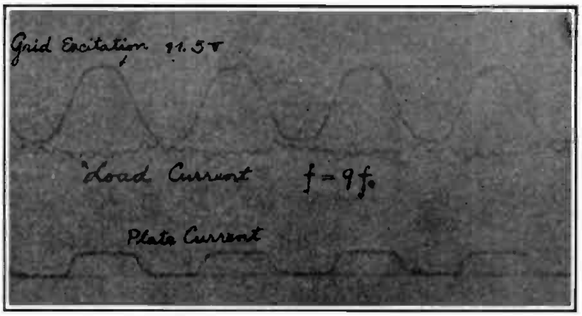


Figure 13

$$F + (n + 1)\epsilon, \ F - (n - 1)\epsilon, \ F + \epsilon.$$

The first two frequencies are the beat frequencies between the m th harmonic of the exciting frequency and the $(n \pm 1)$ th harmonics of the oscillator frequency.

In practice n is ordinarily small, so that the most important components are the last two.

Now, as the exciting voltage increases, these three frequencies cannot coexist independently, and they "attract" each other. But as the value F is fixed by the external force, these three frequencies necessarily become F, thereby reducing the value of ϵ to zero.

It will be sometimes convenient to give a proper name to this new device. It is believed that the role of this apparatus is quite analogous to that of an ordinary step-down transformer. Just as the latter offers a means of obtaining any lower voltage (or current) proportional to the input voltage (or current), so the former provides a means of obtaining a lower frequency which holds any required ratio to the given frequency. From this viewpoint, I propose the name "(Static) Frequency Transformer", for the device. Considered, however, from another point of view, this apparatus gives a frequency having any given ratio to the original frequency, so that the name "(Static) Frequency Changer" is also proposed.

SUMMARY: Alternating current having a frequency which is any fractional value of the frequency of a supplied current, can be obtained by means of a triode oscillator. The phenomenon seems to be due to the "attraction" of two nearly equal frequencies occurring in the triode circuit and to the non-linear characteristic of a triode.

AUDIO FREQUENCY TRANSFORMERS*

By

John M. Thomson

The following assumptions will be made with respect to the characteristics of the transformer.

The primary and secondary inductances are independent of frequency. This is true as long as the flux is uniformly distributed across the core. The inductance will vary with the amount of direct current in the primary, but for each value of direct current the inductance will be constant. Actually the inductance will vary with the strength of the signal. Therefore the amplification will depend on the signal. As the component of the self inductance due to the flux which passes through the coil itself is small the unequal distribution of current across the conductor will have very little effect on the inductance.

The resistances of the primary and secondary coils are independent of frequency. The correctness of this assumption will depend on the size of the conductors used and on the length of the leakage paths. Using the methods developed for power transformers the increase in the resistance of the transformers made by the company with which the writer is associated was found to be less than 1 per cent at 10,000 cycles.

The mutual capacity was taken equivalent to a lumped capacity connected across the terminals. Whether or not this is the correct position for the equivalent capacity is open to discussion.

The self capacity is also assumed to be equivalent to a lumped capacity connected across the terminals of the secondary. The mutual and self capacities are assumed to be independent of frequency but will most likely vary slightly with the frequency if the conductors and the length of the

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leakage paths are such as to give a non-uniform distribution of current.

The exciting current and the core loss have been neglected, but could be represented as in power transformer work by an impedance shunting the primary.

Transient phenomena are neglected.

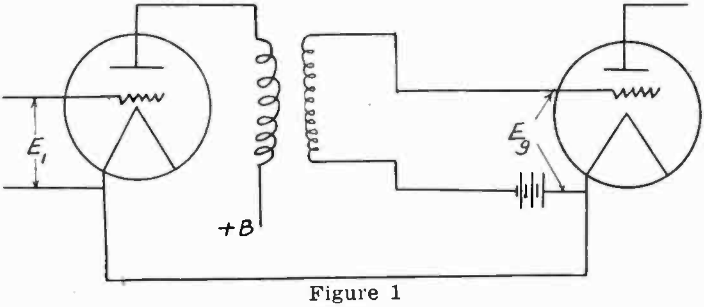

Figure 1

R_p = tube plate impedance in ohms.
U = tube amplification factor = μ, in Figures 5 and 6.
R_1 = resistance of the primary in ohms.
R_2 = resistance of the secondary in ohms.
L_1 = inductane of the primary in henrys.
L_2 = resistance of the secondary in henrys.
M = mutual inductance in henrys.
C_m = mutual capacity between primary and secondary coils in farads.
C_1 = by-pass condenser across the primary plus the self capacity of the primary in farads.

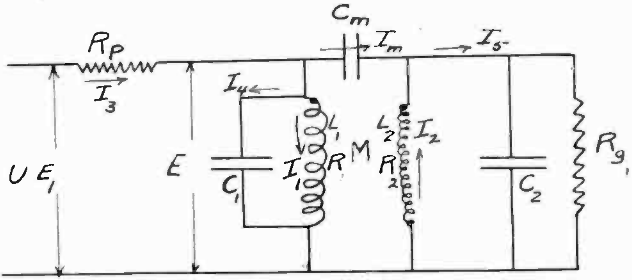

Figure 2

C_2 = Self capacity of the secondary in farads; plus the grid-filament capacity of the tube connected to the secondary of the coil.

$w = 2\pi \times$ frequency in cycles per second.

$$Z_g = \frac{1}{\frac{1}{R_g} + \frac{1}{\frac{1}{j\,w\,c_2}}} = \frac{R_g}{1 + j\,w\,c_2\,R_g}$$

$$Z_1 = R_1 + j\,w\,L_1$$

$$Z_2 = R_2 + j\,w\,L_2$$

The connection of the transformer in the set is given in Fig. 1. Fig. 2 gives the equivalent a-c. electrical circuit.

The following equations were obtained by the use of Kirchoff's Laws:

$$E + I_3\,R_p = u\,E_1 \tag{1}$$

$$I_4 = j\,w\,c_1\,E \tag{2}$$

$$I_3 = I_4 + I_1 + I_m \tag{3}$$

$$I_5 = I_2 + I_m \tag{4}$$

$$E + j\,w\,M\,I_2 = Z_1\,I_1 \tag{5}$$

$$j\,w\,M\,I_1 \quad Z_2\,I_2 + I_5\,Z_g \tag{6}$$

$$I_1\,z_1 + I_2\,z_2 = j\,w\,M\,I_2 + j\,w\,M\,I_1 + \frac{I_m}{j\,w\,C_m} \tag{7}$$

$$I_1\,Z_1 - j\,w\,M\,I_2 - E = 0 \tag{8}$$

$$-j\,w\,M\,I_1 + Z_2\,I_2 + (\,I_2 + I_m\,)\,Z_g - = 0 \tag{9}$$

$$I_1\,(\,Z_1 - j\,w\,M) + I_2\,(\,Z_2 - j\,w\,M\,) - \frac{I_m}{j\,w\,c\,m} = 0 \tag{10}$$

On solving these last three equations for I_1, I_2, and I_3 we get:

$$I_1 = \frac{E}{D}\,\frac{Z_2}{j\,w\,C_m} + \frac{Z_g}{j\,w\,C_m} + Z_2\,Z_g - j\,w\,M\,Z_g, \text{ where} \tag{11}$$

$$D = \left(\frac{Z_1\,Z_2 + W^2\,M^2 + Z_1\,Z_g}{j\,w\,C_m}\right) + Z_g\,(\,Z_1\,Z_2 + W^2\,M^2) \tag{12}$$

$$I_2 = \frac{E}{D}\left(\frac{W\,M}{W\,C_m} + Z_g\,Z_1 - j\,w\,M\,Z_g\right) \tag{13}$$

$$I_m = \frac{E}{D}\,(\,W^2\,M^2 + Z_1\,Z_2 + Z_1\,Z_g - j\,w\,M\,Z_g\,) \tag{14}$$

$$I_3 = I_4 + I_1 + I_m \tag{3}$$

$$I_4 = j\,w\,C_1\,E \tag{2}$$

$$I_3 = j\,w\,C_1\,E + \frac{E}{D}\;\frac{Z_2}{j\,w\,C_m} + \frac{Z_g}{j\,w\,C_m} + Z_2\,Z_g$$
$$-\,2\,j\,w\;\;M\,Z_g + W^2\,M^2 + Z_1\,Z_2 + Z_1\,Z_g\,) \qquad (15)$$

$$I_3 = j\,w\,C_1\,E + \frac{EP}{D} = \frac{HE}{D} \qquad (15a)$$

where $P = \dfrac{Z_2}{j\,w\,C_m} + \dfrac{Z_g}{j\,w\,C_m} + Z_2\,Z_g - 2\,j\,w\,M\,Z_g$
$$+\,W^2\,M^2 + Z_1\,Z_2 + Z_1\,Z_g \qquad (16)$$

and $H = j\,w\,C_1\,D + P$ (17)

$$E + I_3\,R_p = u\,E_1 \qquad (18)$$

$$E = \frac{u\,E_1\,D}{D + H\,R_p} \qquad (19)$$

but $E_g = I_5\,Z_g = (\,I_2 + I_m\,)\,Z_g$ (20)

$$E_g = \frac{E}{D}\left(\frac{W\,M}{W\,C_m} + W^2\,M^2 + Z_1\,Z_2\right)Z_g \qquad (21)$$

Amplification =
$$\frac{E_g}{E_1} = \frac{u\,Z_g}{D + H\,R_p}\left(\frac{W\,M}{W\,C_m} + W^2\,M^2 + Z_1\,Z_2\right) \qquad (22)$$

This expression is rather involved and requires a certain amount of tedious work in using it, but it will give the correct amplification for the assumed conditions.

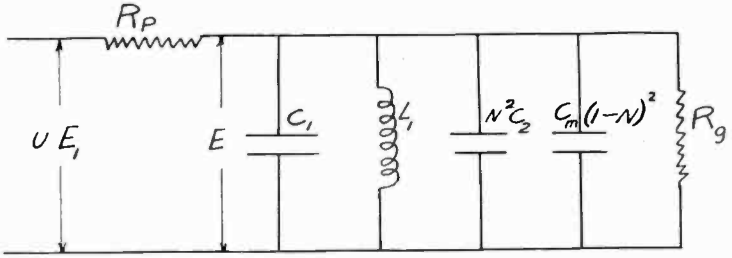

Figure 3

To simplify matters the following assumptions will be made:

Let $R_1 = 0$
$R_2 = 0$

This will not cause any serious error as R_p which is in series with R_1 is usually very large. Similarly R_2 is very small as compared with R_g or $\frac{1}{j\,w\,C_2}$

$$M = \sqrt{L_1\,L_2} \qquad\qquad L_2 = N^2\,L_1$$

This assumption that $M = \sqrt{L_1\,L_2}$ means that the coupling coefficient between L_1 and L_2 is unity, while the leakage in a good transformer will be small (1 per cent or less), the form of the amplification curve at the higher frequencies is largely controlled by the leakage inductance. The amplification obtained by the use of this assumption will therefore not be correct at the high frequencies. Fig. 4 shows the amplification curves obtained by the exact and approximate

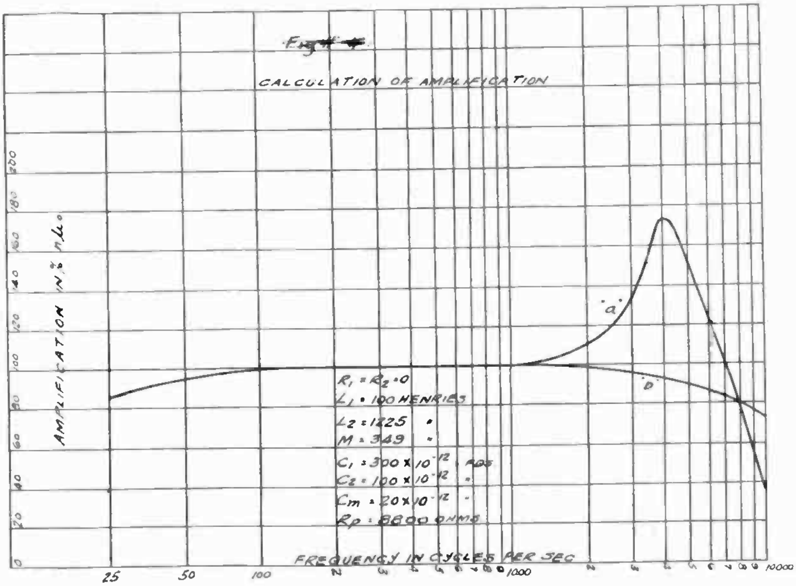

Figure 4—Calculation of Amplification.

methods. The only disagreement is in the upper range. Curve "a" is exact and "b" an approximation.

Since $M = \sqrt{L_1\,L_2}$ and $L_2 = N^2\,L_1$

$$I_3 = j\,w\,c_1\,E + \frac{E}{\frac{R_g}{N^2}} + j\,w\,C_2\,N^2\,E + \frac{E}{j\,w\,L_1}$$

$$+\,j\,w\,c_m\,(\,1 - N\,)^2\,E \text{ and } I_5 = E\,\left(\frac{N}{Z_g}\right) \qquad (23)$$

On examining the equation for I_3 it will be seen that the circuit in Fig. 3 will be equivalent to it.

Let $r_e + j\,x_e$ be the equivalent impedance of the transformer,

then $$I_3 = \frac{E}{r_e + j\,x_e} \tag{24}$$

$$E = u\,E_1 - I_3\,R_p = u\,E_1 - \frac{E\,R_p}{r_e + j\,x_e} \tag{25}$$

$$E = \frac{u\,E_1\,(\,r_e + j\,x_e\,)}{\mathrm{R}_p + r_e + j\,x_e} \tag{26}$$

$$\begin{aligned} E_g = I_5\,Z_g = \frac{E\,N}{Z_g}\,\mathrm{x}\,Z_g = N\,E \\ = \frac{N\,u\,E_1\,(\,r_e + j\,x_e\,)}{R_p + r_e + j\,x_e} \end{aligned} \tag{27}$$

$$\text{Amplification} = \left|\frac{E_g}{E_1}\right| = \left|\frac{n\,u\,E_1\,(\,r + j\,x\,)}{(\,R_p + r_e + j\,x_e\,)\,E_1}\right|$$

$$= \left|\frac{N\,u\,(\,r_e + j\,x_e\,)}{R_p + r_e + j\,x_e}\right| = \frac{N\,u\,\sqrt{r_e^{\,2} + x_e^{\,2}}}{\sqrt{(\,R_p + r_e\,)^2 + x_e^{\,2}}} \tag{28}$$

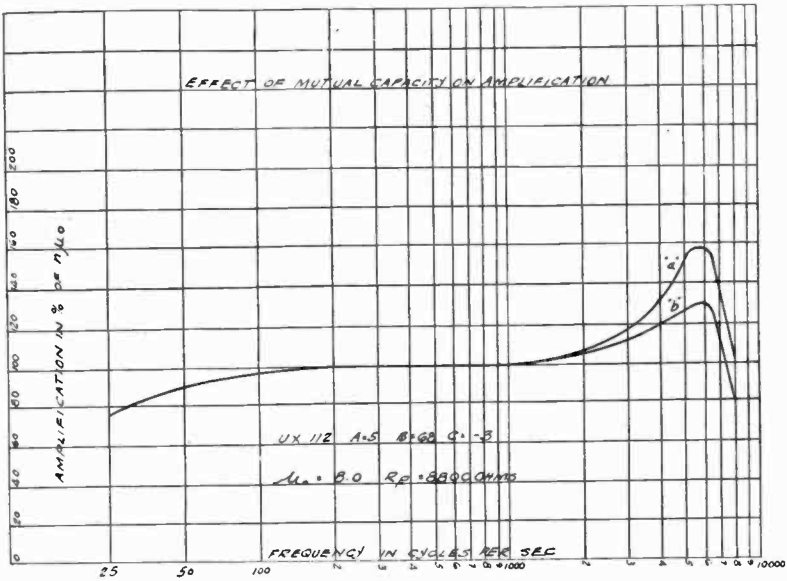

Figure 5—Effect of Mutual Capacity on Amplification.

The greater $r_e + j\,x_e$ is the closer the amplification approaches to $n\,u$.

For maximum amplification $r_e^2 + x_e^2$ must be very large as compared with $R_p^2 + 2\,r_e\,R_p$.

r_e is usually small and if the grid filament resistance is large enough to be neglected r_e will be zero.

The only characteristics of the tube required to give the amplification curve is the plate impedance and the amplification factor of the tube. There is no simple relation between these two factors which would give us a means of comparing the relative merits of tubes.

If the direction of the secondary current is reversed,

$$P = \frac{Z_2}{j\,w\,c\,m} + \frac{g}{j\,w\,c\,m} + (\,Z_2 + 2\,j\,w\,M + Z_1\,)\,Z_g + W^2\,M^2 + Z_1\,Z_2$$

The rest of the proof is similar to the first part.

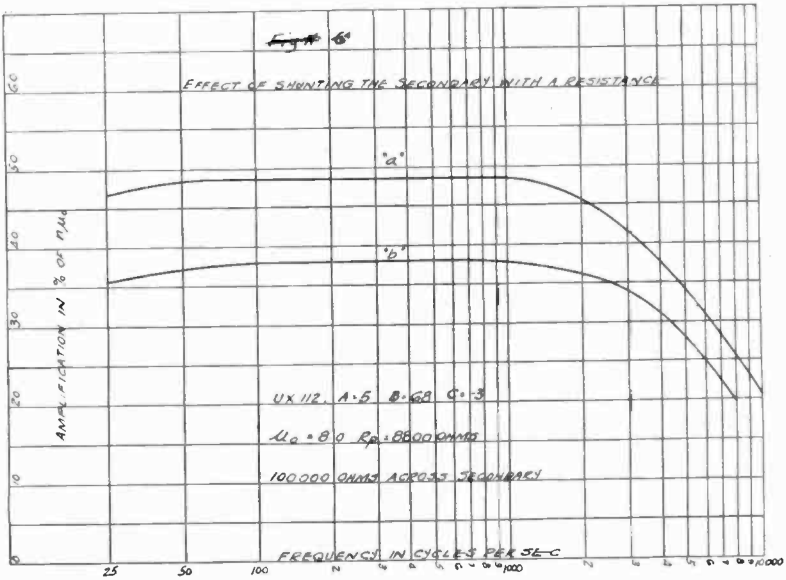

Figure 6—Effect of Shunting the Secondary with a Resistance.

When $R_1 = R_2 = 0$ and $M = \sqrt{L_1\,L_2}$ the same equivalent circuit will be obtained with the capacity C_m now increased to $(\,1 + N\,)^2$ instead of $(\,1 - N\,)^2$.

Fig. 5 shows the effect of the mutual capacity on the amplifications curve of the transformer. Curve "a" was obtained with the mutual capacity of the transformer neglected. Curve "b" shows the effect of the mutual capacity. A, B and C are filament, plate and grid battery voltages.

The curves in Fig. 6 show the effect of a shunted secondary. Curve "a" was calculated from assumed constants while curve "b" is a curve obtained from test. The form of the two curves is very similar.

In conclusion the results obtained for the assumed conditions are as follows:

The maximum amplification is obtained when the equivalent impedance of the transformer is very large as compared with the plate impedance of the tube. There is no simple relation between the plate impedance and the amplification constant of the tube, which will give the performance of the tube when used with audio frequency transformers. The mutual and self capacity of the coils acts as a pass at the higher frequencies and the effect depends on the turn ratio of the transformer.

SUMMARY: In the foregoing paper a method for calculating the amplification curve of an audio frequency transformer is developed in terms of the usual constants of the transformer and of the tube. The distributed capacity of the coils and the mutual capacity between the primary and secondary coils are represented by lumped capacities. The exciting current of the transformer is neglected. As the equation for the amplification in the vector form is rather involved an approximate formula is then developed and its limitations pointed out.

NOTES ON THE TESTING OF AUDIO FREQUENCY AMPLIFIERS*

By
Edward T. Dickey

Fundamentally, the testing of audio frequency amplifiers consists of an examination of the operational and structural features of the amplifier with the view to determining the following points:

1. The voltage amplification, with special regard to the relative amplification at various points along the audio frequency range.

2. Ability to produce the necessary output power for the particular use for which it is intended, without distortion of the impressed wave form.

3. General operational and structural practicality with reference to such points as: freedom from audio frequency howling, susceptibility to howling through lengthening of the input and output leads, correct values of grid and plate potential for the particular tubes used, etc.

Of these points, the most important, and those requiring the most equipment and care in their determination, are undoubtedly the first two. The greater portion of this paper, therefore, will be devoted to consideration of these points.

It is, of course, understood that a test of the audio frequency amplifying system constitutes by no means the entire story on the fidelity of a complete receiving set. The various tuning, amplifying, and detecting stages preceding the audio frequency amplifier, all have their effect on fidelity. Those interested in a discussion of the other factors affecting fidelity and their measurement, are referred to a contemporary paper on "Notes on Receiver Measurements" by Messrs. T. A. Smith and G. Rodwin.

1. Voltage Amplification Measurement

It is not intended in the short space of this paper, to discuss all the methods, or even all of the most important

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methods, of attacking this problem. Discussion will be confined to the methods which have been found by the author to give the best results in combined accuracy, speed of test, and simplicity of operation.

The apparatus necessary consists of the following three essentials:

A. A source of variable audio frequency potential, capable of supplying the required amplifier input potential, and producing oscillations whose wave form differs from the sinusoidal to a negligible degree.

B. Means for accurately measuring the potential impressed upon the input of the audio frequency amplifier under test.

C. Means for accurately measuring the output potential from the amplifier, without appreciably influencing any characteristic of the amplifier.

Description of the apparatus considered most suitable for these three purposes is given below:

A. *The Oscillator.* The source of audio frequency potential found most satisfactory was an oscillator of the beat frequency variety, employing the difference frequency of two radio frequency oscillators. By combining and rectifying these frequencies, and then amplifying the resultant frequency, it was possible to obtain from the oscillator of the type used by the author, audio frequency potentials of the order of 50 volts. The wave shape of the oscillations obtained from a properly constructed oscillator of this type, can be made to contain not more than 1 or 2 per cent of harmonics, and the output potential will remain effectively constant over the frequency range from 30 to 10,000 cycles approximately, without readjustment.

It has become a fairly well established practice in radio laboratories, to use a logarithmic scale for the frequency, in plotting audio frequency characteristic graphs. By properly shaping the plates of the frequency adjusting variable condenser, it is possible to make the frequency produced by the oscillator vary logarithmically with respect to the condenser dial. The advantage of this will be evident when the method of using this oscillator in testing audio frequency amplifiers is described. (See section on General Description of Amplification Test Method).

B. *Input Potential Measurement.* The output of the oscillator used is adjustable internally, and thus for this measurement, it is only necessary to employ a thermo-milliammeter of reliable make, which is accurate over the audio frequency range. Associating this meter with the necessary series resistance, which must be accurate over the range of audio frequencies used, potentials over the desired range may be measured and applied to the amplifier input. As stated previously, the output potential of the oscillator when once set, will remain appreciably constant over the range from 30 to 10,000 cycles, which is sufficient for the usual amplifier measurements. It is considered good practice to read the input voltage again at the end of the test run, to make sure that the oscillator is working properly.

C. *Output Potential Measurement.* Considering the requirement that the measuring device used here shall not affect the characteristics or output of the amplifier, the most obvious instrument to use is a voltmeter of the vacuum tube type. To simplify the operation, and to permit readings to be taken rapidly, a direct-reading type of vacuum tube voltmeter is desirable.

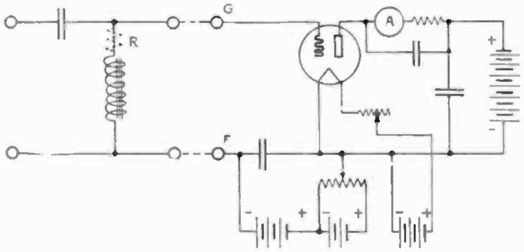

Figure 1—Circuit of Direct-Reading Vacuum Tube Voltmeter

To facilitate operation in connection with the curve drawing mechanism, which will be described later, a tube voltmeter having very nearly a linear characteristic is needed. By the use of a UX-171 tube, it was found possible to construct a voltmeter having a number of very desirable features for amplifier output measurement purposes. This tube has a low plate impedance, and a high grid bias potential, and it is possible, by its use, to obtain a tube voltmeter which is capable of measuring directly potentials up to approximately 70 volts, and which has a reasonably straight characteristic curve from about three volts to the end of the scale. A high plate potential, a high series plate resistance, and a microammeter having a full scale deflection of 100 microamperes, were contributing factors in producing the straight line characteristic.

The circuit used is shown in Fig. 1. Because of the relatively low plate current (maximum approximately 100 microamperes), it is possible to run the filament of the tube below its full rated value (4.5 volts permissible), thus prolonging its life and constancy of calibration. For increasing the accuracy with which the zero reading of the meter might be set, it was found best to use a reading of 10 microamperes as zero, calibrating the meter scale from that point. The tube voltmeter is calibrated using the oscillator and thermoammeter described before. On turning on the filament and grid potentials, it is merely necessary to make a slight adjustment of the potentiometer until the microammeter reads ten. Changes in the zero reading during operation are relatively negligible. The condensers connected across various points of the circuit are used to pre-

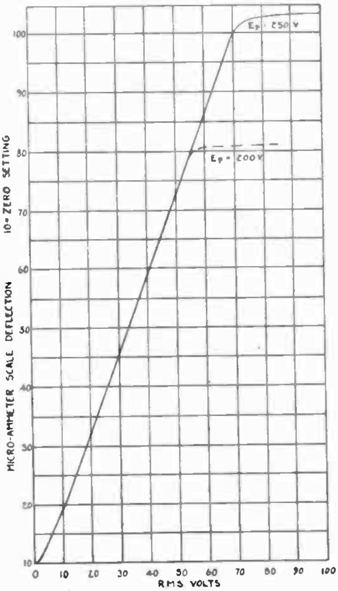

Figure 2—Calibration Curve of Direct-Reading Vacuum Tube Voltmeter (0-70 v. Range).

vent inequalities in reading at the high and low frequencies. As shown, the instrument is equally accurate over the entire audio frequency range within observational error percentage.

At first glance it seems somewhat dangerous to use a plate meter of such high sensitivity in a circuit having such a high applied potential. By a very simple calculation, however, it will be evident that even if the plate and filament of the tube should become short circuited, the high value of series plate resistance would prevent the plate meter from being overloaded. As will be seen from the calibration curve (Fig. 2) the plate meter reading remains practically constant after it has reached a reading of approximately 104 microamperes. While slightly past the end of the scale, this will not injure the instrument.

In amplifiers where there is no direct current flowing through that portion of the output circuit across which the tube voltmeter is connected, the binding posts marked *G* and *F* are connected direct to the amplifier circuit. In cases where the amplifier output circuit contains a direct current component, use is made of the condenser and choke combination shown to the left of the circuit in Fig. 1. An inductance is used for the grid potential return, rather than an ohmic resistance, because of the trouble experienced with the condenser and resistance leak combination, due to the slowness of such a system in returning to normal, after a sudden high potential surge such as is often experienced through induction effects on sets operating on commercial lighting circuits. The resonant period of the inductor and capacitator combination must be outside of the range of audio frequencies used. Actual test has shown this circuit to give appreciable equality of operation over the usual audio range, if the impedance of the choke is kept at least ten times as great as that of the output device. This is quite easy if a loud speaker of usual type is connected to the amplifier output. In cases where the output potential is to be measured across a relatively high ohmic resistance, it is advisable to use an ohmic resistance leak in series with the choke as shown at *R* in Fig. 1. This resistor should have a value approximately 20 times as great as the resistance across which the potential is to be measured.

By a simple alteration of circuit constants, the tube voltmeter may be arranged for the measurement of potentials of the order of a few volts. In Fig. 3 is shown a circuit making available two ranges by the mere throwing of a double-pole double-throw switch. While the low range will not have quite as straight a characteristic as the higher range, it is still quite satisfactory, as will be seen from Fig. 4, which gives the curve for a range having a maximum of about seven volts. A very compact and convenient tube voltmeter of the type described above is shown in Fig. 5.

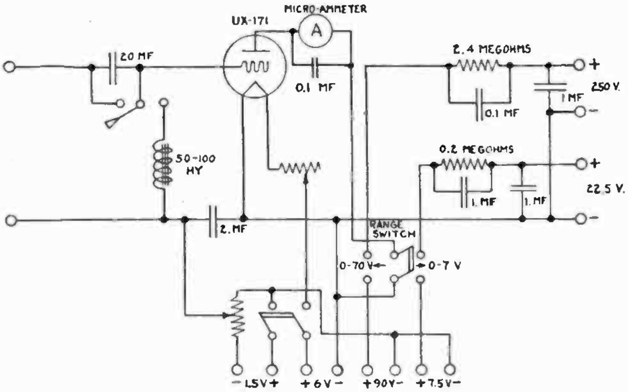

Figure 3—Circuit of Direct Reading Vacuum Tube Voltmeter Arranged for Measurement of Two Ranges of Potentials (0-7 v. and 0-70 v.)

Like all vacuum tube voltmeters using a d-c. indicating instrument, the one described here will have its accuracy affected by the wave form of the potential under measurement. It will be found to be considerably better in this regard than some tube voltmeters, however. Those interested in the effect of wave form on the accuracy of various types of tube voltmeters, are referred to two recent articles on "The Thermionic Voltmeter" by Messrs. W. B. Medlam, and U. A. Oschwald, in the October and November, 1926 issues of *Experimental Wireless and the Wireless Engineer.*

If calibrated by the use of a potential of good wave shape, this tube voltmeter should retain an accuracy of approximately 1 per cent of full scale for a considerable period. Briefly, the points upon which retention of its accuracy depends are, in order of importance:

1. The accuracy of the plate meter, (directly dependent.)

2. Retention of the same value of series plate resistance. This value can be checked quite simply, however. By merely connecting the plate terminal of the tube socket to the negative filament terminal, and lowering the plate potential somewhat, the resistance can be directly obtained by calculation from the impressed potential and plate meter reading.

3. Tube constants. If a tube with normal constants is selected, the low plate current and filament current result in retention of appreciably the same constants for a considerable period.

4. Filament potential. Variation of 10 per cent causes a variation of less than 1 per cent in reading, and therefore accuracy is relatively independent of this factor within these limits.

5. "Zero setting" of the plate meter. Setting for 10 microamperes is important, but is extremely easy to check and, when set, practically never varies over ½ per cent even over a considerable period of continued use.

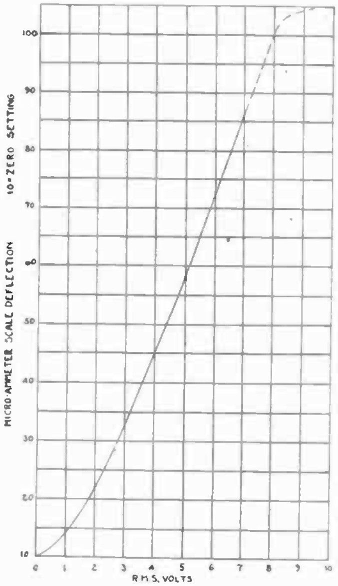

Figure 4—Calibration Curve of Direct-Reading Vacuum Tube Voltmeter (0-7 v. Range)

6. Plate potential. As long as the zero setting is correct,

a variation of as much as 20 per cent in plate potential will cause less than 1 per cent variation in reading, up to the "cut off" point. This point, i. e., the point above which the plate current will not rise with increased input potential, will be lowered. This can be noted in Fig. 2, where the horizontal line marked "200 volts" shows where this point would be for this value of plate potential.

7. Grid potential. It is not necessary to know the actual value, since this is set by watching the plate meter. Its numerical value is of interest only when one wishes to be sure that the tube is not acting as a load across the circuit under measurement, by drawing grid current.

TEST OF MULTI-STAGE A. F. AMPLIFIER

It is important to arrange test conditions so as to approach, as nearly as possible, those which will exist in actual operation. In this connection, the following points are important:

1. Input and output impedances should be the same as in actual use.
2. Long input or output leads, or anything tending to produce interstage couplings which would not be present in normal operation, should be avoided.
3. Values of input potential should be chosen so as to avoid overloading the output tube, and tests should be made preferably using two or more values of input potential, to discover any alteration in operating characteristics with varying inputs.
4. If the amplifier input consists of a transformer winding, provision should be made to have normal direct current flowing through the winding, if the latter is to be connected directly to a tube plate circuit, as is usually the case.
5. Conditions of grounding or shielding should be similar to those of actual use.

In testing a complete amplifier, the types of tubes which are intended to be used in it are generally known. The only points which may be in doubt are usually the characteristics of the output device (loud speaker), and the tube (generally detector) which will precede the amplifier. Usually the amplifier input will be in the form of a transformer primary winding, in which case amplification curves may be taken with several values of series input resistance, and primary direct current.

A simple circuit for use in obtaining the amplification characteristic is shown in Fig. 6. The adjustable resistance R_2 provides means for simulating the plate resistance of the tube preceding the amplifier. R_1 should be small in value compared wih the impedance of the thermometer and oscillator output circuit. In case R_1 is appreciable with respect to R_2, the sum of the two should be made equal to the desired series input resistance. By means of a source of adjustable potential E_1 and a milli-ammeter, a direct current, equal to the plate current of the tube feeding the input, may be set up in the input primary winding.

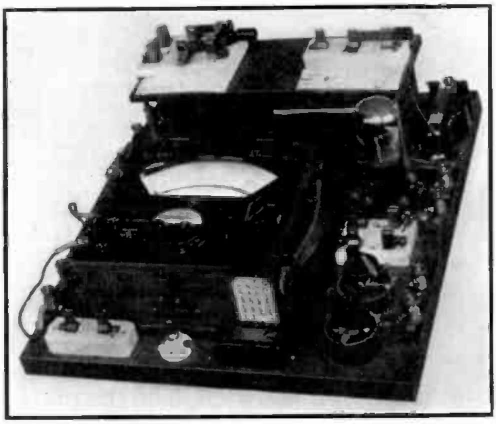

Figure 5—Vacuum Tube Voltmeter.

If the amplifier input is a tube grid circuit, the oscillator potentials may be impressed directly thereon. In the case of resistance coupled amplifiers, the input will probably be in the form of a plate resistance. Various series resistances to imitate tube plate resistance, may be tried between this and the input potential source, but it will generally be unnecessary to put a direct current through the plate resistance. If there is doubt as to the ability of this resistance to carry the necessary current quietly, and without changing its resistance, it is advisable to investigate the resistance in question, separately.

In cases where the amplifier input circuit is not fed from a tube, electrical characteristics simulating as nearly as possible those of actual use, should be arranged at the point where the input potential is impressed on the amplifier.

In case the characteristics of the loud speaker with which the amplifier is to be used are not known, it is generally best to use a resistance in the output circuit of the value for which the output tube is designed. From the curve thus obtained, and a knowledge of the plate impedance of the output tube, it is possible to determine by calculation, the characteristics of the amplifier with any loud speaker whose impedance characteristics are known.

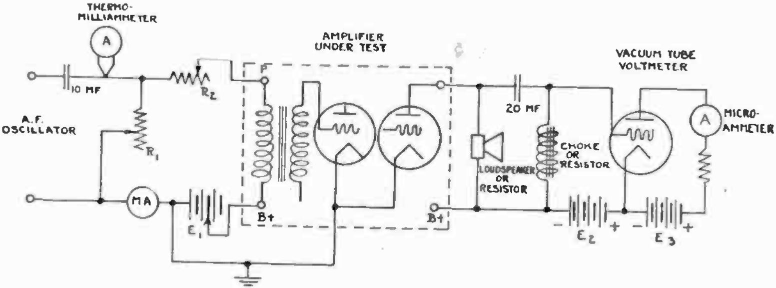

Figure 6—Test Circuit for Transformer-Coupled Amplifier of Several Stages.

TESTING INDIVIDUAL AMPLIFYING TRANSFORMERS

While it is the primary purpose of this paper to discuss the testing of complete amplifiers, the testing of individual audio frequency transformers is believed to be of sufficient importance to warrant a brief separate treatment here. This is a more specialized problem than the testing of completed amplifiers, but is of particular importance for design purposes.

It should be noted that it is practically impossible to construct a test set up which will simulate accurately the circuit reactions to which a transformer will be subjected in a completed amplifier. Thus it is recommended that not too much dependence be placed on results obtained from test of a single transformer, which is intended for use in a multistage amplifier. Features of interstage coupling, input and output coupling, tubes, transformers in other stages, and loud speaker characteristics, all enter into the problem, making it exceedingly difficult to duplicate in the test of a single transformer, the conditions under which it will work in the completed amplifier.

In Fig. 7 will be found a circuit which can be made to simulate to some extent, the characteristics of a single stage of a multi-stage amplifier.

The remarks made previously in connection with Fig. 6 regarding the input circuit adjustments, apply here also. It is recommended that the circuit be grounded at the points indicated, since these points will, in general, be at ground potential in the final amplifier. If the core of the transformer is to be grounded in the final amplifier, it should be grounded during the test. It is important that the grid potential battery of the output measuring tube voltmeter be connected at the point indicated, and not in the grid lead as is sometimes done, otherwise excessive capacity will exist across the transformer secondary. If the grid-filament capacitance of the tube-voltmeter tube is not sufficiently close to that of the tube which is to be used on the secondary side of the transformer, either of two possible expedients may be used:

1. An extra frequency run may be made with a small condenser *C* connected across the transformer secondary. From the two curves obtained, and knowing the grid-filament capacitance of the tube voltmeter, and of the tube which is to be used with the transformer, it will be possible to determine the curve for the tube to be used by extrapolation.

2. Use the type of tube which will be fed from the secondary of this transformer in the final amplifier, and connect the tube voltmeter across a resistance in the plate circuit of this tube. Tested in this way, of course, the voltage amplification measured includes the amplification factor of the tube following the transformer. To get the voltage amplification of the transformer, it is therefore necessary to divide the values obtained by the amplification factor of this tube.

If it is desired to determine the general characteristics of a transformer, so that its behavior with various types of tubes may be deduced, this may be done with fair accuracy by calculations based upon certain measurements. To do this it will be necessary to find:

1. The curves of reactance and effective resistance variation with frequency of the primary winding, with the secondary open. These must be determined for two or more values of secondary shunting capacitance, within usual or desired range of filament to grid tube capacitance, and with several values of direct current through the primary winding. Values of direct current (simulating tube plate current) as high as the greatest which it is proposed to use, should be tried.

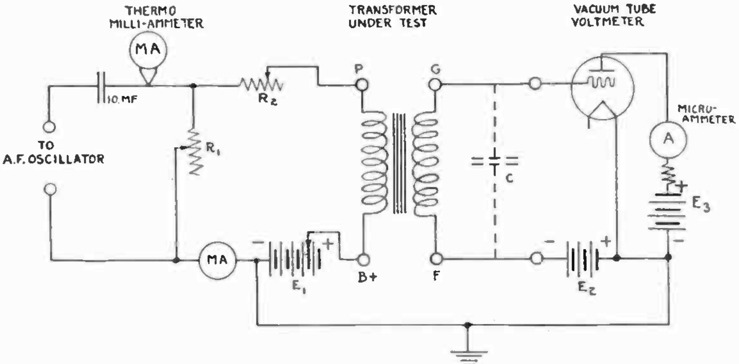

Figure 7—Test Circuit for Single Interstage Transformer.

2. The ratio of potentials across primary and secondary windings should be obtained for the entire frequency range. Here also it will be necessary to take measurements with various values of secondary capacitance, and primary direct current, and curves for the complete frequency range should be plotted.

Then letting e_1 = Potential generated in plate circuit of the tube connected to primary winding,

e_2 = Potential across the secondary, as result of e_1 in tube,

r = Plate resistance of the tube,

R = Effective resistance of Transformer Primary,

X = Reactance of Transformer Primary,

E_2/E_1 = Ratio of potentials measured directly across Secondary and Primary respectively.

The following formula gives the relation between these factors:

$$e_2 = \left(\frac{E_2}{E_1}\right) \frac{e_1}{\sqrt{(r+R)^2+X^2}} \sqrt{R^2+X^2}$$

Thus, with the aid of the curves, and this formula, it is possible to predict with fair accuracy the action of a transformer with any tubes, knowing the plate resistance and plate current of the preceding tube, and the grid-filament capacitance of the tube following.

Two suggested methods of making the measurement of the reactance and effective resistance of the transformer primary winding are given below:

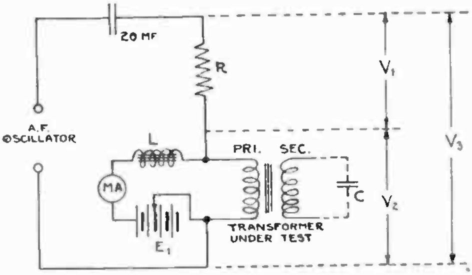

Figure 8—Voltmeter Method for Determining Reactance and Effective Resistance of Transformer Primary.

1. By means of the ordinary inductance bridge, arranged preferably with both the inductance and resistance values of its standards continuously variable.
2. It is often desired to perform this measurement with a definite value of a-c. potential applied to the primary. In this case, the well known three-voltmeter method can be used to advantage. Fig. 8 shows in outline the circuit, with means for applying a d-c. potential to the primary. The reactance of the choke *L* must be sufficiently high with respect to that of the transformer primary to make the effect of the shunting circuit negligible, or its inductance value must be known and its effect allowed for in the calculations.

The use of the tube voltmeter is recommended for obtaining the values of V_1, V_2, and V_3. The calculations involved in getting the values of primary reactance and effective resistance from these potentials is well known, and for the sake of brevity will not be repeated.

GENERAL DESCRIPTION OF AMPLIFICATION TEST METHOD

After the apparatus and device under test is set up, and necessary adjustments to bring about normal operating conditions are made as described in connection with Figs. 6 and 7, there remains the choice of two methods of taking the amplification versus frequency characteristic.

1. The oscillator may be set at various frequencies throughout the desired range, and a reading of the tube voltmeter taken for each frequency. From the data obtained, a curve of the amplifier characteristic may be plotted. This may be plotted either directly in output volts, or, by dividing the output potentials by the constant input potential, a curve showing the voltage amplification at each frequency can be plotted. The closer together the points have been taken, the more nearly will the curve show the operation at all frequencies.

2. A quicker method, and one in which a continuous measurement is made of all points along the frequency range is shown in general outline, with the necessary apparatus arrangement in Fig. 9. Here the shaft of the oscillator variable condenser is geared, by some convenient means, shown here as the gears *M* and *N*, to a drum which is of the proper size to hold a standard sheet of plotting paper. Paper having a logarithmic scale for frequency, and linear scale for amplification is used. The ratio of the gears *M* and *N* is such, that the frequency lines on the plotting paper which come under the pen *P* as the condenser shaft is rotated, may be made to correspond with the frequency produced by the oscillator at the moment. If then the distance of the pen P above the zero axis of the paper, is made to correspond to the output voltage reading of the amplifier, we

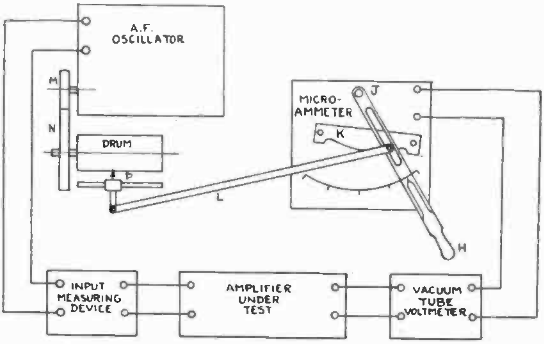

Figure 9—Audio Frequency Amplifier Test Set Up.

have a means for drawing the amplifier characteristic curve directly as the test is being made.

By means of the arm and pointer *H*, which is pivoted at *J*, it is possible to follow the movements of the tube voltmeter microammeter needle. The movements of the arm *H* are communicated to the pen *P* through the lever *L*, and by proper shaping of the guide template *K*, it is possible to have the voltages measured by the tube voltmeter correspond to the deflection of the pen *P* in terms of the ordinate divisions of the plotting paper. Since the templates for the 70-volt and 7-volt ranges differ somewhat in shape, they should be so constructed as to be removable and interchangeable within the curve drawing mechanism. Fig. 10 shows the drum, microammeter, pen, and attendant curve drawing mechanism. Fig. 11 shows this equipment attached to the beat frequency audio oscillator.

If desired, the oscillator condenser may be rotated slowly by motor or clockwork, and the observer needs only to follow the movements of the microammeter needle carefully with the pointer *H*. Upon reaching the upper end of the frequency range, the condenser drive is disconnected, and the paper removed from the drum. Upon the paper, the curve of output voltage vs. frequency for the amplifier under test will be found.

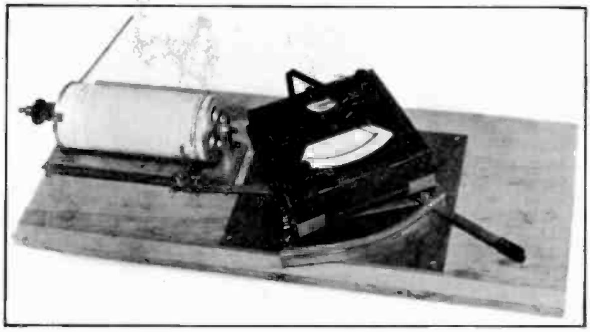

Figure 10—Curve Drawing Mechanism.

2. Test for Wave Form Distortion

Returning now to the second of the tests which are necessary in examining the operation of an audio frequency amplifier, i. e., test of its ability to handle the necessary output energy without distortion of the impressed wave form, we find that test of this point includes determination of the input potential which will cause overloading distortion—and thus the permissible input potential—and for accurate data, involves determination of the percentage of harmonics which are added to the original frequency as it passes through the amplifier.

The principal points in a transformer coupled amplifier where wave shape distortion may occur are, in the magnetic circuit of the transformers, due to excessive flux density, and in the tubes. If the transformers are well designed, the first point can be neglected. Distortion in the tubes will generally be caused by the non-linearity of the grid potential-plate current characteristic, or by the grids of one or more tubes becoming positive with respect to their filaments. On high input potentials, one or more of the grids will become positive, and the resulting grid current and lowered grid filament impedance of the tube, will cause very serious distortion. This type of distortion is by far the more serious of the two caused by the tubes, and is probably the only type which would be noticed by the average person listening to the loud speaker output from the amplifier.

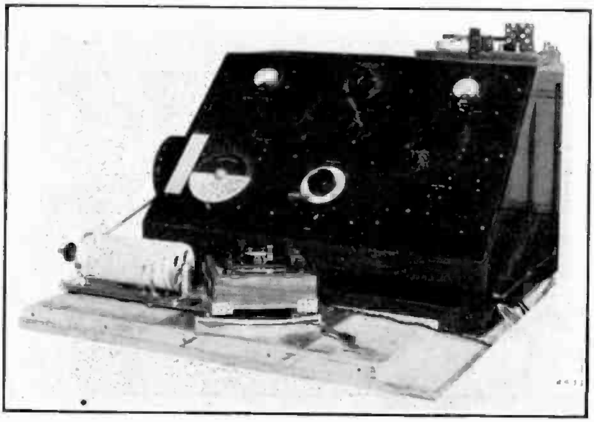

Figure 11—Curve Drawing Mechanism Attached to A. F. Oscillator.

It is often desirable to test an amplifier only for this more obvious type of distortion, and such a test is quite simple. The tube which is generally the first to show signs of grid current is the one in the last stage. By putting a low reading d-c. milliammeter or microammeter in series with the grid of this tube, and gradually increasing the input potential, the maximum permissible input will be just below that value which causes the meter to read in a direction showing that the grid is positive. (Most tubes have a small constant grid current in the opposite direction, which should be disregarded). It is advisable to make the test at several different audio frequencies, and if there is reason to believe that the grids of any of the other tubes in the amplifier are becoming positive, they should be tested in the same way.

For proper and safe operation, the amplifier should be capable of giving more output signal voltage than is normally required, before any of its tubes show signs of grid current at any point of the audio frequency range.

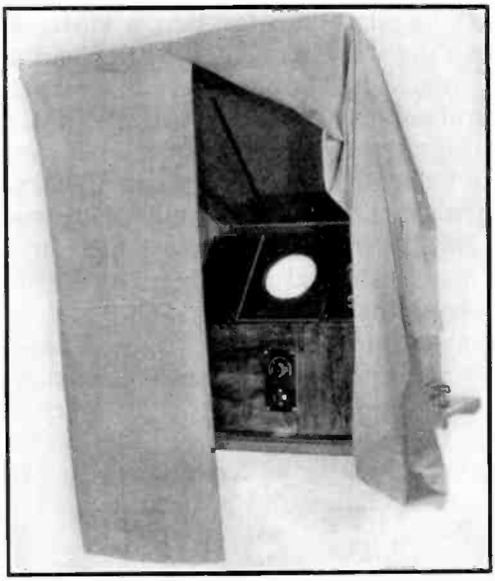

Figure 12—Cathode Ray Oscillograph Tube Mounting.

For those who prefer a method giving more accurate numerical indication of the amount of wave shape distortion present, analysis of the output wave by means of the cathode ray oscillograph is recommended. If potential from the source feeding the input of the amplifier under test, is impressed across the "horizontal motion" plates of the oscillograph, and potential from the amplifier output is impressed across the other set of plates, a figure will be traced on the oscillograph screen, which will show, by its shape, with what fidelity the amplifier reproduces the wave shape impressed upon its input. Fig. 12 shows the mounting used by the author for the cathode-ray oscillograph tube. To facilitate tracing the curves which appear on the screen, a hood was built over the tube mounting cabinet. One side of the hood cloth is shown drawn aside to permit a view of the tube.

If a single straight line is traced, perfect fidelity and no phase shift is indicated. In cases where a phase shift occurs in the amplifier, the figure will assume an elliptical instead of a linear shape. This makes analysis of the figure more difficult, but the elliptical shape is no sign of wave distortion. It merely indicates that a shift in phase has taken place as the wave passed through the amplifier. Oscillograph figures should be obtained for several frequencies, and for several values of input potential. Then by an application of harmonic wave analysis to the figures obtained, it is possible to determine the percentage of harmonics present, for various input potentials and frequencies. Depending on the fidelity of output wave which is desired, the permissible input potential may then be determined.

It is believed that an oscillographic test of this kind should be applied more frequently than it has been in the past, to the testing of audio frequency amplifiers. As the scope of this paper prevents further or more detailed discussion, those interested are referred to a recent article on "Load Carrying Capacity of Amplifiers" by Messrs. F. C. Willis and L. E. Melhuish, in the October 1926 issue of *The Bell System Technical Journal.*

3. General Operation

The general points for test and examination mentioned under the third section are usually different for each amplifier, and no very definite rules for investigation procedure can be given. The method of investigation depends very largely on the particular amplifier, and no general rules can be laid down. The experience and ingenuity of the test engineer must be relied upon generally to determine the method of test. Little if any definite apparatus can be recommended since these tests require, in many cases, merely intelligent and careful inspection. One important point may be mentioned in this connection, however, i. e., in regard to checking whether the proper plate potential is being used on the tubes in a resistance coupled amplifier. The plate potential may be measured with an electrostatic voltmeter at the plate of each tube while the amplifier is in operation, or else its value may be calculated from measurement of the plate current and plate coupling resistance value for each tube.

No attempt has been made in this paper to give anything but the essentials of audio amplifier testing. It will usually be found in practice that each amplifier constitutes to a greater or lesser extent a special problem, since each will differ from the others in certain points. The essential features to be tested for, together with a brief description of the apparatus found most suitable for making the tests, have been given. There are many other methods of taking amplifier characteristics, but it is believed that the methods described here are as good as most of the other methods, and perhaps excel some in points of accuracy, convenience, and speed of measurement.

In conclusion, the author wishes to acknowledge his indebtedness to the following members of the staff of this laboratory:—Dr. Walter Van B. Roberts for valuable suggestions regarding the method of determining the general characteristics of individual amplifying transformers by combined measurement and calculation; Dr. Irving Wolf for the ideas embodied in the method of directly drawing amplifier characteristic curves described in this paper; and Mr. David Grelich for his assistance in connection with the development of the tube voltmeter described.

SUMMARY: The more necessary of the various points which must be tested in examining the performance of audio frequency amplifiers are outlined, and a method of test procedure, found by the author to have desirable characteristics from the points of view of accuracy, speed, and simplicity of operation, is described.

The method used permits the complete curve of amplification vs. frequency for the amplifier under test, to be drawn directly by the test equipment in a very short time. A type of tube voltmeter found convenient for measurement of amplifier output potential is described. Brief discussion is given of the testing of individual audio frequency amplifying transformers, and some test methods applicable thereto are suggested. Amplifier wave form distortion, and overloading are discussed, and methods for the test thereof are recommended.

VARIATIONS IN HIGH-FREQUENCY GROUND WAVE RANGES

By
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(Naval Research Laboratory)

The following information obtained from experimental work made by two naval ships on high frequency ground wave ranges is of interest because it indicates unexpected differences between day and night values, which are not at present predictable from theoretical consideration.

The ground wave of a high-frequency station is usually recognized readily enough if observations are taken well within the normal skip distance area because the ground wave is normally quite steady; that is, free from fading. The observations herein reported were taken on frequencies in the 12,000 and 16,000-kc. bands at distances between zero and 53 miles. The power of the transmitter was approximately 100 watts in the radio frequency circuits. Probably 60 per cent of this power went into the antenna. In agreement with earlier observation it was found that ground waves traveled further over the sea than over the land, so that with daylight conditions on a transmitter of considerable power it was not easy to locate sharply the skip distance which made itself felt rather as a depression in the curve of audibility of the function of the distance rather than an absolute absence of signal, unless frequencies higher than 10,000 kc. were used.

The daylight ground wave ranges of these transmitters working in the 12,000-kc. band were approximately 12 miles, whereas the same transmitters had a satisfactory night range in the same frequency band of something somewhat in excess of 50 miles. The extreme limit of audibility was estimated roughly to be about 75 miles. Actual observations were taken at 53 miles. In the 16,000-kc. band the daylight ground wave ranges were not in excess of 3 miles

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whereas the night ranges extended out to 22 miles in the same frequency band. These results can only be interpreted by supposing that there was some agency at work which produces a markedly greater absorption of the ground wave in the day-time than it does during the dark hours. If this were not true we would be forced to assume that we had to deal with some new kind of a sky wave produced by a very low refracting or reflecting layer. The absence of fading, however, makes the second possibility less likely. The writer is inclined to the opinion that the absorption of these very high frequencies must be greater in the day-time than during the dark hours even for those portions of the wave front which travel close to the earth and which are commonly spoken of as ground waves. No information is available from these particular experiments as to the state of polarization of these ground waves.

It might be that this information is only in line with what we know of somewhat longer wavelengths. A distance of 22 miles for a frequency of 16,000 kc. might be considered as comparable to the distance of 342 miles for a frequency of 1,000 kc. and of course it is well known that 1,000 kc. would suffer enormous variations in intensity at a distance of 342 miles, if one compared the night and day signals; but in the past it has been pretty generally assumed that at 1,000 kc. at a distance of 342 miles, the differences are due to the different character of the sky waves in the two cases.

The Austin-Cohen transmission formula is not to be applied for night reception of 1,000 kc. over the distances mentioned on account of the wave paths having been so utterly different in the two conditions of night and day. In other words, the sky wave, owing to the greater height of the refracting layer, is much more effective at night but in the case of the high frequency ground waves, which are the subject of this note, it is difficult to see how they can have had anything to do with sky waves at all unless they are waves of a variety hitherto unknown or at least unrecognized.

A SUGGESTION OF A CONNECTION BETWEEN RADIO FADING AND SMALL FLUCTUATIONS IN THE EARTH'S MAGNETIC FIELD*

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The fading of radio signals has been referred to time and again as an interference effect. The suggestion that the upper atmosphere may be an important factor in enabling electromagnetic waves to propagate over appreciable distances has been the origin and support of this view. The wave traveling along the ground and the wave arriving by reflection have been supposed to interfere. Within the last few years, however, it has become obvious that more than simple interference is involved. Thus short-wave transmission, i. e., transmission on wavelengths below 100 meters in length, showed in the hands of Taylor and Hulburt[1] that beyond a certain range signals are received entirely by an overhead route. In spite of this fact, the signals show fading. Part of this is explained as a general shift in the height of the reflecting layer. However, this can be by no means all because much of the fading is rapid and does not correspond to the disappearance of the signal in the vicinity of the receiver such as occurs at the skip distance. In fact, the experiments by Mr. Tuve and the writer[2] on reflection of waves from the upper atmosphere showed that fading exists for the reflected waves quite independently of interference with the ground wave. Two suggestions have been made at the time as to a possible cause of the fading. One of them supposed the down-coming wave to be a result of the interference of two or more waves. During a discussion of the subject held at the Department of Terrestrial Magnetism, Dr. A. H. Taylor suggested that the Earth's magnetic field might be the cause of some of the

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[1] For references see bibliography.

fading. On examining the matter, the writer found that since the upper atmosphere must show magnetic double refraction and since, therefore, there must be in general at least two possible rays between the transmitter and receiver, a change in the interference conditions of these two rays must take place whenever the Earth's magnetic field changes. The change in the relative phase of the two rays turns out to be large even if the Earth's field changes slightly. Several gammas (0.00001 Gauss) may turn the plane of polarization of the ray through 30 deg. As a result, the vertical component of the electric intensity may be changed from 0 to its maximum value in a field of 15 gammas. The interference effects here discussed are due to the passage of the ray through an appreciable thickness of ionized air. If the rays are reflected, the interference effects can not be nearly as pronounced. A study of the magnitude of fading in connection with small fluctuations in the Earth's field may lead, therefore, to an understanding of the manner in which the ionization increases with the height.

Some simple calculations are given below in order to show how the interference conditions may change. At the same time the effect of a change of frequency and of the ionization are estimated. The calculations do not pretend to be exact. They are intended to show the order of magnitude of the effects.

Before performing the calculations we shall restate the assumptions about the nature of propagation of radio waves which will be used. We suppose the upper atmosphere to be partly ionized into electrons as well as heavier ions. We neglect the effect of heavy ions and consider the effect of electrons alone. The density of free electrons is supposed to increase with the height up to a certain level. The magnetic field of the Earth affects the motion of electrons in the manner discussed by Appleton[3] as well as Nichols and Schelleng[4]. We calculate below the effects on the intensity of received signals due to changes in the intensity of the Earth's magnetic field and due to small changes in the frequency of the emitted signals. We suppose that while either change takes place, the degree of ionization is constant. On account of certain mathematical difficulties, an exact solution of the problem does not seem to be worth while at this time. However, for some special limiting cases, exact solutions may be obtained and these can be applied probably with a reasonably small error to calculate the order of magnitude in general.

Letting

$$a = \frac{Ne^2}{\pi m}, \quad \nu_0 = \frac{eH}{2\pi me} \tag{1}$$

where

$N =$ number of electrons per cm^3
$e =$ charge of the electron in electrostatic units
$m =$ mass of the electron
$H =$ intensity of the Earth's field

The two velocities of propagation of a beam traveling along H are given by

$$V(v) = c\left(1 + \frac{a}{\nu(v_0 - v)}\right)^{-\frac{1}{2}} \tag{2}$$

$$V(v) = c\left(1 - \frac{a}{\nu(v_0 + v)}\right)^{-\frac{1}{2}} \tag{3}$$

The first of these corresponds to circular polarization in the right-handed sense, while the second corresponds to circular polarization in the left-handed sense if the sense of propagation of the wave is the same as the sense of direction of H. We consider first an idealized condition in which the rays are approximately following the Earth's magnetic field. We may then use formulas (2) and (3) to calculate the refraction of the rays. Let first the ionization be proportional to the square of the height so that

$$a = by^2$$

The two refractive indices corresponding to (2) and (3) are then $\left(1 + \frac{by^2}{\nu(v_0 - v)}\right)^{-\frac{1}{2}}$ and $\left(1 - \frac{by^2}{\nu(v_0 + v)}\right)^{-\frac{1}{2}}$. Both are of the form $\sqrt{1 - a^2 y^2}$. The differential equation of the path is

$$\frac{dy}{\pm\sqrt{\cos^2\theta_0 - a^2 y^2}} = \frac{dx}{\sin\theta_0} \tag{4}$$

$$a^2 = \frac{b}{\nu(v_0 + v)} \quad \text{or} \quad \frac{b}{\nu(v - v_0)}$$

leading to a solution of the type

$$y = \frac{\cos\theta_0}{a}\sin\frac{ax}{\sin\theta_0} \tag{5}$$

The phase of the wave at any point is $\varphi = \int \frac{2\pi v}{V}\,ds$, the integral being taken over the path followed by the ray. If φ changes by π, the interference of the ray with another one of constant phase is changed from constructive to destructive. We are interested, therefore, in the phase difference $\Delta\Phi$ for the two rays if we are to determine the polarization of the wave at the receiving station. A change in the Earth's field δH produces a change $\delta\Delta\varphi$ in the phase difference $\Delta\varphi$. If this change $\delta\Delta\varphi = \frac{\pi}{2}$, the direction of polarization of the resultant linear vibration is turned through $90°$. We must obtain φ for either ray on the supposition that the position of the receiving and the transmitting stations is kept fixed. If the distance between them be l, formula (5) determines $\theta_0 = \frac{al}{\pi}$. Now, in general, if μ is the refractive index

$$\varphi = \frac{2\pi v}{c}\int \mu ds = \frac{2\pi v}{c}\int \sin e_0 \left(1 + \left(\frac{dy}{dx}\right)^2 \right) dx \tag{6}$$

provided the refractive index varies only with the height and provided it is unity at the surface of the ground. Using (5)

$$\varphi = \frac{2\pi v}{c}\, l\, \frac{1+\sin^2\theta_0}{2\sin\theta_0} = 2\pi \frac{1+\frac{a^2l^2}{\pi^2}}{2\left(\frac{al}{\pi}\right)}\,\frac{l}{\lambda} \tag{7}$$

$$\left(\lambda = \frac{c}{v} = \text{wavelength}\right)$$

Keeping the range l and the frequency v fixed, we suppose the Earth's field H to have undergone a variation δH. This changes a^2 on account of a change in v_0 (see (4)). The change in φ due to δa is

$$\delta\varphi = \pi\frac{l}{\lambda}\,\frac{\frac{a^2l^2}{\pi^2}-1}{\frac{al}{\pi}}\,\frac{\delta a}{a} = \pi\frac{l}{\lambda}\,\frac{A^2l^2-1}{Al}\,\frac{\delta A}{A} \tag{8}$$

where $\pi A = a$, and

$$\delta\,\Delta\varphi = \delta\,(\,\varphi_2 - \varphi_1\,) = \pi\,\frac{l}{\lambda}\left(\frac{A_2 l^2 - 1}{A_2 l}\,\frac{\delta A_2}{A_2} - \frac{A_1 l^2 - 1}{A_1 l}\,\frac{\delta A_1}{A_1}\right)$$

It follows from (4) that $2\,\frac{\delta a_1}{a_1} = -\,\frac{v_0}{v + v_0}\,\frac{\delta v_0}{v_0}$, $2\,\frac{\delta a_2}{a_2} = \frac{v_0}{v - v_0}\,\frac{\delta_0}{v_0}$. Remembering also that $H^{-}\,\delta H = v_0^{-1}$ we have

$$\delta\,\Delta\,\varphi = \frac{\pi\,l}{2\,\lambda}\left(\frac{A_2^{\,2}\,l^2 - 1}{A_2 l}\,\frac{v_0}{v - v_0} + \frac{A_1^{\,2}\,l^2 - 1}{A_1 l}\,\frac{v_0}{v + v_0}\right)\frac{\delta\,H}{H} \tag{9}$$

Let $l = 0$, then

$$\delta\,\Delta\varphi = -\,\frac{\pi}{2}\,\frac{1}{\lambda}\left(\frac{1}{A_2}\,\frac{v_0}{v - v_0} + \frac{1}{A_1}\,\frac{v_0}{v + v_0}\right)\frac{\delta H}{H}$$

The quantity A_2 may be obtained in this case from the height h to which the wave goes. It is easily shown to be for this case $A_2 = \frac{1}{\pi h}$ so that for zenith reflection

$$\delta\,(\,\Delta\varphi\,) = -\,\frac{\pi}{2}\cdot\frac{\pi h}{2}\,\frac{v_0}{v - v_0}\left(1 + \left(\frac{v - v_0}{v + v_0}\right)^{1/2}\right)\frac{\delta H}{H} \tag{10}$$

If $h = 100$ kilometers and $\lambda = 70$ meters, then $\frac{v_0}{v - v_0} \cong 1/2$ and $\delta\,(\,\Delta\varphi\,) \cong -\,3.8 \times 10^3\,\frac{\pi}{2}\,\frac{\delta H}{H}$. Hence a change in H by $\frac{1}{3800}$ part of its whole amount would turn the plane of polarization of the downcoming beam through 90 deg. This would cause very severe fading even at a short distance from the transmitter for a change of $13\gamma\;(= 13 \times 10^{-5}$ *Gauss* $)$ in the Earth's field. A field one-third of that amount would cause appreciable fading. Variations of that order of magnitude in the Earth's field are very frequent and may, therefore, cause an appreciable amount of fading.

Although formula (10) applies strictly to the case of zenith reflection, it is a good approximation to (9) for considerable ranges because $A_2 l$ becomes equal to unity [for the number chosen above] if $l = 300$ kilometers. For 150 kilometers (10) is still a reasonable approximation. This suggests a connection with the distance of maximum fading for broadcast wavelengths. This may be determined in addition to the proportion of ground and reflected waves by the fact that for a certain l, $\delta\Delta\varphi$ in accordance with (9) must vanish. Since in formula (10) the phase velocities (2), (3) have been used and since these velocities cannot be applied well for short-range transmission, we should expect the fading to be greatest some place between $l=0$ and $l=300$ kilometers.

Formula (7) gives also the change in the condition for interference if the magnetic field of the Earth stays constant and the frequency of the signal is changed. Using a^2 as given in (4) we get $2\,\frac{\delta a_2}{a_2}=-\frac{2v-v_0}{v-v_0}\,\frac{\delta v}{v}$, $2\,\frac{\delta a_1}{a_1}=-\frac{2v+v_0}{v+v_0}\cdot\frac{\delta v}{v}$ and hence

$$\left.\begin{aligned}
\delta\varphi_2 &= \frac{\pi l}{\lambda}\left(\frac{2}{A_2 l}-\frac{A_2^2 l^2-1}{2A_2 l}\,\frac{v_0}{v-v_0}\right)\frac{\delta v}{v}\\
\delta\varphi_1 &= \frac{\pi l}{\lambda}\left(\frac{2}{A_1 l}+\frac{A_1^2 l^2-1}{2A_1 l}\,\frac{v_0}{v+v_0}\right)\frac{\delta v}{v}\\
\delta(\varphi_1-\varphi_1) &= \frac{2\pi l}{\lambda}\left(\frac{1}{A_2 l}-\frac{1}{A_1 l}+\frac{1-A_2^2 l^2}{4A_2 l}\,\frac{v_0}{v-v_0}+\right.\\
&\left.\frac{1-A_1^2 l^2}{4A_1 l}\,\frac{v_0}{v+v_0}\right)\frac{\delta v}{v}
\end{aligned}\right\}\quad(11)$$

For short range transmission this becomes

$$\delta(\varphi_2-\varphi_1)=\frac{\pi}{2}\,\frac{\pi h}{\lambda}\left[-4\left(\sqrt{\frac{v+v_0}{v-v_0}}-1\right)+\frac{v_0}{v-v_0}\left(1+\sqrt{\frac{v-v_0}{v+v_0}}\right)\right]\frac{\delta v}{v}\qquad(12)$$

With $\lambda=70$ meters and $h=100$ kilometers $\delta(\varphi_2-\varphi_1)=\frac{\pi}{2}\times 1400\,\frac{\delta v}{v}$. Hence if $\frac{\delta v}{v}=\frac{1}{1400}$ the direction of polarization is rotated through 90°. This corresponds to a frequency change of 3000 cycles per second. Hence even a frequency change of 1000 should produce an appreciable effect on the intensity of the received signal.

Finally, we estimate for the above case the change in $\varphi_2-\varphi_1$ which is produced by a change of ionization. We find in a way quite similar to that just used that

$$\delta(\varphi_2-\varphi_1)=\frac{\pi}{2}\,\frac{l}{\lambda}\left(\frac{A_2^2 l^2-1}{A_2 l}-\frac{A_1^2 l^2-1}{A_1 l}\right)\frac{\delta N}{N}\qquad(13)$$

where δN is the change in the number of electrons per unit volume. For short-range transmission this becomes

$$\delta\,(\varphi_2 - \varphi_1) = -\frac{\pi}{2}\cdot\frac{\pi h}{\lambda}\left(-1 + \sqrt{\frac{v + v_0}{v - v_0}}\right)\frac{\delta N}{N} \qquad (14)$$

With the numbers used above, a rotation through 90° is produced if $\frac{\delta N}{N} = \frac{1}{1800}$. Since changes in the ionization are presumably responsible for changes in the Earth's field, it is reasonable to suppose that (9) and (14) should be combined in any actual case in order to give the resultant change.

In the above we have constantly referred to the short wavelength region $(v > v_0)$. If the wavelength is long, i.e., if $v < v_0$, only the ray defined by (3) returns to the Earth. If this ray interferes with the ground, wave effects quite similar to those just discussed are produced. Formulas (9), (11), (13) must be modified, however, by omitting the terms in A_2. The order of magnitude of the effects is still the same.

Even though $\delta\,(\varphi_2 - \varphi_1)$ is the same for a considerable range according to (9), (11), (13), the fading produced may vary considerably over that range. Thus the value of $\varphi_2 - \varphi_1$ is very important because it determines whether the electric vector of the downcoming wave is horizontal or not. The same fluctuation of H may produce, therefore, a reenforcement of the signal in one place and a diminution in another.

Small frequency changes are, of course, known to produce an effect on the intensity of the received signal. It is also known that interrupted continuous waves show less fading on short wavelengths than continuous waves[5]. It is somewhat difficult to point out definitely what has been happening in the experiments just mentioned. For short wavelengths, however, it seems certain that the fading must be entirely due to the condition of the wave returned by the upper atmosphere. The high "resolving power" of the interference apparatus in that case demonstrates a behavior similar to the one discussed above. It cannot be interpreted on the basis of interference with the ground wave. It must be also remembered that experiments by Mr. Tuve and the writer (loc. cit.) showed that the reflected wave shows fading which is sufficiently great to explain an appreciable amount of the general fading.

The influence of magnetic storms on transmission has given opportunity for considerable discussion. However, it would be dangerous to maintain that the simple calculations given above explain the effects observed. Presumably magnetic storms are accompanied by large changes in the ionization of the reflecting layer. A complete explanation must, therefore, be based on formulas corresponding to both (9) and (13).

SUMMARY: The state of polarization of a radio wave returned by the reflecting layer is studied as a function of (a) small changes in the intensity of the Earth's magnetic field, (b) small changes in the frequency of the wave, and (c) small changes in the ionization of the atmosphere.

It is shown that for the special electron distribution considered appreciable effects on the intensity of the signal are to be expected for fluctuations in the Earth's field of the order of 5 gammas, for changes in the frequency of the order of 1,000 cycles, and for changes in the ionization of the order of one part in 5,000. [The number applies to $\lambda = 70$ meters.]

The dependence of fading on range indicates that there is a certain range of maximum fading which is of the general order of 100 miles.

APPENDIX

It seems that a certain amount of justification is needed for using Snell's law and for discussing two rays only. This is obtained by the following considerations.

1. *Equations of propagation in a homogeneous magnetically active medium.* The equations of propagation in a magnetically active medium are generally well known. However, it seems that in their applications to radio some misconceptions have arisen. For this reason it will be well to summarize the important facts. It is supposed that the wave propagating through the layer is monochromatic and has a frequency $v = \frac{\omega}{2\pi}$. The medium through which it propagates is supposed to contain N electrons per cm^3. Each electron has a charge e and a mass m. The forces acting on

each electron are as follows: (a) The electric intensity $(E_x, E_y, E_z)\, e^{iwt}$ of the wave, (b) the force $\frac{4\pi}{3}\,(P_x, P_y, P_z)$ where P is the polarization of the medium, (c) the force due to the Earth's magnetic intensity (O, O, H_o) supposed here to be constant and directed along the axis OZ, the magnitude of this force being $\frac{e}{c}\,(y, -x, O)\, H_o$, and (d) the forces which are brought about by collisions.

Denoting the displacement of an electron from its position of equilibrium by (x, y, z) we have

$$(P_x, P_y, P_z) = Ne\,(x, y, z)$$

Thus the equations of motion are neglecting (d).

$$\begin{aligned} x &= \frac{e}{m} E_x e^{iwt} + \frac{e}{mc}\, y H_o + \frac{4\pi}{3}\frac{Ne^2}{m}\, x \\ y &= \frac{e}{m} E_y e^{iwt} - \frac{e}{mc}\, x H_o + \frac{4\pi}{3}\frac{Ne^2}{m}\, y \qquad (1) \\ z &= \frac{e}{m} E_z e^{iwt} + \frac{4\pi}{3}\frac{Ne^2}{m}\, z \end{aligned}$$

For most purposes we can neglect the terms in $\frac{4\pi}{3}\frac{Ne^2}{m}$ with the result that Maxwell's equations become

$$\begin{aligned} \frac{\partial H_z}{\partial y} - \frac{\partial H_y}{\partial z} &= \frac{1}{c}\left[(1+\alpha)\frac{\partial E_x}{\partial t} - i\beta\frac{\partial E_y}{\partial t}\right] \\ \frac{\partial H_x}{\partial z} - \frac{\partial H_z}{\partial x} &= \frac{1}{c}\left[i\beta\frac{\partial E_x}{\partial t} + (1+a)\frac{\partial E_y}{\partial t}\right] \qquad (2) \\ \frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y} &= \frac{1+\gamma}{c}\frac{\partial E_z}{\partial t} \end{aligned}$$

where $\alpha = \frac{4\pi Ne^2}{m(\omega'^2 - \omega^2)}$, $\beta = \frac{4\pi Ne^2}{m}\;\frac{\omega'}{\omega(\omega'^2 - \omega^2)}$,

$\gamma = -\frac{4\pi Ne^2}{m\omega^2}$, and $-\frac{1}{c}\frac{\partial H}{\partial t} = curl\ E.$

If a plane wave is possible, all the quantities involve $exp\ \frac{2\pi i}{\lambda}\,(lx + my + nz - Vt)$. Substituting in Maxwell's equations we have:

$$\frac{V}{c} H_x = mE_z - nE_y;\ mH_z - nH_y = -\frac{V}{c}\left[(1+\alpha)E_x - i\beta E_y\right]$$

$$\frac{V}{c} H_y = nE_x - lE_z;\; nH_x - lH_z = -\frac{V}{c}\,[i\beta E_x + (1+\alpha)E_y] \tag{3}$$

$$\frac{V}{c} H_z = lE_y - mE_x;\; lH_y - mH_x = -\frac{V}{c}\,(1+\gamma)\,E_z$$

whence

$$\frac{E_x}{n\,(lA - imB)} = \frac{E_y}{n\,(mA + ilB)} = \frac{E_z}{A^2 - B^2 - (l^2 + m^2)\,A}$$

$$= \frac{\frac{V}{c} H_x}{m\,(A^2 - B^2 - A) - in^2 lB} = \frac{\frac{V}{c} H_y}{l\,(B^2 + A - A^2) - in^2 mB}$$

$$= \frac{\frac{V}{c} H_z}{in\,(l^2 + m^2)\,B} \tag{4}$$

where

$$A = 1 - \frac{V^2}{c^2}\,(1+\alpha)\,;\; B = \beta\,\frac{V^2}{c^2};\; C = 1 - \frac{V^2}{c^2}\,(1+\gamma)$$

These equations determine the polarization of the wave if l, m, n, V are known. In order to find V, eliminate E_x, E_y E_z, H_x, H_y, H_z in (3). Then we find

$$(A^2 - B^2)\,(C - n^2) + (n^2 - 1)\,AC = 0 \tag{5}$$

or letting

$$u = \frac{V^2}{c^2} \tag{6}$$

we have

$$u\,(pu^2 + qu + r) = 0 \tag{7}$$

where

$p = (1+\gamma)\,[\,(1+\alpha)^2 - \beta^2 = \quad (1+\gamma)\,[\,1 + \alpha + \alpha\,(1+\gamma)\,]$

$q = -2\,(1+\alpha)\,(1+\gamma) + (n^2 - 1)\,[\,(1+\alpha)\,(\alpha - \gamma) - \beta^2\,] = -2\,(1+\alpha)\,(1+\gamma) \quad + (n^2 - 1)\,(\alpha - \gamma)$

$r = 1 + \gamma + (n^2 - 1)\,(\gamma - \alpha) = 1 + \gamma + (n^2 - 1)(\gamma - \alpha)$

For every n there are two roots of (7) which are not identically zero. These roots correspond to two possible velocities of propagation in a direction making a given angle $\cos^{-1} n$ with the field H. The roots are real if $q^2 > 4\,pr$ and complex if $q^2 < 4\,pr$. Thus either both waves are possible or none.

2. *Propagation of waves in a non-homogeneous medium.* We shall suppose next that the concentration of electrons

varies from point to point in the medium in a known manner. Thus N is a known function of x, y, z. We shall further suppose that within a wavelength $N(x, y, z)$ is practically constant. Under these conditions we can discuss the propagation of waves in the medium by means of geometrical optics to a certain extent.

Let us suppose first for simplicity that the magnetic field is absent so that the medium is isotropic but non-homogeneous. This case is well known and forms the basis of Hamilton's theory of dynamics. The fundamental equation to solve is the wave equation

$$\frac{\delta^2 u}{\delta x^2} + \frac{\delta^2 u}{\delta y^2} + \frac{\delta^2 u}{\delta z^2} - \frac{1}{[V(x, y, z)]^2} \frac{\delta^2 u}{\delta t^2} = 0 \qquad (8)$$

subject to the condition that u is a periodic function of the time having a frequency $\nu = \frac{\omega}{2\pi}$. This reduces (8) to

$$\Delta u + \frac{\omega^2}{V^2} u = 0 \qquad (8^1)$$

The statement that geometrical optics give an approximate solution of the problem is equivalent to saying that

$$u = C \, exp \, 2\pi i v \left(t - \int \frac{ds}{V} \right) \qquad (9)$$

is a sufficiently exact solution of (8^1) where $\int \frac{ds}{V}$ is the path difference expressed in seconds, the integral being taken over a possible ray, i. e., subject to the condition

$$\delta \int \frac{ds}{V} = 0 \qquad (10)$$

by Fermat's Principle. Substituting (9) into (8) and letting

$$S = \int \frac{dx}{V} \qquad (11)$$

we have

$$\frac{i}{2\pi v} \Delta S + \left[\left(\frac{\delta S}{\delta x} \right)^2 + \left(\frac{\delta S}{\delta y} \right)^2 + \left(\frac{\delta S}{\delta z} \right)^2 \right] - \frac{1}{V^2} = 0 \qquad (12)$$

Fermat's Principle and the construction of Huygens being equivalent, it is found without difficulty that

$$\left(\frac{\delta S}{\delta x} \right)^2 + \left(\frac{\delta S}{\delta y} \right)^2 + \left(\frac{\delta S}{\delta z} \right)^2 - \frac{1}{V^2} = 0 \qquad (13)$$

so that (12) is satisfied provided ΔS may be neglected in comparison with $\frac{2\pi v}{V^2}$ or provided the change in $\frac{1}{V}$ per unit length when multiplied by $\frac{V^2}{2\pi V}$ is negligible compared to unity [because $\frac{\partial^2 S}{\partial x^2}$ is of the order of $\frac{\partial}{\partial x}\left(\frac{1}{V}\right)$]. This means that $\left|\frac{1}{2\pi v}\frac{\partial V}{\partial x}\right| << 1$ or $\left|\frac{\lambda}{2\pi}\frac{\partial \log V}{\partial x}\right| << 1$. Thus the relative change in V in $\frac{1}{2\pi}$ fraction of a wavelength must be negligible in comparison with unity.

The above reasoning is well known and has been made use of recently by Louis de Broglie[6] in a quite different connection. So far it applies only to non-polarized waves. Our first problem is to see what happens in the case of polarized electromagnetic waves. The equations of propagation are

$$curl\ H = \frac{\epsilon(x, y, z)}{c}\frac{\partial E}{\partial t} \qquad (14)$$

$$curl\ E = -\frac{1}{c}\frac{\partial H}{\partial t} \qquad (15)$$

where ϵ is the effective dielectric constant of the medium, viz, $1 + 4\pi\frac{P}{E}$. This ϵ is supposed to vary from point to point. However, the variation in ϵ is supposed to be slow so that $\frac{\partial \log \epsilon}{\partial x} << \frac{\partial Ex}{\partial x}$. Since the components of E and H satisfy the wave equation with $V^2 = \frac{c^2}{\epsilon}$ we could try to use (9) for the components directly. However, we must satisfy not only (14) and (15) but also the relations

$$div\ H = C \qquad (16)$$

$$div\ D = div\,(E + 4\pi P) = div\,(\epsilon E) = 0 \qquad (17)$$

as well as (14) and (15). Solutions given by (9) will not necessarily satisfy either of these equations. It is known, however, that a scalar φ and a vector a each of which satisfies (9) can be used to express E and H as

$$\vec{E} = -\frac{1}{c}\frac{\partial \vec{a}}{\partial t} - \Delta\varphi \qquad (18)$$

$$\vec{H} = curl\ \vec{a} \qquad (19)$$

provided

$$\Delta\varphi + \frac{1}{c}div\frac{\partial \vec{a}}{\partial t} = 0 \qquad (20)$$

and provided ϵ satisfies the condition of varying slowly [as defined above]. The proof of this is quite similar to that in the well-known case of constant ϵ and need not be given. An equation equivalent to (20) is

$$\frac{c}{V^2}\frac{\delta\varphi}{\delta t} + div\,\vec{a} = 0 \tag{20'}$$

which is well known for $V = c$.

The case of a plane polarized wave in a homogeneous medium can be obtained by letting

$$a_x = -\frac{cE_0}{\omega} sin\,\omega\left(t - \frac{z}{c}\right),\ a_y = a_z = 0;\quad \varphi = 0 \tag{21}$$

This wave travels along OZ, and has its electric vector along OX, while the magnetic vector is along OZ.

A plane wave making an angle θ with 03 and polarized with its electric vector along OX is similarly represented by

$$a_x = -\frac{cE_0}{\omega} sin\,\omega\left(t - \frac{z\cos\theta + y\sin\theta}{c}\right),\ a_y = a_z = 0;\ \varphi = 0 \tag{22}$$

The wave normal for this wave is the line $\frac{z}{cos\,\theta} = \frac{y}{sin\,\theta}$. The solution (22) applies to a homogeneous medium only. If the medium is not homogeneous, we apply solution (9) to a_x so as to satisfy (8). Equations (14), (15), (16), (17) are then also satisfied if (18), (19) are used. Thus this wave travels without change of polarization. The amplitude of the electric intensity is found to remain constant along the path, while the magnetic intensity at any point is $\frac{c}{V}$ of the original. This corresponds also to the fact that the ratio of the electric and magnetic intensities in the wave is $\frac{V}{c}$.

Suppose next that the wave normal is in the xz plane. For a homogeneous medium

$$\begin{aligned} a_x &= -\frac{c\,E_0\cos\theta}{\omega} sin\,\omega\left(t - \frac{z\cos\theta + x\sin\theta}{c}\right) \\ a_y &= 0 \\ a_z &= +\frac{c\,E_0\sin\theta}{\omega} sin\,\omega\left(t - \frac{z\cos\theta + x\sin\theta}{c}\right) \end{aligned} \tag{23}$$

In order to find the solution for the non-homogeneous medium [the non-homogeneity being confined to the direction z]

we suppose s to be determined by (10) and the quantity $-\frac{c\,E_o}{\omega}\,sin\,\omega\left(t-\frac{z\,cos\,\theta+x\,sin\,\theta}{c}\right)$ to be the value of u for $V=c$. We then try

$$a_x = u\,cos\,(\,sz\,),\ a_y = 0,\ a_z = -u\,sin\,(\,sz\,) \qquad (24)$$

$$u = -\frac{c\,E_o}{\omega}\,sin\,\omega\left(t-\int\frac{ds}{V}\right)$$

Introducing (24) into (8) we are concerned with expressions of the type $\frac{\partial^2 u}{\partial x^2}\,cos\,(\,sz\,)+2\,\frac{\partial u}{\partial x}\,\frac{\partial}{\partial x}\,(\,cos\,(\,sz\,)\,)+u\,\frac{\partial^2}{\partial x^2}\,cos\,(\,sz\,)$. The particular non-homogeneity assumed makes $\frac{\partial}{\partial x}\,cos\,(sz\,)=0$. However $\frac{\partial}{\partial z}\,(\,cos\,sz\,)\neq 0$. This contributes a term of the type $\frac{\partial u}{\partial z}\,\frac{\partial\,(\,cos\,sz\,)}{\partial z}$. But $cos\,(\,sz\,) = V\,\frac{\partial S}{\partial z}$ and $\frac{\partial}{\partial z}\,cos\,(\,sz\,)=\frac{\partial V}{\partial z}\,\frac{\partial S}{\partial z}+V\,\frac{\partial^2 S}{\partial z^2}$. So we have to consider in (12) additional terms $\frac{i}{2\pi v}\left[\frac{\partial V}{\partial Z}\left(\frac{\partial S}{\partial z}\right)^2+V\,\frac{\partial S}{\partial z}\,\frac{\partial^2 S}{\partial z^2}\right]$. A consideration quite similar to the one which made us disregard ΔS makes us disregard these terms as well.

The propagation of a wave in an isotropic medium thus presents no difficulty. If the medium is rendered non-isotropic through the influence of a magnetic field and if the terms in $\frac{4\pi}{3}\,\frac{Ne^2}{m}$ occurring in (1) may be neglected, the two possible velocities of propagation in a given direction $(\,l,\,m,\,n\,)$ may be determined by means of (7) and the states of polarization corresponding to each of these velocities are obtained by (4). If it is at all possible to discuss the propagation of waves by means of rays, then the rays must obey Fermat's Principle of least time. Therefore, we arrive at the following synthetic solution of the problem. While the plane wave is in a region of the atmosphere in which the electron density is zero, we may consider it resolved into two plane waves with such polarizations that on entering the refracting portions of the atmosphere each of these has a polarization satisfied by (4). From there on each wave is supposed to be bent in accordance with Fermat's Principle keeping its polarization determined by the particular root of (7) chosen for it in the beginning.

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NOTE ON PIEZOELECTRIC GENERATORS WITH SMALL BACK ACTION

By

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A number of circuit arrangements suitable for automatic piezo control have been devised. Three such arrangements are shown in the Figures.

In Fig. 1, the piezoelectric element is connected between the grid and a copper cylinder which is free to move over the coil in the plate circuit. The copper cylinder can be closed (forming a single turn) or open (axial cut.) The latter is more efficient since no appreciable power dissipation can take place in the cylinder, while the first method provides good tuning over a considerable frequency range. Experiments have shown this arrangement to be useful. The back action is very small since the piezoelectric element is coupled through distributed capacity to the plate branch.

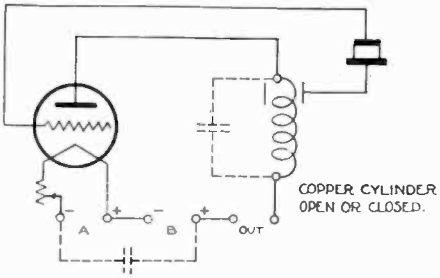

Figure 1

In the arrangement of Fig. 2, two tubes are used. Tube 1 contains the ordinary piezo oscillator. Tube 2 delivers the

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load. The same grid variations take place as on the grid of tube 1 but without loading it appreciably.

In Fig. 3, the same scheme is utilized with the exception that a coupling condenser C is used between the two grids and a grid path *r* and a voltage in series are inserted. Additional plate voltage taken from B_2 can be used for increasing the output. With this circuit the last tube can be adjusted to the proper point of the dynamic characteristic while the leak ρ of the first tube is used for producing proper amplitude of the current in that circuit.

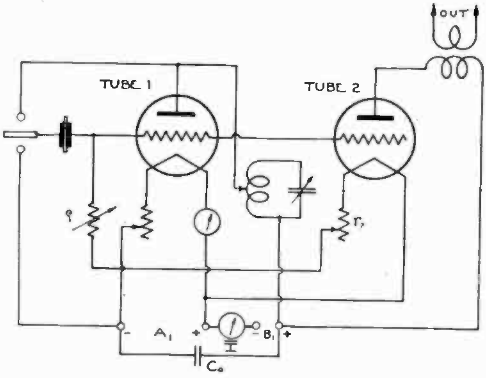

Figure 2

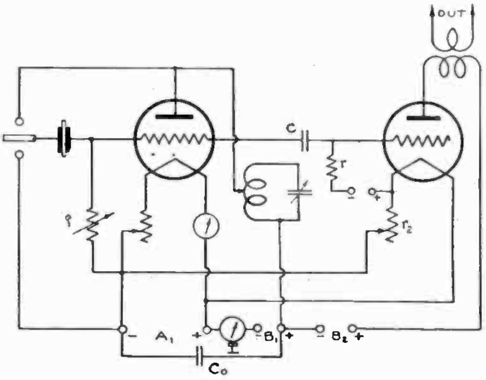

Figure 3

STANDARD FREQUENCY DISSEMINATION *

By

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In the radio laboratory of the Bureau of Standards, the term "standard frequency dissemination" designates that phase of the work devoted to making frequency standards widely available. The Bureau does this in a number of ways including the calibration of piezo oscillators and frequency meters, but this paper deals with the standard frequency disseminations through the medium of radio transmission.

The basis of this scheme depends upon the fact that a standard of radio frequency may, neglecting the effects of interference, be transmitted over great distances and reproduced at the receiving station with an accuracy equal to that attainable at the source of the frequency. Hence, if a radio transmitting station is operating on a constant frequency of accurately known value, that station serves as a disseminator of the frequency over an area determined by the effective range of transmission and reception.

The Bureau of Standards has three avenues for this means of standard frequency dissemination. First, standard frequency transmissions; second, selection by actual frequency measurements of certain transmitting stations which are termed "standard frequency stations;" third, the selection of certain "constant frequency stations" which maintain their frequencies close to the licensed values.

The standard frequency signals are transmitted from the Bureau's station WWV and from other stations which are equipped to make the transmissions with accuracy and regularity and which employ frequency standards in agreement with those used at the Bureau of Standards. Announcements of these transmissions are made from time to time in the Radio Service Bulletin, in the newspapers and in

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radio magazines. The transmissions from the Bureau of Standards have been maintained approximately once each month since 1922. Prior to October, 1924, the range of frequencies covered was from 125 to 2,000 kilocycles, but since that date this has been extended to 6,000 kilocycles. Evidence of the importance and general use of these transmissions was given by the fact that an announcement of their possible discontinuance in April, 1926, brought many objections from various laboratories.

In cooperation with the transmissions from WWV, similar transmissions were established from station 6XBM at Stanford University, California, in September, 1924. A standard frequency meter was built at the Bureau of Standards, given an initial calibration, and shipped to the University. A final calibration in terms of the Bureau's frequency standard was obtained by means of harmonics based upon a few fundamental frequency values. The values were determined by simultaneous measurements of transmitting stations by observers at the Bureau and Stanford University and from a small frequency meter having a number of fixed points determined in the Bureau's laboratory. The transmissions from 6XBM were terminated in June, 1926.

During the past year and in cooperation with the Bureau of Standards, standard frequency transmissions have been established from station 1XM, Massachusetts Institute of Technology, and from station 9XL, Gold Medal Flour Co., near Minneapolis, Minn. These transmissions cover the higher frequencies of direct value to amateurs.

Recently the accuracy of the standard frequency transmissions from WWV has been somewhat increased. Heretofore the transmission frequencies were determined by reference to primary standard frequency meters having high precision and constancy in calibration. The transmissions are now based upon a single frequency value of a quartz plate used in a piezo oscillator, checked during transmission by a standard frequency meter. A special advantage of the piezo oscillator as compared with the frequency meter is that it is unaffected by any power fluctuations which may occur in the transmitting set, but a change in the transmitted frequency is manifested in the head phones connected to the piezo oscillator by a deviation from zero beat.

The measurements upon standard frequency stations have been regularly maintained since November, 1923, and the results of these measurements are published monthly in the Department of Commerce Radio Service Bulletin. Only stations sufficiently close to the Bureau of Standards laboratory to be regularly and reliably received are measured. The number of stations that can be listed is necessarily limited. At the present time there are thirteen standard frequency stations, and these lie within the frequency range from 17 to 1,000 kilocycles.

For the measurements of the frequencies of the standard frequency stations the Bureau of Standards now has two completely equipped laboratories. One of these is located at the Bureau, while the other is situated near Kensington, Maryland, in a location which is almost entirely free from interference from electrical machinery and electrical apparatus.

The apparatus for measuring the frequencies of distant transmitting stations comprises a receiving set, a local generator, and a frequency meter. The generator is tuned to zero beat, and it is then exactly reproducing the frequency of the distant station subject to a slight error due to an audible beat in the band of audio frequencies approximately 32 cycles in width. The frequency of the local generator is then measured by means of the frequency meter. In the case of measurements of transmitting stations having frequencies of approximately 25 kilocycles or less, the width of the audio frequency band in the region of zero beat is sufficiently wide to introduce objectionable error in the measurements. To overcome this, a method is employed which involves adjusting the variable condenser of the local generator so that an audible beat of equal pitch is heard each side of the condenser setting corresponding to exact zero beat. A type of variable condenser is employed such that the changes in dial setting are directly proportional to the change in its capacity. The true zero-beat setting of the condenser is therefore located half-way between the settings giving the beat of equal pitch. Another method is to adjust the local generator until the beat produced matches the beat from an audio tuning fork having a frequency of accurately known value.

Improvements in the accuracy with which station frequencies can be measured have recently been made. One of these improvements is merely a refinement of the original method and involves the use of a frequency meter so constructed as to give high precision. A method of securing increased accuracy in the measurements which is somewhat new in its application is through the use of a heterodyne frequency meter in conjunction with a piezo oscillator. A heterodyne frequency meter is a generator constructed to give high precision and provided with calibration curves plotted to a large scale. In making a station measurement, the zero-beat adjustment is obtained in the same manner as before. The setting of the condenser of the heterodyne meter is then noted and although this is read to a high degree of precision, the frequency corresponding to it is not highly accurate. This is due to the fact that the calibration of the heterodyne meter does not remain constant. A correction for this frequency is obtained by utilizing harmonics from the piezo oscillator and the heterodyne meter. These points are then located upon the curve sheet and determine a new curve of high accuracy which is exactly similar in shape to the corresponding portion of the original curve. The intersection of this new curve with the station setting of the heterodyne meter is the true frequency of the station. The accuracy of this method may be extended to a very high order, the principal limitations being the accuracy of calibration of the piezo oscillator and the effects of temperature and humidity upon this calibration and the scale chosen for the calibration curves of the heterodyne meter.

Since the number of stations upon which frequency measurements can be made is limited, an additional scheme for compiling a list of stations that may serve as standards was adopted by the Bureau. These are the constant frequency stations listed each month in the Radio Service Bulletin. In April the list included approximately 5 per cent of the broadcasting stations in the United States. The selection of these stations depends upon the nature of the transmitting equipment which must not be subject to sudden changes in frequency, upon care and diligence on the part of the station operators, and the use of a special device for determining the station frequency. The special device may be a frequency indicator or a piezo oscillator.

The standard frequency dissemination work herein described is furthered by the issuance of mimeographed publications on methods of utilizing station frequencies, and apparatus for station frequency regulation. The Letter Circulars, obtainable by request from the Bureau of Standards, are named below.

Letter Circular 171, "Methods and apparatus for measurement of the frequencies of distant radio transmitting stations."

Letter Circular 180, "Specifications for frequency indicator, Bureau of Standards Type B, for use in radio transmitting stations."

Letter Circular 186, "Specifications for portable piezo oscillator, Bureau of Standards Type N."

Letter Circular 214, "Requirements of constant frequency stations."

Letter Circular 223, "Use of the piezo oscillator in radio broadcasting stations."

FORMULAS FOR THE CALCULATION OF THE CAPACITY OF ANTENNAS*

By
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A knowledge of the capacity of an antenna is of importance in calculating the natural wavelength of an antenna, for purposes of design and in the comparison of different types of antenna. Only the electrostatic capacity will be here considered. If the wavelength used is large, compared with the length of the antenna, this may be regarded as giving the effective capacity of the antenna.

The capacity of a conductor is defined as the quotient of the quantity of its charge by its potential. The potential is the sum of the potentials given it by its own charge and by the charges on the earth and on all the other conductors of the system. The effect of the charges on the earth is taken into account by including in the system image wires, one for each of the actual conductors. Each image wire is supposed to be placed as far below the surface of the earth as the conductor to which it corresponds is above the surface of the earth, and to carry a charge equal and of opposite sign to that upon the conductor, to which it corresponds.

Exact formulas for the capacity of a conductor are known in only a few instances, and unfortunately even the simplest case of a single horizontal cylindrical wire antenna would seem to offer great mathematical difficulties.

The known exact formula for the capacity of an isolated ellipsoid has sometimes been used, the longest dimension of the ellipsoid being taken equal to the length of the wire and the other two axes each equal to the radius of its cross section. However, the surface of such an ellipsoid does not coincide at all closely with that of the cylindrical wire, as the ends are approached, and in addition no account is taken of the effect of the earth.

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The formula for the capacity of two parallel wires of infinite length is well known, and thus, by considering one of these to represent the image wire, the solution is at hand for a horizontal wire whose length is very great compared with its height above the earth. This condition is, however, not satisfied with the usual single horizontal wire antenna, and thus the assumption that the capacity per unit length is the same as for an infinite wire of the same diameter, placed at the same height above the earth, gives only an approximate value for the capacity.

The charge distribution on an infinite horizontal wire is uniform in the direction of the axis. That is, the charge on the cylindrical surface intercepted between planes perpendicular to the axis and taken one cm. apart, is everywhere the same. Furthermore, the potential at an external point, due to the charge on the wire is very closely the same as though the same charge were distributed with a uniform linear density along the axis of the wire. In the case of a finite horizontal wire, on the other hand, the charge density is manifestly greater at the ends than at the center.

Supposing, however, that a charge be distributed uniformly over its cylindrical surface, the resulting potential may be calculated for different points of its surface. These potential values differ very little except for points quite close to the ends, and only at the extreme ends does the potential drop appreciably. The potential given to the wire by a uniform distribution on its image wire is practically constant over its whole length.

We may assume the value calculated for the middle point of the wire as fairly representative, and assume it to give an approximation to the true value which is better than that obtained from the formula for the infinite wire. This procedure was used in obtaining the formulas derived by the author and published in 1917.*

The variation along the wire of the potentials calculated on the assumption of a uniform charge density, shows that such a distribution could not be in equilibrium. The fall of potential toward the ends would cause charge to flow from the center toward the ends thus lowering the potential at the center, and raising it at the ends, until the whole wire would reach a uniform potential. This would have a value

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*Bureau of Standards Circular 74, "Radio Instruments and Measurements" which would evidently be less than the value previously calculated for the center, and greater than that calculated for the ends.

We may assume, as was done by Howe,* that the average of the potentials calculated for a uniform charge density, taken over the conductor, is a good approximation to the true equilibrium potential. The following summary shows the differences between the various approximations above enumerated in the case of a horizontal wire 50 ft. long, 0.01 ft. in diameter, and placed 25 ft. above the earth.

	RELATIVE POTENTIALS	
	Isolated	Earth Effect Included
Infinite Wire Assumption	———	18.420
Isolated Ellipsoid	18.420	———
Uniform Charge Density, Center	18.420	17.458
Howe's Assumption	17.807	16.874

An idea of the accuracy of the Howe assumption has been obtained by the author of the present paper by a method of successive numerical approximations to the true distribution of charge and the equilibrium potential, in two special cases. For the wire above assumed, isolated in space, it appears that the Howe's approximation to the potential is 0.2 to 0.3 per cent too large. For the same wire, arranged in a vertical position, with its lower end only one foot from the ground, the error is about 1 per cent, and in the same direction as before.

Such calculations are laborious and time consuming, and each new combination of wires has to be treated as a special case. Thus, no general statement can be made as to the accuracy of the Howe approximation in the case of more complicated antennas. The author believes, however, that with the long thin wires usual in antennas, the Howe method of approximation gives an accuracy sufficient for engineering requirements.

Accordingly this method of solution has been used to derive the formulas of this paper. All the more common forms of antenna have been included. In those cases where a number of wires have been joined together, two methods may be followed. First, the Howe potential may be calculated for each wire, on the supposition that a uniform linear charge density has been given to the whole system. The

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Lond. Elec. 73, p. 829; 1914.

equilibrium potential is taken as the mean of the Howe potentials of all the wires, weighting each value according to the length of the wire to which it corresponds. Or, as an alternative method, the various elements may each be supposed uniformly charged, but with a different value of charge density for each wire. The relation between these different charge densities is determined so as to give the same Howe potential to each wire. The two methods are in close agreement for long thin wires.

A collection of formulas found by the methods described, covering different antenna types, together with tables of constants to aid in the calculations, and tables of the capacity itself for certain simple antenna systems, has been issued as a letter circular by the Bureau of Standards.

DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued May 24, 1927---July 26, 1927

By

JOHN B. BRADY

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

1,629,685—RADIO TELEPHONY—W. T. DITCHAM, of Twickenham, England. Filed Dec. 18, 1920, issued May 24, 1927. Assigned to Radio Corporation of America.

1,629,727—ELECTRIC WAVE PRODUCING AND CHANGING DEVICE—H. J. MURRAY, of Brooklyn, N. Y. Filed April 1, 1921, issued May 24, 1927.

1,629,825—RADIO APPARATUS—J. J. W. KENAN and WILLIAM M. CADY, of Newark, New Jersey. Filed March 28, 1925, issued May 24, 1927.

1,629,826—RADIO APPARATUS—J. J. W. KENAN and WILLIAM M. CADY, of Newark, New Jersey. Filed March 28, 1925, issued May 24, 1927.

1,629,867—ELECTRICAL CONDENSER—J. F. HERMAN, of Chappaqua, New York. Filed March 26, 1925, issued May 24, 1927. Assigned to Straitline Radio Corp., of New York.

1,629,882—CONDENSER—W. H. PETIT, of Dayton, Ohio. Filed May 26, 1922, issued May 24, 1927. Assigned to Miami Radio Corp., Co.

1,630,227—APPARATUS FOR RECORDING ELECTRICAL SIGNALS—ALBERT H. TAYLOR, of Washington, D. C. Filed Jan. 2, 1924, issued May 24, 1927.

1,630,349—ARRANGEMENT FOR SUPPRESSING IDLE CURRENT IN RADIO TRANSMITTERS—A. MEISSNER, Berlin, Germany. Filed April 6, 1925, issued May 31, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie M. B. H.

1,630,354—RADIO SIGNALING APPARATUS—T. D. PARKIN, Chelmsford, England. Filed Dec. 2, 1921, issued May 31, 1927. Assigned to Radio Corp. of America.

1,630,383—DETECTOR—E. J. HAVERSTICK, Oakmont, Pa. Filed June 28, 1923, issued May 31, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,630,431—ELECTRON DISCHARGE DEVICE—W. G. HOUSEKEEPER, New York, N. Y. Filed July 29, 1922, issued May 31, 1927. Assigned to Western Electric Co.

1,630,443—MULTIPLE TRIODE VACUUM TUBE—F. S. McCULLOUGH, Wilkinsburg, Pa. Filed Mar. 15, 1923, issued May 31, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,630,859—CIRCUIT SYSTEM FOR RADIO FREQUENCY CURRENTS—R. S. MINER, West Hartford, Conn. Filed April 2, 1923, issued May 31, 1927. Assigned one-half to The C. D. Tuska Co.

1,630,885—VARIABLE CONDENSER—J. R. BARNHART, Lakewood, Ohio. Filed April 9, 1924, issued May 31, 1927. Assigned to Walter M. Scott.

1,630,980—RADIO INTERFERENCE ELIMINATOR— F. W. STEIN, Atchison, Kansas. Filed April 30, 1925, issued May 31, 1927.

1,630,021—THERMIONIC INDICATING MEANS RESPONSIVE TO LIGHT VARIATIONS—J. J. DOWLING, Bathgar, Dublin, Ireland. Filed Mar. 2, 1926, issued May 31, 1927.

1,631,035—RADIO TUBE—H. K. HUPPERT, San Francisco, Calif. Filed Dec. 21, 1925, issued May 31, 1927.

1,631,100—VIBRATION DAMPING MEANS FOR VACUUM TUBES—C. C. LAURITSEN, St. Louis, Mo. Filed Feb. 25, 1926, issued May 31, 1927. Assigned to The Dyal Co.

1,631,226—VARIABLE CONDENSER—E. J. SCHRODER, Lowell, Indiana. Filed April 20, 1925, issued June 7, 1927.

1,631,360—VARIABLE CONDENSER—G. H. CLARK, Brooklyn, N. Y. Filed June 27, 1922, issued June 7, 1927. Assigned to Radio Corp. of America.

1,631,453—VARIABLE CONDENSER—R. L. ASPDEN, Chorley, England. Filed May 29, 1926, issued June 7, 1927.

1,631,599—VARIABLE CONDENSER—A. LEIB, Berlin, Germany. Filed Sept. 14, 1923, issued June 7, 1927. Assigned to Gesellschaft fur Drahtlose Telegraphie M. B. H.

1,631,636—VARIABLE CONDENSER—E. B. LEWIS, Waterbury, Conn. Filed June 17, 1925, issued June 7, 1927.

1,631,738—VERNIER CONDENSER—F. KOCH, Newark, New Jersey. Filed April 28, 1924, issued June 7, 1927. Assigned to U. S. Tool Company, Inc.

1,631,672—VACUUM DISCHARGE APPARATUS—S. DUSHMAN, Schenectady, N. Y. Filed Mar. 4, 1919, issued June 7, 1927. Assigned to General Electric Co.

1,631,917—THERAPEUTIC OSCILLATOR—C. C. COOK and G. S. WALKER, Saginaw, Michigan. Filed Jan. 28, 1925, issued June 7, 1927.

1,632,039—ANTENNA STRUCTURE—A. A. OSWALD and E. L. NELSON, East Orange, N. J. Filed Jan. 4, 1924, issued June 14, 1927. Assigned to Western Electric Co.

1,632,054—OSCILLATION GENERATOR—J. T. L. BROWN, New Brighton, N. Y. Filed Dec. 3, 1923, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,069—WAVE TRANSMISSION SYSTEM—R. V. L. HARTLEY, South Orange, N. J. Filed Aug. 14, 1924, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,074—CONTROL APPARATUS—W. G. HOUSKEEPER, New York City. Filed Mar. 30, 1922, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,080—ELECTRIC DISCHARGE DEVICE—J. B. JOHNSON, Elmhurst, N. Y. Filed Dec. 27, 1921, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,130—CONDENSER—A. HADDOCK, East Orange, N. J. Filed Dec. 7, 1925, issued June 14, 1927. Assigned to Bell Telephone Laboratories, Inc.

1,632,135—ELECTRODE STRUCTURE FOR ELECTRON DISCHARGE DEVICES—W. F. HENDRY, New York, N. Y. Filed April 1, 1920, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,150—RADIO RECORDER—H. P. SPARKES, Edgewood Park, Pa. Filed Sept. 23, 1921, issued June 14, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,632,252—TWO-WAY WAVE TRANSMISSION SYSTEM—J. F. FARRINGTON, Flushing, N. Y. Filed Dec. 29, 1923, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,369—RADIO SIGNALING SYSTEM—A. CROSSLEY, Washington, D. C. Filed Oct. 12, 1925, issued June 14, 1927. Assigned to Wired Radio, Inc.

1,632,389—OSCILLATION GENERATING AND SUPPLY SYSTEM—A. M. CURTIS, East Orange, N. J. Filed Dec. 8, 1923, issued June 14, 1927. Assigned to Western Electric Co., Inc.

1,632,487—VARIABLE CONDENSER—R. H. MANSON, Rochester, N. Y. Filed Feb. 16, 1923, issued June 14, 1927. Assigned to the Stromberg-Carlson Telephone Mfg. Co.

1,632,488—VARIABLE CONDENSER—R. H. MANSON, Rochester, N. Y. Filed Nov. 24, 1924, issued June 14, 1927. Assigned to the Stromberg-Carlson Telephone Mfg. Co.

1,632,615—VACUUM TUBE—L. E. MITCHELL and A. J. WHITE, East Cleveland, Ohio. Filed June 5, 1924, issued June 14, 1927. Assigned to General Electric Co.

1,632,649—VARIABLE CONDENSER—F. O. HARTMAN, Mansfield, Ohio. Filed Nov. 23, 1925, issued June 14, 1927. Assigned to Hartman Electrical Mfg. Co.

1,632,650—VARIABLE CONDENSER—F. O. HARTMAN, Mansfield, Ohio. Filed Dec. 8, 1925, issued June 14, 1927. Assigned to Hartman Electrical Mfg. Co.

Re. 16,651—ADJUSTMENT OF WAVELENGTHS—G. H. CLARK, Washington, D. C. Filed (original) Jan. 18, 1915; reissue filed April 20, 1920, issued June 14, 1927. Assigned to Radio Corp. of America.

1,632,870—THERMIONIC VALVE—A. C. BARTLETT and R. LE ROSSIGNOL, of Waterford, and Harrow, England, respectively. Filed Nov. 17, 1926, issued June 21, 1927. Assigned to General Electric Co. Ltd.

1,632,878—RADIO CONDENSER COUPLING—H. A. BREMER, Chicago, Ill. Filed May 24, 1926, issued June 21, 1927.

1,632,982—WAVE METER—J. O. MAUBORGNE and GUY HILL, of Washington, D. C. Filed Aug. 6, 1920, issued June 21, 1927.

1,633,019—ELECTRICAL CONDENSER—W. H. HUTH, Chicago, Ill. Filed Nov. 19, 1925, issued June 21, 1927.

1,633,059—CRYSTAL DETECTOR—H. C. ADAM, St. Louis, Missouri. Filed Mar. 12, 1923, issued June 21, 1927.

1,633,100—PLURAL CHANNEL SIGNALING—R. S. HEISING, East Orange, N. J. Filed Mar. 3, 1916, issued June 21, 1927. Assigned to Western Electric Co., Inc.

1,633,285—HIGH POTENTIAL ELECTRICAL CONDENSER—J. A. PROCTOR, Lexington, Mass. Filed Feb. 1, 1921, issued June 21, 1927. Assigned to Wireless Specialty Apparatus Co.

1,633,481—FREQUENCY REDUCING ARRANGEMENT—J. G. FALLOU, Paris, France. Filed Nov. 8, 1926, issued June 21, 1927. Assigned to L'Union Generale D'Electricite.

1,633,516—MULTIPLE VARIABLE CONDENSER—W. M. BROWER, Palo Alto, Calif. Filed Jan. 17, 1927, issued June 21, 1927. Assigned to Federal Telegraph Co.

1,633,566—ELECTRICAL CONDENSER—W. A. BOCKIUS, of Park Ridge, Ill. Filed Feb. 13, 1925, issued June 28, 1927.

1,633,775—ANTENNA ARRANGEMENT FOR SHORT-WAVE APPARATUS—A. ESAU, of Jena, Germany. Filed Oct. 8, 1926, issued June 28, 1927.

1,633,870—CONDENSER—E. F. POTTER, of Glencoe, Ill. Filed Jan. 26, 1923, issued June 28, 1927. Assigned to Kellogg Switchboard and Supply Co.

1,633,915—CONDENSER—G. A. YANOCHOWSKI, of Chicago, Ill. Filed Jan. 25, 1923, issued June 28, 1927. Assigned to Kellogg Switchboard and Supply Co.

1,633,932—RADIO RECEIVING SYSTEM—F. B. FALKON, of Wilkinsburg, Pa. Filed April 26, 1923, issued June 28, 1927. Assigned to Westinghouse Electric & Manufacturing Company.

1,634,390—RADIO TRANSMITTING SYSTEM—V. K. ZWORYKIN, of Wilkinsburg, Pa. Filed March 17, 1924 issued July 5, 1927. Assigned to Westinghouse Electric and Mfg. Co.

1,634,627—SPARK GAP—H. H. OSBORN, of Chicago, Ill. Filed June 10, 1921, issued July 5, 1927. Assigned to H. G. Fischer & Co.

1,634,896—RADIO APPARATUS—L. B. COGSWELL, of West Springfield, Mass. Filed March 25, 1926, issued July 5, 1927.

1,634,930—VARIABLE CONDENSER—C. P. BROCKWAY, of Toledo, Ohio. Filed Nov. 22, 1923, issued July 5, 1927. Assigned to Arcturus Radio Co.

1,634,962—RADIO RECEIVING SYSTEM—R. SACHTLEBER, of Brooklyn, New York. Filed Sept. 22, 1924, issued July 5, 1927.

1,635,117—SIGNAL RECEIVING SYSTEM—F. W. DUNMORE, of Washington, D. C. Filed Feb. 27, 1922, issued July 5, 1927.

1,635,152—RADIO BROADCAST SELECTING AND DISTRIBUTING SYSTEM—EDWARD E. CLEMENT, of Washington, D. C. Filed Feb. 29, 1924, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,153—SUBDIVIDED SERVICE SYSTEM OF RADIO BROADCAST DISTRIBUTION—E. E. CLEMENT of Washington, D. C. Filed Oct. 28, 1924, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,156—RADIO BROADCAST DISTRIBUTING SYSTEM—E. E. CLEMENT, of Washington, D. C. Filed Oct. 21, 1925, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,157—RADIO BROADCAST DISTRIBUTING SYSTEM—E. E. CLEMENT, of Washington, D. C. Filed Oct. 28, 1924, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,158—RADIO BROADCAST DISTRIBUTING SYSTEM—E. E. CLEMENT, of Washington, D. C. Filed Oct. 28, 1924, issued July 5, 1927. Assigned to Edward F. Colladay.

1,635,459—RADIO CONDENSER—P. A. CHAMBERLAIN, of Chicago, Ill. Filed Sept. 18, 1925, issued July 12, 1927. Assigned to Mohawk Electric Corp.

1,635,556—CONDENSER—R. E. MARBURY, of Wilkinsburg, Pa. Filed Jan. 16, 1924, issued July 12, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,635,990—ELECTROSTATIC CONDENSER—JOHN O. GARGAN, Brooklyn, N. Y. Filed Aug. 26, 1924, issued July 19, 1927. Assigned to Western Electric Co., Inc.

1,635,992—ELECTRON DISCHARGE DEVICE—A. HADDOCK, East Orange, N. J. Filed Oct. 4, 1924, issued July 19, 1927. Assigned to Western Electric Co., Inc.

1,635,999—ELECTRON DISCHARGE DEVICE—W. G. HOUSKEEPER, New York, N. Y. Filed Sept. 22, 1922, issued July 19, 1927. Assigned to Western Electric Co.

1,636,015—ELECTRON DISCHARGE DEVICE—V. L. RONCI, Brooklyn, N. Y. Filed April 27, 1923, issued July 19, 1927. Assigned to Western Electric Co., Inc.

1,636,094—ARC. GENERATOR—R. HERZOG, Berlin, Germany. Filed Dec. 29, 1925, issued July 19, 1927. Assigned to C. Lorenz Aktiengesellschaft.

1,636,239—VACUUM TUBE—H. E. METCALF, San Leandro, Calif. Filed Oct. 13, 1924, issued July 19, 1927. Assigned to The Magnavox Company.

1,636,261—VACUUM OR THERMIONIC TUBE—A. G. THOMAS, Lynchburg, Virginia. Filed Aug. 2, 1923, issued July 19, 1927.

1,636,269—CRYSTAL DETECTOR—E. A. WRIGHT, Chicago, Ill. Filed Aug. 22, 1925, issued July 19, 1927. Assigned to F. H. Noble Co.

1,636,328—CONDENSER—L. B. SAUER, Western Springs, Ill. Filed May 23, 1924, issued July 19, 1927. Assigned to Kellogg Switchboard and Supply Co.

1,636,503—VARIABLE CONDENSER—H. G. DORSEY, Gloucester, Mass. Filed Nov. 11, 1922, issued July 19, 1927. Assigned to Submarine Signal Company.

1,636,570—RADIO RECEIVING APPARATUS—F. A. KOLSTER, Palo Alto, Calif. Filed Jan. 24, 1926, issued July 19, 1927. Assigned to Federal Telegraph Co.

1,636,921—PIEZO AUDION—A. McL. NICOLSON, New York City. Filed Nov. 11, 1926, issued July 26, 1927. Assigned to Wired Radio, Inc.

1,637,045—RADIO RECEIVING SYSTEM—EDWARD W. KELLOGG, Schenectady, N. Y. Filed April 15, 1922, issued July 26, 1927. Assigned to General Electric Co.

1,637,119—ELECTROMAGNETIC SOUND REPRODUCER—F. A. KOLSTER and S. A. SOLLIE. Filed Jan. 18, 1927, issued July 26, 1927. Assigned to Federal Telegraph Co.

1,637,310—TRANSMISSION SYSTEM FOR RADIANT ENERGY—J. H. HAMMOND, JR., Gloucester, Mass. Filed originally Sept. 28, 1917, renewed Aug. 1, 1924 and issued July 26, 1927.

Des. 72,871—CONE SPEAKER—M. C. RYPINSKI, L. W. STAUNTON and S. BOURNE, of Yonkers, N. Y.; Jackson Heights, N. Y.; and New York City, respectively. Filed Mar. 10, 1926, issued June 14, 1927. Assigned to Brandes Laboratories, Inc.

1,637,119—ELECTROMAGNETIC SOUND REPRODUCER—F. A. KOLSTER and S. A. SOLLIE, of Palo Alto, Cal. Filed Jan. 18, 1927, issued July 26, 1927. Assigned to Federal Telegraph Company.

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED JULY 6, 1927

Elected to Associate Grade

Arkansas,	Conway, State Teachers College	Cordrey, E. E.
California,	Berkeley, 1615 Bonita Avenue	White, Harold D.
	Marshall, Radio Corp. of America	Wissmann, J. T.
	Santa Ana, 1017 North Ross St.,	Adamson, Wm. S.
	Wilmington, c-o Union Oil Company	Rang, George N.
Connecticut,	Fairfield, 180 Grasmere Avenue	Hurd, Ralph G.
Dist. Columbia,	Anacostia, Bellevue, Naval Res. Lab.	Fromer, Edw. J.
Georgia,	Atlanta, 1458 S. Gordon St., N. W.	Brewin, R. R.
	Atlanta, 956 Hurt Bldg.	Duncan, J. A. Jr.
	Atlanta, 8 East Wesley Avenue	Hillegas, J. W.
	Atlanta, 615 Rhodes Bldg.	Reed, R. I.
	College Park, 230 W. John Calvin Ave.	Evarts, C. W.
	Palmetto,	Cochran, B. W.
Illinois,	Chicago, 5517 Cornell Ave.,	Borgeson, Carl A.
	Chicago, 2139 Roscoe St.,	Douglas, James M.
	Chicago, 4439 Clifton Ave.,	Koch, Wm. A.
	Chicago, 3553 N. Paulina St.,	Marshall, C. E.
	Chicago, 1826 Diversey Parkway,	Poore, H. E.
	Harrisburg, Box 80,	Tate, J. R.
	Lake Forest,	Haviland, T. T.
	Mt. Pleasant, Station WJAZ	Gustafson, G. E.
	Springfield, 400 S. Illinois St.,	Whannel, R. I.
Indiana,	Albion, 609 East Hazel St.,	Evans, Harry R.
	Connersville, 629 Central Avenue,	Hanes, F. B.
	Huntington, 938 Henry St.,	Walter, I. V.
Iowa,	Le Mars,	Kime, D. O.
	Webster City,	Daniels, P. H.
Maryland,	Baltimore, R. C. A., 1508 Eutaw Place	Rhead, R. G.
Massachusetts,	Boston, 154 State Street,	Rosenwald, E. D.
	Melrose, 51 Malvern St.,	Phibrick, L. S.
Michigan,	Ann Arbor, Univ. of Michigan,	Holland, L. N.
Minnesota,	Excelsior,	McKesson, L. J.
Missouri,	Kansas City, 2505 Harrison,	Frisco, Ernest
	Kansas City, 5208 Prospect Ave.	Young, Wm. K.
	St. Joseph, 1118 Lincoln St.,	Watts, H. F.
	St. Louis, 210 N. Tenth St.,	Van Sickle, G. W.
Nebraska,	Omaha, 2656 Douglas St.,	James, G. W.
	Wayne, Wayne Hospital,	Shum, J. M.
New Jersey,	East Orange, 99 North 22nd St.,	O'Donohue, J. P.
	Hillside, 268 Conklin Ave.,	McCauley, J. H.
	Jersey City, 344 Forrest St.,	Shimshak, S. J.
	Paterson, 79 Auburn St.,	Coddington, R. L.
	West New York. 648 Palisade Ave.,	Machiorleti, L.
New York,	Binghamton, 215 Washington St.,	Adams, W. E.
	Brooklyn, 77 Bay 32nd St.,	Coburn, R. M.
	Brooklyn, 845 East 13th St.,	Glaser, E. M.
	Buffalo, 195 St. James Place,	Cavileer, R. P.
	Buffalo, 241 Reed St.,	Englan, Geo.
	Buffalo, 86 Crescent Ave.,	Huntsinger, F. J.
	Buffalo, 781 Ellicott St.,	Johnson, J. A.
	Buffalo, 303 Waltz Ave.,	Miller, H. A.
	Buffalo, 106 West Parade Ave.,	Nowak, V. V.
	Buffalo, 147 Condon Ave.,	Olson, Carl
	Buffalo, 184 Grape St.,	Rackle, Clarence
	Buffalo, 2407 Niagara St.,	Ritchie, W. G.
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	Buffalo, 76 Downing St.,	Windrath, R. V.
	Lockport, 245 West Ave.,	Lerch, H. W.
	Mitchel Field, 1st Observation Squadron,	Arnold, Thos. E.
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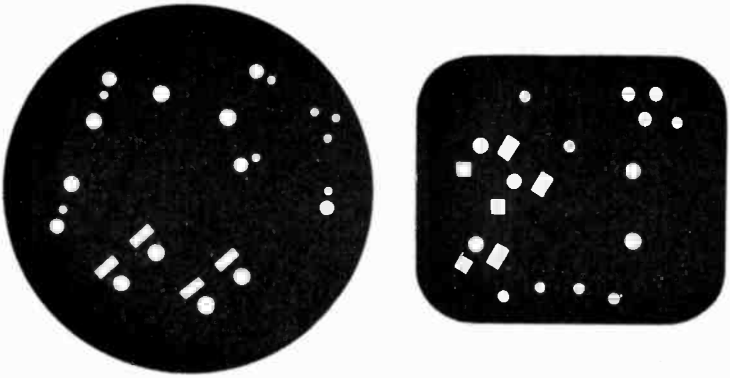
	New York, 557 West 124th St.,	F. C. Lee
	New York, 520 West 163rd St.,	Marron, Geo. P.
	New York, 523 East 147th St.,	Meyer, H. W.

	New York, 37 West 84th St.,	Stern, Irving
	New York, Battery Park, Barge Office, Rm. 10,	Wortman, F. W.
	Niagara Falls, 2454 Whitney Ave.,	Brown, O. R.
	Niagara Falls, P. O. Box 619	Lidbury, F. A.
	N. Tonawanda, 309 Division St.,	Freck, R. H.
	Staten Island, Dongan Hills,	Simons, Chas. D.
	Rochester, 35 Douglas Road,	McCanne, Lee
	Rochester, 72 Bedford St.,	Miller, R. G.
North Carolina,	Greensboro, c-o Y. M. C. A.,	Worrall, W. W.
North Dakota,	Fargo, State College,	Swisher, C. L.
Ohio,	Cleveland, 1326 Hayden Ave.,	Crabb, Robert
	Cleveland, 9007 Detroit Ave.,	Dorn, H. P.
	Cleveland, 1864 West 45th St.,	Sturtevant, Mark
Pennsylvania,	Harrisburg, 1926 Park St.,	Beachley, V. E.
	Nanticoke, 250 E. Main St.,	Brader, N. H.
	Philadelphia, 2901 N. Front St.,	Grunstein, Philip
	Philadelphia, 5107 N. Eleventh St.,	Miller, Chas. C.
	Pittsburgh, 1530 Rockland Ave.,	Sparr, H. J.
	South Mountain, Franklin County,	McIver, W. R.
	Wilkinsburg, 513 Holmes St.,	Cunningham, T. D.
Virginia,	Richmond, 3008-D Ellwood Ave.,	Selph, O. M.
Washington,	Auburn, P. O. Box 815,	Nagata, C. N.
	Ellensburg, 1006 East Second St.,	Sloan, D. H.
	Seattle, 6616 Fauntleroy Ave.,	Agner, G. M.
	Seattle, 2816-14th St.,	Richardson, Gordon
	Tacoma, Hotel Winthrop, Sta. KMO,	Haymond, Carl E.
Wisconsin,	Madison, 1702 W. Lawn Ave.,	Austin, O. C.
	Milwaukee, 991-3rd St.,	Bogenberger, J. Jr.
	Milwaukee, 780-29th St.,	De Land, R. E.
	Milwaukee, 1506 Capitol Drive,	Holst, E. W.
	Milwaukee, 678 Astor St.,	Leche, O. P.
	Milwaukee, 425 East Water St.,	Manney, G. C.
	Milwaukee, c-o School of Engineering,	Ruckle, Ernest
	Milwaukee, 772 Second St.,	Russell, Murray
	Milwaukee, 914 Center St.,	Seibel, D. E.
	Milwaukee, 467 Jackson St.,	Vaughn, F. A.
	Oshkosh, 668 Algoma Blvd.,	Shambeau, Wm. R.
Br. Columbia,	Vancouver, 458-30th Ave. E.	Way, Chas. W.
Br. West Indies,	San Fernando, 66 St. James St.,	Mallalieu, K. J.
Canada,	New Brunswick, Fredericton,	McCormack, E. F.
	Ontario, Toronto, 63 Balmoral Ave.,	Hebbel, C. T.
China,	Hong Kong, P. O. Box 491,	Diggle, A. J.
Costa Rica,	Cartago, Box 115,	Bond, Joseph
England,	Berks., Winnersh, "Kynance"	Weight, H. A.
	Chester, 69 Hoole Road,	Benn, H. L.
	Lancs., Southport, 67 Virginia St.,	Storry, T. G.
	Surrey, Surbiton, S. Bank House,	Harmer, L. B.
	Stoke-on-Trent, 30 Liverpool Road,	Watchurst, W. D.
New Zealand,	Ashburton, 263 Wills St. E.,	Dawson, W. M.
	N. Auckland, Parenga, Cape Maria Radio,	Tyler, G. E.

Elected to Junior Grade

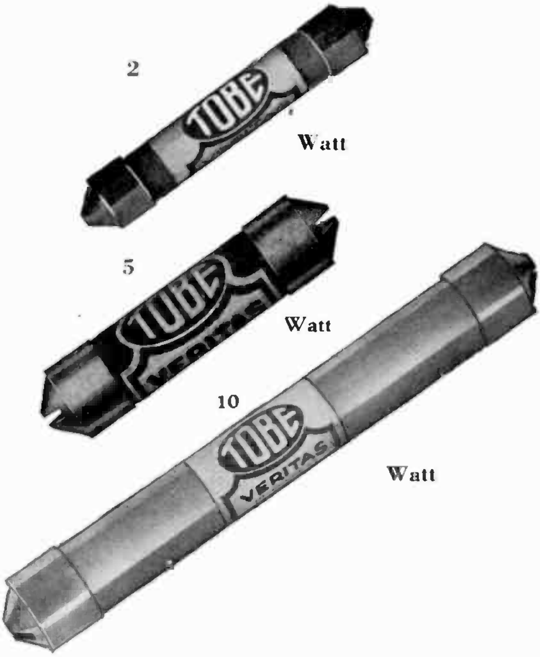
California,	Fresno, 827 Divisadero St.,	Foin, O. F. Jr.
Kentucky,	Custer,	Pyle, R. C.
Minnesota,	Duluth, 527 N. 54th St.,	Conrad, N. L.
Nebraska,	Omaha, 4112 N. 23rd St.,	Stultz, E. P.
New York,	807 St. Ann's Ave.,	Nicholides, E.
Wisconsin,	Milwaukee, c-o School of Engineering,	Gainer, Wm. P.
	Milwaukee, 1599 Port Washington Ave.,	Woodward, G. E. Jr.
Canada,	Ontario, Toronto, 2 Spadino Road,	Knapman, Jack
Scotland,	Banffshire, Buckie, 41 West Church St.,	George, F. R.



TOBE

TRADE MARK REG'D U. S. PAT. OFFICE

VERITAS RESISTORS



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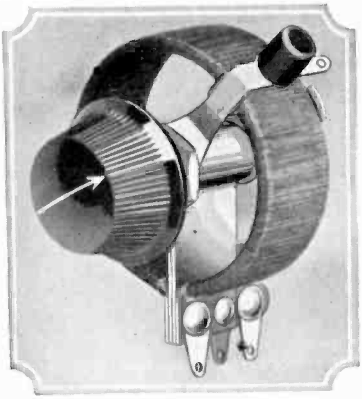

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Central Radio Laboratories
16 Keefe Avenue Milwaukee, Wisconsin

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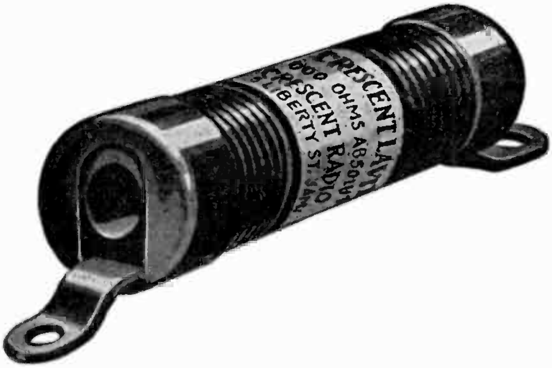

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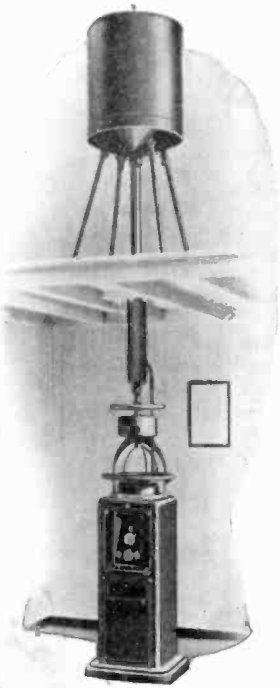

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CARTER

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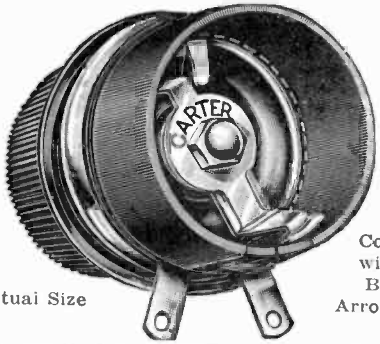

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Arrow-Pointer
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The frame is one piece all metal; will not break or crack.

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This Heavy Duty Type also made in Potentiometer form.

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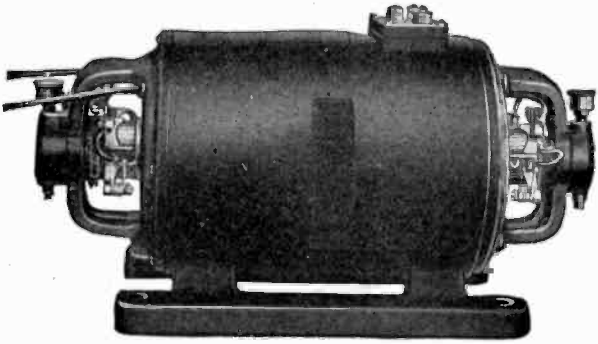

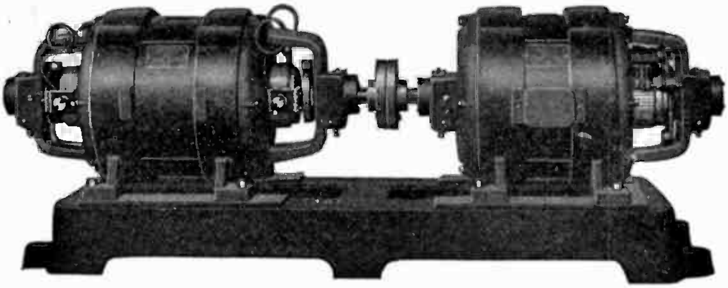

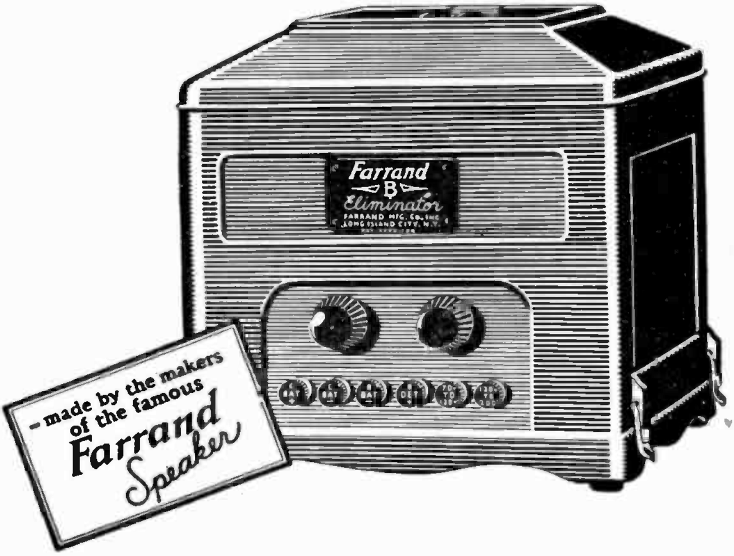

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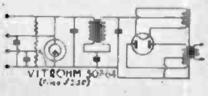

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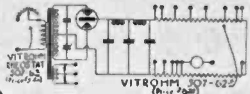
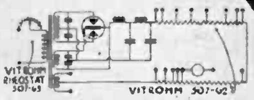
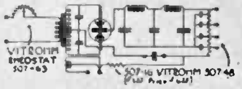
*Uses Vitrohm Resistor 507-62**
and Vitrohm Rheostat 507-63



Q R S A-B-C
400 m.a. CURRENT SUPPLY UNIT

*Uses Vitrohm Resistor 507-62**
*and Vitrohm Rheostat 507-63**





Q R S and RAYTHEON
85 m.a. CURRENT SUPPLY UNIT

Uses Vitrohm Resistors 507-16 and 507-48**

35 years of research and experience in the manufacture of resistors is incorporated in Vitrohms for radio.

There are available Vitrohms to give you noiseless, dependable service wherever resistance is indicated in a current or power supply circuit.

Vitrohms do not age or change in resistance value after use. Ten, twenty or thirty years of constant use under all conditions are every-day records of Vitrohms.

Vitrohms are "pre-aged" wire wound on porcelain tubes and protected by fused-on vitreous enamel for the permanent protection of resistance wire and terminals.

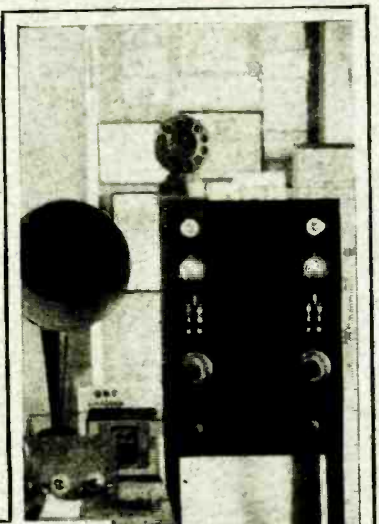
Per square inch of surface, Vitrohms have greater watt dissipation than any other resistor.

Send 15c for "How To Use Resistance in Radio." It contains many circuits of interest to all experimenters. Bulletin 507 describing Vitrohms for radio is sent without charge upon request.

* Approved by the Raytheon and Q R S Laboratories.

Ward Leonard Electric Company
3741 South Street Mount Vernon, N. Y.

Batteries can't be beat for plate power



Station 2TL
Port Jervis, N. Y.

Mr. Jos. W. Gibbons

When you put on the cans and hear the buzzes, rattles and jazz of raw or partly rectified AC; and when in the midst of the racket you suddenly come on a smooth pure DC note, strong, steady, sharp, and easily readable—that's the station you start to copy.

How about yourself, OM? Are you giving the other fellow the sort of sigs you prefer to read? Eveready Layerbilt "B" Batteries are putting on the air every night the prettiest DC notes you can hear. And as for DX—well, just read this, from Jos. W. Gibbons, 2TL, Port Jervis, N. Y.: "Short waves travel far on low power . . : 7PH of Everett, Washington, heard my signals when using 90 volts of Eveready 'B' Battery for plate power."

2TL adds: "Having followed up radio since the old spark days, I find that Eveready Batteries have stood the test of time on both transmitter and receiver. The old timers know that 'B' batteries can't be beat for plate power for a good report."

As far as we have been able to discover, the Eveready Layerbilt "B" Battery No. 486 is the longest-lasting, most economical "B" battery ever built, for both transmitters and receivers.

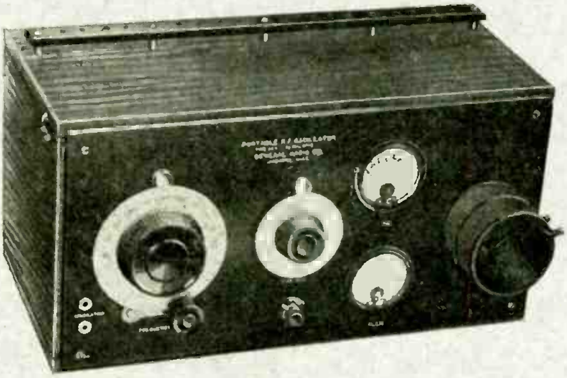
National Carbon Company, Inc.
New York San Francisco
Unit of Union Carbide and Carbon Corporation

Tuesday night is Eveready Hour Night—9 P. M., Eastern Standard Time, through the WEAF network.

EVEREADY
Radio Batteries
-they last longer

Portable R. F. Oscillator

with range of 15 to 30,000 meters



TYPE 384 R. F. OSCILLATOR

The Type 384 oscillator has a wide range of utility in the radio laboratory. It may be used as a source in high frequency measurements of coils and condensers, or for checking radio receivers. It is particularly useful for checking over-all receiver characteristics when combined with the General Radio Type 413 Beat-Frequency Oscillator. By using the Type 384 R. F. Oscillator and Type 413 Beat Frequency Oscillator radio frequency and audio frequency tests may be made simultaneously.

The Type 384 R. F. oscillator covers the range from 15 to 30,000 meters by means of nine plug-in coils.

A single UX-199 tube is used which permits an entirely self-contained instrument. A plate milliammeter is provided to indicate oscillation.

Type 384 R. F. Oscillator, without Coils . . .	$80.00
Type 384-D Figure 8 Coil (200-600 meters)	$4.00
Prices on coils covering other ranges on request.	
Type 413 B. F. Oscillator	$210.00

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Write for Bulletin 6150-I

General Radio Co., Cambridge, Mass.

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