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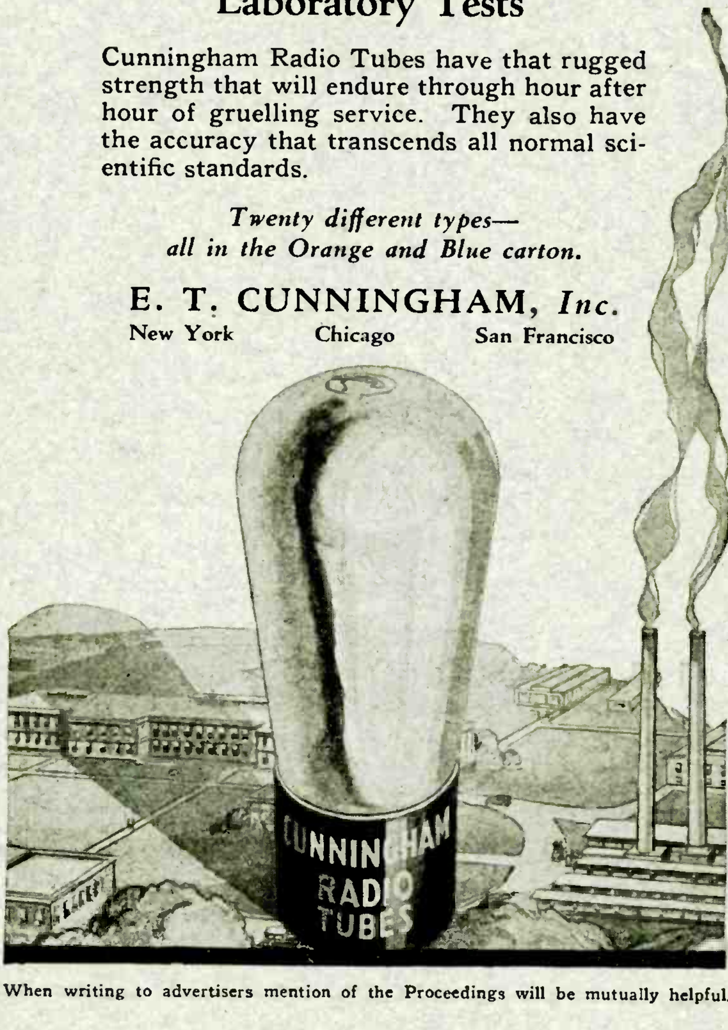

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New Bulletin No. K-420 covers R-S Types TA (3½ in.) and FA (4 in.) small, panel type, alternating current ammeters, voltmeters and wattmeters in capacities from 100 milli-amperes to 200 amperes and from 10 volts to 300 volts. The Type TA is illustrated.

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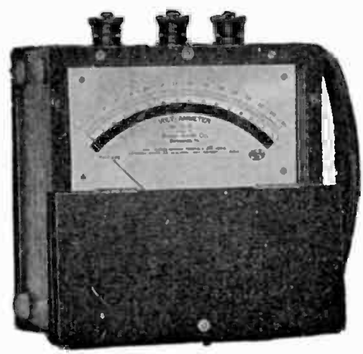

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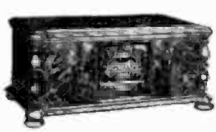

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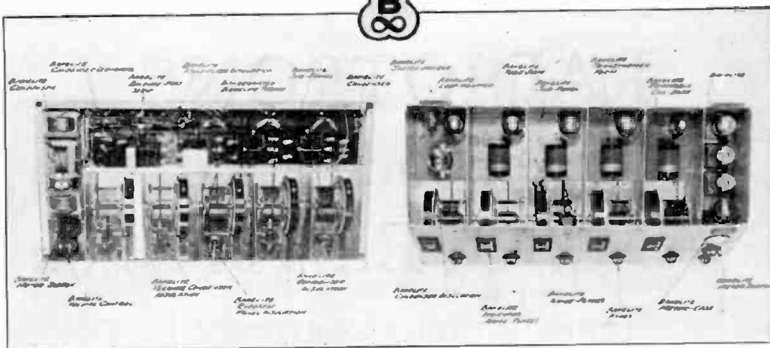

Two views of Transoceanic "Phantom" Receiver, showing the extensive use of Bakelite Materials. Made by Golden-Leutz, Inc., Long Island City, N. Y.

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

WARREN FETTER HUBLEY
Late Treasurer, Institute of Radio Engineers, 1915-1927

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 15 November, 1927 Number 11

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INSTITUTE ACTIVITIES

October Meeting of the Board of Direction

At the meeting of the Board of Direction of the Institute, held on October 5, 1927 in the Institute Offices, the following were present: Dr. Ralph Bown, President; Frank Conrad, Vice-President; Dr. A. N. Goldsmith, Secretary; L. A. Hazeltine, R. A. Heising, J. V. L. Hogan, R. H. Marriott and J. M. Clayton, Assistant Secretary.

Mr. Melville Eastham was elected Treasurer of the Institute to fill the unexpired term caused by the death of Mr. Warren F. Hubley.

Upon recommendation of the Committee on Admissions, the following applications for transfer or election to the higher grades of membership in the Institute were approved: Transfer to the grade of Fellow: M. C. Batsel and L. E. Whittemore; transfer to the grade of Member: Carlton D. Morris and Conway Taylor; election to the grade of Member: John F. Grinan, E. B. Moullin and G. McB. Salt.

One hundred and two Associate and ten Junior members were elected.

In the absence of nominations by petitions, the ballot for 1928 officers of the Institute will be as follows:

For President, Professor E. M. Bennett and Dr. A. N. Goldsmith.

For Vice-President, A. H. Grebe and L. E. Whittemore.

For Managers, Dr. W. G. Cady, Dr. J. H. Dellinger, W. G. H. Finch, and R. H. Manson.

Liebmann Memorial Prize

The Liebmann Memorial Prize for this year was awarded to Dr. A. Hoyt Taylor for his work in connection with the investigation of radio transmission phenomena. The award will be made at one of the sessions of the 1928 convention of the Institute early in January.

1928 Year Book

With the ballots for 1928 Officers of the Institute, which will be in the mail shortly after November 1st, there will be found a form to be used by all members in notifying the Institute of any changes of business or residential address for inclusion in the 1928 Year Book. These forms should be executed and returned with the ballots. All members are urged to return both promptly in the envelope provided therefor.

The 1928 Year Book will contain a catalog of the membership of all grades, as of December 31, 1927. Changes in address or unpaid members cannot be included in the membership list if the Institute is not notified prior to that date.

Membership Cards

At the October meeting of the Board of Direction it was decided that membership cards for *all* grades of membership shall be available next year. These cards will be issued only upon request.

Institute Meetings

New York Meeting

At the New York meeting of the Institute, held on October 5, 1927 in the Engineering Societies Building, 33 West 39th Street, New York City, a paper on "Method of Reducing The Effect of Atmospheric Disturbances" was presented by Major E. H. Armstrong.

In the discussion which followed the paper, the following took part: Dr. A. N. Goldsmith, Carl T. Englund, Roy Weagant, R. H. Marriott, D. K. Martin, A. H. K. Morse, A. E. Reoch, and others.

The attendance at this meeting was over six hundred.

A limited number of copies of the paper, in pamphlet form, are available to members of the Institute upon application to the Institute Offices.

Atlanta Section

A meeting of the Atlanta Section of the Institute was held in the Atlanta Athletic Club on September 7, 1927. H. L. Wills, Vice-Chairman, presided. Mr. Oliver, of the Southern Bell Telephone Company, read the paper of Messrs. Diamond and Webb on "The Testing of Audio-Frequency Transformer-Coupled Amplifiers."

Following the paper and discussion, the motion picture "Silent Drama of Telephony" was shown.

The attendance at this meeting was twenty-five.

Buffalo-Niagara Section

At the invitation of the local Section of the American Institute of Electrical Engineers, the Buffalo-Niagara Section of the Institute of Radio Engineers attended an A.I.E.E. meeting on September 23, 1927 at which E. C. Markley of the General Electric Company delivered a paper on "Carrier Current Telephony." This paper was discussed by C. A. Boddie and others.

A meeting of the Buffalo-Niagara Section was held on October 12, 1927 in Townsend Hall, University of Buffalo. L. C. F. Horle was the presiding officer.

A paper by Francis H. Engel, of the Radio Corporation of America, on "Notes on A-C. Tubes" was presented by Mr. Engel. Messrs. Hector, Stone, Horle, Jones and others participated in the discussion which followed.

The attendance was fifty-five members.

The next meeting of the Section will be held on November 17, 1927 in the Science Hall, University of Buffalo, at which time Dr. Ralph Bown, President of the Institute, will present a paper on "Transatlantic Telephony."

Canadian Section

At the September 14th meeting of the Canadian Section, held in the Electrical Building of the University of Toronto, A. M. Patience presided.

C. I. Soucy, of the University of Toronto, presented the first of a series of "Junior Lectures" prepared for the benefit of the younger members of the Section. The first lecture was entitled "Inductance Coils." It is planned that at each meeting of the Section, for a period of about twenty minutes, one of these Junior Lectures will be given. The proposed subjects include Coils, Condensers, Application of Coils and Condensers, Tubes, Radio-Frequency Circuits, Audio-Frequency Circuits, Sensitivity and Selectivity, Loudspeakers, Batteries and Rectifiers, Installation and Service Problems and other subjects.

The paper of the evening, entitled "Rectox, The Oxide-copper Rectifier", was presented by A. L. Atherton. In the discussion which followed, the following took part: Messrs. C. I. Soucy, C. L. Richardson, D. Hepburn, J. Leslie, G. E. Pipe, C. A. Lowry, A. R. Starr and others.

Cleveland Section

On September 23, 1927 a meeting of the Cleveland Section, J. R. Martin presiding, was held in the Auditorium of the new School of Lighting, General Electric Company, Nela Park. Wm. T. L. Cogger, Manager of the Vacuum Tube Department of the Nela Works, presented a paper on "The New Vacuum Tube (A-C.)".

Seventy-five members of the Section attended the meeting.

On November 4, 1927, Dr. Ralph Bown, President of the Institute, will present a paper "Transatlantic Telephony" in the Case School of Applied Science, Cleveland.

Philadelphia Section

The September meeting of the Philadelphia Section was held on the twenty-third of the month in the Bartol Laboratories. J. C. Van Horn presided.

A paper by Dr. Ralph Bown on "Transatlantic Radio Telephony" was presented by Dr. Bown.

Messrs. Miller, Arnold, Snyder and others took part in the discussion which followed.

The attendance was fifty-two members.

San Francisco Section

The San Francisco Section is planning a renewal of its meetings beginning with a session in November, to continue throughout the winter and spring months.

At a luncheon in the Engineers' Club, San Francisco, on October 7, 1927 the affairs of the Section were discussed by Dr. L. F. Fuller (member of the Committee on Sections), D. B. McGown, B. H. Linden, A. Y. Yuel, and Past-President Donald McNicol.

The large growth of radio on the Pacific Coast now makes possible the presentation of a wide range of papers on radio and related subjects of interest to members of the Institute.

Seattle Section

On September 17, 1927 a meeting of the Seattle Section was held in the Telephone Building, 3rd and Senaca Streets, Seattle. Mr. T. M. Libby was Chairman.

Two papers were presented. The first, by J. R. Tolmie, was a discussion of the Preliminary Drafts of Reports of the Institute's Sub-committees on Vacuum Tubes, Receiving Sets and Electro-Acoustic Devices. The following participated in the discussion of these preliminary drafts: Messrs. Libby, Willson, Burleigh, and others.

The second paper, by Howard F. Mason, was entitled "Short Wave Transmitter Operation in the Arctic." Messrs. Libby, Burleigh, and Tolmie discussed this paper.

Twenty-eight members of the Section were present at the meeting.

Proposed Sections

Continued correspondence is being held with members of the Institute looking to the formation of Sections of the Institute in the following localities: Milwaukee, New Orleans, Pittsburgh, Denver, and Minneapolis.

It is anticipated that Sections will be formed in several of these cities before the first of next year.

Committee Work

Sub-Committee of the Standardization Committee

A meeting of the Sub-committee on Determination of Circuit Constants was held at Institute Headquarters on September 23, 1927. The following members of the Sub-committee were present: H. M. Turner, Chairman; C. T. Burke, G. C. Crom, and G. S. Southworth.

Miscellaneous questions pertaining to symbols and nomenclature which have been referred to the Committee were discussed and for the most part agreed upon. Certain of these matters are to be taken up with the American Institute of Electrical Engineers with the hope of securing uniformity of action.

The greater portion of the day was devoted to a consideration of a partially completed preliminary draft of a report. One of the major problems is that of obtaining convenient and reliable methods of measuring the inductance of iron-cored coils with and without superposed direct current. The Chairman has spent considerable time studying the various methods in general use and devising others and comparing them. Some of this material was submitted to the Committee members present at this meeting for their comments and suggestions. Other problems discussed included determining the characteristic curves of audio-transformers, measuring the capacity of large and small condensers, the distributed capacity of coils, and measuring the resistance of grid leaks.

Committee on Admissions

At its meeting held on October 5, 1927, the following members of the Committee on Admissions were present: Vice-President, Frank Conrad, Chairman; Professor L. A. Hazeltine, Dr. L. M. Hull, F. H. Kroger, and H. F. Dart. The Committee acted upon eleven applications for transfer or election to the higher grades of membership in the Institute.

Committee on Membership

A meeting of the Committee on Membership was held in the Institute Headquarters on September 23, 1927. At this meeting H. F. Dart, Chairman, E. R. Shute, and J. M. Clayton were present.

The Committee has many plans before it looking to circularization and other activities which will result in an increase in the membership in the Institute.

OBITUARY

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

Warren Fetter Hubley

Mr. Hubley's interest in the development of The Institute of Radio Engineers goes back to the early days of the organization. He first became interested in the meetings of the Wireless Institute. When that parent organization combined with the Society of Wireless Telegraph Engineers to form The Institute of Radio Engineers in May, 1912, he continued to attend the meetings and was elected to associate membership November 4th, 1912. He was elected to the grade of Member February 8, 1926. In 1915 he was elected Treasurer and continued in that capacity, except for the year 1917, until his death.

Mr. Hubley, throughout his long and helpful association with The Institute of Radio Engineers, gave unsparingly of time and effort in the interest of the Institute. His unremitting attendance at meetings of the Board of Direction of the Institute rendered available his wise counsel and careful analysis of the problems of Institute management, to which he gave his undivided and thoughtful attention.

The Board of Direction as a group has lost a capable and conscientious associate, and as individuals they have lost a close friend.

Mr. Hubley was born May 9th, 1880. He died at 4 A.M. September 19th, 1927.

ON THE VALUES AND THE EFFECTS OF STRAY CAPACITIES IN RESISTANCE-COUPLED AMPLIFIERS*

By

Manfred von Ardenne and Wolfgang Stoff

(Berlin)

It is a well-known fact that among the stray capacities in a vacuum-tube amplifier the capacity between grid and plate of the tube is of primary importance, especially since this capacity is materially increased by the amplification of the tube. In Germany this effect first was explained by H. Barkhausen.† The corresponding equations, however, have very often not been properly applied. For example, the effective direct grid-plate capacity is given as $(1+A_v)$ times larger than the statically measured C_{gp}. A_v is the actual voltage amplification of the tube arrangement in question. In this case the fact is overlooked that the phase-difference existent between ΔE_g and ΔE_p is generally not 180 deg., but some other considerably different angle. This fact, of course, is immaterial in respect to mere amplification, but it is of first importance in regard to the charge and actual amount of grid-plate capacity increase. The fact that the actual increase of C_{gp} is, according to the phase angles, smaller than $(1+A_v)$ times, is naturally of the greatest influence on the judgment of amplifier-dimensions such as are given by the writer since the beginning of 1925, namely, resistance-coupled audio-amplifiers with tube-plate resistances and outer plate circuit resistances of the order of megohms. In the following, the general theory of stray capacities will first be developed with due consideration to the phase conditions mentioned above, and then the values and effects of stray capacities will be discussed for a three-stage amplifier of practical dimensions.

Fig. 1 shows the simplified circuit and Fig. 2 the corresponding symbolic diagram employed to explain the calculation of the effective stray capacities in a single-stage amplifier with optional plate load. Here the tube is considered to be an alternating current generator with the e.m.f. $=\mu\cdot\Delta E_{go}$ and the internal resistance

* Original Manuscript received by the Institute, July 6, 1927. Revised Manuscript received by the Institute, August 12, 1927.

† See also Manfred von Ardenne and Wolfgang Stoff: "Die Berechnung der Scheinkapazitäten bei Widerstandsverstärkern," Jahrbuch der drahtlosen Telegraphie und Telephonie, Vol. 30, No. 13.

r_p, according to a well-known definition. Parallel to the plate-circuit impedance load Z_b are the plate-filament capacity C_{pf} and the grid-plate capacity C_{gp}, which is connected in series with the

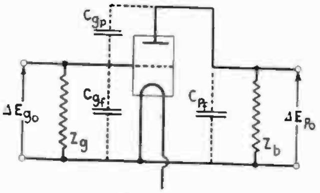

Figure 1

e.m.f. $=\mu \cdot \Delta E_{go}$. The alternating grid-potential ΔE_{go} should be considered as given; ΔE_{po} is the potential across $Z_{b_{tot}}$, which is the total plate-circuit impedance formed by Z_b and the parallel stray capacities. The potential ratio $\dfrac{\Delta E_{po}}{\Delta E_{go}}$ is given by the following well-known equation:

$$\frac{\Delta E_{po}}{\Delta E_{go}} = -\mu \cdot \frac{1}{1+\dfrac{r_p}{Z_{b_{tot}}}} \tag{1}$$

Here $\dfrac{1}{Z_{b_{tot}}}$ is equal to: $\dfrac{1}{Z_b}+j\omega C_{pf}+j\omega C_{gp}\cdot\left(1-\dfrac{\Delta E_{go}}{\Delta E_{po}}\right)$ (2)

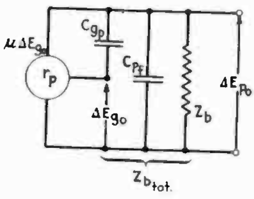

Figure 2

After a mathematical operation with these two equations the following results:

$$\frac{\Delta E_{po}}{\Delta E_{go}} = -\mu \cdot \frac{1-j\omega C_{gp}\cdot\dfrac{r_p}{\mu}}{1+r_p\left(\dfrac{1}{Z_b}+j\omega C_{pf}+j\omega C_{gp}\right)}$$

Just as in the plate circuit the stray capacities must also be considered in the grid-circuit. Fig. 3 shows the corresponding symbolic diagram for the grid-circuit. Parallel to the grid circuit impedance Z_g of optional type are the grid-filament capacity C_{gf} and the grid-plate capacity C_{gp}, which is in series with the e.m.f. $=\Delta E_{po}$. The total grid-circuit impedance $Z_{g_{tot}}$ may then be calculated according to the following equation:

$$\frac{1}{Z_{g_{tot}}}=\frac{1}{Z_g}+j\omega C_{gf}+j\omega C_{gp}\left(1-\frac{\Delta E_{po}}{\Delta E_{go}}\right) \tag{3}$$

In order to make use of equation (3) properly the phase-angle ϕ between ΔI_{po} and ΔE_{po} and ψ between ΔI_{po} and ΔE_{go} have to be introduced. The potential ratio $\frac{\Delta E_{po}}{\Delta E_{go}}$ may then be divided into the real component $\frac{\Delta E_{po}}{\Delta E_{go}}\cos(\phi-\psi)$ and the imaginary one $j\frac{\Delta E_{po}}{\Delta E_{go}}\sin(\phi-\psi)$:

Equation 3 then becomes the following:

$$\frac{1}{Z_{g_{tot}}}=\frac{1}{Z_g}-j\omega C_{gf}\cdot j\frac{\Delta E_{po}}{\Delta E_{go}}\sin(\phi-\psi)+j\omega C_{gp}\left(1-\frac{\Delta E_{po}}{\Delta E_{go}}\cos(\phi-\psi)\right) \tag{3a}$$

In this formula the total effective stray capacity corresponds to:

$$C_{eff}=C_{gf}+C_{gp}\left(1-\frac{\Delta E_{po}}{\Delta E_{go}}\cos(\phi-\psi)\right) \tag{4}$$

For the effective stray capacity only the real component in the ratio $\frac{\Delta E_{po}}{\Delta E_{go}}$ is of influence.

According to the type of plate circuit load three distinct cases can be distinguished for the reaction on the grid-circuit:

First, Z_b is a pure ohmic resistance,
Secondly, the inductive component in Z_b is dominant,
Thirdly, Z_b is mainly capacitive.

For a pure ohmic load on the plate circuit the maximum value of effective stray capacity results, since the first-mentioned formula is correct in this special case:

$$C_{eff_{tot}}=C_{gf}+C_{gp}(1+A_v) \tag{4a}$$

The condition of a pure ohmic resistance load, however, is only met in practice for the lower frequencies (below about 500 cycles), especially in resistance-coupled amplifiers. Furthermore, it should be noted that even if a pure ohmic resistance is connected in the plate circuit, for radio frequencies the stray capacities already existent according to equation (2) form a considerable capacitive load.

Generally the effective stray capacities should always be calculated according to equation (4); it then is of no consequence

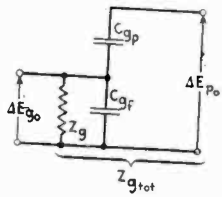

Figure 3

whether the plate circuit has a capacitive or inductive load. It is, however, of importance that, as formula (3a) shows, the imaginary component of the ratio $\dfrac{\Delta E_{po}}{\Delta E_{go}}$ gives rise to a real component in the grid-circuit. This real component is negative in the case of an inductive plate circuit load, and can therefore produce self-oscillation of the amplifier. For a capacitive plate circuit load this component is positive and therefore damps the grid-circuit somewhat.

In dimensioning audio frequency amplifiers with resistance-capacity coupling, it is of primary importance to determine the frequency response toward the higher audio frequencies, where the capacitive load on the plate circuit is no longer negligible[1]. Correct results are then only obtained if the effective values of the stray capacities are calculated according to formula (4). Besides, the real component appearing in the grid-circuit must be taken into consideration as well, as was mentioned above. This real component has, since all the phase-angles in the formulas given are defined with respect to the direction of the grid-potential as a normal, no phase-angle difference from a pure ohmic resistance in the grid-circuit.

This real component thus effects a reduction of the total grid-circuit impedance $Z_{g_{tot}}$ for the higher frequencies; in practice, it therefore is necessary to reckon with slightly smaller grid-leak values in order to take account of this effect. If both effects are taken into consideration: First, the reduction of the effective value

[1] A sufficiently uniform response of the amplifier to impulses of frequencies near the lower limit of audibility may easily be obtained by suitably dimensioning the grid-coupling capacities.

of the stray capacities because of the phase-difference, and, secondly, the reduction of the total grid-circuit impedance due to the real resistance component introduced into the grid circuit at higher frequencies, then the actual frequency response of the amplifier at the higher audio-frequencies turns out to be considerably more uniform than should be expected from formula (4a).

The practical appliance of the theoretical considerations given above can best be shown by means of an example. Fig. (4) shows the simplified circuit diagram of a three-stage resistance-capacity-coupled amplifier, as will be considered given for the calculations. The dimensions of the components correspond to measurements taken by the writer on an amplifier of this type.

$$R_{b1} = R_{b2} = 3 \text{ megohms}$$
$$R_{b3} = r_{p3}$$
$$R_{g2} = R_{g3} = 8.5 \text{ megohms}$$
$$\mu_1 = \mu_2 = 31$$
$$\mu_3 = 5$$
$$r_{p1} = r_{p2} = 0.7 \text{ megohms.}$$

In a normal circuit arrangement the direct inter-electrode tube capacities for German tubes were measured and found to be from

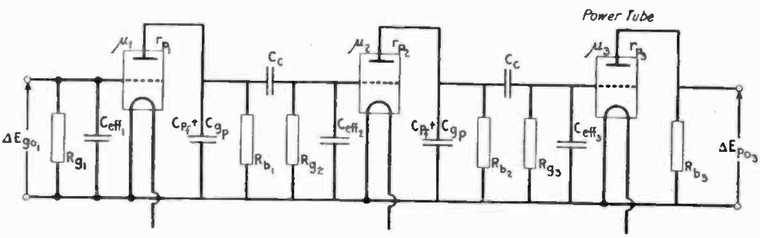

Figure 4

1 to 2 micro-microfarads. In the following a mean value of 1.44 micro-microfarads will be assumed for C_{gp}, C_{gf}, and C_{pf}.

For the simplified assumption of a pure ohmic plate circuit resistance of the power-tube $R_{b3} = r_{p3}$ the following equation exists:

$$\frac{\Delta E_{po3}}{\Delta E_{go3}} = -\frac{\mu}{2} = -2.5$$

The further calculations will be carried out at the frequency of 10,000 cycles which is near the upper limit of audibility.[2] At this frequency the following may be assumed:

$$Z_{b_{tot1}} = Z_{b_{tot2}}; \quad Z_{b_{tot2}} = Z_{g_{tot3}};$$

$$\Delta E_{po1} = \Delta E_{go2}; \quad \Delta E_{po2} = \Delta E_{go3}.$$

[2] At this frequency, the capacitive load of course has the greatest influence.

According to formula (4a): $C_{eff}=1.44+1.44(1+2.5)=6.5$ micro-microfarads. Accordingly, the total effective stray capacities are found to be $C_{eff_{tot}}=C_{pf}+C_{gp}+C_{eff}=9.4$ micro-microfarads. If the factor $j\omega C_{gp}\dfrac{r_r}{\mu}$ is disregarded in comparison to (1), then formula (1a) is changed into:

$$\frac{\Delta E_{po}}{\Delta E_{go}}=-\mu\frac{1}{1+r_p\left(\dfrac{1}{R_b}+\dfrac{1}{R_g}+j\omega C_{eff_{tot}}\right)}$$

After the correct values for the different constants are introduced:

$$\frac{\Delta E_{po2}}{\Delta E_{go2}}=-31\frac{1}{1+0.32+j\,0.41}.$$

$$\frac{\Delta E_{po2}}{\Delta E_{go2}}\cos(\phi-\psi)=-21.6;\quad \frac{\Delta E_{po2}}{\Delta E_{go2}}\sin(\phi-\psi)=6.8;\quad A_{v2}=22.6.$$

The static amplification in this stage is $A_v=23.8$; the dynamic amplification therefore is practically equal to the static even at the high frequency of 10,000 cycles. The frequency response in this second stage therefore may be assumed to be uniform. After examining the curvature of the dynamic working characteristic it will be found that here too the amplitude-distortion is negligible, which is of primary importance in this stage because the amplitude is large.

For the first stage the following values must be assumed:

$$C_{eff_{tot}}=3\times 1.44+1.44\times(1+22.6)=38.5 \text{ micro-microfarads}$$

$$\frac{1}{R_{g_{tot1}}}=\left(\frac{1}{8.5}+0.62\right)10^{-6}=0.715\times 10^{-6}$$

Just as above, the following values result:

$$\frac{\Delta E_{po1}}{\Delta E_{go1}}\cos(\phi-\psi)=-9.6;\quad \frac{\Delta E_{po1}}{\Delta E_{go1}}\sin(\phi-\psi)=8.9;\quad A_{v1}=13.1.$$

In spite of the fact that the phase-difference $(\phi-\psi)$ is rather large in this stage (about 140 deg.), the only pure resistance-capacity coupled stage, the dynamic amplification for the frequency of 10,000 cycles does not drop even to one-half of the static amplification. The same of course applies to the three-stage amplifier as a whole:

$$A_{vst_{tot}} = 1420$$

$$A_{vdyn_{10000}} = 730 = 0.51 A_{vst_{tot}}.$$

These voltage amplification factors given here are those determined by measurement of the input alternating grid potential and of the output alternating potential across the output resistance. The amplification factors calculated here practically exactly comply with measurements taken on a three-stage amplifier of the same dimensions. According to these measurements the dynamic amplification at the frequency of 10,000 cycles falls off to 0.516 times the maximum amplification. Therefore, the effective stray capacity values calculated here may be assumed to be practically equal to those actually present.

As the above calculations have shown, the actual amplification factors of such resistance-capacity-coupled amplifiers and all other amplifiers with appreciable reaction of the plate-circuit load on the preceding grid-circuit can only then be correctly calculated, *if the different stages are calculated proceeding backward, that is, beginning with the plate circuit of the last stage.* Wrong results are obtained, if founded on some assumptions one stage only is examined, and the result gained simply multiplied by the number of stages involved, as in the case of the frequency-response calculation of a three-stage amplifier with resistance coupling. As has been shown by the calculations carried out above, with resistance-capacity-coupled audio-frequency amplifiers the frequency response of the *first stage only* is of material importance. Three-stage resistance-coupled amplifiers of these dimensions with such high amplification factors, however, are seldom used as pure audio-frequency amplifiers. As a rule the grid of the first stage is directly connected to a tuned radio-frequency circuit, which is then coupled either to the antenna or to the output of a radio-frequency amplifier. By the influence of the considerable capacitive load on the plate-circuit of the first tube, a very sensitive "anode-bend" detection effect for radio frequency only is then introduced.[3] Two-stage amplifiers of this description, containing one voltage-amplification stage followed by a power-tube, have a practically complete uniform frequency-response, as is apparent from the calculations given above.

[3] Compare the first writer's publication "Über Anodengleichrichtung," Jahrbuch der drahtlosen Telegraphie und Telephonie, Zeitschrift für Hochfrequenztechnik, Berlin. Vol. 29. No. 3.

MOUNTING QUARTZ OSCILLATOR CRYSTALS*

By

R. C. Hitchcock

(Research Department, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania)

Quartz oscillators are becoming more widely used to control the frequencies of radio stations. Under fairly constant temperature conditions, and with good regulation of filament current and plate voltage on the controlled tube, a quartz master oscillator may readily be kept to within 50 parts in a million of an assigned frequency.

There are several methods of cutting a piezo-electric plate from a quartz crystal. Fig. 1 shows a right prism of quartz with the optic axis perpendicular to the bases. The usual orientation[1] for a quartz piezo-electric plate is that of method 2, giving a thickness wavelength of 105 to 115 meters per millimeter. The crystals used

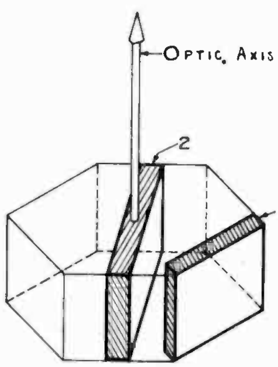

Figure 1—Methods of cutting piezo-electric plates.

for all the tests to be described in this article are cut according to method 1, having a thickness wavelength of 140 to 150 meters per millimeter. For a given wavelength, the crystal of method 1 is two-thirds as thick as a crystal of method 2. For some wavelengths, the thinner crystal of method 1 has definite advantages, due to the stronger electric field which is applied.

* Original manuscript received by the Institute, June 21, 1927.

[1] A. Hund: Uses and Possibilities of Piezo-electric Oscillators. Proc. I.R.E. Aug. 1926, p. 448. A. Crossley: Piezo-electric Crystal-controlled Transmitters. Proc. I.R.E. Jan. 1927, p. 13.

Of the several factors which govern the action of a piezo-electric crystal when controlling a radio-frequency circuit, the mounting of the crystal is quite important. It will be assumed that the circuit is properly designed, that a good piece of piezo-electric material is available, and that the only variable to be considered is the mounting of the crystal. The main feature of a crystal-controlled master oscillator is its ability to regulate a radio frequency very closely. In addition, it is quite desirable to have the crystal-controlled circuit radiate as much power as possible, in order to reduce the number of radio-frequency power amplifiers.

In using crystal-controlled tubes radiating more than a few watts, several problems arose which did not occur when the power was only a fraction of a watt. The circuit of Fig. 2 is a typical crystal-controlled circuit in which the grid bias is supplied through a

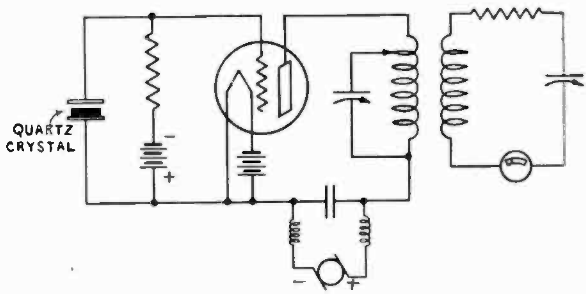

Figure 2—Piezo oscillator with power-absorbing circuit.

grid-leak, this being found to give better control of the radio frequency than when a choke coil was used. In addition to the crystal circuit, an absorbing circuit is used (see Fig. 2 at the right) to measure the amount of power radiated. Using this circuit the various mountings were tried and their effect on the output determined. The frequency characteristics were found by using a crystal-controlled receiving tube circuit coupled very loosely to the power circuit which was being tested. The crystal which controlled the receiving tube circuit was ground to give a frequency several hundred cycles above or below the power crystal circuit, so that the audio beat could be heard by receivers which were inserted in the receiving tube plate circuit.

When the crystal-controlled circuit was arranged so that as much as 10 watts could be absorbed, it was found that the piezo-electric reaction of the crystal was so great as to cause sparking at the electrodes. The sparking caused electrode corrosion which soon interfered with the free action of the crystal, and also spoiled the clear-cut character of the radio-frequency output. High power also caused the crystal to heat badly, often resulting in cracking, which of course stopped the oscillations.

In order to compare directly the various types of mountings the frequencies of the crystals were ground to less than 3 percent of one million cycles per second (300 meters wavelength). From a survey of mountings for crystals of this frequency, the possibilities at other frequencies can be considered.

All mountings for piezo-crystals consist of electrodes which are smooth and plane on the side next to the crystal. One of the simplest mountings is that of a flat metal plate on which the crystal rests, and an upper plate resting on top of the crystal, (see Fig. 3). This mounting as shown is one in which the upper plate rests on the crystal, the only pressure being its weight, while others use a spring to press the upper electrode more firmly on the

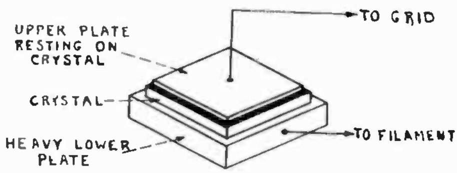

Figure 3

crystal. At high radio frequencies, perhaps above two million cycles per second (less than 150 meters wavelength) the simple type of mounting shown in Fig. 3 is fairly satisfactory.

For frequencies in the vicinity of one million cycles another type of mounting, in which the upper electrode is supported a definite distance away from the crystal, has certain advantages which the simpler mounting does not have. The disadvantages of the simpler holder for frequencies of the order of a million cycles per second are shown when a large amount of power is taken from the crystal-controlled circuit. An ordinary crystal-controlled UX-210 tube (7-1/2 watt rating) can radiate 5 to 7 watts at a million cycles with nearly any type of mounting. When using a UV-211 tube (rated at 50 watts) with 1,000 volts on the plate and absorbing 50 to 80 watts from the plate circuit, crystals were often broken when using the above simple method of upper electrode resting on the crystal. Using the spaced mounting, in an oil bath at constant temperature, which will be described later, tests of several hours were run successfully at outputs of 50 to 80 watts without breaking the crystal.

Spaced Upper Electrode

When mounting the upper electrode a fixed distance away from the crystal, it is essential to have the electrodes parallel. If the electrodes were not parallel, the electric field would be unevenly distributed, tending to strain the crystal. Various mountings were constructed which provided a space above the crystal, yet which were simple in construction. One type, which proved useful for several sizes of crystals, is shown in Figs. 4 and 5. In this mounting an accurately turned hard rubber ring spaces the upper electrode a small distance away from the crystal, and when a different crystal

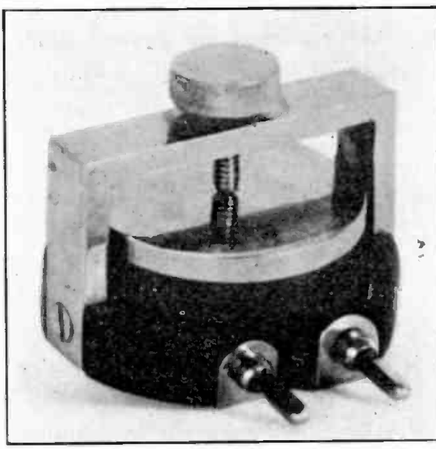

Figure 4

is to be used an appropriate ring is substituted in the holder. A definite spacing which could always be duplicated was afforded by keeping the ring and its crystal in the same envelop.

The crystal vibrations are not damped by the pressure of the upper plate in the spaced type of mounting; this accounts for the increased power which can be obtained. A further advantage of a spaced holder is that the intensity of the electrostatic field across the crystal is reduced, due to what is effectively a series condenser. This reduced field helps prevent breakage of the crystal due to dielectric failure of the quartz.

There is a certain phenomenon which occurs when using a spaced upper electrode which may give trouble if its characteristics are not known. The position of the upper electrode for maximum power and accurate frequency control is somewhat critical. If a holder with an adjustable upper electrode is available, an interesting experiment can be performed. If the upper electrode is continuously moved away from the crystal, the radio frequency emitted by the crystal-controlled circuit is raised, but not at a uniform rate. There will be successive points where the frequency will "jump" a few hundred cycles, or occasionally as much as a few thousand cycles per second. If the power is measured at the same time that the frequency is being investigated, it will be found to be a minimum at the same points where the frequency jumps occurred. This test would make it seem that the use of a spaced upper electrode was undesirable, but there is a simple explanation and solution to this phenomenon. The crystal when vibrating strongly sets up compressional air waves, which when reflected back from the upper plate in certain definite positions exert a

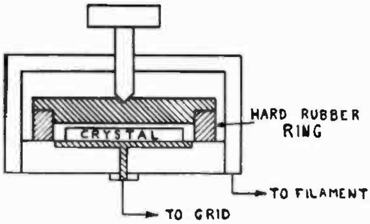

Figure 5

damping effect on the crystal. These are simply standing waves in air. For a first approximation it can be assumed that the speed of this air wave is the same as that of an ordinary sound wave in air. The air wavelength in millimeters would then be equal to the speed of sound in meters per second divided by the frequency in kilocycles per second. That is

$$\lambda_{mm} = \frac{331_{m/sec}}{F_{kc/sec}}$$

It was found that the assumption of the ordinary speed of sound in air did not give the correct distance of air wavelength for minimum power and frequency shift. In fact several experimental curves showed that the speed of the sound wave emanating from a vibrating crystal was nearly twice that of an ordinary sound wave. This makes the wavelength twice as long as when using the usual speed of sound in air, and a more correct formula is

$$\lambda_{mm} = \frac{662_{m/sec}}{F_{kc/sec}}$$

It should be noted that this equation is simply a further approximation, and that the speed of sound is really a function of the energy with which the crystal vibrates and varies with the amount of power radiated. The reason for the faster speed of propagation is to be found in the very energetic crystal vibration when radiating large amounts of power from the controlled circuit. It is well known that explosive waves travel faster than ordinary sound waves which have much smaller amplitudes.[2] A strongly vibrating quartz oscillator crystal can be regarded as a source of rapidly recurring explosive waves, therefore having a velocity greater than ordinary sound waves.

A simple experiment with a screw top holder serves to locate the correct position of the upper electrode for any piezo-electric

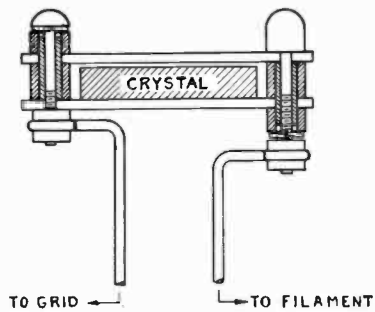

Figure 6—Piezo crystal mounting for low power.

crystal. An absorbing circuit is coupled to the crystal-controlled circuit, and as the upper electrode is raised, the first point of minimum power is noted. The best operating position of the upper electrode is to be found between the upper surface of the crystal and the first position of minimum power. For crystals vibrating at a million cycles per second the spacing of the upper electrode above the crystal can be from two-tenths to four-tenths of a millimeter, the smaller value being preferable.

In mounting electrode plates a fixed distance apart, a reliable spacer should be used. Also, due to the fact that the assembly for permanent use has to be mounted inside a glass bulb, it is seen that any material which contains gases would not be satisfactory. This excludes the use of molded materials such as micarta and bakelite. The heat which is applied to outgas the crystal electrodes is sufficient to rule out other materials which become plastic at

[2] The Detonation Wave from Solid Explosives: W. C. Holmes, *Jour. Fr. Inst.* Apr. 1927, p. 549.

fairly low temperatures, for instance, hard rubber. In the assembly which is shown in Figs. 6, 7, and 8, the spacers are made of fused quartz tubing. This material has a very low coefficient of expansion, contains no more gases than the glass of the bulb itself, and does not become plastic at ordinary temperatures. Sections of glass tubing can also be used for spacers, although their coefficient of expansion is much higher than for fused quartz. The feature of low expansion coefficient is not of such importance where the

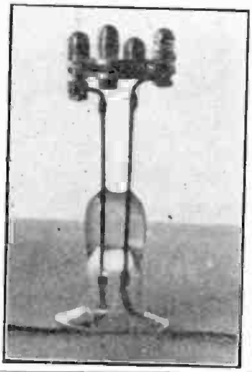

Figure 7

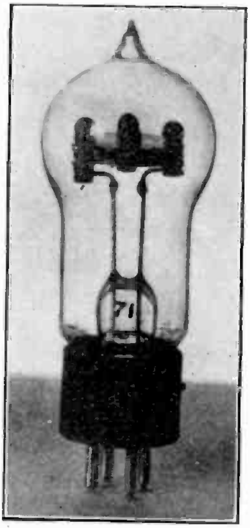

Figure 8

crystal mounting is immersed in a temperature-controlled bath. The oil bath will be described later in detail.

Electrode Material

Having determined the best spacing of the upper electrode above the crystal, the next problem was to find the best material of which to make the electrode plates. When large amounts of power were absorbed from the crystal-controlled circuit, the periodic reaction of the crystal produced such large potential differences between the upper face of the crystal and the lower surface of the upper electrode, that sparks passed between them, as mentioned at the beginning of this article. These sparks are detrimental in two ways. In the first place they are strongly oxidizing sparks which rapidly corrode the electrode plates. Using brass electrodes, it was found necessary to clean off the corrosion nearly every day when a crystal was in continuous operation. The other objectionable feature is that air is a fairly good insulator; the spark does not pass until the potential difference has become quite large, and as the spark leaves it creates a disturbance in the form of a transient vibration of the crystal, which is added to the emitted radio frequency. When a crystal plate is sparking, the rough pitch of the sound can be heard by listening near the crystal holder, and this pitch is the same as that which is added to the radio frequency, making it unsuitable for master oscillator control of radio stations. To reduce the effect of oxidizing the electrodes, metals which were somewhat resistant were adopted, such as

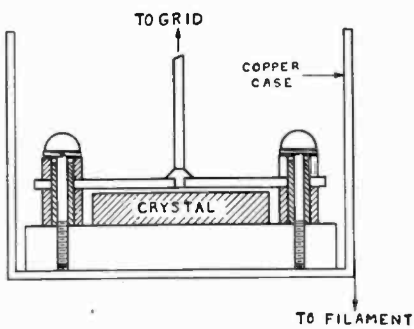

Figure 9—Piezo crystal mounting for high power.

monel metal, pure nickel, and stainless iron. Of these three, pure nickel is perhaps the easiest to machine and makes a very good crystal electrode.

There are two methods of procedure which can be used to prevent sparking. The original procedure was to reduce the plate voltage of the crystal-controlled tube. The output power was thereby reduced, and in turn the voltage feedback through the tube was diminished, which eliminated the sparking at the crystal. The disadvantage was obvious; the number of radio amplifier stages had to be increased to compensate for the decreased power in the crystal-controlled tube circuit.

Glow Discharge at Crystal

The second method is the one which is now being used successfully for control crystals at the radio station KDKA. This method substitutes a silent glow discharge for the rough spark. The glow discharge takes place at much lower potential differences than the spark, and accomplishes the desired result. A glow discharge is best accomplished by using some gas which ionizes readily, which means that the crystal holder must be sealed, as in Fig. 8. The glow discharge is preferably carried by an inert gas, so that it will not attack the electrode plates when in its ionized state. Neon is an inert gas which permits glow discharges at very low potential differences, and a pressure of a few millimeters of neon has proved satisfactory for quartz oscillator plates.

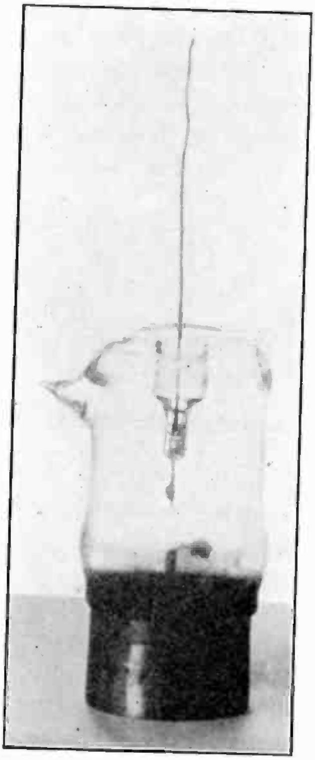

Figure 10

UX-210 tubes with 400 volts on the plate can radiate 5 to 7 watts of clear-cut radio-frequency output when using a crystal plate mounted in a neon-filled tube. Under the same conditions of clear radio output, the same crystal in air would only be able to supply one or two watts and the tube would have to be operated at a reduced plate voltage of 200 or 300 volts.

When piezo crystals mounted in neon are controlling power circuits, there can be seen a pink glow surrounding the crystal. If the plate circuit is tuned to a frequency much higher than that of the crystal, a brilliant glow will fill the entire crystal tube. As the frequency of the plate circuit is lowered, the glow becomes more concentrated, until when the plate circuit is tuned to its best point (slightly higher than the crystal frequency) the crystal itself seems to be permeated with the pink glow, but there is very little glow in the rest of the tube.

The effect of varying temperature is a very important one for quartz plates. Depending on the method of cutting the quartz plate, the frequency-temperature coefficient is from 25 to 50 parts in a million per centigrade degree. For quartz plates oscillating in the thickness direction the latter figure should be used. For a constant radio frequency the temperature of the quartz plate should be kept as uniform as possible. Quartz crystals are not symmetrical; it is due to this fact that they are piezo-electric. This asymmetry is a disadvantage in that the crystal when heated expands unevenly in its thickness and length directions. It expands very nearly twice as fast in the thickness direction as it does in one of its longer dimensions. Any use of a crystal to control more than 10 watts causes so much dielectric displacement current to flow through the crystal that it heats up. If the crystal is operated continuously without cooling the heat finally causes such large stresses that the crystal cracks. Sometimes only a corner cracks and again the crack may occur through the middle of the crystal. If only the corner has been cracked, the crystal can be made to oscillate again by carefully grinding away all the portion that shows any cracks. If a large crystal cracks in the middle, it may be cut into smaller plates which are all right for radio oscillators. In other words, the failure seems to be a local one and the rest of the crystal is unaffected.

Cooling Crystals for Power Work

A means of cooling the crystal is necessary when continuous operation at high powers is desired. The holder which is used to do this is shown in Figs. 9 and 10. This mounting retains the features of the glass bulb type, in that the upper electrode is spaced above the crystal, and an inert gas at low pressure is used for a glow discharge. By using direct metallic conduction from the lower electrode to the copper case, which is sealed to a glass superstructure, the cooling is easily effected by immersion in a tank of oil kept at the desired temperature. Fig. 11 shows the tank with the crystal unit immersed, and Fig. 12 gives the connections for the thermostat T, the heating element R, also the lamp L and condenser C which serve to protect the thermostat contacts on breaking the circuit. For a 3-gallon oil tank, R was 85 ohms, and L was a 100-watt tungsten lamp. The working temperature of the crystal unit is chosen to be above that of the highest summer room temperature. The temperature chosen for a series of tests was 58 deg. C. A good bimetallic thermostat of the helical type

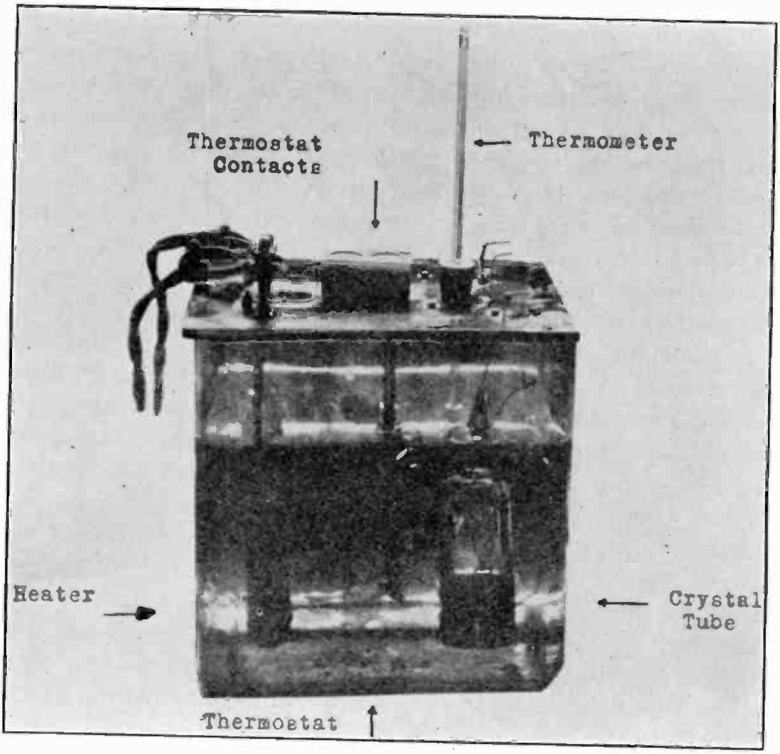

Figure 11

was found to regulate the temperature of this 3-gallon oil bath within two-tenths of a centigrade degree. This variation should keep a piezo-electric quartz crystal constant to better than 10 parts in a million. Frequency differences are difficult to measure to better than 20 cycles when using a million cycles per second, but it can be said definitely that the control is better than 20 parts in a million.

It should be noted that both the heater and the crystal are placed below the center of the oil bath. This is to insure adequate cooling of the crystal and to provide good circulation in the case of the heater. If the crystal heats up above the oil temperature, the thermostat cuts off the major part of the current through the heater, and the coolness of the room serves to bring the oil temperature down. If the crystal does not heat up very much, the regular heater is turned on and off by the thermostat, in keeping the oil bath at the desired temperature.

Using a crystal unit in the oil bath mentioned, and feeding into the grid a UV-211 tube using the circuit of Fig. 2, and applying 1000 volts on the plate, a test of 8 hours was run while absorbing 50 watts in the auxiliary circuit. Longer tests would be illuminating and are in progress, but it is felt that if there were a flaw in this manner of operating a crystal, that it would show up

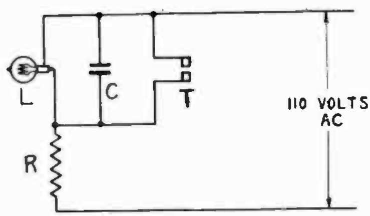

Figure 12—Thermostat and heater connections.

in a severe test of the kind just described. The temperature-controlled oil bath can also be used for crystals mounted as in Fig. 8 which are not used primarily for maximum power output.

For crystals of natural thickness frequencies of about one million cycles per second, a spaced electrode holder, mounted in an inert gas, and properly cooled, is recommended for power work. For crystals of higher frequencies, there may be some difficulty in making the spacing of the upper electrode as small as is necessary to avoid the resonant air wave. In this case, it might prove advisable to use a harmonic of a lower frequency having a large power output. For crystals of lower frequencies than a million per second there should be no trouble in using exactly similar mountings to those of Figs. 6 and 9, with a small distance between the crystal and upper electrode.

Summary

Various types of mountings are described for quartz oscillator crystals having frequencies in the vicinity of one million cycles per second. For low power work a spaced upper electrode is recommended, mounted in a neon-filled tube. For radiating up to 50 watts from the crystal-controlled tube, a similar mounting placed in a copper-based tube is advised. This latter tube can be cooled by immersion in an oil bath. Accurate temperature control of the oil bath keeps the radio frequency within narrow limits.

THE SHORT WAVE† LIMIT OF VACUUM TUBE OSCILLATORS*

By

C. R. Englund

(Bell Telephone Laboratories, Inc.)

The following paper contains the results obtained while investigating the practical short wavelength limit of the vacuum tube oscillator. While considerable work at short wavelengths has been published by various investigators, no conclusive data determining the practical short wave limit have come to the author's notice.

Short waves may be obtained from vacuum tubes in at least two quite distinct manners, the first being the orthodox oscillator

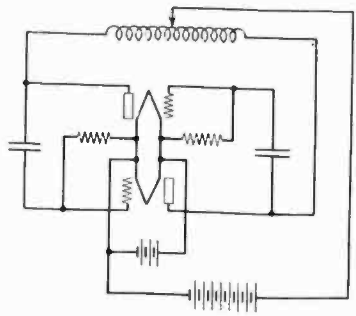

Figure 1

operation where a coupling between grid and plate circuits by regeneration builds up resonance oscillations of frequencies determined by these circuit constants, and the second (the Barkhausen type) being an oscillation within the tube itself in the electron atmosphere between grid and plate. The frequencies of the second type of oscillation are apparently determined by the time of transit of an electron between the tube electrodes and are thus dependent upon the electrode voltages. Little power has so far been obtained in the second manner. The work here reported refers to the first type or normal oscillator operation.

It was found that shorter waves could be obtained by means of a double tube symmetrical oscillator, than could be got by a single oscillator, due to the fact that the total length of conductor carry-

* Original manuscript received by the Institute, July 24, 1927.

† In the present state of the short wave art, it is a distinct inconvenience to use "frequency" rather than "wavelength," since all the measurements are made in terms of lengths.

ing oscillating currents could be reduced, in the double tube connection, below what could be reached by means of a single tube. For wavelengths above five meters the tubes available in commercial use were found adequate, and except for the limitations imposed by the necessity for low resistance condensers and inductances of very small values, the oscillator technique differed little from that at considerably longer waves. Small inductances of low resistance at these wavelengths require considerable copper, but commercial tubing has been found adequate.[1] This paper therefore concerns only the range below five meters.

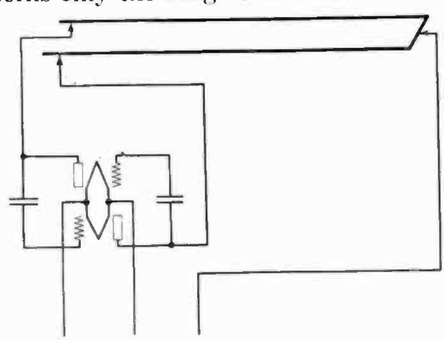

Figure 2

In a cursory consideration of experimental procedure it might seem that the connection of a vacuum tube to a "Lecher System" of parallel conductors offers greater possibilities of reaching shorter waves than does the use of lumped circuits. This has not been borne out by our experience. The first type of circuit tried was a symmetrical connection of standard *215-A* tubes (*N* tubes) Fig. 1, connected to a quarter wave Lecher system as shown in Fig. 2. Results showed that the waves actually obtained were much longer than four times the length of the Lecher system. The following simple theory, while not exact, is adequate to explain the results.

The impedance of a Lecher wire system of length l and shorted at the far end, is given by the expression

$$S_o = \sqrt{\frac{Z}{D}} \tanh \sqrt{DZ}\, l$$

where $Z = R + i\omega L$ is the complex series impedance per unit length and $D = A + i\omega C$ the complex shunt admittance per unit length

[1] The ordinary 1/8 in. (0.318 cm.) commercial copper tubing was chiefly used here.

of the system. Since the tube reactances are capacities we may say that oscillations will occur when the Lecher wire impedance is an inductance and these oscillations will be approximately of the frequency given by $\omega^2 L'K = 1$ where L' is the effective inductance of the Lecher wires and K the capacity of the tube electrodes in shunt on the Lecher wires. Or

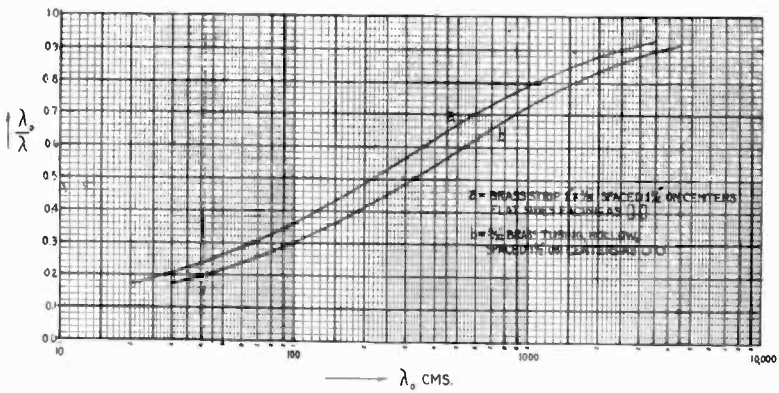

Figure 3

$$\text{imag. part of}\left[\sqrt{\frac{R+i\omega L}{A+i\omega C}}\tanh\sqrt{(R+i\omega L)(A+i\omega C)}\,l+\frac{1}{i\omega K}\right]=0$$

is the resonance condition. Neglecting R and A, since approximate values will suffice, we have the transcendental equation

$\sqrt{\frac{L}{C}}\tan\omega\sqrt{LC}\,l=\frac{1}{\omega K}$ or if λ be the true wavelength obtained and $\lambda_0 = 4l$, $X=\frac{\pi}{2}\times\frac{\lambda_0}{\lambda}$ we have $X\tan X=\frac{C\lambda_0}{4K}$ which may be readily solved graphically. In order that $\lambda=\lambda_0$ we must have $\frac{Cl}{K}=\infty$ or only at relatively long wavelengths does the ratio $\frac{\lambda_0}{\lambda}$ (for given tubes) approach unity. If $\frac{\lambda_0}{\lambda}$ be determined experimentally for a given wavelength it can be calculated for other wavelengths and Fig. 3 gives two curves thus obtained for two Lecher systems, markedly different in shunt capacity. No essential distinction can be drawn between such Lecher systems and a lumped inductance at the shorter wavelengths; it is only at longer wavelengths that space wave resonance can occur. Naturally the addition to the Lecher system of a length of additional wire equal to half a wavelength (or 1/4 wavelength if open circuited at the end) need not affect the wavelength generated and, conversely, by damping out the fundamental wave by means of appropriate shunts at the current antinodes harmonic oscillations should be produced. The additional energy loss is apparently pronounced, however, for

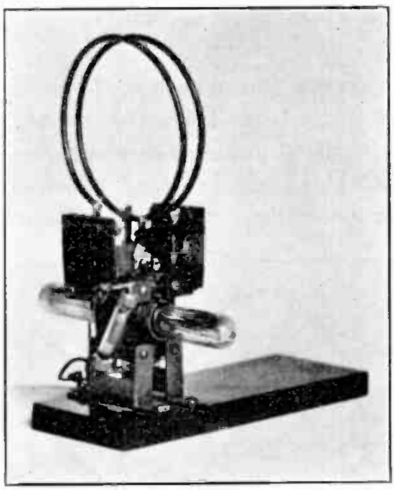

Figure 4—4.2-meter wavelength

while it has been possible to produce such harmonic waves, only the fundamental occurs without shunts. A little consideration will show that no harmonic wavelength is obtainable which is not more easily gotten as the fundamental of a shorter length of line. Accordingly, no further attempts to excite Lecher systems by means of *direct* connected oscillators were made and simple inductances were solely used for the remaining work.

Fig. 4 is a photograph of a balanced *215-A* tube oscillator using the circuit of Fig. 1 and giving a wavelength of 4.2 meters. By reducing the size of the coil this wavelength was considerably reduced. The tube capacity was the only tuning capacity used. Realizing that too much metal was used in this oscillator it was rebuilt with very compact blocking condensers of tinfoil and paraffined paper and a wavelength of 1.9 meters was reached.

In casting about for some scheme of connecting the tubes so as to reduce their effective capacity the circuit of Fig. 5 was devised and tested. It operated well, usually without a grid leak and could be pushed down to 1.67 meters with ordinary based

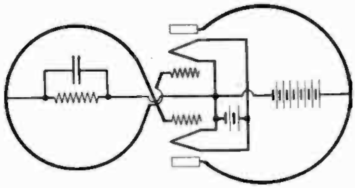

Figure 5

215-A tubes. Fig. 6 is a photograph of the actual oscillator. The two coils (each a single turn) had to be connected geometrically as shown; if their mutual inductance was reversed no oscillations occurred. The lead in wires in *215-A* tubes are about as short as it is possible to make them, and when unbased tubes were used,

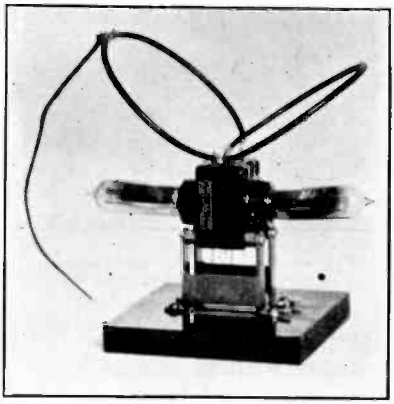

Figure 6—1.67 meters. $E_b = 180$ volts.

as in Fig. 7, the wavelength was reduced only to 1.42 meters.

Since this represented the limit attainable with commercial *215-A* tubes attempts were made to push down in wavelength by increasing plate voltage and filament current. A plate voltage increase is of no use alone if the filament does not have a corresponding emission of electrons, in fact a very active filament is the backbone of a short wave oscillator. The limit is reached when the plate overheats. No marked shortening in obtainable wavelength occurred by pushing the plate current per tube to 17 milliamperes momentarily (E_b app. 200 volts, normal I_b 2 milliamps.) and had such shortening been obtained it would not have been of practical value.

It therefore appeared that the only prospect of further reduction in wavelength was in the construction of a special tube

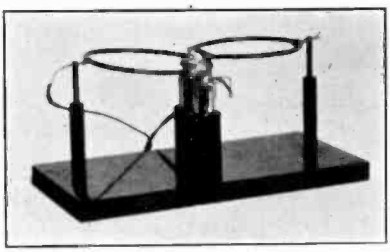

Figure 7—1.42 meters. $E_b = 180$ volts.

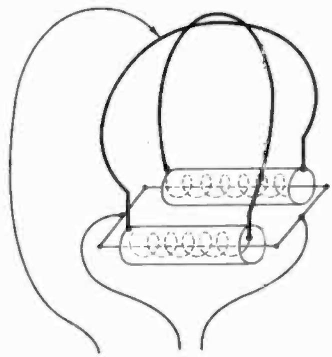

Figure 8

and the assembly adopted was that shown schematically in Fig. 8. Five of these tubes were tested, the last one with doubled leads inside the tube and one-ampere filaments. With this a wavelength of 1.05 meters was finally reached, as a rather unstable

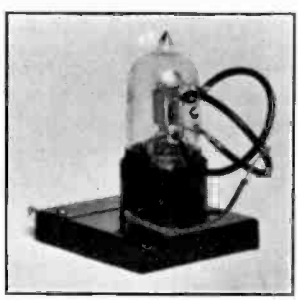

Figure 9—1.05 meters. $E_b = 180$ volts.

oscillation, by pushing the tubes hard. Fig. 9 is from a photograph of the complete oscillator. At two meters the operation of these tubes was quite satisfactory and Fig. 10 shows such an oscillator arranged to vary the frequency by changing the length of the oscillating coils through sliding contactors. The circuit is the same, electrically, as that of Fig. 8.

As is evident from Fig. 9 the physical dimensions have been reduced to the point where it is necessary, for further progress, to put the resonant circuits inside the tube itself. The inter-electrode capacity of such a tube would have to be kept very small also, all further experiments having indicated that this is more important than low plate impedance. So far no limitations except those of

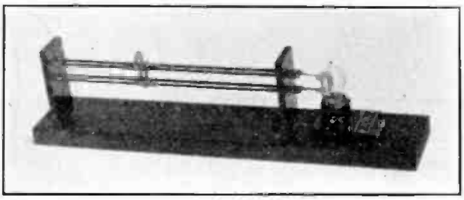

Figure 10—2 meters.

ordinary energy loss have been apparent. But to operate below one meter a tube must have the properties of a low loss condenser with terminal leads of very low inductance. Unnecessary inclusion of glass dielectric and the use of nickel electrodes are inadvisable and the tube geometry must be rigidly worked out. The construction of such a tube is necessarily a double assembly to keep the leads short, and the wavelength is not variable.

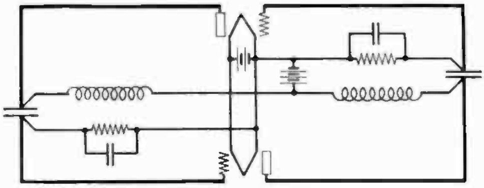

Figure 11

At the shortest wavelength obtained a grid leak stopped the oscillations, at longer wavelengths, say three meters, the oscillations were so vigorous that if a leak was used it had to be a low resistance one or the tube "blocked" at a rate varying from 4 to 10 kilocycles per second. The oscillating current during blocking was very rich in harmonics, and these could be demonstrated as of wavelengths well below one meter. If necessary, therefore, the use of harmonics will enable the production of small amounts of energy at wavelengths below one meter and it is possible that this will be technically useful sometime. It might appear that such harmonics could be amplified after selection but apparently the technique which will produce amplifier tubes will give oscillator tubes, therefore the need for amplification brings the problem back to the oscillator itself. Possibly a square law detector pushed hard with an input from a short wave oscillator will make a useful source of shorter waves.

Another double-tube oscillator circuit which seems feasible is given in Fig. 11. It is complicated by requiring blocking condensers and for the shortest waves needs a different geometrical arrangement from that in Fig. 8. It was tried once and not giving any apparent improvement over the circuit of Fig. 8 was not used further.

The one-meter wave not being controllable, it is pertinent to determine what wavelength limit can be reached with reasonable

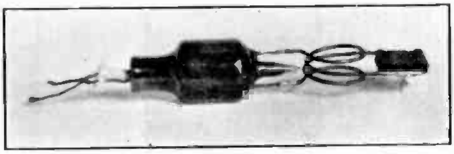

Figure 12—3.1 meters. $E_b = 700$ volts.

power and some frequency control, with the present vacuum tube technique. With the double *215-A* tube a two-meter wave is available with frequency control, with a reservation to be referred to later. While the power thus available is small, it is sufficient for wavelength, capacity, and inductance measurements, a subject which will be deferred for the moment. Some rough tests on commercial tubes, operated at normal voltages are added in the table below.[2] These values are not necessarily the shortest attainable but represent wavelengths which can be obtained with reasonable ease.

Tube	Connection	Meters
230-D (60 mil. fil.)	push pull	2.0
205-D ("E" 5 watt)	single (200 volts E_b)	3.2
221-D (1/4 amp. fil.)	push pull	3.3
211-D ("G" 5 watt)	push pull	3.5

No frequency control was left at these wavelengths. It therefore appeared possible to design a *5-watt* tube which should have some frequency control left at 3.5 meters and a photograph of such a tube is shown in Fig. 12. With the oscillating circuit shown

[2] A recent measurement has given Radio Corporation UX-852, single tube, 1,000 volts 2.6 meters.

attached, the wavelength is 3.1 meters, and by means of a small variable shunting capacity the wavelength can be controlled within a reasonable range. With such a tube (at 3.66 meters) it has been possible to transmit, without noticeable attenuation, three watts for a distance of 5.2 meters (the room would not allow a longer distance) over a brass pipe transmission line, and radiation experiments, using a simple vacuum tube voltmeter as receiver, have been made up to a distance of 5 1/2 wavelengths. At a four-meter wavelength, reception up to distances of a mile has been possible with a double detection receiver.[3] It seems certain, therefore, that wavelengths down to 3.5 meters are readily available for engineering

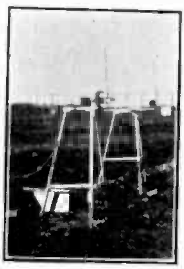

Figure 13

Figure 14

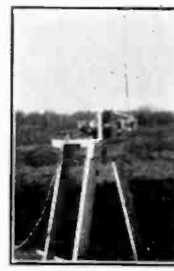

Figure 15

purposes. Figs. 13, 14, and 15 show transmitter, receiver, and transmitter reflector antenna systems operating at 3.66 meters. It is a curious fact, possibly worth mentioning, that the human body forms a very fair Hertzian resonator, and at 3.66 meters is so nearly in tune with the radiation that an operator becomes a mobile parasitic antenna seriously interfering with radiation experiments.

The results given above are somewhat at variance with the data published by a number of investigators[4] since only by the use of special tubes was it possible to shorten the wavelength below that reported for (apparently) ordinary tubes by these investigators. Especial care was therefore taken to make certain that the values here reported are fundamental wavelengths and not harmonics of a fundamental but unobserved oscillation. It is not difficult to assure certainty in this matter; it is merely necessary to have a resonant Lecher system which is longer than a full wavelength and which has a low attenuation. Brass tubes of

[3] I am indebted to Mr. H. C. Baumann of this laboratory for the construction of this receiver.

[4] See list of references attached.

0.8 cm. diameter spaced approximately 7 cm. give such a system and with this it should be difficult to mistake harmonic minima for the practical zeros of the fundamental. Various Lecher frames have been made and compared, both as open end and short circuit end frames. The check between $\lambda/4$ and $\lambda/2$ frames was always found good to 0.2 percent although it would seem that a short circuit end, when in the form of a plate, would be more nearly ideal than an ordinary open end. Figs. 16 and 17 show two short and adjustable $\lambda/4$ frames used, one having a meter in it. The presence of the meter necessitates a small correction. The method of using these frames was found to be very simple. For the shorter

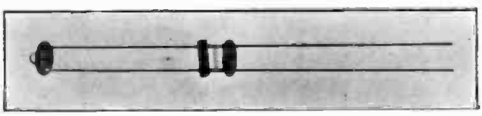

Figure 16—Range 1.2–3.2 meters.

wavelengths it was sufficient to place one end of the wave frame near the oscillator and adjust by observing the tube plate current. At space resonance the energy absorption from the tube circuit was so marked that it could not fail of observation and would stop the oscillation when increased. For longer waves where more power was available a meter in the Lecher frame could be used for the rough setting and for the fine setting a sensitive thermocouple and micro-ammeter connected to a coil of several turns of wire could be used as a loosely coupled exploring unit to indicate maxima and minima on the frame while its length was being varied. Almost equally sensitive is a vacuum tube voltmeter with tube mounted in the handle holding the wire coil. If the experiment is one involving radiation, a Lecher frame in proximity to a sharply tuned receiving antenna and detector gives very distinct and accurate wavelength settings.

The measurement of small inductances and capacities by means of Lecher frames has been mentioned. The theory, when great accuracy is not required, is simple. Thus, suppose a $\frac{\lambda}{4}$ Lecher frame be cut off a distance X from the closed end, the length of open line $\frac{\lambda}{4}-X$ being rejected. Obviously, this frame will be again in resonance to the wavelength λ if an impedance equal to the effective impedance of the piece cut off is placed across the open end. The impedance of a short length of Lecher frame with open end is

$$S_\infty = \frac{\sqrt{\frac{Z}{D}}}{\tanh\sqrt{DZ}\left(\frac{\lambda}{4} - X\right)}$$

and if we neglect the resistance and conductance of the frame

$$S_\infty = \frac{\sqrt{\frac{L}{C}}}{i\tan\omega\sqrt{LC}\left(\frac{\lambda}{4} - X\right)}$$

or as long as $\frac{X}{\lambda} < \frac{1}{4}$ the impedance is capacitative. Hence,

$$\frac{1}{i\omega K} = S_\infty = \frac{\sqrt{\frac{L}{C}}}{i\tan\omega\sqrt{LC}\left(\frac{\lambda}{4} - X\right)}$$

and

$$K = C\left(\frac{\lambda}{4} - X\right)\frac{\tan\omega\sqrt{LC}\left(\frac{\lambda}{4} - X\right)}{\omega\sqrt{LC}\left(\frac{\lambda}{4} - X\right)}$$

is the capacity necessary to reterminate the Lecher frame for space resonance. For small values of $\left(\frac{\lambda}{4} - X\right)$,

$$K = C\left(\frac{\lambda}{4} - X\right)$$

a very convenient expression. When $X > \frac{\lambda}{4}$, or the wave frame has been lengthened instead of shortened, K becomes negative and the impedance which will effectively retune the wave frame to a $\frac{\lambda}{4}$ length is an inductance. This is a particularly neat method of adjusting choke ciols for short wave work, as was experimentally verified.

A reservation was mentioned in connection with the two-meter oscillation obtained with the apparatus of Fig. 10, with reference to the frequency control. The frequency was readily adjusted to the required value on an ordinary percentage basis, but when a heterodyne reception with two oscillators was attempted it failed. Two meters is the equivalent of 150,000,000 cycles and 1 k.c. is therefore only one part in 150,000 and a setting to this accuracy would be considered difficult under any circumstances. Here, with the tube elements themselves constituting the major part of the tuning capacities, the natural frequency variations were so pronounced that a heterodyne note did not exist; only a hiss could be heard as the oscillator frequencies passed through the

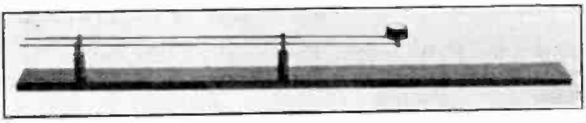

Figure 17—Range 3.4–6.2 meters.

region of synchronism. Later a beat was looked for at 3.6 meters, and this was successful. But the resulting note was extremely variable. Some of this variability was traced to the a-c. excitation of the power tube (full wave rectified alternating current on the plates and straight alternating current on the filaments) where the mechanical tractions set up inside the tube were apparently great enough to wobble the tube elements markedly. But also on battery excitation the frequency variation was considerable, and even when the coupling between the oscillators was sufficient to pull the frequency of one about, as the frequency of the other was varied, a clear note was difficult to obtain. Possibly, with great care, heterodyne reception will be feasible, but it will not be an engineering proposition without something like crystal control.

A list of short wave vacuum tube oscillator references, which it is hoped is complete, is attached.

Summary

An investigation of the short wave limit of vacuum tube oscillators of the normal type, where regeneration builds up oscillations in a resonant circuit, has indicated that the physical limits for ordinary commercial tubes lie between 3.5 and 1.5 meters, depending on the type of tube used. By means of a small special tube a wavelength of 1.05 meters was reached. No frequency control

was left at this wavelength. With a small power tube not deviating greatly from ordinary vacuum construction it was possible to operate at 3.5 meters with adequate frequency control, and it is concluded that the 3.5–5 meter range is available for technical purposes. Apparatus for measuring wavelength is discussed together with its application to the measurement of capacity and inductance. A bibliography of short wave vacuum tube oscillator work is attached.

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DIRECTIONAL RADIATION WITH HORIZONTAL ANTENNAS*

By

A. MEISSNER

In continuation of the investigation with horizontal antennas and reflectors[1] the attempt was made to concentrate the radiation energy by combining several antennas. If several horizontal

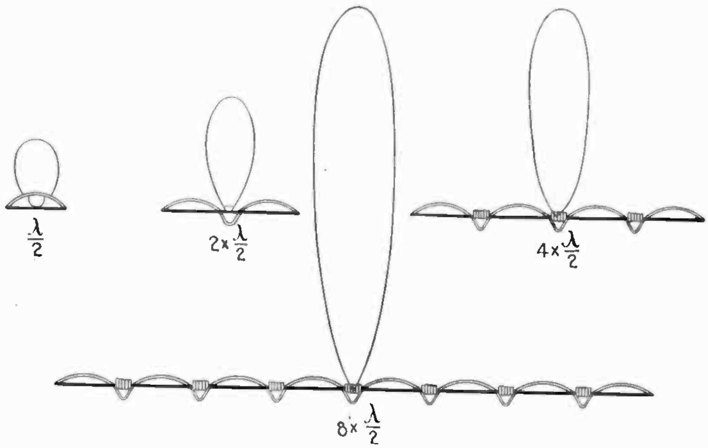

Figure 1—Concentration of radiation in the horizontal plane.

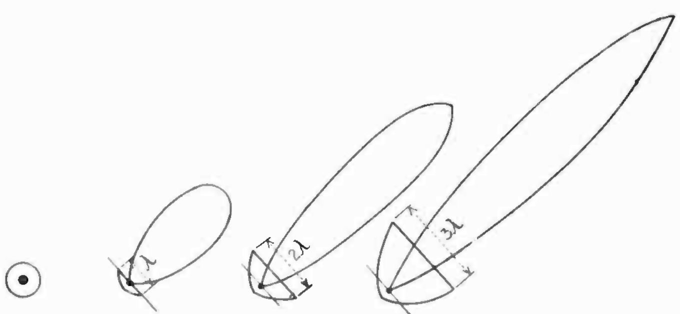

Figure 2—Concentration of radiation in the vertical plane.

antennas are so arranged as to oscillate together in the same phase, radiation patterns are obtained as shown in Fig. 1. The radiation angle as measured in the horizontal plane (angle determined by the

* Original manuscript in German received by the Institute, July 16, 1927. Translated manuscript received by the Institute, August 22, 1927.

[1] Jahrbuch für drahtlosen Tel. & Tel., Vol. 28, 1926, page 78.

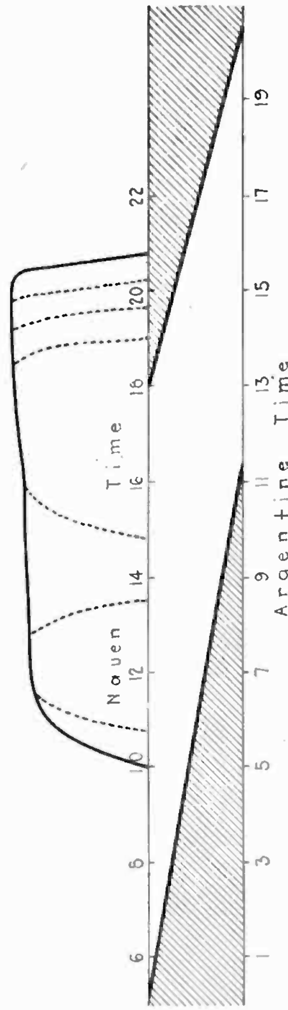

Figure 3—Communication periods with the Argentine on 11 meters.

half-amplitude of the radiation on both sides of the maximum) diminishes with increasing number of antennas in the ratio of 72 deg.:42 deg.:32 deg. to 14 deg. for 1, 2, 4, and 8 half-wave radiators respectively. The equal phasing of the radiation is obtained by the use of a non-radiating half-wave system[2] (non-radiating loops, coils or combination of coils and condensers) connected between two adjacent antennas. The concentration of the radiation in the vertical plane (Fig. 2) is obtained by an arrangement of a parabolic reflector about the horizontal antennas.

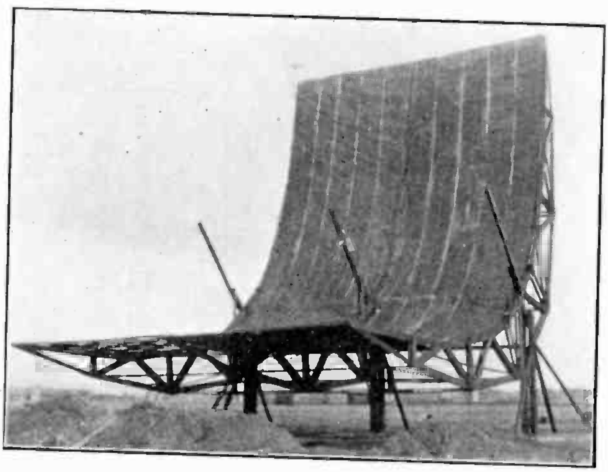

Figure 4

The concentration of the beam is sharper the greater the ratio of the width of the opening of the parabola to the wavelength.[3] The first reflector used was built of copper sheet and other reflector systems were compared with this "standard system." It was found that the metal-sheet surface could be entirely replaced by wire with equally good results; the individual wires of the wire reflector being one-half wavelength long.

With a beam concentrated in this way in both the horizontal and vertical planes, the determination of the optimum angle for the radiation from the transmitter was next attempted. In order to reduce the dimensions and the cost of the entire beam transmitting system as much as possible, the shortest possible wavelength was used at which reception over the distance 10,000 km. (South

[2] Franklin, British Patent.

[3] Tatarinoff, Jahrbuch für drahtlosen Tel. & Tel., Vol. 28, 1926, page 117. It seemed more suitable to deviate somewhat from the parabola form.

America) could still be expected. According to computations and theories[4] no reception would in general be possible with 11 meters over greater distances, as according to theory the beam is bent away from the earth at such high frequencies. Contrary to theory, the investigation showed that not only did the wavelength of 11 meters work extraordinarily well over great distances, but that this wave when horizontally polarized by reflectors was for the most part even the best for daylight transmission with South America. This showed that the wavelength range for short wave communication could be extended considerably downward. In connection with reception in South America, several interesting

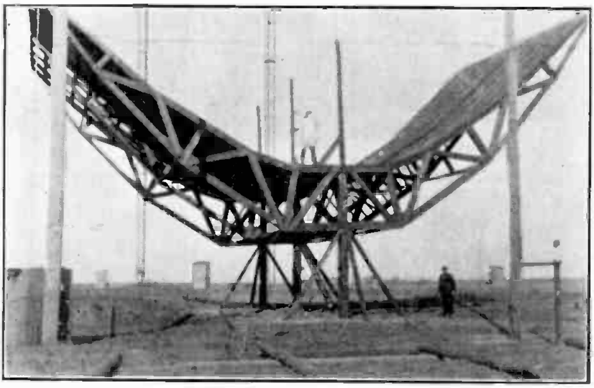

Figure 5

things were noted. The reception on the wavelength of 11 meters was also possible during the period in which a part of the intervening path lay in darkness. Fig. 3 shows the time of reception; underneath, the period of daylight for the month of April (above for Nauen and below for Rio de Janeiro). The reception began usually about 10 o'clock Central European Time. The period from 13 to 15 o'clock is often one of uncertainty. From then until 21 o'clock reception is very good. It was further characteristic of 11-meter transmission that on certain days reception failed completely. The following values of audibility ratios were obtained in comparing transmission from the 11-meter transmitter *AGK* antenna energy less than 1 kw. and the 15-meter transmitter *AGA*, antenna energy 8 kw. (beam radiator with vertical polarization;

[4] A. H. Taylor and E. O. Hulburt, Phys. Rev., Vol. 27, 1926, page 189. Lassen, Jahrbuch für Drahtlosen Tel. & Tel., Vol. 28, 1926, page 109.

8 parallel wires half a wavelength long and half a wavelength separation):

Date	Time	*AGK* (1 kw.) =11 m.		*AGK* (8 kw.) =15 m.
May 17, 1927	14–19 o'clock	1	:	2–6
	19–21 "	1	:	1
May 18, 1927	14–20:30"	5	:	1
May 19, 1927	14–19 "	6	:	1
	19–20:30"	2	:	1
May 5, 1927	none	normal		

The transmitter *AGK* is, in other words, better than the transmitter *AGA* but at times fails completely. It appeared as if the

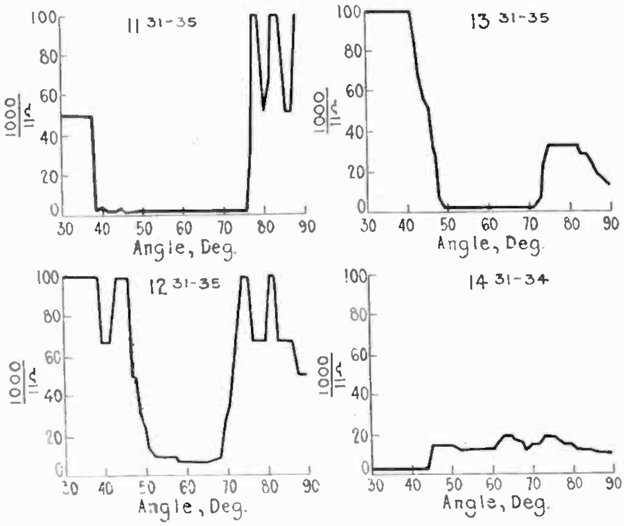

Figure 6—Signal strength for various reflector angles.

latter was somehow connected with a depression approaching the transmitter. Should this be the case, that the influence of the lower atmosphere (up to 20 km.) is of significance, a high reflector angle (80 deg.-second optimum as shown later) may bring the signal back in again on such days that the signal with low reflector angle goes out completely. The investigation so far had only been carried out with a radiation angle of the reflectors of 38 deg.

Reception on 11 meters shows comparatively less fading than on 15 meters. How much of this decrease in fading might be due to using 500-cycle plate supply on the 11-meter transmitter is not known. Fig. 4 shows the beam reflector system which was used. The antenna consists of two half-wave sections connected by a phase reversal coil. To this coil are connected the conductors leading from the transmitter which is located 28 meters distant. The reflector is 11.5 meters long and 19 meters wide and rests on three supports 3.9 meters high. At the right in the figure are shown the two windlasses with which the reflector is turned. Two men, one on each windlass, can easily raise the reflector from 30 deg. to 90 deg. in four minutes. Fig. 5 shows the reflector in the 90 deg. position. (The transmitting arrangements were in charge of Mr. J. Pohl, while the reception observations were made by

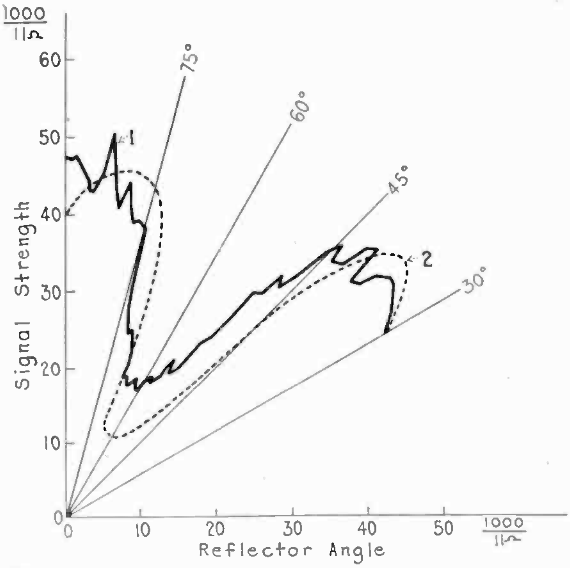

Figure 7—Signal strength for variation of reflector angle.

Mr. A. Ellerbrock in Rio de Janeiro and later in Buenos Aires). As the reflector turned, observations in Rio de Janeiro indicated a maximum with the reflector at an angle of about 38 deg. A second maximum, usually 10 percent less, was observed at about 80 deg. In between lay a deep minimum. Fig. 6 shows a few typical reception curves. The turning of the reflector took four minutes. While the first three curves show definite maxima, the last curve does not. This lack of definite maxima occurred in about 10 per cent of the observations. There was no difference between forenoon and afternoon. In Fig. 7 the mean value of the field intensity is plotted for the various angular settings. The investigation was carried on in such a way that at first measurements were made for several days and mean values obtained for each setting of the reflector (mean value Fig. 6, Curve 2). Later measurements were made for several days in which the reflector was rotated between 30 deg. and 90 deg. during each four minutes. (Mean value for three days, Fig. 6, Curve 1). Frequent comparisons were made with a simple vertical antenna (3/4 wavelength long). The vertical antenna was seldom heard and when it was heard, the intensity was only a fraction of that obtained from the reflector (approximately 1/10).

From these results one would assume that on all other wavelengths as well as on 11 meters, an optimum radiation angle exists for transmission in which space radiation plays a part. At present investigations are in progress to determine this angle for 20 meters.

MAKING NORMAL COORDINATES COINCIDE WITH THE MESHES OF AN ELECTRICAL NETWORK*

By

E. A. Guillemin

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The theory of normal coordinates is perhaps best known and certainly most effectively interpreted physically in connection with a mechanical system of elastically-coupled mass points which carry out small oscillations about a point of equilibrium. Here a normal coordinate is such a direction in space upon which the projection of the oscillation of any point of the system contains only one of the natural frequencies of the system. There are as many such normal coordinates as there are normal frequencies; and since these directions are in general neither normal to each other nor coincide with the axes of the fundamental frame of reference, each has projections upon all coordinates of the reference system. Mathematically this is evidenced by the fact that the component oscillation of any mass-point of the system upon any reference coordinate contains in general all of the natural frequencies of the system. Furthermore, we can easily visualize a process of rotation of the reference coordinates which will bring at least one of them to coincide with a normal direction or normal coordinate. And when carried out mathematically, such a transformation will cause the normal frequency, belonging to the particular normal coordinate in question, to be confined entirely to the reference coordinate which has been brought into coincidence. The projected oscillations upon the remaining reference coordinates then lack this normal frequency, i.e., it has been eliminated from the rest of our reference coordinates.

In setting up the differential equations for the normal mode of oscillation of an electrical network, we find an exact mathematical analogy to those of a mechanical system having the same number of degrees of freedom. Hence, from a mathematical standpoint at least, there also exist in the electrical system certain normal coordinates. The physical interpretation of them is, however, somewhat more puzzling than in the case of the mechanical system, and it is the object of this paper to point out in as direct a way as possible what, in the electrical network, is the equivalent of the normal coordinate of a mechanical system; and how it is

* Original manuscript received by the Institute July 26, 1927.

possible, by properly fixing the circuit constants, to bring about coincidence with certain meshes of the network, thereby confining the corresponding natural frequencies to these meshes and eliminating them from the rest.

We will begin with a brief review of the solution to the homogeneous system of equations describing the free oscillations of an electrical network. If we number the meshes of our system from 1 to n, and let $\lambda_{ii}, \rho_{ii}, \sigma_{ii}$ represent the total inductance, resistance, and elastance respectively present in mesh i, and further let $-\lambda_{ik}, -\rho_{ik}, -\sigma_{ik}$ represent the total inductance (here the algebraic sum of mutual and common self-inductance), resistance and elastance respectively common to meshes i and k, then obviously the countervoltage in mesh k due to current i_k in it is given by:

$$\lambda_{kk}\frac{di_k}{dt}+\rho_{kk}i_k+\sigma_{kk}\int i_k dt,$$

and the countervoltage induced into mesh k from mesh i will be given by:

$$\lambda_{ik}\frac{di_i}{dt}+\rho_{ik}i_i+\sigma_{ik}\int i_i dt.$$

In order to bring the analogy to the equivalent mechanical system still closer, we define:

$$x_k=\int i_k dt$$

as the corresponding "mesh charge," it being the equivalent of displacement in the mechanical sense just as electric current is the equivalent of velocity in the mechanical system. If we further use the abbreviation:

$$a_{ik}=\lambda_{ik}\frac{d^2}{dt^2}+\rho_{ik}\frac{d}{dt}+\sigma_{ik} \tag{1}$$

then the homogeneous system of simultaneous linear differential equations, which are an expression of Kirchoffs' e.m.f. law for the force-free network, become:

$$\left.\begin{array}{l} a_{11}x_1+a_{12}x_2+\cdots+a_{1n}x_n=0 \\ a_{21}x_1+a_{22}x_2+\cdots+a_{2n}x_n=0 \\ \cdot\quad\cdot\quad\cdot\quad\cdot\quad\cdot\quad\cdot \\ a_{n1}x_1+a_{n2}x_2+\cdots+a_{nn}x_n=0 \end{array}\right\} \tag{2}$$

We assume as solutions, the normal functions:

$$x_k=Y_k\epsilon^{pt} \tag{3}$$

Substitution into (2) gives the algebraic system of condition equations:

$$\left.\begin{aligned} b_{11}Y_1+b_{12}Y_2+\cdots+b_{1n}Y_n=0\\ b_{21}Y_1+b_{22}Y_2+\cdots+b_{2n}Y_n=0\\ \cdot\quad\cdot\quad\cdot\quad\cdot\quad\cdot\quad\cdot\\ b_{n1}Y_1+b_{n2}Y_2+\cdots+b_{nn}Y_n=0 \end{aligned}\right\} \tag{4}$$

where:

$$b_{ik}=\lambda_{ik}p^2+\rho_{ik}p+\sigma_{ik} \tag{5}$$

are now simple algebraic polynomials in p. If the conditions (4) can be satisfied in agreement with arbitrary initial conditions, then our assumed solution will be valid and useful. Obviously this eliminates the trivial solutions:

$$Y_1=Y_2=\cdots=Y_n=0.$$

Before going on with the solution in the usual manner, let us give the conditions (4) a more definite interpretation. In the first place it will be noted that the electrical equivalent to a physical dimension in the mechanical system is a mesh; and that the number of meshes in the electrical system is equivalent to the number of dimensions of the corresponding mechanical system. For the latter this number is limited to three in our actual space. In the electrical system, this number is physically unlimited—and fortunately mathematically also, the system (2) being written for an n-dimensional space. Now let us define the vector $\boldsymbol{b}_1$ with the components:

$$\boldsymbol{b}_{11},\ \boldsymbol{b}_{12},\ \cdots,\ \boldsymbol{b}_{1n}.$$

Similarly, the vector $\boldsymbol{b}_2$ with components:

$$\boldsymbol{b}_{21},\ \boldsymbol{b}_{22},\ \cdots,\ \boldsymbol{b}_{2n},$$

and so on down to a vector $\boldsymbol{b}_n$ with the components:

$$\boldsymbol{b}_{n1},\boldsymbol{b}_{n2},\ \cdots,\ \boldsymbol{b}_{nn},$$

the b_{ik}'s being those defined by (5). Then if we consider:

$$Y_1,\ Y_2\cdots Y_n$$

of equations (4) as components of a vector Y in our n-dimensional space, the condition equations (4) may be put in the vector form:

$$\left.\begin{aligned} (\boldsymbol{b}_1,\ Y)=0\\ (\boldsymbol{b}_2,\ Y)=0\\ \cdot\qquad\cdot\quad\\ (\boldsymbol{b}_n,\ Y)=0 \end{aligned}\right\} \tag{6}$$

where the round parentheses indicate scalar product. But if the scalar product of two vectors vanishes, it means that these are orthogonal to each other. Hence we see that the conditions (4), or their equivalent (6), demand that the vector Y be determined in such a way as to be *simultaneously orthogonal* to the given system of vectors $b_1, b_2, \cdots, b_n$, assuming for the moment that the p's are known. But the components of Y we know to be the transient mesh charge amplitudes corresponding to the natural angular velocity p which fixes the b_{ik}'s. Hence it is clear that the direction of our vector Y is the direction of the normal frequency. And we see at once that the condition equations (4) fix the directions of these normal coordinates.

We have so far said nothing about the possibility of fulfilling these condition equations. We shall give this a physical interpretation as well. Let us confine our argument for the moment to a three-dimensional problem so that the process of visualization becomes easier. Our coordinate axes are numbered 1, 2, 3 instead of being lettered x, y, z as is the usual case. Suppose then that the vectors b_i all have three components different from zero and are linearly independent. The conditions (4) or (6) require us to find such a direction in space as will be simultaneously orthogonal to all three of these vectors b_i. Obviously this is impossible, since all the available dimensions are consumed by the given system b_i. However, if the latter all lie in a plane, i.e., consume only two out of the available three dimensions, then the vector Y can be drawn normal to this plane to satisfy (4) or (6). And if the given system of vectors b_i all lie along the same direction, two independent solutions to (4) are possible. In general, if the given system of vectors b_i consume $(n-m)$ of the n available dimensions, then there are m possible independent solutions to the system (4). The determinant of the system is then said to have the rank $(n-m)$. If we demand a unique solution to the system (4), and this is the case with our network problem, then the rank of the determinant

$$D(p) = \begin{vmatrix} b_{11} & \cdots & b_{1n} \\ b_{21} & \cdots & b_{2n} \\ \cdot & \cdot & \cdot \\ b_{n1} & \cdots & b_{nn} \end{vmatrix} \tag{7}$$

should be $(n-1)$. This means that the determinant shall vanish, but that at least one of its first minors shall not vanish. The vanishing of the determinant:

$$D(p) = 0 \tag{8}$$

is the well-known determinantal equation which fixes the natural angular velocities p.

The problem of making a normal coordinate coincide with one of the meshes of the given network is now clear. If for a given root $p=p_\nu$ of (8), an entire column in the determinant (7) vanishes, then obviously the normal coordinate will occupy the coordinate, i.e., mesh, thus vacated by the b_i's. Incidentally, if we make a column in (7) vanish, the corresponding row vanishes also due to the relation:

$$b_{ik}=b_{ki}.$$

To make a column vanish for one of the roots of (8) it is necessary that:

$$b_{i2}=\alpha b_{i1};\ b_{i3}=\beta b_{i1};\ \cdots b_{in}=\gamma b_{i1} \tag{9}$$

where α, β, $\cdots$, γ are any proportionality factors, not necessarily integers. For, expanding (7) by minors we have for instance for the first row:

$$D(p)=b_{11}B_{11}+b_{12}B_{12}+\cdots+b_{1n}B_{1n} \tag{10}$$

where the B_{ik} are the minors of b_{ik}. But by (9) we have for $i=1$:

$$D(p)=b_{11}\{B_{11}+\alpha B_{12}+\beta B_{13}+\cdots+\gamma B_{1n}\}. \tag{11}$$

The determinantal equation is factorable, and one of the natural frequencies is given by:

$$b_{11}=0. \tag{12}$$

It is this frequency which goes with the normal coordinate coinciding with mesh one in this case. Similarly we could have made any other mesh a normal coordinate. Mesh one in the above case will be the only one in which the frequency given by (12) can be present. The transient solutions for the other meshes will contain only $(n-1)$ terms. Obviously we can use this process to eliminate objectionable natural frequencies as will be shown later.

In general, mesh k becomes a normal coordinate when:

$$b_{ik}=\alpha_{ik}b_{kk} \tag{13}$$

where the α_{ik}'s are any proportionability constants. If (13) is satisfied for all the b_{ik}'s then all meshes become normal coordinates.

The determinantal equation for this case becomes:

$$D(p)=(b_{11}\,b_{22}\cdots b_{nn})\begin{vmatrix} 1 & \cdots & & \cdots\alpha_{1n} \\ \alpha_{21} & 1 & & \cdots\ \alpha_{2n} \\ \alpha_{31} & \cdots & 1 & \cdots\ \alpha_{3n} \\ \cdots & & & \cdots \\ \alpha_{n1} & \cdots & & \cdots\ 1 \end{vmatrix}=0. \qquad (14)$$

But:
$$b_{ii}=\frac{\alpha_{ik}}{\alpha_{ki}}\,b_{kk},$$

so that (14) is equivalent to

$$b_{ii}=0. \qquad (14a)$$

The entire network possesses only one natural frequency.

Suppose we have a primary mesh (mesh No. 1), coupled to $(n-1)$ other meshes and fix the circuit constants so that:

$$b_{1k}=\alpha_{1k}b_{kk}. \qquad (15)$$

The other meshes are assumed not coupled to each other so that

$$b_{ik}=0 \text{ for } i\neq k\neq 1.$$

Then the determinantal equation becomes:

$$D(p)=\begin{vmatrix} b_{11} & \alpha_{12}b_{22} & \cdots & \cdots\ \alpha_{1n}b_{nn} \\ b_{21} & b_{22} & 0\cdots & \cdots\ 0 \\ \cdot & \cdot & \cdot & \cdot \\ b_{n1} & 0 & \cdots & \cdots\ b_{nn} \end{vmatrix}$$

$$=(b_{22}b_{33}\cdots b_{nn})\begin{vmatrix} b_{11} & \alpha_{12} & \cdots & \cdots\ \alpha_{1n} \\ b_{21} & 1 & 0\cdots & \cdots\ 0 \\ \cdot & \cdot & \cdot & \cdot \\ b_{n1} & 0 & \cdots & \cdots\ 1 \end{vmatrix}=0. \qquad (16)$$

The roots are obtained from:

$$b_{22}=0;\ b_{33}=0;\ \cdots b_{nn}=0; \qquad (17)$$

and:

$$\begin{vmatrix} b_{11} & \alpha_{12} & \cdots & \cdots\ \alpha_{1n} \\ b_{21} & 1 & \cdots & \cdots\ 0 \\ \cdot & \cdot & \cdot & \cdot \\ b_{n1} & 0 & \cdots & \cdots\ 1 \end{vmatrix}=0. \qquad (18)$$

All meshes except mesh No. 1 are normal coordinates, and the frequencies given by (17) are confined to their respective meshes. Mesh No. 1 contains only the one frequency defined by (18), which appears also in all the other meshes, since mesh No. 1 is the only one which is not a normal coordinate. If we made mesh No. 1 a normal coordinate also, the case would revert to the previous one, and all normal frequencies would coincide as given by (14a). In our present case, however, we have present in the network n different natural frequencies, but $(n-1)$ of these are pigeon-holed in their respective normal meshes and cannot appear in the rest of the network in the form of free or transient oscillations. When the network is subjected to suddenly applied forces in all of the meshes, the transient current in mesh No. 1 will contain only one frequency, while those in the other meshes will contain this one and also their own normal frequency as given by (17).

An interesting point in connection with the transient solution can be brought out if we consider a network where only one of the meshes, say mesh No. 1, is made a normal coordinate. The pair of conjugate complex roots, which define the one frequency in this normal coordinate, are given by:

$$b_{11}=0.$$

Let us indicate this root by $p=p_1$, and evaluate the corresponding transient current amplitude by Heaviside's formula. In the notation used in this paper, this formula becomes, for suddenly applied d-c. voltages:

$$i_{ik}=E\sum_{\nu}\frac{B_{ik}(p_\nu)\epsilon^{p_\nu t}}{\left(\dfrac{dD}{dp}\right)_{p=p_\nu}} \tag{19}$$

where the summation is to extend over all the roots of:

$$D(p)=0.$$

the subscripts ik on the current indicate that the latter belongs to mesh i with the voltage impressed in mesh k or vice versa. Suppose we assume the voltage impressed in mesh No. 1. Then the one term of (19) for $p=p_1$ becomes:

$$(i_{11})_{p=p_1}=E\frac{B_{11}(p_1)\epsilon^{p_1 t}}{\left(\dfrac{dD}{dp}\right)_{p=p_1}}. \tag{20}$$

But by the rule for differentiating determinants we have:

$$\left(\frac{dD}{dp}\right)_{p=p_1}=\left(\sum_1^n{}_{ik}\frac{\partial D}{\partial b_{ik}}\frac{db_{ik}}{dp}\right)_{p=p_1}=\left(\sum_1^n{}_{ik}B_{ik}(2\lambda_{ik}p+\rho_{ik})\right)_{p=p_1}$$

But since for $p=p_1$ the first row and column of $D(p)$ vanish, we see that:

$$B_{ik}(p_1)=0 \text{ for } i=k\neq 1, \tag{21}$$

so that:

$$\left(\frac{dD}{dp}\right)_{p=p_1}=B_{11}(p_1)\cdot(2\lambda_{11}p+\rho_{11}),$$

and substituting into (20) we get:

$$(i_{11})_{p=p_1}=\frac{E\epsilon^{p_1t}}{2\lambda_{11}p_1+\rho_{11}} \tag{22}$$

which shows us that the amplitude of this normal oscillation, which is confined to mesh No. 1, depends only upon the constants of this mesh and not at all upon those of the rest of the network. That the frequency given by $p=p_1$ is actually confined to mesh No. 1, is immediately evident from (21) and (19) and needs no further elaboration.

Suppose now that we impress our voltage in any of the other meshes, and proceed to calculate the transient current amplitude for $p=p_1$ by formula (19). Here again, due to (21) we find that this amplitude vanishes, even for mesh No. 1. Hence we note the very interesting fact that, not only is the normal frequency $p=p_1$ confined to the mesh which is its normal coordinate, but it will not come into action unless the actuating force is applied in that mesh. In the case where all meshes except one are normal coordinates, the transients in all meshes, due to a force applied in the non-normal mesh, contain only one term in spite of the fact that the whole network possesses n natural frequencies. In other words, a normal frequency which coincides with a mesh can only be actuated by a force applied in that mesh, and does not respond to disturbances which occur in the other meshes. Obviously the complexity of transients in complicated networks may be reduced considerably by making use of this fact.

The point of chief interest which this analysis has in connection with steady state operation is the elimination or reduction of the number of resonance maxima when the network is subjected to sustained variable frequencies. In simple coupled radio circuits containing only two meshes, the double resonance peak in the secondary is an objectionable feature and gives rise to "broad" waves. By making the primary a normal coordinate, one of these can be confined there, and the secondary will then contain only one resonance peak. Or by making both meshes normal coordinates, the network will contain only one frequency and hence only one resonance peak in primary or secondary. The second case would, however, be harder to maintain under variable conditions of secondary tuning and radiation;—but preference here is merely a matter of design.

In general, if a steady state e.m.f. of the form:

$$e(t) = Re\left\{E\epsilon^{\mu t}\right\} \tag{23}$$

(where μ may be real, imaginary, or complex, and *Re* denotes "real portion of") is impressed in mesh k, the resulting steady state charge in mesh i is given by:

$$x_{ik} = Re\left\{\frac{EB_{ik}(\mu)\epsilon^{\mu t}}{D(\mu)}\right\}, \tag{24}$$

and hence the corresponding mesh current is given by:

$$i_{ik} = \dot{x}_{ik} = Re\left\{\frac{E\mu B_{ik}(\mu)\epsilon^{\mu t}}{D(\mu)}\right\} \tag{25}$$

Again let mesh No. 1 become a normal coordinate, and let its normal frequency be represented by:

$$p_1 = -\delta \pm j\omega \tag{26}$$

Then for $i = k = 1$, and $\mu = j\omega$ we see that:

$$\left[\frac{B_{11}(\mu)}{D(\mu)}\right] \rightarrow \infty \text{ for } \rho_{ik} \rightarrow 0, \tag{27}$$

and hence it is apparent that mesh No. 1 resonates at this point. However, keeping the e.m.f. in mesh No. 1 and calculating the current in any other mesh we have:

$$i_{i1} = Re\left\{\frac{E\mu B_{i1}(\mu)\epsilon^{\mu t}}{D(\mu)}\right\}. \tag{28}$$

But

$$\left[\frac{B_{i1}(\mu)}{D(\mu)}\right]_{i \neq 1} \rightarrow \frac{0}{0} \text{ for } \rho_{ik} \rightarrow 0. \tag{29}$$

We can get at the behavior of this indeterminate form for large values of μ by evaluating in the usual manner. For the denominator we have:

$$\left[\sum_1^n ik\frac{\partial D}{\partial b_{ik}}\frac{db_{ik}}{d\rho_{ik}}\right]_{\rho_{ik}=0} = \sum_1^n ikB_{ik}^* \cdot \mu$$

where the * indicates the limit for $\rho_{ik}=0$. But again:

$$B_{ik}^*=0 \text{ for } i\neq k\neq 1,$$

so that the coefficient of the linear term in the Taylor expansion of the denominator becomes:

$$B_{11}^* \cdot \mu.$$

For the numerator we have:

$$\left[\sum_1^n rs\frac{\partial B_{i1}}{\partial b_{rs}}\frac{db_{rs}}{d\rho_{rs}}\right]_{\rho_{rs}=0} = \sum_1^{n}{}' rsB_{\left\{ {i1 \atop rs} \right\}} \cdot \mu \qquad \dagger$$

where the prime on the sum indicates that the terms for $r=i$ and $s=1$ are to be omitted. But here we see that:

$$B^*_{\left\{ {i1 \atop rs} \right\}}=0 \text{ for } r\neq 1,$$

so that the corresponding Taylor coefficient becomes:

$$\sum_2^n sB^*_{\left\{ {i1 \atop 1s} \right\}} \cdot \mu,$$

and our indeterminate (29) evaluates to:

$$\frac{\sum_2^n sB^*_{\left\{ {i1 \atop 1s} \right\}}}{B_{11}^*}. \qquad (29a)$$

Here the denominator is two powers higher in μ than the numerator, and hence the factor (29) varies as $1/\mu^2$ for large μ. Hence it is clear that all meshes except No. 1 do not resonate to the normal frequency p_1; and that for large values of the latter, the steady state oscillations in all but mesh No. 1 are very much suppressed. Mesh No. 1 tends to confine its normal frequency even in the steady state.

† The minor $B_{\left\{ {ik \atop rs} \right\}}$ is obtained from $D(\mu)$ by striking the rows i and r and the columns k and s, and prefixing the sign $(-1)^{i+k+r+s}$.

If now we put the source of e.m.f. in any other mesh except No. 1 and try to feed the normal frequency $\mu = j\omega$ into it, we have:

$$i_{1i} = Re\left\{\frac{E\mu B_{1i}(\mu)\epsilon^{\mu t}}{D(\mu)}\right\}.$$

But due to the symmetry of $D(\mu)$, this is identical with (28), and hence we should arrive at (29a) again for the limit (29). In other words, it is just as hard to feed the resonance frequency into a normal mesh as it is to get it out of one. The normal mesh resonates to its own frequency only when the e.m.f. acts in it, and in that case tends to confine the oscillation there.

Obviously many other interesting combinations of normal and non-normal meshes may be gotten up and investigated both in the transient and steady states. The results of such investigations are often both interesting from the purely theoretical standpoint, as well as useful from the practical. It is the hope of the writer that the principles involved have been made sufficiently clear so that the reader may carry out his own extensions of the idea by himself.

Summary

The theory of normal coordinates in oscillatory systems is briefly reviewed, and their significance in connection with the electrical network pointed out. The relation which must hold between the circuit constants in order that normal coordinates may be made to coincide with certain meshes is derived. The effect of having made a certain mesh a normal coordinate of the system is then shown for the transient state, first, to confine the corresponding normal frequency to that mesh; secondly, to cause its amplitude to be independent of the other circuit constants; and, thirdly, to cause that normal frequency to remain inactive unless the actuating e.m.f. is impressed in that mesh. The effect so far as the steady state is concerned, is to confine resonance with the normal frequency in question to the normal mesh, and in general to present a means for eliminating or localizing resonance peaks in more complicated networks.

VOLTAGE DETECTION COEFFICIENT*

By

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In a previous paper[1] it was shown that the current of modulation frequency which flows in the plate circuit of a triode detector is given by the expression

$$\Delta^2 \mathbf{I}_{pl} = (\mathbf{Det.\ I})_l \sqrt{2} m (\Delta \underline{E}_{oh})^2 \qquad (1)$$

In the above equation and those that follow the Δ sign indicates small but finite values. Bold-face type indicates complex quantities. The subscript p refers to the plate of the triode, and the subscripts l and h refer to low or modulation and high or radio frequencies, respectively. m is the degree of modulation and $\Delta \underline{E}_{oh}$ is the maximum value (indicated by the underscore) of the unmodulated radio voltage impressed in the grid circuit of the triode detector. The notation used here and in the previous paper is fully described in a paper entitled "Vacuum Tube Nomenclature" published in these Proceedings by the author.[2]

The complex factor $(\mathbf{Det.\ I})_l$ is defined as the total current detection coefficient, but reference to the previous paper shows that the value of $(\mathbf{Det.\ I})_l$ depends not only upon the tube characteristics, but also upon the impedance of the particular plate load used during the measurement of $(\mathbf{Det.\ I})_l$. It is then not easy to determine the detection coefficient if the tube is used with some other plate impedance. If, however, the plate current $\Delta^2 \mathbf{I}_{pl}$ be multiplied by the total plate circuit impedance a voltage is obtained which is a fictitious voltage but which, although non-existent, may be used for the calculation of the plate current of modulation frequency when the particular plate load is known. It was suggested in the previous paper[1] that a voltage detection coefficient may be defined according to the relation

$$\Delta^2 \widehat{\mathbf{E}}_l = (\mathbf{Det.\ E})_l \sqrt{2} m (\Delta \underline{E}_{oh})^2 \qquad (2)$$

which is more fundamental and a better quantity in terms of which to express the detection obtained with a particular triode.

* Original manuscript received by the Institute July 9, 1927.
Paper presented before the American Section, International Union of Scientific Radio-Telegraphy, April 21, 1927.

[1] A Theoretical and Experimental Investigation of Detection for Small Signals by E. L. Chaffee and G. H. Browning, I.R.E. Feb. 1927, Vol. 15, No. 2.

[2] Vacuum Tube Nomenclature by E. L. Chaffee. I.R.E. March, 1927, Vol. 15, No. 3.

It is convenient to think of $(Det.\ E)_1$ as the fictitious voltage introduced into the plate circuit by a 70.7 o/o modulated signal of one-volt maximum amplitude if the detection followed the same law for so large a signal. It is probably better to consider one ten-thousandth of $(Det.\ E)_1$ as the equivalent audio voltage in the plate circuit due to a 70.7 o/o modulated signal of .01-volt maximum amplitude.

In the paper referred to above a method is given for measuring $\Delta^2 I_{pl}$ from which, knowing m and $(\underline{\Delta E_{oh}})$, the value of $(Det.\ I)_1$ can be calculated. The purpose of the present paper is to present a method of directly measuring $\widehat{\Delta^2 E_l}$ from which $(Det.\ E)_1$ can be calculated. Some experimental results are here presented to show the application of the method.

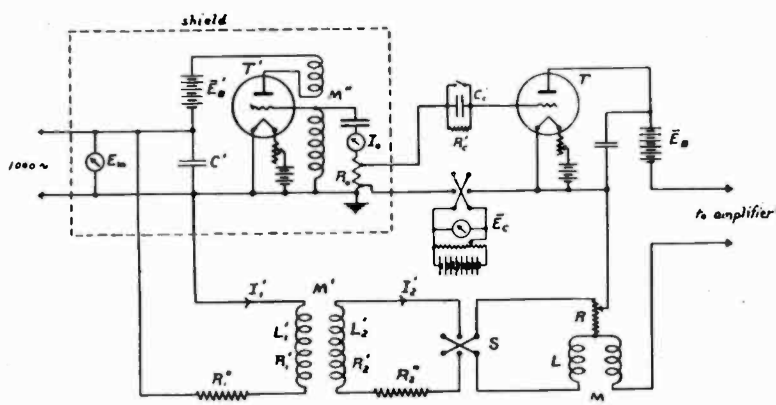

Figure 1

The diagram of connections for measuring the fictitious voltage $\widehat{\Delta^2 E_l}$ is shown in Fig. 1. Tube T is the triode under test. T' is a small radio-frequency oscillator, the output of which is modulated by an external source of power at 1,000 cycles. The oscillator is preferably enclosed in a shield shown by the dotted rectangle surrounding T' and its circuits. The oscillator has the oscillatory circuit connected to its grid. The plate circuit of the oscillator is regeneratively coupled to the oscillatory circuit through the mutual inductance M''. The plate voltage of the oscillator is pulsating and is obtained by adding to the steady voltage $\overline{E}_B'$, the modulating 1,000-cycle voltage E_m. The condenser C' serves only as a by-pass around the source E_m for the high-frequency currents in the plate circuit. The modulated oscillatory current passes through a small four-terminal resistance R_o, consisting of a short piece of fine resistance wire on which are soldered the potential taps. The voltage drop across the two potential terminals is introduced into the grid circuit of the tube T. Also in the grid circuit are the potential-changing means to provide polarizing potential $\overline{E}_c$, and the grid condenser and resistance C_c' and R_c'. The latter may be short-circuited when only plate-circuit rectification is considered.

Rectification taking place in tube T, in effect, introduces in the plate circuit a 1,000-cycle voltage $\Delta^2 E_1$. This voltage is opposed by an equal and opposite known voltage introduced into the plate circuit, balance being determined by silence in telephones connected to a two-stage amplifier. The known opposing voltage is obtained by passing through a variable resistance R and the coil L of the variable mutual inductance M a 1,000-cycle current derived from the modulating source E_m. An air-core, closely-coupled transformer, having mutual inductance M', is introduced and the circuit constants so adjusted that I_2' is in phase with E_m. To make possible any phase of the balancing voltage the current I_2' can be reversed by switch S, and M can be reversed by rotation of the secondary coil of M.

The condition under which the current I_2' is in phase with the modulating voltage E_m is simply stated below. Consider first, apart from the particular circuits of Fig. 1, a primary circuit which has a resistance R_1 and inductance L_1 coupled by a mutual inductance M to a secondary circuit having resistance R_2 and inductance L_2. Then the complex expression for the secondary current when an e.m.f. of E_1 volts is impressed in the primary circuit is

$$I_2 = \frac{jM\omega E_1}{(R_1+jL_1\omega)(R_2+jL_2\omega)+M^2\omega^2} \tag{3}$$

$$= \frac{jM\omega E_1}{(R_1R_2-L_1L_2\omega^2+M^2\omega^2)+j(R_1L_2\omega+R_2L_1\omega)}$$

If the first bracket of the denominator is made equal to zero then the secondary current is in phase with the e.m.f. E_1 and given by the expression,

$$I_2 = \frac{E_1 M}{R_1L_2+R_2L_1} \tag{4}$$

Placing the first bracket of the denominator of (3) equal to zero gives as the condition necessary for no phase difference between E_1 and I_2—

$$R_1R_2 = L_1L_2\omega^2 - M^2\omega^2 \text{ or} \tag{5}$$

$$\eta_1\eta_2 = 1-\tau^2$$

where $\eta_1 = \dfrac{R_1}{L_1\omega}$, $\eta_2 = \dfrac{R_2}{L_2\omega}$ and $\tau = \dfrac{M}{\sqrt{L_1 L_2}}$.

Applying the above theory to the present case as given by the diagram of Fig. 1, the condition to be fulfilled in order that the current I_2' be in phase with E_m is, by equation (5)

$$(R_1'+R_1'')(R_2'+R_2''+R) = L_1'(L_2'+L)\omega^2 - M'^2\omega^2 \quad (6)$$

The value of the secondary current I_2' is then obtained by applying equation (4) to the present case.

$$I_2' = \frac{M'E_m}{(R_1'+R_1'')(L_2'+L)+(R_2'+R_2''+R)L_1'} \quad (7)$$

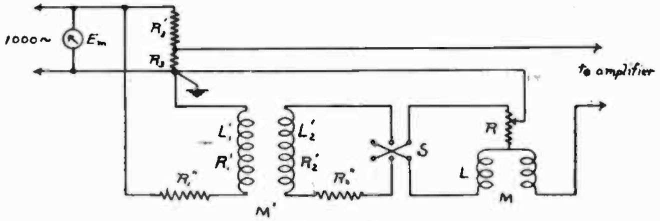

Figure 2

Although the proper values of the resistances of the circuit can be determined by calculation, usually it is easier to determine their proper values by experiment as follows: Referring to Fig. 2, two resistances, R_3 and R_3', are connected in series across E_m and the junction between the two resistances is connected to the terminal of the amplifier which in Fig. 1 is shown connected to the plate battery $\overline{E}_B$. R_3 may conveniently be one-thousandth of R_3'. Then R_2'' and R_1'' can be so adjusted that with M equal to zero, balance is obtained with some value of R. The value of R gives the number of ohms which introduces into the plate circuit a voltage equal to $\dfrac{E_m R_3}{R_3+R_3'}$. It is usually desirable to make R_2'' about 10,000 ohms so that the variation of R in normal operation causes only a small percentage variation in the total secondary resistance. R is shown in the diagram as a potentiometer on the high-dial end. In the experiments to be described R was a three-dial box having dials of tenths, units and tens of ohms. As just stated, the connections were changed so that all of the 10-ohm coils were in the secondary circuit, the switch arm being represented by the arrow in the diagram. The total maximum change of resistance due to varying the two lower dials was then only 11 ohms.

Before measurements of detection coefficient can be made the oscillator T' must be adjusted to give sinusoidal modulation. The oscillator is tested by short-circuiting the alternating voltage E_m and reading I_o for various values of $\overline{E}_B'$. The value of coupling M'' and the grid polarizing potential of the oscillator should be so chosen that the plot of I_o against $\overline{E}_B'$ is straight over a considerable range of $\overline{E}_B'$, as shown in Fig. 3. A value of $\overline{E}_B'$ is then chosen corresponding to a point about midway on the straight portion of

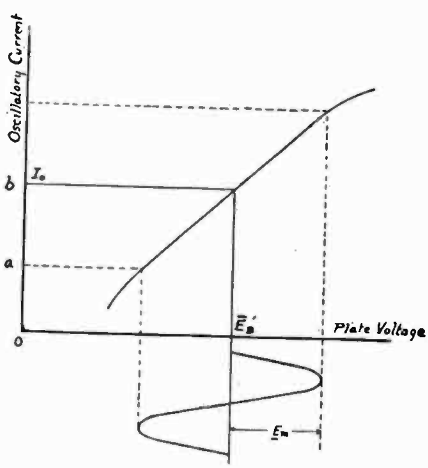

Figure_3

the plot. The peak value of E_m or $\underline{E}_m$ is then chosen to vary the plate voltage over the straight line portion of the plot. The degree of modulation m is then given by the ratio $\frac{\overline{ab}}{\overline{ob}}$ where $\overline{ob}$, as indicated in Fig. 3, is equal to I_o, the r.m.s. value of the unmodulated radio current, and $\overline{oa}$ is the minimum value of the r.m.s. radio current as it is modulated. It is assumed that the amplitude of the radio-frequency current changes as its plate voltage is varied at 1,000-cycles according to the plot of Fig. 3 which was obtained by a slow variation of plate voltage.

The value of $(\Delta \underline{E}_{oh})$ to be used in equation (2) is obviously

$$\Delta \underline{E}_{oh} = \sqrt{2} I_o R_o \tag{8}$$

where I_o is the value of the radio current when there is no modulation. When the modulating e.m.f. E_m is superposed upon the steady plate-battery voltage $\overline{E}_B'$, the radio frequency current will increase to $I_o\sqrt{1+\frac{m^2}{2}}$.

Some tests were made upon two tubes of the same size and make, except one has an amplification factor of 8.5 and the other a μ of 20. The low-amplification-factor tube corresponds to the type 201-A Radiotron.

The constants of the circuits during the tests described in this communication are given below:

$\overline{E}_B' = 45$ volts	$M' = 1.36$ h.
$E_m = 10$ volts at 1000 cycles	$L_1' = 0.841$ h.
$m = 0.51$	$L_2' = 3.63$ h.
$I_o = 74.7$ milliamperes	$R_1' = 576.$ ohms
$R_o = 0.504$ ohms	$R_2' = 1490.$ ohms
$(\underline{\Delta E_{oh}}) = 0.0533$ volts	$R_1'' = 3310$ ohms
	$R_2'' = 10000$ ohms

When calibrating the transformer by the method of Fig. 3 it was found that 87.5 ohms in R balanced $\frac{5}{1005} \times E_m$. Then

$$\widehat{\Delta^2 E} = \frac{5}{1005 \times 87.5} \cdot \sqrt{R^2 + M^2\omega^2}$$

$$= 0.000568\sqrt{R^2 + M^2\omega^2} \tag{9}$$

and $$Det.\ E = \frac{0.000568}{\sqrt{2} \times 0.51 \times 0.0533^2} \cdot \sqrt{R^2 + M^2\omega^2}$$

$$= 0.278\sqrt{R^2 + M^2\omega^2}$$

The curves of Figs. 4 and 5 which exhibit observed points give the values of (*Det. E*) for plate-circuit rectification plotted against the grid polarizing potentials. The three curves of Fig. 4 are for the tube possessing an amplification factor of 8.5 and the two curves of Fig. 5 give the results for the high-μ tube. The values of (*Det. E*) increase as the plate current, plotted against grid volts, approaches zero due to the increasing resistance of the tube. The curves end when the current has become so small that balance is indefinite.

Not all of the audio voltage given by (*Det. E*) is available as

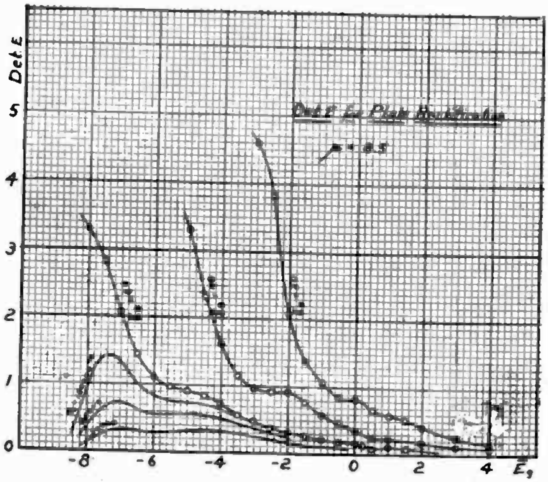

Figure 4

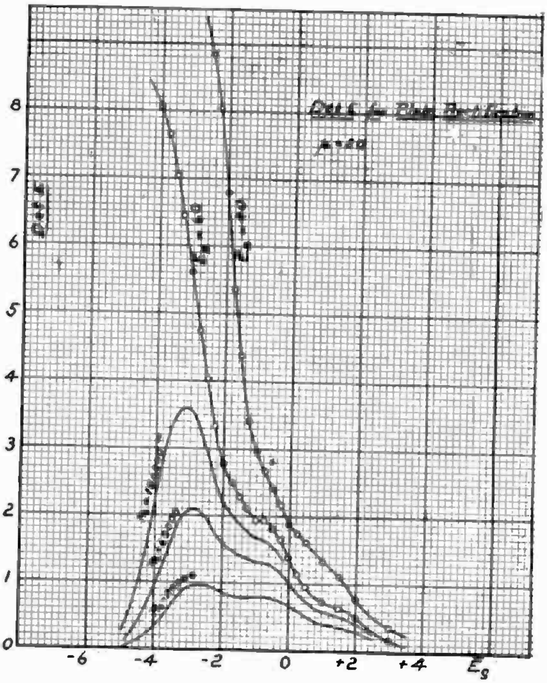

Figure 5

output voltage to act upon the next amplifier tube. The available fraction of (*Det. E*) is obtained by multiplying by the ratio

$$\frac{\sqrt{R_b^2+X_b^2}}{\sqrt{(r_p+R_b)^2+X_b^2}},$$ where R_b and X_b are the resistance and reactance of the load in the plate circuit of the detector. It is apparent then that, given the curves of (*Det. E*) and of r_p for a tube, the output voltage can be obtained for any plate load. If the plate load has no reactance then the ratio becomes $\frac{R_b}{r_p+R_b}$. The

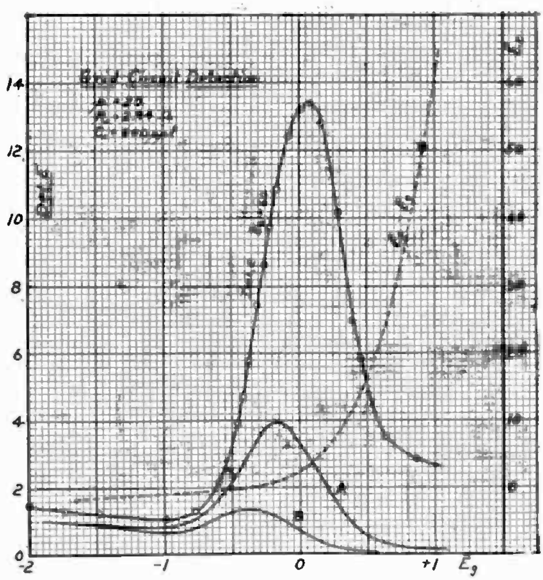

Figure 6

three full-line curves of Figs. 4 and 5 were obtained by multiplying the curve of (*Det. E*) for 60 volts by the above fraction for three values of R_b given on the curves. It is worthy of note that the values of k_p for the two tubes at the grid voltage which gives maximum detection are not very different. The value of k_p at -7 volts for the low-μ tube is 8.2×10^{-6} mhos. and for the high-μ tube at -3 volts is 9.9×10^{-6} mhos. Examination of the curves shows that the high-μ tube is the better detector when used to give plate-circuit rectification but in either case it is desirable to use a plate load of high impedance.

The curves of Figs. 6 and 7 which show the observed points give the values of (*Det. E*) for grid-circuit rectification, when the grid resistance is about 3 megohms and the grid condenser 200 micro-microfarads. Curves of Fig. 6 are for the low-μ tube and those of Fig. 7 are for the high-μ tube, both for a plate potential of 60 volts. The curves are plotted against *actual* grid volts, but since the grid takes current when the grid voltage is positive there is a

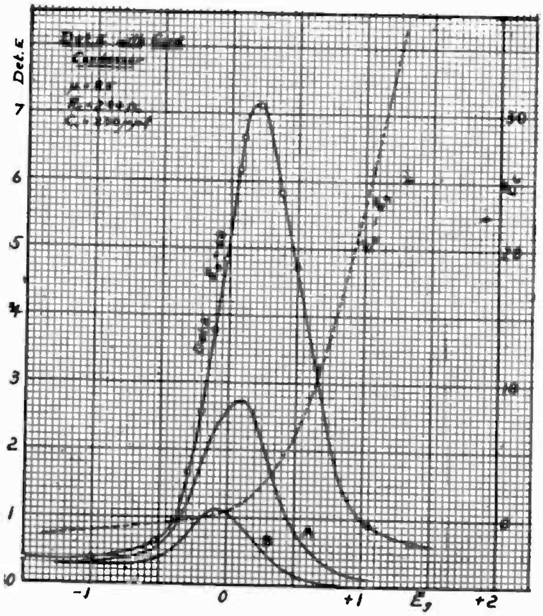

Figure 7

large voltage drop through the external grid resistance necessitating sometimes a very large applied polarizing potential $\overline{E}_c$ to obtain even a slightly positive value of $\overline{E}_g$. The dotted curves of Figs. 6 and 7 give the relations between $\overline{E}_g$ and $\overline{E}_c$.

In the case of grid-circuit rectification there are two effects which operate to cause the actual available voltage output to be less than that given by (*Det. E*). In the first place only a certain fraction of the fictitious voltage acting in the plate circuit is available as output voltage as explained in the case of plate-circuit rectification. The ratio $\dfrac{\sqrt{R_b^2+X_b^2}}{\sqrt{(r_p+R_b)^2+X_b^2)}}$ reduced to apply to a pure resistance plate load for the two cases shown in Figs. 6

and 7, are plotted in Fig. 8. The full-line curves are for the low-μ tube; the dotted curves apply to the high-μ tube.

The second cause operating to reduce the output voltage in the case of grid-circuit rectification is the damping of the input oscillatory circuit due to the conduction between grid and filament. Let ΔE_{oh} be the radio voltage across the oscillatory circuit when there is no grid conduction as in plate-circuit rectification, and let

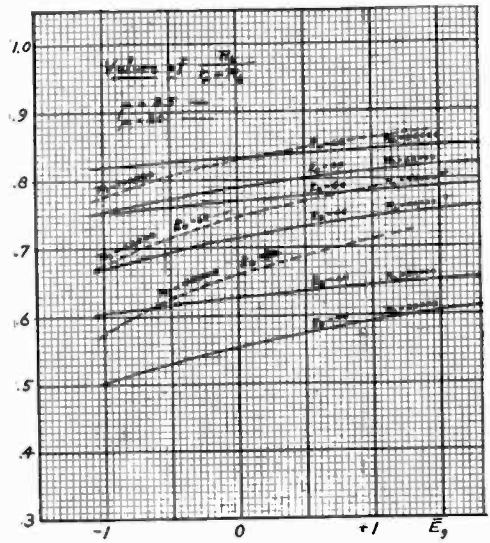

Figure 8

$\Delta E_{oh}'$ be the radio voltage across the same circuit when the grid conductance is different from zero and equal to k_g. The second reducing factor is then $\left(\frac{\Delta E_{oh}'}{\Delta E_{oh}}\right)^2$. The ratio is squared because detection varies as the square of the radio amplitude. It can be easily shown that this ratio has, to a sufficient accuracy, the value given in (10) below.

$$\left(\frac{\Delta E_{oh}'}{\Delta E_{oh}}\right)^2=\left(\frac{\eta}{\eta+L\omega k_g}\right)^2=\left(\frac{1}{1+\frac{L^2\omega^2}{R}\cdot k_g}\right)^2 \tag{10}$$

where η is the ratio $\frac{R}{L\omega}$ for the oscillatory circuit. The ratio $\frac{L^2\omega^2}{R}$ is the equivalent resistance of the oscillatory circuit at resonance.

Referring again to Figs. 6 and 7, the full-line curves marked A were obtained from the detection-coefficient curves by multiplying by the product of the factor from equation (10) and the factor taken from Fig. 8 for the value of plate-load resistance R_b equal to five times the plate resistance r_p of the tube at zero grid volts. The values of R_b in the two cases are, respectively, 60,000 ohms and 100,000 ohms. For the two A-curves the value of $\frac{L^2\omega^2}{R}$ was taken as 100,000 ohms. The two curves marked B in Figs. 6 and 7 are similar to the A-curves except that the value of $\frac{L^2\omega^2}{R}$ was chosen to be 360,000 ohms. This value is a reasonable value because it corresponds to an η of 0.005, a value of L of 286 microhenrys, and a wavelength of 300 meters.

The enormous reduction in detection due to the factor of equation (10) is apparent. No account is taken of the effect of regeneration if it exists either solely because of the capacity between grid and plate of the tube, or if regeneration is purposely provided. Of course the effect of regeneration is to increase the voltage across the oscillatory circuit due to an e.m.f. induced in the coil, but such regeneration is probably more effective in increasing the signal strength with plate-circuit rectification than with grid-circuit rectification, for in the latter case, although regeneration decreases η, the factor of equation (10) is at the same time reduced. However, it is well known that since regeneration in both types of detection can bring the system to the point of oscillating, the signal strength then should be approximately the same in both cases. This is borne out by experimental observation.

The presence of the grid conduction acting across the oscillatory circuit very much broadens the resonance curve to the sacrifice of selectivity. The difference in selectivity in the two cases with no regeneration is very apparent in actual practice. In fact, taking into account all of the considerations, it seems to the author that grid-circuit rectification has little to justify its use either from the point of view of sensitivity or from the stand-point of quality.

Finally, it may be worth noting that the method of measurement outlined in the paper can easily be applied to the measurement of the voltages across a plate-load resistance or reactance, or across the secondary of a plate-circuit transformer, and in order to include the effects of damping of the oscillatory circuit by the grid conductance and to include the effects of natural or purposed regeneration, the radio voltage could conveniently be introduced in series with the inductance of an oscillatory circuit connected in the grid circuit of the detector. These changes might seem more satisfactory because the tube would be used as in practice. Such measurements are not here given because they are less definitive of the tube itself, but depend upon the particular impedances used.

Summary

The advantages of the use of voltage detection coefficient as a means of expressing the sensitivity of a detector and for the comparison of detectors are discussed. An experimental method is given for the measurement of the fictitious equivalent audio voltage which, if acting in the plate circuit of a detector, would give the demodulated audio plate current. Some experimental results are given for both plate-circuit detection and grid-circuit detection. The experimental results show that a high-μ tube is usually more sensitive as a detector than a low-μ tube.

RADIO VISION*

By

C. Francis Jenkins

(Jenkins Laboratories, Washington, D. C.)

In speaking to you this evening on the subject of the electrical transmission of pictorial representations, may I say that in our laboratory we have found it convenient and informative to use the words radiogram, radiophone, and radio vision when we speak of radio-carried service; and to say telegram, telephone, or television whem we speak of wire-carried service.

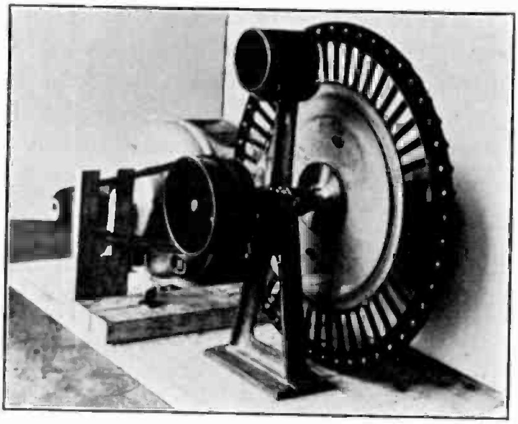

Figure 1—Laboratory receiver for radio vision. The black box contains a neon gas lamp; the slotted lens-disc sweeps the light-image spot across the screen in lines while the overlapping prism-disk distributes the lines from top to bottom of the screen.

The art of electrical picture-transmission is very old, relatively, for more than fifty years ago successful demonstrations were made in sending pictures by wire.

And there have been many workers, too, but the attainment of each was given but passing notice until the stamp of approval was put thereon by the great laboratories of the Bell Telephone Company, when, in April last, they made their spectacular demonstration between Washington and New York. This demonstration gave a great impetus to the development of electrical transmission of all kinds of pictorial representation.

* Original manuscript received by the Institute, July 6, 1927.
Presented before the Philadelphia Section of the Institute of Radio Engineers, June 24, 1927.

In general there have been but two types of mechanisms employed, the cylinder and the flat surface. The cylinder has been used most often in the transmission of still pictures.

But obviously a flat receiving surface is the only possible type for radio vision and television, for the eye must seem to see the whole picture all the time, and that can only be done on a flat screen.

However, whatever type of machine is used, the only method employed to the present time consists in a linear analysis of the picture, scene, or object, and the instantaneous synthesis of each line on a distantly located receiving surface.

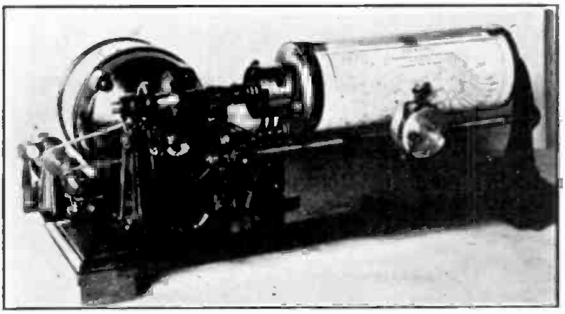

Figure 2—The weather map ship's receiver. The pen-box is moved across the base-map on the rotating cylinder by the screw, while the incoming radio signals touch the ink pen to the paper to build up the weather map.

The lights and darks of each successive line are changed into electrical current of corresponding strengths, which, carried to distant receivers, is there changed back into like light intensities and assembled on a suitable surface, for example, a sheet of paper, a photo film, or a flat picture screen.

It is quite evident that if this synthesis in the received picture is to result in an exact likeness to the transmitted subject, perfect synchronism of all receiving mechanisms with that of the transmitting mechanism must exist.

To attain synchronism, the synchronous motor is the simplest, but it is limited in application.

Other synchronizing methods have consisted of clock-controlled motors, where elapsed time is the standard against which all the motors were regulated.

Tuning forks for controlling the motors have been extensively employed, but this is but a modification, for it is only a smaller and more frequent division of time than the clock method.

Because pendulum clocks and most tuning forks soon stop when installed aboard a rolling ship in a rough sea, an oscillator, resembling the escapement of a ship's chronometer, has been developed, and has been found quite a successful motor controller.

A simple method of synchronizing consists in maintaining the speed of the motor constant with a governor, together with means for automatically setting the receiving cylinder at zero on the beginning of each revolution of the transmitter cylinder.

We employed this method in the "three-cornered experiment" recently conducted by the U. S. Navy, the Weather Bureau, and the writer's laboratory.

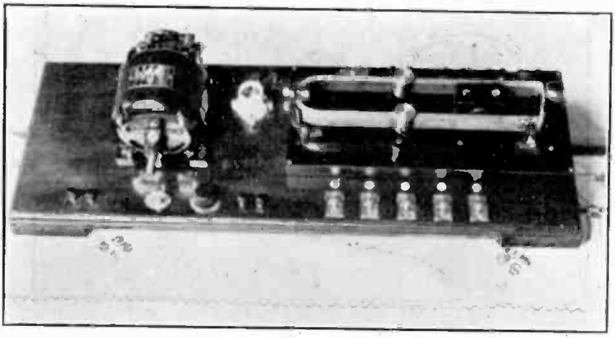

Figure 3—Fork-controlled motor unit. The vibration of the free ends of the fork arms cuts out armature resistance to keep the motor up to a definite speed. The fork continues to vibrate by reason of the current pulses in the electro-magnet between the fork arms.

Each morning the Weather Bureau made up a weather map, gave it to us, and we put it on our transmitter in the Navy building, which was connected by wire to the radio broadcast transmitters at Arlington, Virginia.

Weather map receivers were set up at certain land stations, and also on board the *U. S. S. Trenton*, and the *U. S. S. Kittery*. The latter made experimental cruises between Naval Operating Base at Hampton Roads and Caribbean Sea ports.

This territory was chosen for its well-known static disturbances, coincident with hurricane-forming zones. And it was found that weather maps could be recorded on the receivers with certainty even when weather information could not be received by code. The Florida hurricane of September provided a severe test of the system, demonstrating its worth and dependability.

While these "side-uses" are doubtless valuable, my premier ambition was radio vision, and so we do these other things only in "breathing spells" between attacks on the main problem.

Our first public demonstration of radio vision occurred on June 13, 1925, when we showed in the laboratory in Washington, in the presence of Navy Secretary Wilbur, Admirals Taylor and Robinson, and many others, what was happening at the time at the Naval Air Station at Anacostia, some miles distant. It was the first radio vision demonstration ever made, I believe, and quite an historical event to many of us.

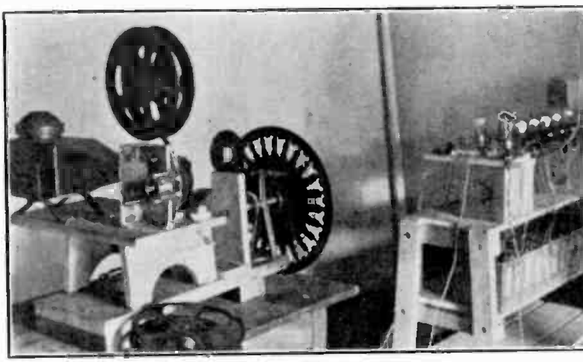

Figure 4—The simple mechanism by which motion pictures are transmitted. The rotation of the slotted lens-disk sweeps the image of each picture "frame" on the film across the light-sensitive cell in the box shown at the extreme right above the tubes.

The possibilities of radio vision for home entertainment, and of television in business, have been told so repeatedly in the public press that I hardly need restate here the promises of views of distant inaugural ceremonies, flower festivals, and baby parades. Being technical men I rather think you are more interested in details of methods and mechanisms.

As you know, the general scheme is to analyze the object or scene by a rotating scanning disk which in the transmitter permits the light reflected from the subject to fall on a light-sensitive cell, which, just as in still pictures, changes these light values into like current values; and which at the receiver permits light from a given source to be seen, directly or by reflection from a screen.

Such a scanning disk was shown in a patent as early as 1884. It consisted of a disk with one-fiftieth of an inch (1/50 in.) holes therein, arranged in a spiral.

The holes were an inch apart in the spiral, and the ends of the spiral had an inch offset; therefore, the contemplated picture was an inch square made up of fifty lines.

The light intensity, as in a pin-hole camera, is limited to the amount which can pass through this minute aperture, that is, 1/50 of 1/50, or only 1/2,500 part of the whole light.

To overcome this limitation it was proposed by a Frenchman that an arc lamp of high-intensity be focused on these minute apertures, after passing which the spot of light is swept across the subject by the rapid rotation of the scanning disk. I hardly think

Figure 5—Strip device by which news copy, or like matter, can be transmitted by radio or by wire automatically as a continuous process. The overlapping prismatic rings in rotation sweep each typewritten line across a light-sensitive cell as the message moves longitudinally.

he expected to radio-transmit a baseball game by sweeping thereover a point of light from an arc lamp.

We, in the laboratory, think that limiting the light by passing it through these apertures is not the best plan; the available light is too limited. So, to get a greater value of the light, we usually make the openings in the scanning disk 1-1/2 inches in diameter and put lenses over the openings; and get the required tiny light-spot by focusing the light-source as a tiny flying spot on the receiving screen, to build up the moving picture.

While we have made a variety of mechanisms, this has been a fundamental principle in all of them.

Television would doubtless have been attained as early as the telephone if a suitable light-sensitive cell had been available for the transmitter, and an adequate light-source for the receiver.

Such tools are available today, and marked progress has been made in recent months toward a radio vision receiver acceptable to the public, and with the many hands and minds now engaged on the problem, I confidently believe its completion is the work of but a few months more.

In the transmitter we need more sensitive light-cells, or perhaps I should say, light-cells giving greater current output, a current output which will more dependably start the first tube of an amplifier.

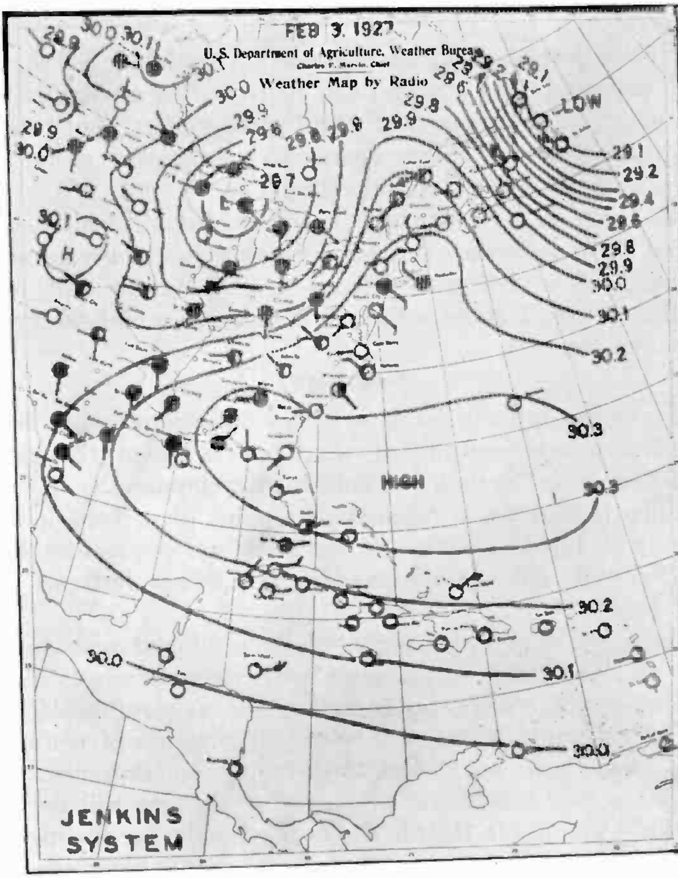

Figure 6—One of the weather maps as received aboard the *U. S. S. Kittery*. The base map is printed in brown and the isobars and other weather information received by radio is received thereon in red ink.

With light reflected from outdoor objects, the potassium cell gives but a very small current, only a few microamperes.

The present day potassium cells do not give current output proportional to the light intensity, but rather to the cell-surface covered by the light.

Resistance cells act too slowly for the required speed of light-reaction, which is of the order of 250,000 per second.

Our problem in the receiver is a light-source more intense than the neon lamp, but having approximately its high-speed light-change.

The other possibility is a steady light source and a light-valve between the source and the screen for modulating the light to build up the picture.

For this purpose many have proposed a bisulphide cell through which polarized light is passed and controlled by a potential (the Kerr effect) or a current field (the Faraday effect), but this cell requires such a large energy change to produce any useful light change as to put it out of consideration.

Of course, we have our own proposed solutions, which I shall tell you about if they are successful. Radio vision for home entertainment will revive radio interest as nothing else will, for it combines the reach of the radio and the fascination of the story told in pantomime.

Summary

To see by radio what is actually happening at a distant place is now an accomplished scientific attainment, though the mechanism is not yet a merchandising development.

When it shall have reached that point of perfection, then one may sit in one's home and see inaugural ceremonies, baseball, football, polo games, mardi gras, flower festivals, and baby parades.

Radio vision is also experimentally combined with audible radio, and when these instruments are made generally available, then radio pictures and music and speech at the fireside, sent from distant world points, will be the daily source of news; the daily instructional class, and the evening entertainment; and equally the long day of the sick and the shut-ins will be more endurable, and life in the far places less lonely, for the flight of radio is not hindered by rain or storm, or snow blockades.

The electrical transmission of still pictures of photographs, sketches, maps, and the like, is now an every day affair, applicable to newspaper illustration and other usefulness.

But for the research worker it has lost its interest for him, it is too easy. To increase the speed of picture presentation ten thousand times, as required in radio movies, makes it a sporting proposition, and really worth while.

The apparatus emplayed in transmitting pen-and-ink sketches, photographs, and movies by radio are all simple in construction, though modified to best suit each particular use.

BOOK REVIEWS

Radio Theory and Operating, BY MARY TEXANNA LOOMIS. LOOMIS PUBLISHING CO., THIRD EDITION, 886 Pages with approximately 700 illustrations, flexible covers. Price $3.50.

In the main this is a textbook for courses training students for commercial operating (although it might prove of value as a reference book for those engaged in other radio lines) and is written by the President of the Loomis Radio College.

A short historical section gives details of the experiments of Dr. Mahlon Loomis, of Washington, D. C., in 1865 et seq., who devised what is claimed to be a practical wireless telegraph system and to whom later a patent was granted. While definite proof and details of the system are unavailable, still many of the ideas indicated in the diagrams and notes reproduced anticipated modern practice and it would seem that Dr. Loomis should occupy a more important place in radio history than has been heretofore accorded him.

In common with many other textbooks it is used to best advantage in connection with lectures, since many of the terms used in the first few chapters are not adequately explained until later. Many cases will be noted where the use of definitions apparently from dictionaries do not disclose the exact distinctions usually accorded the terms.

A large number of inaccurate statements decrease the value of the book as a reference work. These are mainly confined to the first two hundred pages. These include such statements as: "A henry is defined as the energy which is induced by the cutting of one hundred million lines of force per second."

"A coil of high resistance wire passes all of the current in the circuit but consumes a portion of it in heat."

"The back e.m.f. due to inductance in a circuit is called reactance."

"These (referring to an illustration of a logarithmic condenser) are known as 'Straight Line' condensers, because when a graph is made of the capacity or of wavelengths in meters in a circuit in which they are used the 'curve' will be a straight line."

It is unusual that so many of these items should have survived to the third edition.

On the other hand, the major portion of the book is devoted to detailed description of apparatus circuits and procedure in broadcast transmitting, commercial operating, amateur operating and receiving fields. Very little mathematical theory is included, but a number of useful tables are to be found.

The book is excellent guide for studying for a commercial operating examination. Pertinent questions and problems appear in each section, and a complete index of 27 pages completes the book.

R. R. BATCHER

Properties and Testing of Magnetic Materials, BY THOMAS SPOONER, MCGRAW-HILL, 1927. FIRST EDITION, 385 Pages. Price $5.00.

The author of this book has admirably succeeded in gathering and arranging in useful form the more important properties of commercial ferromagnetic materials, and modern methods of testing and inspection. The book is singularly free of the accumulated "deadwood" of the early years of magnetic investigation, and although intended primarily for the use of the engineer it will also find a useful place in the library of the research worker.

GREENLEAF W. PICKARD

DIGESTS OF UNITED STATES PATENTS RELATING TO RADIO TELEGRAPHY AND TELEPHONY

Issued September 20, to October 11, 1927

By

JOHN B. BRADY

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

1,642,688—FIXED CONDENSER—ALFRED MOSS, of New York, N. Y., Filed July 14, 1925, issued September 20, 1927. Assigned to Electrad, Inc.

1,642,861—ELECTRIC RELAY—L. B. TURNER, of Cambridge, England. Filed Feb. 13, 1919, issued Sept. 20, 1927. Assigned to Westinghouse Electric & Manufacturing Co.

1,643, 015—RADIO RECEIVING CIRCUITS— E. W. HOUSE, of Birmingham, Alabama. Filed August 27, 1925, issued Sept. 20, 1927.

1,643,323—DIRECTIVE ANTENNA ARRAY—JOHN STONE STONE of San Diego, Calif. Filed Jan. 4, 1921, issued Sept. 27, 1927. Assigned to American Telephone and Telegraph Co.

1,643,781—RADIO SENDING SYSTEM—DONALD G. LITTLE of Wilkinsburg, Pa., Filed Jan. 8, 1924, issued Sept. 27, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,643,782—DEVICE FOR ALTERING THE WAVE LENGTH—SIEGMUND LOEWE, Berlin, Germany. Filed Sept. 2, 1921, issued Sept. 27, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,644,242—CONDENSER—H. P. DONLE, Meriden, Conn. Filed Sept. 22, 1923, issued Oct. 4, 1927. Assigned to The Connecticut Telephone & Electric Co., Inc.

1,644,266—ANTENNA CONSTRUCTION—FREEMAN ORNE, Dubuque, Iowa. Filed Dec. 21, 1925, issued Oct. 4, 1927.

1,644,601—VACUUM DISCHARGE TUBE—A. P. HANS-GERD NICKEL, Charlottenburg, and JOHANNES J. SPANNIER, Berlin, Germany. Filed Feb. 24, 1927, issued Oct. 4, 1927.

1,644,744—ELECTRON TUBE—H. M. PINGEN, of Toledo, Ohio. Filed March 21, 1922, issued Oct. 11, 1927.

1,644,796—AMPLIFYING AND DETECTING TUBE—H. P. STUART, of Seattle, Washington. Filed March 27, 1926, issued Oct. 11, 1927.

1,644,906—FRAME AERIAL AND THE LIKE—P. W. WILLANS, of Pattishall, Towcester, England. Filed April 5, 1926, issued Oct. 11, 1927.

1,645,280—PHOTOELECTRIC CELL—T. R. GOLDSBOROUGH, of Wilkinsburg and OTTO H. ESCHHOLZ, of Pittsburgh, Pa. Filed Sept, 28, 1926, issued Oct. 11, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,645,291—POLYPHASE PLATE CIRCUIT EXCITATION SYSTEM—A. NYMAN, Swissvale, and FRANK CONRAD, of Pittsburgh, Pa. Filed August 18, 1921, issued Oct. 11, 1927. Assigned to Westinghouse Electric & Mfg. Co.

1,641,749—VARIABLE CONDENSER—JACOB M. ENDERS, Schenectady, N. Y. Filed Sept. 22, 1923, issued Sept. 6, 1927. Assigned to General Electric Company.

1,645,462—INSULATED SUPPORTS FOR INDUCTANCE COILS—LOUIS STEINBERGER of Brooklyn, N. Y., and GUY HILL, of Washington, D. C. Filed Dec. 11, 1922, issued Oct. 11, 1927. Assigned to Wired Radio, Inc.

1,645,231—ELECTROMAGNETIC SOUND REPRODUCERS—FREDERICK DIETRICH of N. Y. and WILLIAM H. GERNS of East Orange, N. J. Filed March 5, 1926, issued Oct. 11, 1927. Assigned to Brandes Laboratories, Inc.

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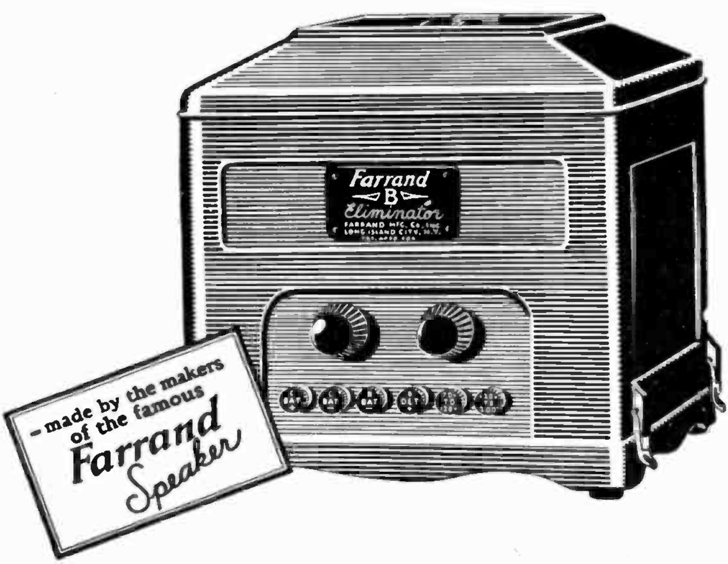

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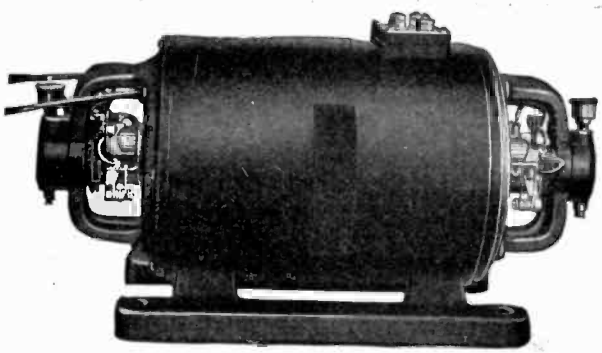

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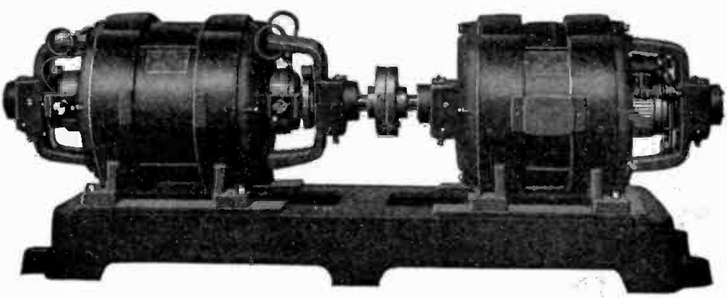

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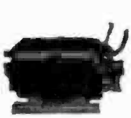

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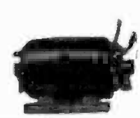

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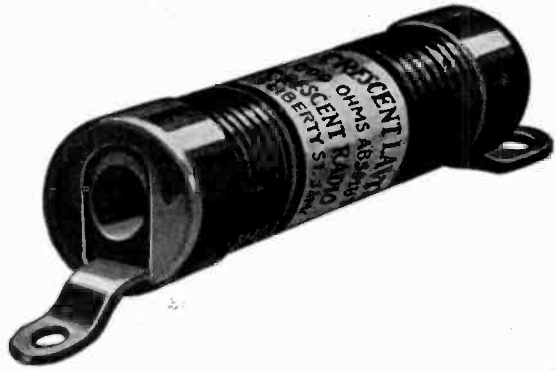

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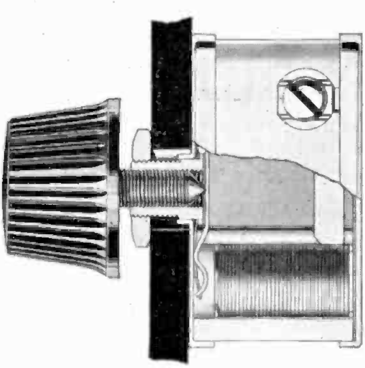

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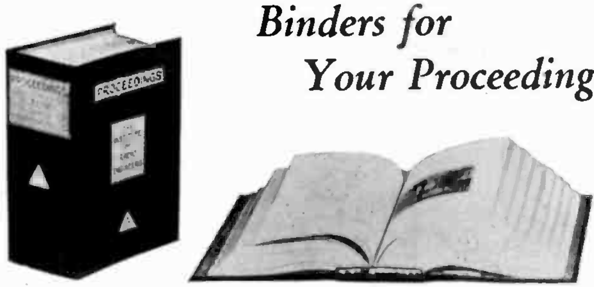

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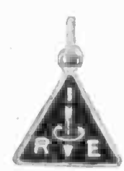

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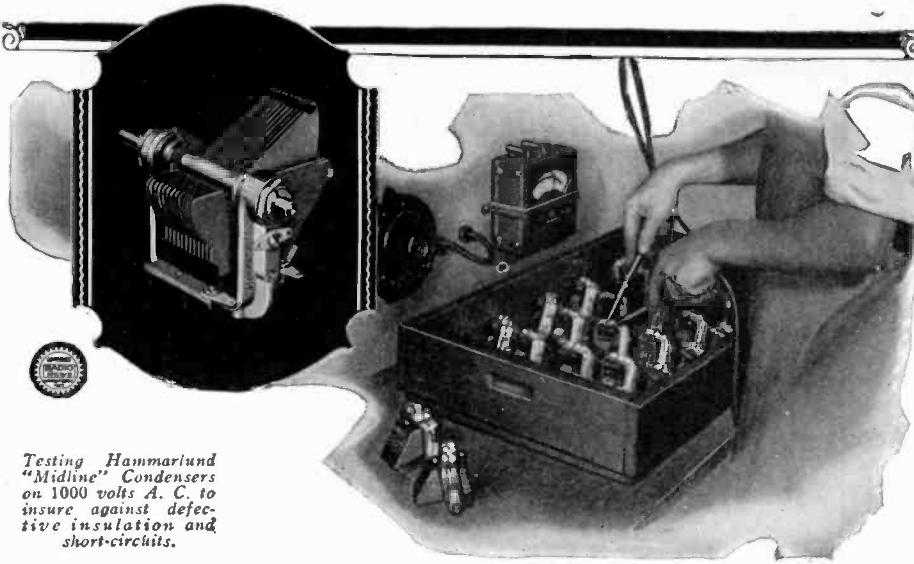

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