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The Institute of Radio Engineers

Volume 13 DECEMBER, 1925 Number 6

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GENERAL INFORMATION

The Proceedings of the Institute are published every two months and contain the papers and the discussions thereon as presented at the meetings in New York, Washington, Boston, Seattle, San Francisco, or Chicago.

Payment of the annual dues by a member entitles him to one copy of each number of the Proceedings issued during the period of his membership.

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INSTITUTE ACTIVITIES

Institute News and Notes

The first Fall meeting of the Section was held on the evening of October 8. In the next issue of the PROCEEDINGS it is hoped we shall be able to announce the meetings schedule for the Winter months. Members visiting San Francisco should plan to be present at these meetings when convenient. Meetings are held on the second Thursday of each month at the Engineers' Club.

Los Angeles Section

Preliminary work is now under way looking to the organization of a Section of THE INSTITUTE at Los Angeles, California. Col. J. F. Dillon, of the San Francisco Section, is representing headquarters in the organization plans.

Standards Committee

At the October 14 meeting in New York of the Standards Committee, the following members of the committee were in attendance: Ralph H. Bown, chairman, R. H. Marriott, J. V. L. Hogan, Donald McNicol, C. B. Joliffe, L. E. Whittemore, L. A. Hazeltine, E. H. Armstrong and H. M. Turner. The published Report should be ready for distribution to members early in the coming year.

Chicago Section

The Chicago Section of THE INSTITUTE held a meeting in the rooms of the Western Society of Engineers, Chicago, on the evening of October 9. Hugh A. Brown, assistant Professor of Electrical Engineering, University of Illinois, presented a paper on "Voice-Controlled Transmission in Radio Telephony."

Inaugural Meeting in Toronto

The new Canadian Section of THE INSTITUTE, organized in Toronto, Ontario, held an inaugural banquet and get-together meeting on the evening of October 2. The banquet was served in the dining-room of the Ward St. works of the Canadian General Electric Company. Eighty radio engineers and workers attended the meeting. Among those present were: Prof. T. R.

Rosebrugh, Prof. W. H. Price, H. C. Don Carlos, C. L. Richardson, W. J. Hevey, Keith Russel, C. S. Mallett, Prof. A. R. Zimmer, Prof. Pacent, J. J. Ashworth, F. A. Gaby, O. E. Forrest, A. S. Edgar, W. L. Amos, T. Rogers, H. Swift, W. B. Cartmel. The meeting was attended also by Donald McNicol, vice-president, New York, and by C. P. Edwards, Director of Radio Service, Department of Marine and Fisheries, Ottawa, Canada.

Washington Section

A meeting of the Washington Section of THE INSTITUTE was held in the Department of Commerce Building, Washington, on the evening of October 6. The annual election of officers took place, and a talk was delivered by Chairman A. Hoyt Taylor, on the subject: "High Frequency Phenomena."

Washington Conference

THE INSTITUTE's representatives appointed to attend the Radio Conference held in Washington, November 9, under the auspices of Secretary Hoover, were: R. H. Marriott, J. V. L. Hogan and Donald McNicol.

Membership Elections

At a recent meeting of the Board of Direction the following transfers were approved: To Fellow, Arthur F. Van Dyck. To Member, Austin Bailey, Powel Crosley, Jr., R. D. Duncan, Jr., E. H. Felix, C. M. Jansky, Jr., E. T. Jones, J. Marsten, E. L. Nelson, Arthur Nilson, R. S. Ohl, J. M. Thorburn, H. M. Wilby.

At the same meeting the following direct elections to grade of Member were approved: A. S. Blatterman, C. H. Burkhead, W. B. Cartmel, E. M. Deloraine, L. B. Henson, C. C. Jackson, H. L. Kirke, H. Levinson, L. H. Mansell, James Nelson, R. R. Pecorini, A. W. Peterson, R. J. M. Raven-Hart, F. Reichmann, E. R. Shute.

Applications Committee

The Applications Committee, of which Prof. J. H. Morecroft is chairman, holds regular monthly meetings, at which all applications in hand at the time are acted upon.

Annual Meeting in New York

President Dellinger, at the September meeting of the Board of Direction, appointed a committee to study the proposal for an annual meeting of THE INSTITUTE. Mr. J. V. L. Hogan is chairman. Report and decision in the matter shall be announced in an early issue of the PROCEEDINGS.

Membership Committee

The Membership Committee, of which Mr. L. E. Whittemore is chairman, held a meeting on October 21. Each member of THE INSTITUTE will shortly receive a circular on the subject, accompanied by an application form and postal card form, which it is hoped will receive personal attention to the end that all competent radio workers may be given an opportunity to apply for membership in THE INSTITUTE.

Institute Award

At the October meeting of the Board of Direction the Liebmann Memorial Prize for the year 1925 was awarded to Mr. Frank Conrad in recognition of the value of research work in short wave signaling carried on by him. This prize is awarded annually by THE INSTITUTE, and amounts to five hundred dollars in cash. The money award is accompanied by an appropriate letter of transmittal. Other radio engineers who have received this prize in past years are: R. A. Weagant, R. A. Heising, L. F. Fuller, C. S. Franklin, H. H. Beverage, and John R. Carson.

Radio Historical Museum

In the past year or two there has been occasional discussion on the subject of a historical museum into which might be gathered the various private collections of early relics of the art, large or small. The subject is one which should receive attention in the near future so that something may be done before the scattered, individual collections are dissipated. Members of the INSTITUTE who possess radio relics, or who know of the location of collections, are invited to forward information relative thereto to INSTITUTE headquarters. After an inventory has been made of available collections, a committee will be appointed by the President to take up the subject and see what can be provided in the way of suitable housing.

Technical Papers in Pamphlet form

Until such time as the PROCEEDINGS of the Institute are published monthly, instead of bi-monthly, as at present, some papers will be printed in pamphlet form in addition to those which appear in the PROCEEDINGS. One paper now available is "Recent Advances in Marine Radio Communication," by T. M. Stevens, of the Radio Corporation of America, New York. Copies of this paper may be obtained by writing to the Secretary of the INSTITUTE, 37 West 39th St., New York.

AN INVESTIGATION OF TRANSMISSION ON THE HIGHER RADIO FREQUENCIES*

By

A. Hoyt Taylor

(Superintendent of Radio, Naval Research Laboratory, Anacostia, District of Columbia)

The object of this paper is to present graphically and in systematic form information which is a summary of the range data for various frequencies so far as it can be estimated from the extensive experiments carried on by the Naval Research Laboratory at Anacostia, supplemented by considerable information from outside sources which has come in various ways to our knowledge.

Two further objects of this paper are, first, to indicate the regions which require further exploration, and second, to show the places where certain transition phenomena of a more or less abrupt character occur as the frequencies are varied from 100 kilocycles to 20,000 kilocycles. This should bring out the peculiarities of high frequency transmission and serve as a guide, in a general way, in formulating policies looking forward to the possible wider adoption of high frequency communication.

The attached range chart is based on the following considerations:

(a) 5 kilowatts in the antenna.

(b) Average antenna installation.

(c) Communication between points on the same meridian.

The chart is nevertheless generally applicable to east-and-west communication or any communication where there is considerable time difference between the points involved, provided due accounts are taken of this time difference. Nevertheless it must be admitted at the start that the problem is much more complicated for such a condition, especially where there is a very large number of hours of time difference between two points.

There is, of course, considerable difference between the daylight ranges, summer and winter, but not anything like the differences which occur in the night range. Therefore the line on

*Received by the Editor May 9, 1925.

the chart indicating daylight ranges must be considered as giving average ranges, summer and winter, but for the night range, the lower dotted line indicates the winter night range and the upper dotted line indicates summer night range.

A cross entered on the line indicates the limit of actual exploration and the extension of the line beyond the cross indicates the probable ranges. A "U" entered on the line, indicates, if bounded on the right and left by two arrows, a region (generally a short region) within which communication is uncertain—or in case the "U" stands further out to the right hand side of the diagram, it indicates that for ranges longer than those corresponding to the position of the "U," communication begins to become uncertain.

As an illustration, take the 4,000 kilocycle band. The daylight communication is set at about 750 miles. The summer night communication at about 7,000, but uncertain after 3,000. The winter night communication extends to 10,000 miles, but is subject to some uncertainty after 6,000. Another instance of the use of the "U" is shown in the 6,000 kilocycle band, in which there is an uncertain region between 150 and 400 miles beyond which the range again becomes certain and extends to 10,000 miles (probably), but is uncertain after 7,000. That is, this frequency has two regions of uncertainty: one at close regions and the other at very distant regions. The use of the question mark ("?") is for the purpose of indicating unexplored regions. The use of the "M" on the diagram means that the radiation "skips over" or "misses" entirely the region indicated; therefore the "M" is always bounded by arrows right and left. An instance of this is the 15,000 kilocycle band which skips over the region from 75 miles to 700.

Starting with 100 kilocycles, the ranges of which are fairly well known from the performance of certain ship transmitters, we see that the daylight range is about 1,200 miles, the summer night range 2,000 but uncertain after 800 on account of heavy strays in the summer time, and the winter night range extends to 2,500, but becomes uncertain after 2,000 on account of the strays and fading. At 200 kilocycles, we find the daylight range shortened to 800 miles, the summer night range good on the whole for a greater distance than the 100 kilocycles. This is true because there are less strays at 200 kilocycles in the summer time than there are at 100. The winter night range, however, overlaps that for 100 kilocycles, going to 3,000 miles, altho uncertain after 1,800 on account of fading. At 500 kilocycles the daylight range is still further shortened and the summer night range is not certain for any greater distance than the daylight range, but the winter night range is certain for much greater distance and the extreme winter night limit (2,000 miles) is more than three times the normal daylight range. At 1,000 kilocycles we see the daylight range still further shortened, but the night range considerably exceeding it even in the winter, whereas in the winter the extreme ranges are very greatly in excess of the normal daylight range, with, however, about half of the winter night range in the region of uncertainty. It should be stated at this point that the table is based entirely on continuous wave telegraphic communication. At 2,000 kilocycles the daylight range is cut to 125 miles and the summer night range is not a great deal better; but the winter night range is enormously greater than the daylight range; with, however, a great region of uncertainty in the winter night range, due to fading. The performance here is based on the Fleet's report of certain transmitters built by this Laboratory, and tested on shipboard. It is also based on amateur data to a certain extent.

Between 2,000 and 3,000 kilocycles, the phenomena show a rather abrupt change. At 3,000 kilocycles the daylight range is much greater than at 2,000, which is a reversal of ordinary behavior at lower frequencies. The summer night range is enormously extended and the winter night range still more so. We see that the reliable night ranges for summer jump to 2,000 miles, and as a matter of fact this figure is probably considerably underestimated. It is, however, desired to make the chart conservative at least in its application to the higher frequencies. The night ranges in the winter time, however, are certain up to 6,000 miles. Comparing 3,000 kilocycles with 100 kilocycles, we find the 100 kilocycles excels in daylight range, but that the 3,000 kilocycles greatly excels in possible night ranges. At 4,000 kilocycles we see the daylight range extended to 750 miles, the summer night range certain to 3,000, with a possible night range to 7,000; while the winter night range is certain to 6,000 with possible ranges to 10,000. At 6,000 kilocycles the daylight range extends to 1,000 miles and the summer night range to 4,000 with possible ranges to 7,000, whereas the winter night range probably goes to 10,000, but has not been fully explored. We may consider it uncertain at least after 7,000.

A new and interesting phenomenon makes its presence felt for the first time in the diagram regarding an uncertain period within short distances during the winter night transmission;

namely, between 150 and 400 miles. In the next line on the diagram for 7,500 kilocycles, the daylight range has been further extended to 1,200 miles, thus nearly equalling the 100 kilocycle transmitter, but an uncertain region not far from the transmitter has been introduced between 100 and 350 miles during the summer night range and a skip, or entirely-missed, region, occurs in winter night ranges between 100 and 350 miles. This frequency has been explored to 4,000 miles, but it is believed that it will carry very much further. We may say that it is uncertain, however, after 8,000 and probably will carry to 10,000 on the winter nights. At 10,000 kilocycles we see that the "jump" or "miss" occurs both summer and winter and also by daylight, but the daylight jump is only about 500 miles, whereas the summer night jump is very great indeed. Very little exploration has been made of this frequency for nocturnal transmission, but there is good reason to suspect that the summer night jump is in the neighborhood of 2,000 miles and the winter night jump possible 4,000 miles. There is also good reason to believe that frequencies not far from 10,000 kilocycles can be successfully used for extreme night ranges. (In this connection, there should be noted Samoa's successful reception and practically steady intensity of tests from Schenectady, New York, on tests of 35 meters during night hours.)

Fifteen thousand kilocycles: Here the daylight jump is increased to between 600 and 700 miles and an uncertain region follows this to 1,000 miles, but beyond this, as far as the exploration has gone (4,000 miles), results are excellent. It is impossible to say what the daylight range will come to beyond 4,000 miles. It is known, however, that the missing region or the jump in this frequency is very great at night, both summer and winter. Very little exploration has been made here, but there are some data indicating that this frequency can be successfully used at 10,000 miles even at night. This statement is based on the establishment of two-way communication with Australian 2-CM (Sydney). He used 21 meters and this station used 20.8 meters, between 1 A. M. and 2.30 A. M. Eastern Standard Time. It was broad daylight in Sydney when the test commenced. It should be noted at this point that this Laboratory has so far not used more than 750 watts in the antenna in the twenty-meter band. Australian 2-CM was using still lower power. It is evident, then, that figures estimated for 5 kw. in the antenna, but actually based on experiments with less than 1 kw., ought to be fairly conservative.

At 20,000 kilocycles the exploration is very scanty indeed, but there is information at hand giving the distance of the daylight jump as in the neighborhood of 1,500 miles with an uncertain region extending to 2,500 miles and which probably is followed by a certain region for a considerable distance beyond. Absolutely nothing, however, is known of night ranges on these frequencies. The ranges of the direct, or earth-bound components are well enough known; they are about 60 to 70 miles for 15,000 kilocycles and in the neighborhood of 40 to 50 miles for 20,000 kilocycles.

It is not the purpose of this particular paper to go into details of the vast amount of information upon which the range chart is based, nor to divulge at this particular time the theory which is gradually forming in the minds of the engineers of this Laboratory, which we believe will account for these curious effects. The purpose is rather to serve as a practical guide to indicate what ranges may be covered at different frequencies and what ranges remain to be explored, and what we hope to get in the un-explored regions.

It may be stated at the present time, however, that some of the most valuable information confirming earlier data on the matter of the "skip" or "miss" region was obtained from the daily reports made by Major J. O. Mauborgne, United States Army, who took observations from an Army Transport, on 16, 32, 20.8, and 41.7 meters, all the way from New York to Panama. It is believed that the information concerning the uncertain regions and "skip" regions is fairly definitely known for daylight work. If any special criticism could be made of the range chart it would be that it underestimates the summer night ranges on 3,000 kilocycles.

When one considers the chart as a whole, the high frequencies show clearly their enormous superiority from a point of view of economy on power consumption and general cost, and further it is possible to obtain ranges with high frequencies which we cannot hope to equal with almost any practical amount of power on lower frequencies.

To apply the chart to east-and-west communication, we must, for the present, consider that during the hours when daylight obtains over the entire stretch, we apply the daylight range data. For the hours which night obtains over the entire stretch, we apply the night data. In the intermediate hours when part is sunlight and part dark, much further exploration will have to be made, but we do know that a sort of compromise condition

does exist and it does appear further that a 5 kw. transmitter, equipped with about four frequencies, would be in a position to obtain highly creditable ranges at any time of either day or night, and whether for north-and-south communication, or for east-and-west. We must, however, at present, until further information comes in from stations like Samoa, Guam, and Cavite, be forced to believe that the east-and-west problem is more difficult of practical solution.

It must be understood in referring to the range chart that estimates on range and references to the "missing" of "skipped" areas on the higher frequencies refer in the case of the daylight ranges to conditions existing in the middle of the day, and for the night ranges, to conditions existing in the middle of the night. For west-and-east work this must be interpreted as meaning conditions when the sun is half-way between the two meridians under consideration. It is well known, of course, that there is a more or less gradual transition from daylight to dark conditions; in fact it is not nearly as abrupt as one would anticipate it to be, especially in the summer time.

Since the first part of this paper was written, Samoa has reported successful reception of the 20.8 meter wave from the Naval Research Laboratory as early as 8 P. M. zone-plus-five time, which means that 6,000 out of the 7,000 miles between Washington and Samoa were traversed in daylight. Also it is well known that 20-meter signals from amateurs on the West Coast are now received as late as midnight or 1 A.M. zone-plus-five time, which could not have been done during winter nights. This is interesting as showing a gradual change in the skipped region for 20 meters. The skipped region is less in the daytime, gradually increases to somewhat less than 2,000 miles in the summer nights and very likely is considerably in excess of this in the winter nights, altho it is not known with certainty whether it ever comes down to earth again in the winter nights. One must therefore, conceive of the skipped distance on the higher frequencies undergoing a lengthening process as the night wears on, followed by a shortening process as daylight approaches. Most of the information on the higher frequencies must, of course, be considered as incomplete subject to future revision. Nevertheless certain fundamental things in the behavior of these frequencies seem to be quite definitely established.

Balboa also reports satisfactory reception of the 20.8-meter wave thruout the 24 hours at this time of the year, but it is not anticipated that this will be possible in the winter

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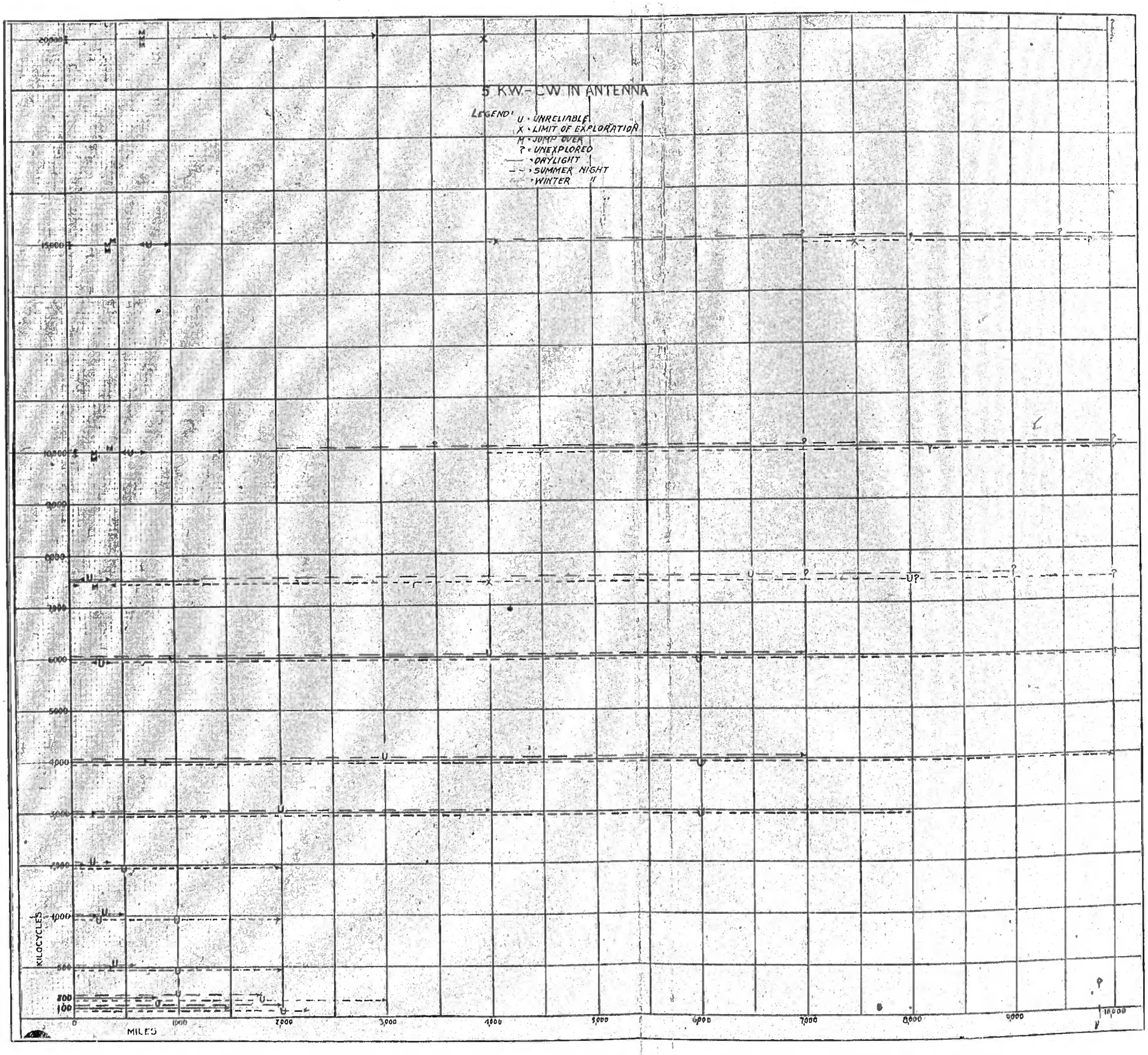
5 K.W.-C.W. IN ANTENNA
LEGEND:
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M = JUMP OVER
? = UNEXPLORED
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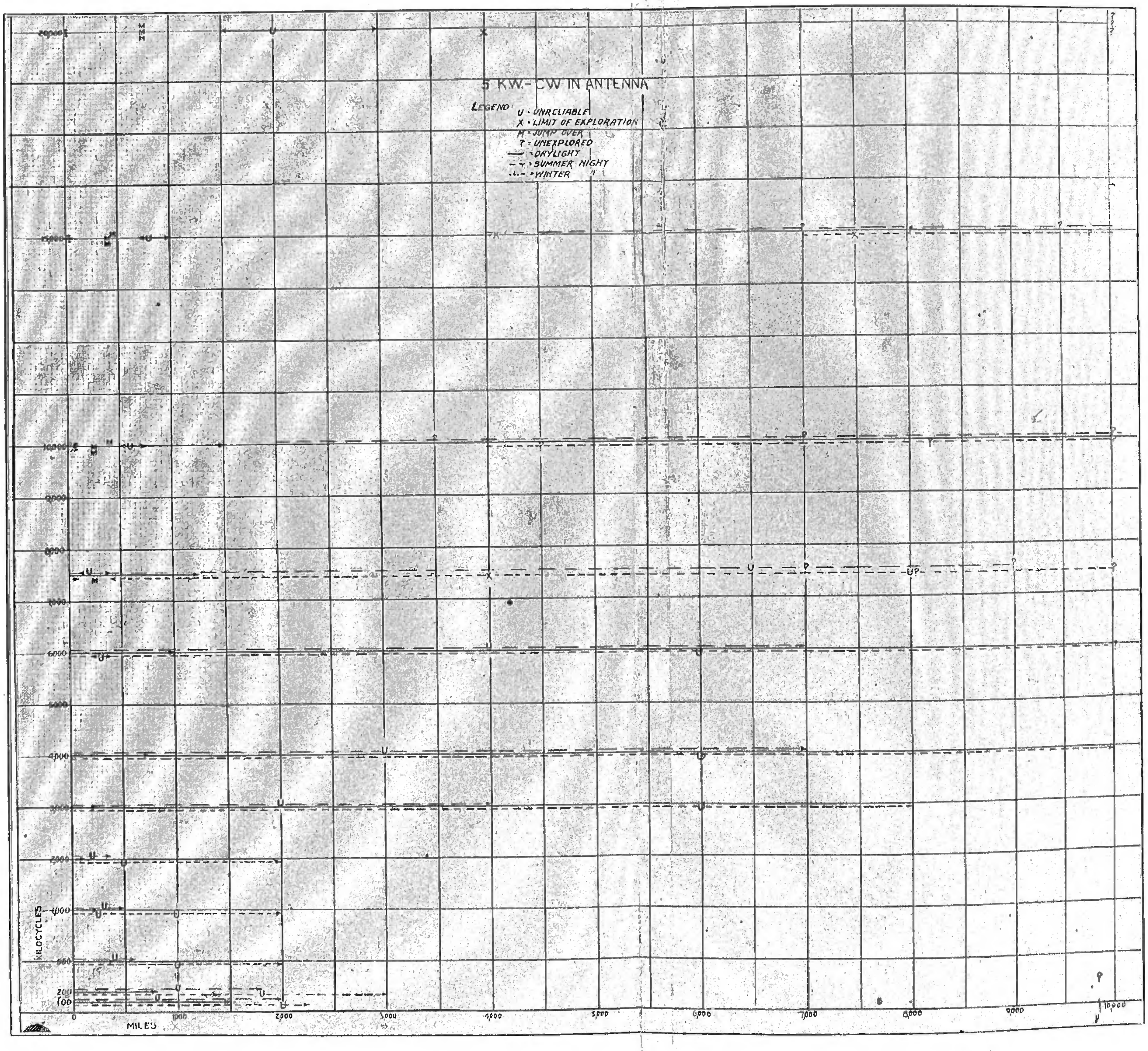
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April 21, 1925.

SUMMARY: A preliminary range chart has been constructed for telegraphic communication, 5 kw. in the antenna, at various frequencies. The conclusions upon which the range chart is based are derived from experiments made by the Naval Research Laboratory, from experiments made by amateurs, and upon such data as the Laboratory has had access to from commercial and Government sources at home and abroad.

An attempt has been made to indicate in a general way the advantages and disadvantages of high frequency telegraphic transmission. Various critical regions are pointed out where new phenomena make their appearance: in particular these regions are (1) the region between 2,000 and 3,000 kilocycles where daylight ranges begin to increase with increasing frequency at the same time that the night ranges show extremely great increase and a degree of reliability which would be wholly unanticipated from obervations made at frequencies lower than 2,000 kilocycles. (2) A region around 6,000 kilocycles where an uncertainty develops during the winter nights at relatively very short ranges. (3) The development at successively higher frequencies of this uncertain range into the missing region which is most pronounced in the winter nights, and finally as the frequency is increased makes itself felt in the summer nights and at still higher frequencies even in the daytime.

The development of this missing region to extensive areas is shown to take place with frequency rise to 20,000 kilocycles. The chart also attempts to indicate, in a general way, the region of uncertain communication and the regions where further exploration is urgently needed. It is quite evident that the range data are far from complete and that many individual cases will be found in contradiction to the chart, but it does represent a sort of general average of the situation as it presents itself to the Engineers in the Naval service.

It is hoped that the publication of this data will promote useful discussion and collaboration in this new and interesting field. The data would no doubt have to be modified materially to make it apply to any highly directive system of transmission.

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SUMMARY: A preliminary range chart has been constructed for telegraphic communication, 5 kw. in the antenna, at various frequencies. The conclusions upon which the range chart is based are derived from experiments made by the Naval Research Laboratory, from experiments made by amateurs, and upon such data as the Laboratory has had access to from commercial and Government sources at home and abroad.

An attempt has been made to indicate in a general way the advantages and disadvantages of high frequency telegraphic transmission. Various critical regions are pointed out where new phenomena make their appearance: in particular these regions are (1) the region between 2,000 and 3,000 kilocycles where daylight ranges begin to increase with increasing frequency at the same time that the night ranges show extremely great increase and a degree of reliability which would be wholly unanticipated from obervations made at frequencies lower than 2,000 kilocycles. (2) A region around 6,000 kilocycles where an uncertainty develops during the winter nights at relatively very short ranges. (3) The development at successively higher frequencies of this uncertain range into the missing region which is most pronounced in the winter nights, and finally as the frequency is increased makes itself felt in the summer nights and at still higher frequencies even in the daytime.

The development of this missing region to extensive areas is shown to take place with frequency rise to 20,000 kilocycles. The chart also attempts to indicate, in a general way, the region of uncertain communication and the regions where further exploration is urgently needed. It is quite evident that the range data are far from complete and that many individual cases will be found in contradiction to the chart, but it does represent a sort of general average of the situation as it presents itself to the Engineers in the Naval service.

It is hoped that the publication of this data will promote useful discussion and collaboration in this new and interesting field. The data would no doubt have to be modified materially to make it apply to any highly directive system of transmission.

A NEW DIRECTIONAL RECEIVING SYSTEM*

By

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Introduction

Reduction of static interference, or to state it more correctly, reduction of the ratio of static to signal, has been, almost since the beginning of the radio art, the most important problem in radio engineering. It is now well known that static disturbances have definite points of origin and that the impulses which are detected at a receiving station have definite directions of propagation. A receiving system having no directional selectivity is, therefore, affected by static impulses from all directions and, in spite of many inventions, it has not yet been possible to improve its signal-static ratio except by limiting the frequency band transmitted. A system which, however, is so designed as freely to receive waves arriving from a limited range of directions is susceptible only to static disturbances propagated within that range, and large improvements in signal static ratio have been claimed for different types of directive antenna systems during the past few years.[1]

A directional receiving system for radio telephony in which directional selectivity is obtained by combining the output voltages from two antennas is described in this paper. The main feature of the system is the arrangement for controlling the output voltages of the antennas, so that they may be combined to neutralize each other or to reinforce each other as desired. A double detection (super-heterodyne) receiver is employed and the output voltages, which are combined so as to produce the directional characteristic, are the intermediate frequency currents due to the waves received by the antennas and the beating oscillator currents. The control of these output voltages is effected by operating upon the beating oscillator currents.

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[1] See "The Wave Antenna," by Beverage, Rice and Kellogg, "Journal of the American Institute of Electrical Engineers. This paper contains an excellent list of all publications on directional systems up to date.

Requirements of Directive Antenna Systems

The impulsive nature of most static makes it impossible to discriminate against it by frequency selective means, the effect of the impulses being to set up in the selective circuits free oscillations having a frequency and damping determined by the natural periods and the time constants of the circuits. It follows, then, that static effects in a plurality of antennas can be made to neutralize each other only if the *natural periods* and *time constants* of the antenna circuits are alike. It may be emphasized that this paper deals with reception of speech signals which require a much wider frequency band of the receiving set than telegraph signals. In a receiving set for telegraph reception, which responds only to a very narrow frequency band, perhaps a few hundred cycles, it is not so important that the antenna circuits be exactly alike.

In a multi-antenna system, it is necessary to control the phase and intensity of the currents delivered from each independent antenna circuit, in order that their mutual reinforcement or neutralization may be properly accomplished. Hitherto this has been done by incorporating phase and amplitude control apparatus directly in the antenna circuits, and the addition of such apparatus has made it extremely difficult to achieve the requisite equality of the antenna circuits. For this reason, the use of directional receiving systems has been very limited and has been almost entirely restricted to the long wave signaling systems. Alexanderson[2] in his paper on the Barrage receiver recommends always making the antenna circuits aperiodic, but such high resistance circuits are very inefficient, and they will increase the "set noise"[3] very much.

By placing the means for controlling the output voltages in the beating current circuit which is substantially independent of the antenna circuits, these may be reduced to a simpler form and may readily be made accurately alike and more efficient, thereby making it possible to receive waves of any radio frequency with good directional selectivity.

It should be observed that directional discrimination against continuous waves of frequencies, different from the frequency band to which the antenna circuits are tuned, is not necessary as the frequency selection effected by these circuits and the inter-

[2] Proceedings of The Institute of Radio Engineers, volume 7, page 363, August, 1919.

[3] The "set noise" is defined as the field strength of the weakest signal that can be received by the set when there is no static, and is determined by the amplifier noises.

mediate frequency circuits is sufficient to suppress all interference from such waves. Further, since the effect of static disturbances is to produce free oscillations in the antenna circuits at the frequencies to which the circuits are tuned, it is evident that the directional discrimination against continuous waves will be effective also against static impulses. This has been checked by experiments.

The receiving system of Figure 1 comprises two similar an-

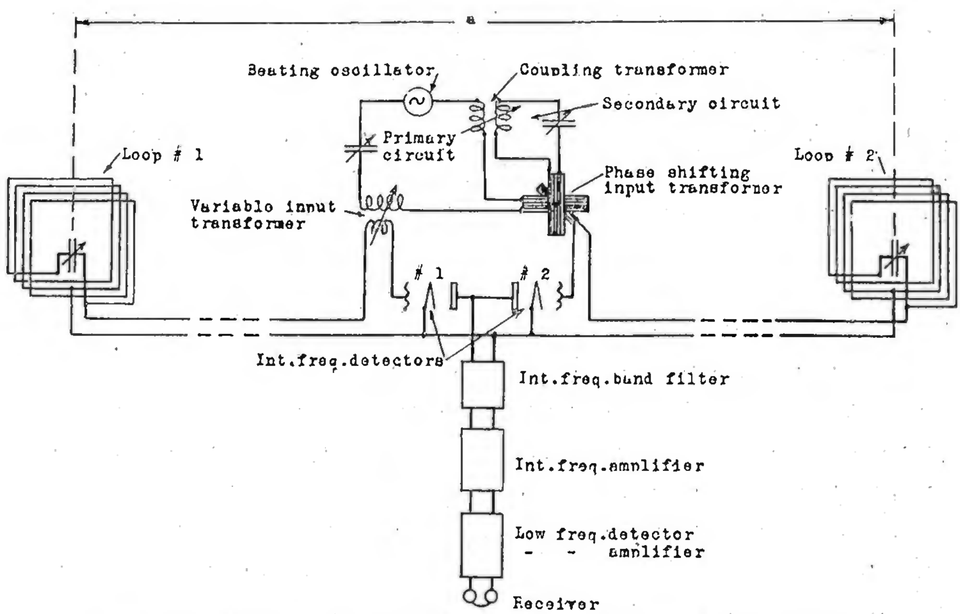


FIGURE 1—Schematic Circuit Diagram of "Two-Loop System"

tenna circuits, 1 and 2, the antennas (each of which is a loop) being so disposed as to produce in the two circuits emf's which are, in general, different in phase. The mid-points of the loops are connected to the "ground" of the set in order to reduce open antenna effects. The loops are arranged to have their planes substantially coincident and perpendicular both to the ground and to the plane of the desired wave. Their centers are spaced apart a distance a, and it will be shown later that this distance should preferably be one-twelfth of the wave-length it is desired to receive. An intermediate frequency detector tube is used for each antenna, the plate of these tubes being connected in parallel to the intermediate frequency filter. The secondary winding of a variable coupling transformer is inserted in series with the lead connecting the end of loop number 1 to the grid of its detector tube. This transformer supplies the beating oscillator current to the antenna circuit 1. The beating oscillator current is supplied to antenna circuit 2 thru a phase-shifting transformer which comprises two fixed coils having their planes mutually perpendicular and a third coil which can rotate inside the fixed coils. One of the fixed coils, the primary windings of the beating current input transformer to antenna circuit 1 and of a coupling transformer, and a tuning condenser make up a primary circuit tuned to the beating current frequency. The other fixed coil is part of a secondary tuned circuit which is loosely coupled to the primary circuit. At resonance, the currents in the primary and the secondary circuit are 90° out of phase and any desired phase angle of the beating current input to antenna circuit 2 can therefore be obtained by rotating the phase coil. The rest of the set, namely, the intermediate frequency filter mentioned above, the intermediate frequency amplifier, the low frequency detector and low frequency amplifier are well-known apparatus of the type usually employed in double detection sets and no detailed description is necessary.

The Operation of the System

It will be assumed that the system is intended to receive waves of frequency f propagated horizontally and in the direction from right to left in the diagram, Figure 1. The adjustments that should be given to the system and the preparations that should be adopted for the antenna spacing will be determined by considering the reception of horizontally propagated waves of the frequency f arriving in a direction at any angle β to the desired direction of reception (see Figure 2). As a measure of the directional effect will be taken the intensity of the resultant intermediate frequency current thru the intermediate frequency filter. The manner in which this varies with the direction of propagation, assuming a constant field intensity, defines the directive selectivity.

Two factors enter into the variation of the emf's produced

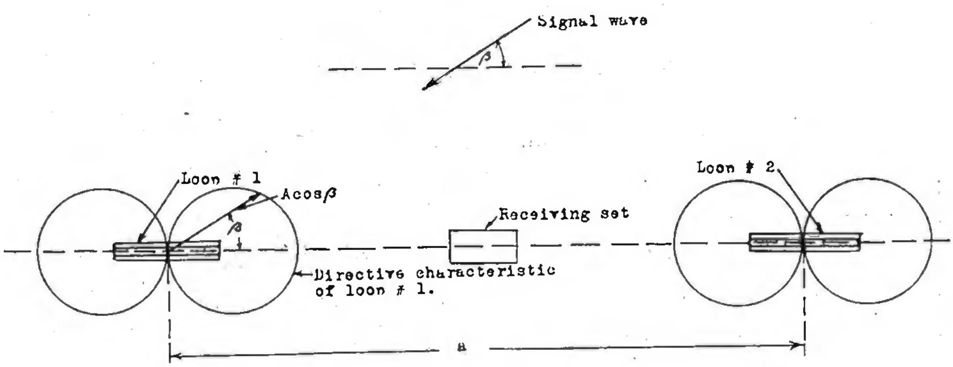


Figure 2—Top View of "Two-Loop System"

in the antenna circuits 1 and 2 by the waves. These are, first, the amplitude variation due to the directive properties of the individual loops, and second, the phase difference between the two emf's which depends upon the effective separation of the loops in the direction of propagation. The instantaneous values of the two emf's impressed upon the intermediate frequency detectors from the antennas may, therefore, be expressed by the equations:

$$\begin{aligned} e_1 &= A_1 \cos\beta \cos 2\pi f t \\ e_2 &= A_2 \cos\beta \cos\left(2\pi f t + \frac{2\pi a \cos\beta}{\lambda}\right) \end{aligned} \tag{1}$$

where A_1 and A_2 are the amplitudes corresponding to zero value of the angle β, that is, for the favored direction of propagation, and λ is the wave length corresponding to the frequency f. It is preferable to use two loops exactly alike, in which case the values A_1 and A_2 are the same. The subscripts 1 and 2 here, and in what follows, refer to the circuits 1 and 2, respectively.

From the beating oscillator there will also be impressed upon the detectors, emf's of a frequency f_o and of relative phases and amplitudes determined by the adjustments of the input transformers. The values of these emf's may be expressed by the equations:

$$\begin{aligned} e_1' &= B_1 \cos 2\pi f_o t \\ e_2' &= B_2 \cos (2\pi f_o t + \varphi) \end{aligned} \tag{2}$$

in which φ is the phase difference determined by the adjustment of the phase shifting input transformer, and B_1 and B_2 are the amplitudes of the emf's.

The important components of the currents in the output filter following the intermediate frequency detectors are those of the difference frequency $f-f_o$. These components are given by the equations:

$$\begin{aligned} i_1 &= K_1 A_1 B_1 \cos\beta \cos 2\pi(f-f_o)t \\ i_2 &= K_2 A_2 B_2 \cos\beta \cos\left[2\pi(f-f_o)t + \frac{2\pi a \cos\beta}{\lambda} - \varphi\right] \end{aligned} \tag{3}$$

the first of which represents the output current as a result of the intermodulation of the emf's e_1 and e_1' and the second of which represents the output currents due to the emf's e_2 and e_2'. The factors K_1 and K_2 involve the detecting efficiencies of the detectors and the total impedance of their output circuits. They will, in general, not be exactly alike but their ratio will be practically constant and they will not affect the relative phases of the emf's, due to the fact that the two detectors have the same output circuit. The resultant effect in the signal reproducer is proportional to the sum of the two currents of equation (3). Since it is required that the two currents neutralize each other for certain directions of propagation, it is obviously necessary that their amplitudes must be made equal, that is

$$K_1 A_1 B_1 = K_2 A_2 B_2 = I \tag{4}$$

This may be done by varying the ratio of the beating current inputs B_1 and B_2 by means of the variable input transformer to antenna circuit 1. The amplitude of the resultant current is now given by the equation:

$$I_\beta = 2\, I \cos\beta \cos\left[\frac{1}{2}\left(\varphi - \frac{2\pi a \cos\beta}{\lambda}\right)\right] \tag{5}$$

It is evident that the resultant current will be zero under two conditions, first, when β equals $\frac{\pi}{2}$, that is, when the wave is propagated at right angles to the common plane of the two loops and second, when

$$\varphi = \frac{2\pi a \cos\beta}{\lambda} + \pi \tag{6}$$

One of the most useful adjustments, altho not necessarily the best under all circumstances, is that under which the signals from waves propagated in the opposite direction from the favored are neutralized. The angle β corresponding to this case is equal to π and the requisite phase displacement of the two beating current inputs is $\varphi = \pi - \frac{2\pi a}{\lambda}$[4]. The directive characteristic of a system having this adjustment is given by the equation

$$I_\beta = 2\, I \cos\beta \sin\left[\frac{\pi a}{\lambda}(1 + \cos\beta)\right] \tag{7}$$

which is derived directly from equation (5). The signal producing current corresponding to waves received in the favored direction is denoted by I_o, which is the value of I_β for the special case of $\beta = 0$ or

$$I_o = 2\, I \sin\frac{2\pi a}{\lambda} \tag{8}$$

Equation (8) indicates that the distance between the loops

[4] Note that φ is a function of wave length λ, which means that a complete balance can be obtained only for one frequency in the signal. The maximum relative changes of the frequencies in a speech signal will, however, always be so small that the amount of unbalance will be negligible.

should not exceed a quarter of the wave length $\left(\frac{2\pi a}{\lambda} = \frac{\pi}{2}\right)$. For small separations, the signal current I_o increases proportionally to the separation of the loops and reaches half the maximum possible value (or the same value as one loop alone would give) when the separation is one-twelfth of the wave length. For separations greater than this, the gain increases more slowly than the separation.

The ratio of amplitude of the current produced by a wave propagated in the direction β to that of the current produced by a wave of equal intensity propagated in the favored direction, is expressed by the equation

$$R = \cos\beta \frac{\sin\left[\frac{\pi a}{\lambda}(1+\cos\beta)\right]}{\sin\frac{2\pi a}{\lambda}} \tag{9}$$

As the separation of the loops is reduced, the ratio R approaches the limiting value

$$R' = \frac{\cos\beta}{2}(1+\cos\beta) \tag{10}$$

R' has maximum values for β equal to 0°, 120°, and 240° and has zero values at angles of 90°, 180°, and 270°. The ratio is naturally equal to unity when β is zero. The maximum ratios recurring at 120° and 240° are equal to 0.125. For one-quarter wavelength the ratio R is 0.1923 for angles of 120° and 240°.

The general form of the directive characteristic is shown in Figure 3, in which polar co-ordinates are used. The dotted figure represents the characteristic corresponding to an antenna separation of one-quarter wave length, and the full line that for the limiting condition which is closely approximated for all antenna spacings of less than one-twelfth wave lengths. It is evident that a distinct advantage in respect to the relative amounts of signal and interference received is to be gained by reducing the antenna separation to one-twelfth of a wave length, altho this is accompanied by a reduction of signal strength. To decrease the spacing still further will economize on the land required for the system, but will require, first, more amplification with a corresponding increase in "set noise" and, second, a higher degree of stability of the system.

It has already been pointed out that neutralization of the interference may be secured in other directions than that considered in the foregoing analysis, and it may happen in practice

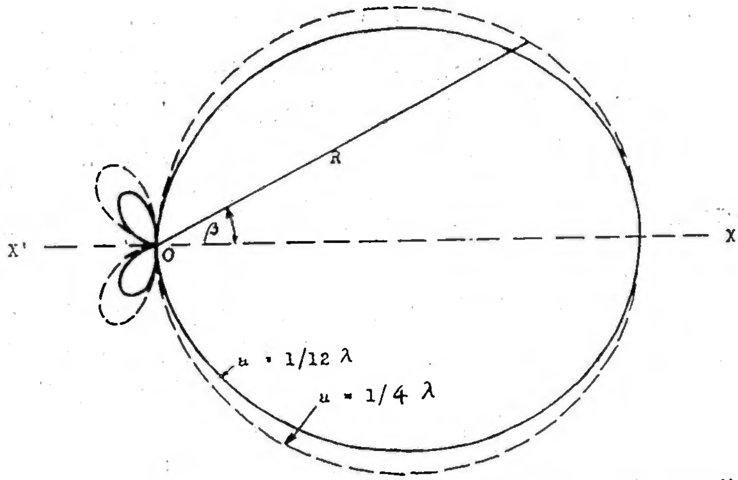


FIGURE 3—Directive Characteristic of "Two Loop System"

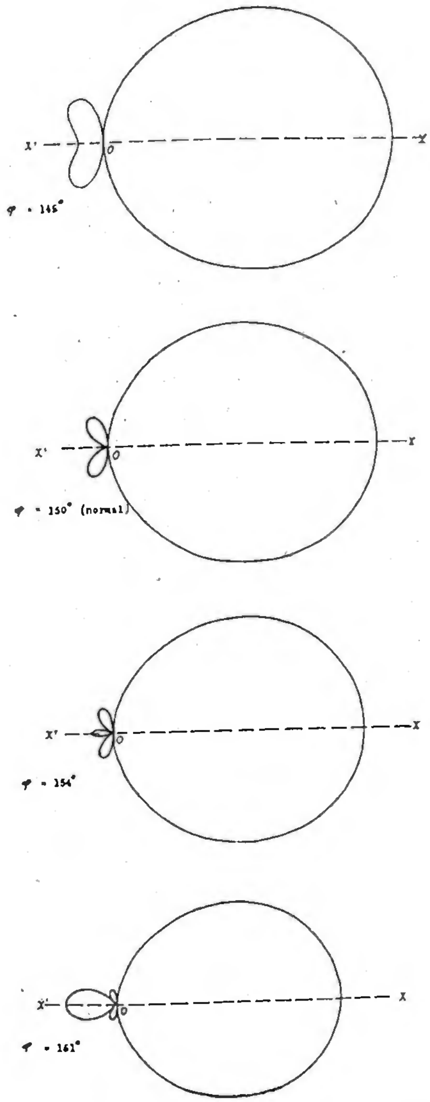


FIGURE 4—Directive Characteristics of "Two-Loop System." ($a = 1/12$)

that the most troublesome interference may be that arriving from some other general direction. The simplicity of the phase control arrangements and the fact that the control can be exerted without disturbing the equality of the antenna circuit constants, make it a very simple matter to secure the most efficient suppression of interference under any circumstances. All that it is necessary to do is to adjust the secondary coils of the two beating current input transformers until the signal is heard with the least amount of interference.

The effect of changing the phase angle φ is shown in Figure 4, in which directive characteristics are calculated for φ equal to 145°, 150°, 154°, and 161°, all corresponding to a loop separation of one-twelfth of the wave length. The curves show that the setting of the phase angle is not very critical where it is desired to reduce fairly evenly distributed static interferenice.

EXPERIMENTAL VERIFICATION

Short Wave Two-Loop System

The system described in the foregoing was tried out at Cliffwood, New Jersey, with signals in the broadcast frequency range because the short wave length of these signals make it possible to mount the two loops on a construction similar to a turn-table which is capable of rotation in a horizontal plane. The photographs in Figures 5 and 6 show the general construction, and the receiving set itself is shown in the illustrations, Figures 7 and 8. The receiving set is installed in the 6×6-foot house, shown between the two bridges in Figure 5. The whole construction rotates around a bolt in the center of the house, the house being mounted on four 8-inch truck casters and the outside ends of the bridges carrying the loops being supported by ordinary wheelbarrow wheels. The distance between the loops is 34.4 meters, giving an upper limit of the wave lengths of signals to be received of approximately 600 meters. The loop diameter is 1.76 meters and each of the loops are wound with 6 turns of bare number 16 copper wire. One loop can be turned a few degrees as it is quite important that the loops point in the same direction.

The circuit is the same as that shown on Figure 1, except that the tuning condensers for the loops are mounted in the receiving set in order to improve the facility of tuning adjustments. The ends and mid-point of each loop are connected respectively to the tuning condensers and ground of the set by 3 wires suspended in a horizontal plane and 30 cm. apart.

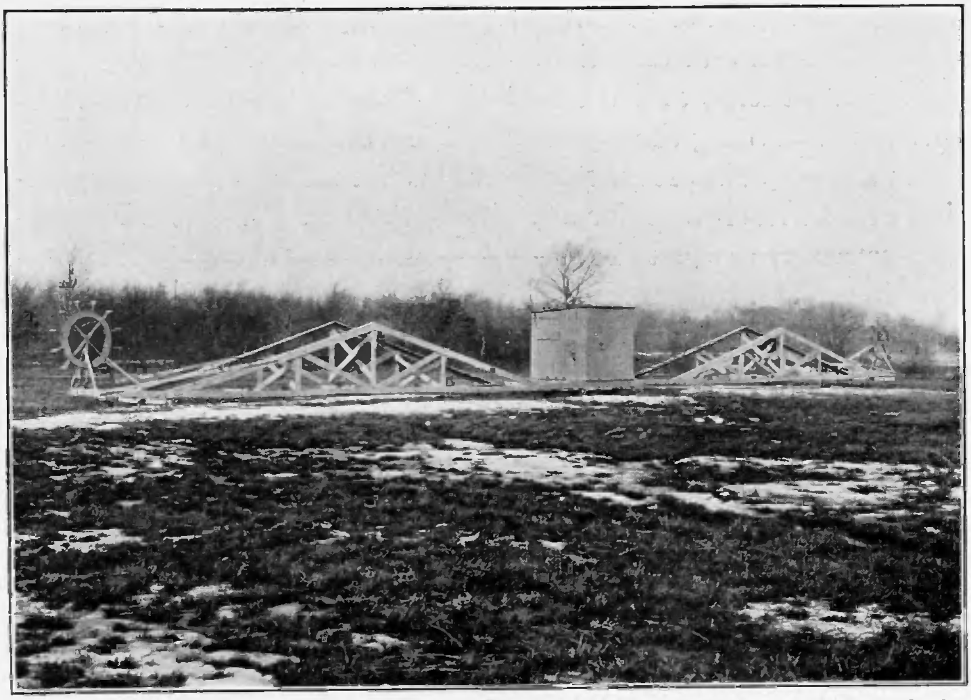


FIGURE 5—Short Wave Two-Loop System. Shows the Two Loops and the House Containing the Receiving Set

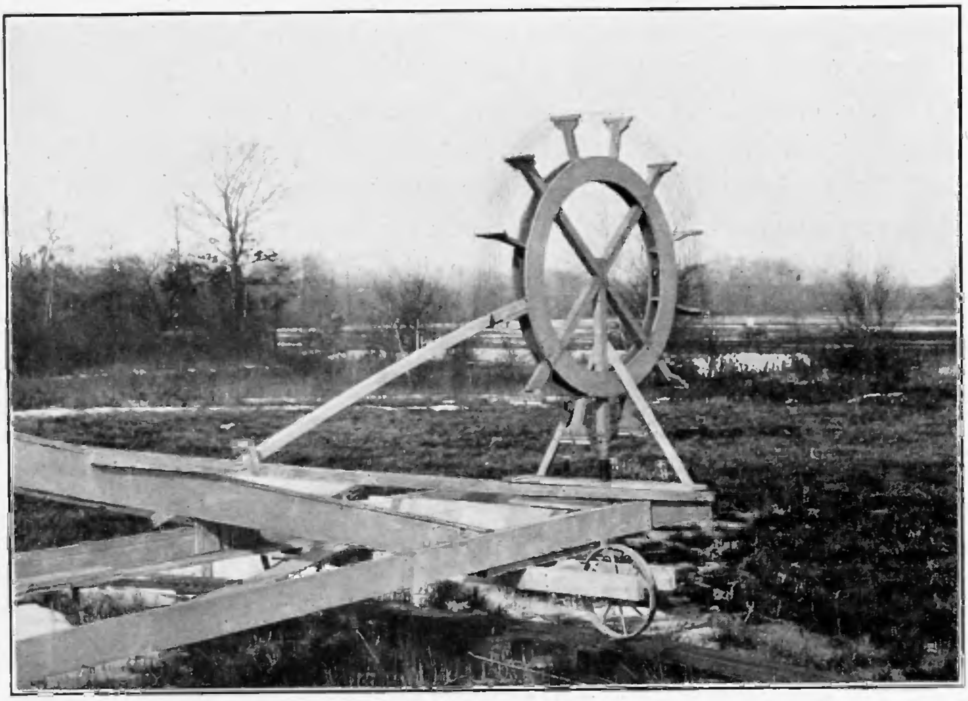


FIGURE 6—Short Wave Two-Loop System. One of the Loops

Figure 9 shows an experimental directional characteristic of this system. The incoming signal, in this case from station WEAF ($\lambda = 492$ meters), was first balanced out with the loops pointing towards WEAF ($\beta = 180°$), and then the whole system was turned. The radius R in the diagram is proportional to the input voltage on the low frequency detector and there is seen to be a good agreement between the experimental curve and the theoretical curve shown in Figure 3.

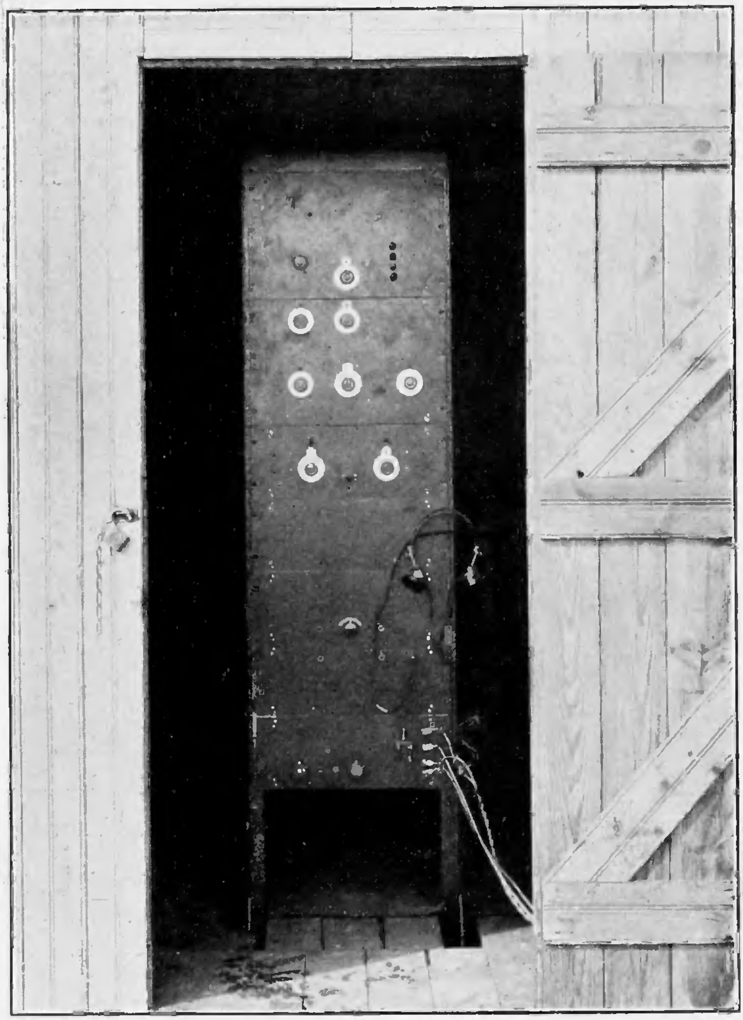


FIGURE 7—Short Wave Two-Loop System. Front View of Receiving Set

It may be pointed out that the tuning of the system to a signal is quite simple. First, one of the loops is short-circuited and the other loop tuned and the beating oscillator frequency adjusted as for an ordinary double detection receiver. Then the beating oscillator circuits are tuned up and finally the previously short-circuited loop is tuned. The set is now ready for the two adjustments of beating oscillator inputs giving a minimum of interference. It is quite convenient, especially at long wave

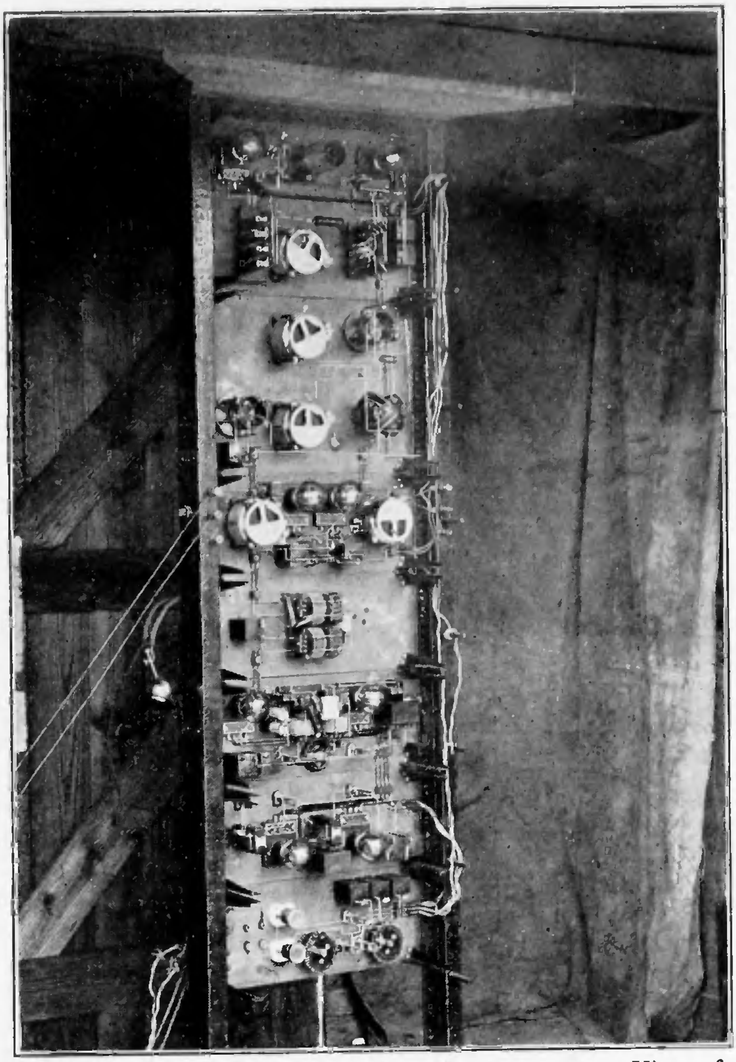


FIGURE 8—Short Wave Two-Loop System. Rear View of Receiving Set

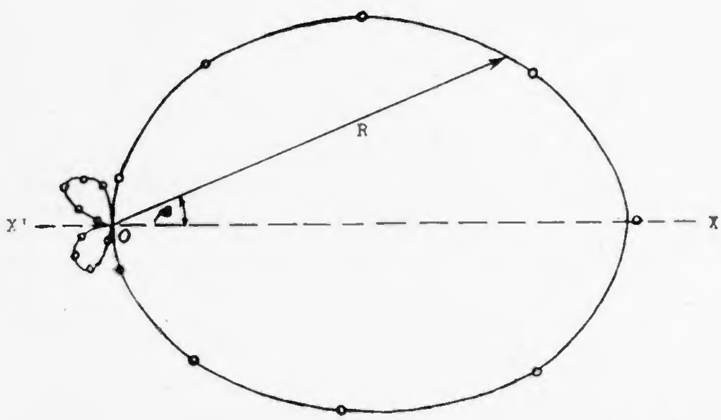


FIGURE 9—Experimental Directive Characteristic of "Two-Loop System"

lengths, to tune the set on a local oscillator, the frequency of which is adjusted to zero beat with the desired signal.

The short wave system described above was tested during the summer months, 1924, at Cliffwood, New Jersey, and found to verify all conclusions derived from the shape of its directional characteristic. The reduction in spark interference when receiving signals from broadcast stations in Philadelphia was especially noticeable. The spark interference at Cliffwood, New Jersey, is mainly due to stations along the shore and on ships around New York Harbor and the New Jersey coast, so that the interfering waves are coming from "behind" when the system is adjusted for receiving Philadelphia stations. On many occasions it was possible to reduce the summer static interference so much that talk from broadcasting stations which was absolutely unintelligible when received on one loop alone, was made clearly intelligible by the two-loop system. It may here be pointed out that static interference at broadcast frequencies, in the summer, is mainly due to local thunderstorms and it is therefore generally directive, but the direction is quite arbitrary. A gain in regard to reduction of static interference can, therefore, not always be expected since the static may come from the same direction as the signal wave. At long waves the direction from which static waves arrive is generally southwest, so that a considerable reduction in static may be expected when receiving signals from Europe.

Long Wave Two-Loop System

A receiving system suitable for reception of single sideband speech signals of 5,000-6,000 meters wave lengths was constructed and tested out during the fall of 1924 at Cliffwood, New Jersey. The locations of the receiving set and the two loops were permanent, as it is obviously not practicable to rotate a system 400 meters long like a turn-table. However, this limitation is not so important because long wave receivers are generally used for reception of signals from one direction only. The main object was to receive signals from Europe and the two loops were therefore located in a vertical plane at an angle of 60° east of north and 400 meters apart. It was found later on, however, that it is equally important, when the loops are located, to consider the direction from which the interfering waves arrive. If, for instance, the angle between the average direction of the interfering waves and the signal waves is only 90°, then quite an improvement can be obtained by making the angle β between the signal wave direction and the plane of the two loops 30° instead of 0°

(see diagram in Figure 10). During the winter months it was found that static interference was coming from the southwest during the night and from south-southeast during the day, in which case it is preferable to have two sets of loops and switch from one set to the other according to the direction of the static.

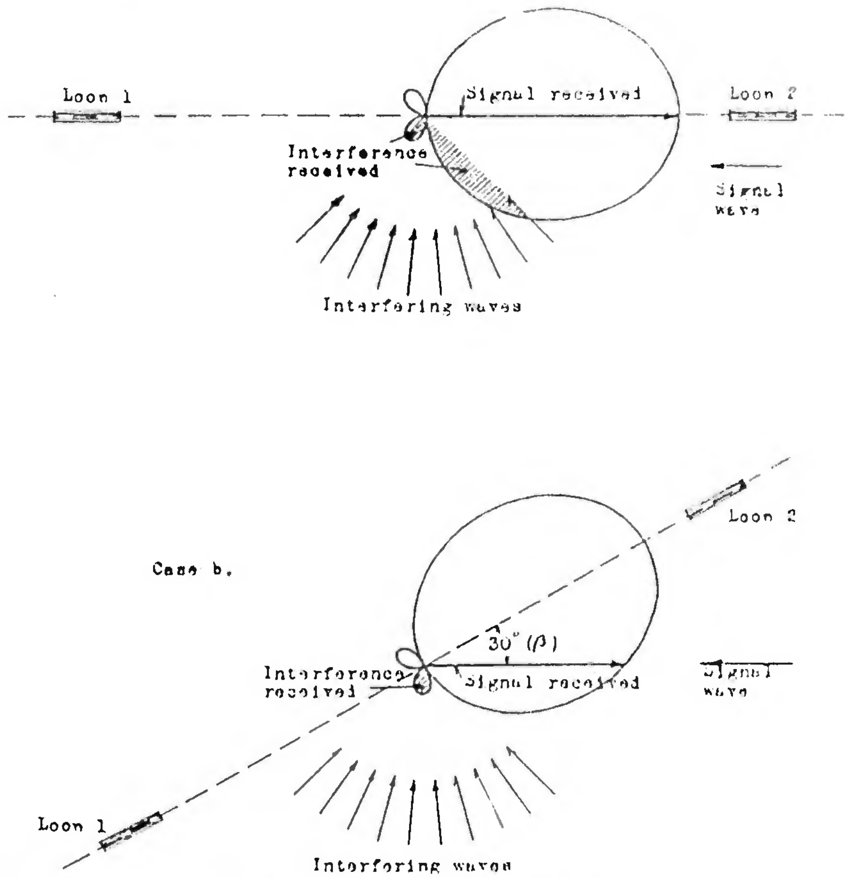


FIGURE 10—Effect of Changing Location of Loops. Location in Lower Figure Improves Signal-noise Ratio in Case of Distribution of Interference as Shown

The circuit diagram is the same as that shown on Figure 1, with the exception of the antenna circuits which are shown in Figure 11. Double tuned antenna circuits are used here at long waves, due to the fact that a low resistance single loop circuit is too selective for speech signal reception. The loop circuit and the secondary circuit are coupled together electromagnetically by means of a variable coupling coil. Electrostatic coupling was tried also, but it was found much more difficult to obtain a good balance of impulse interference with this kind of coupling. The reason for this is as follows: Let it be assumed that both the electrostatically and the electromagnetically coupled circuits are tuned, using weak coupling, to a certain frequency and that then the couplings are increased in order to get the well-known resonance curves with two peaks. Now, in the case of electrostatic coupling the amplitudes of these two peaks will only be alike if the time constants are the same for the primary and the secondary circuit, or if the inductances and capacities have the same values in both circuits, while in the case of electromagnetic coupling the two peaks will always have the same amplitudes.

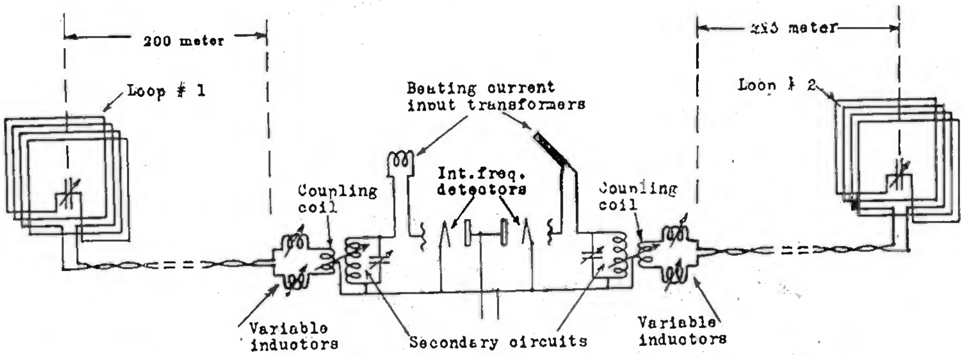


Figure 11—Antenna Circuits of "Long-Wave Two-Loop System"

The loops are installed in 10×10-foot wood houses in order to make the system independent of the weather conditions. Each loop is 8 feet square and has 40 turns of bare number 14 copper wire spaced ¾ inch. The loops can be rotated because it is quite important to measure the "loop minimum"[5] and also to have the two loops pointing in the same direction. The loop tuning condenser is mounted in the loop, but the loop circuit tune can be varied a little in the building where the set is installed by means of two variable inductances inserted in series with the two wires that connect the loop with the coil coupled to the secondary circuit. The arrangement of the antenna circuits shown in Figure 11 makes it possible to use an ordinary weatherproof pair of twisted wires lying on the ground for the 200-meter connection between the loop and the set without introducing appreciable losses in the antenna circuits. This is due to the low impedance of the variable inductors and the coupling coil terminating the line.

The illustration, Figure 12, shows one of the loop houses. The total weight of house and loop is less than 2,000 lbs., so that a team of horses can readily move the loop to any desired location. Such flexibility is very desirable in a "long wave two-loop system" which is still in its experimental stage.

[5] By the "loop minimum" is understood the ratio of maximum to minimum emf. induced in the loop as this is rotated. This ratio ought to be at least 25 times.

By placing the two loops close together it can be determined whether the constants of the two antenna circuits are sufficiently alike. For a small distance "a" between the loops, it should be possible to balance out signals from all directions because the emf's in the loops are then always in phase. The "balance," that is, the decrease in interference when receiving with the combined loops, as compared to the interference when receiving on one loop alone, depends only upon the constants of the two antenna circuits, and it ought to be at least 40 times. The full

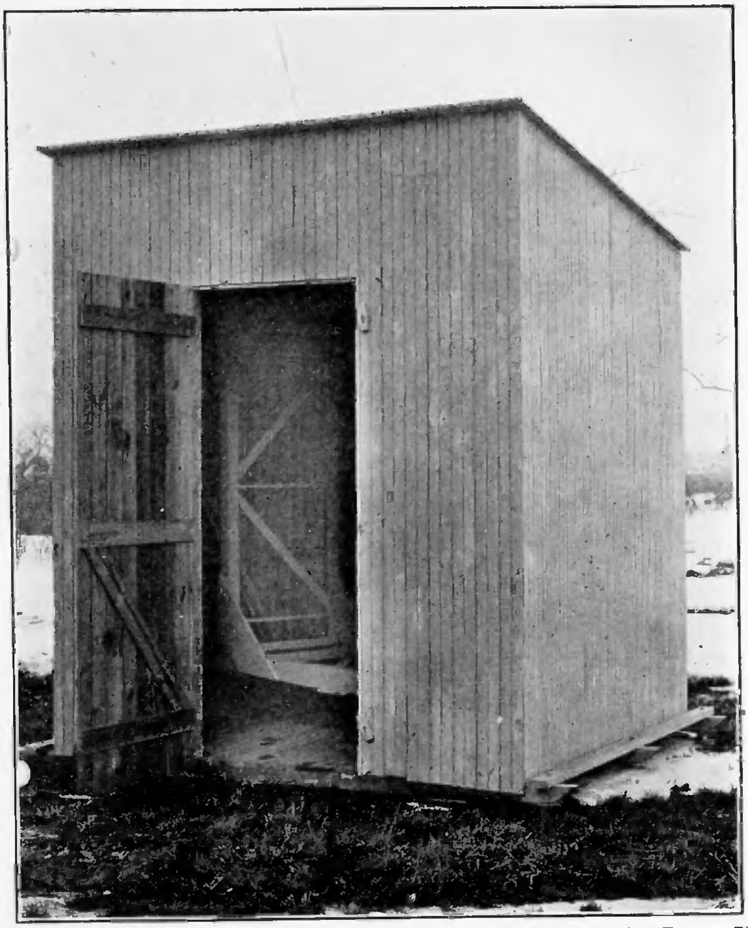


FIGURE 12—"Long Wave Two-Loop System." One of the Loop Houses

length of twisted wire is naturally used when experimenting with the loops close together, and while the loops must point in the same direction, this direction is chosen so that the coefficient of coupling between the loop circuits is at a minimum. After the antenna circuits are thus thoroly tested, the loops can be moved to their right location.

In order to compare this system with other directional systems it is necessary to measure its signal-to-noise ratio and compare it with the signal-noise ratio of some standard antenna system. So far the loop antenna has been our "standard" for comparison. Such measurements require that both the selectivity and the "set-noise" be the same for the standard system and the system the improvement of which is to be determined. In the case of the two-loop system it is easy to satisfy these requirements because the standard system may be obtained by merely short-circuiting one of the loops. The signal amplitudes received by a single loop and by the two balanced loops are equal when the distance between the loops is one-twelfth of a wave length, in which case the improvement in signal-noise ratio is equal to the increase in the noise when one loop is short-circuited. It is thus seen that it is not necessary actually to receive signals in order to determine the improvement of the system. If the distance between the loops is not one-twelfth of a wave length, then a correction factor must be added. The best method of measuring the increase in noise when one of the loops is short-circuited is to decrease the beating oscillator input, at the moment the loop is short-circuited, to such a value that the noise level remains the same. The decrease in beating oscillator input measures the improvement of the system.

If the character of the interfering noise has not changed when one of the loops is short-circuited, it is not difficult for experienced operators to adjust the beating oscillator input for equality of noise level and check results. But often static interference is so variable that several measurements have to be taken. The measurements described above are generally checked by measurements which employ a local warbler signal[6] introduced in one of the loops, and adjusted until it can just be heard thru the noise. The improvement is then determined by the ratio of the warbler signal inputs corresponding to reception with one loop short-circuited and with two balanced loops, respectively.

It is desirable to measure the improvement in the signal-to-noise ratio of this long wave two-loop system. Such measurements would include continuous data on signal-to-noise ratios, strength of static, and direction of static. So far the system has been available only in the winter time, so that it has been impossible to make systematic measurements. However, valuable information has already been obtained and the results are, that, whenever the improvement was only 2 to 3 times, it was found

[6]Warbler signal oscillators and their use is described in a paper on "Radio Transmission Measurements," by Bown, Englund, and Friis, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 11, number 2.

that the general direction of static was south-southeast, that is, the static direction was at right angles to the plane of the two loops while static from south, southwest, and west was always reduced 4-6 times, often 8-10 times, and sometimes even 20 times. On one day only could no improvement be noticed, but static was that day coming from the northeast. It may be emphasized that a 7-times reduction in static corresponds to a 7^2 or approximately 50 times saving in power at the transmitting station.

Condenser Antenna—Loop System

The receiving set of the "long wave two-loop system" described above can obviously be used to combine the output voltages of other types of antenna. A system consisting of a condenser antenna and a loop will thus give the well-known cardioid directional characteristic shown in Figure 13, and as it is unidirectional, such a combination is especially suited for the determination of the direction of static interference. The long wave set was therefore equipped with switches so that the two loop circuits ordinarily used could be interchanged with a loop and condenser antenna both located close by the set. The illustration, Figure 14, shows the condenser antenna which is made up of two 5×5-meter frames covered with copper netting and spaced 15 cm. apart, and the loop used can be seen towards

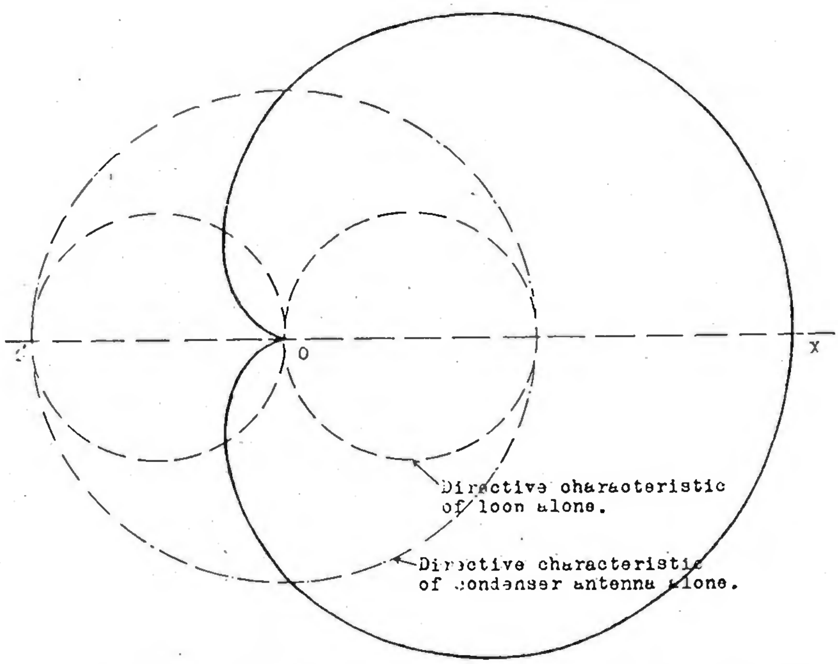


Figure 13—Directive Characteristic of "Condenser Antenna-Loop System"

FIGURE 14—Condenser Antenna

left in the illustration, Figure 15, of the long wave receiving set. This system was used to determine the general directions of the static mentioned earlier in this paper and has naturally been of great help in the interpretation of the different improvement values obtained for the "two-loop system." So far the condenser antenna-loop system has shown that static at long waves in general has some directional characteristic. Sometimes the ratio of maximum to minimum static received when turning the loop was as high as fifteen times, that is, static was very directional.

The mathematical treatment necessary to derive the equation for a cardioid directional characteristic shown in Figure 13 is very simple and the procedure is similar to that used in the analysis of the two-loop system. The maximum and minimum signals are produced by waves propagated in the plane of the loop and they may be reversed by simply reversing the loop connections. Discrimination against interference from a specific direction may be secured by directing the plane of the loop towards the interfering source, the polarity of the loop being such that signals of that direction are suppressed. An experimental directive characteristic which was obtained by measuring the strengths of a continuous wave signal for different positions of the loop checked the cardioid curve in Figure 12 exactly.

The emf's induced by the signal in the loop and condenser antenna circuits are always in quadrature and the beating oscillator tuning circuits can therefore be made somewhat simpler than

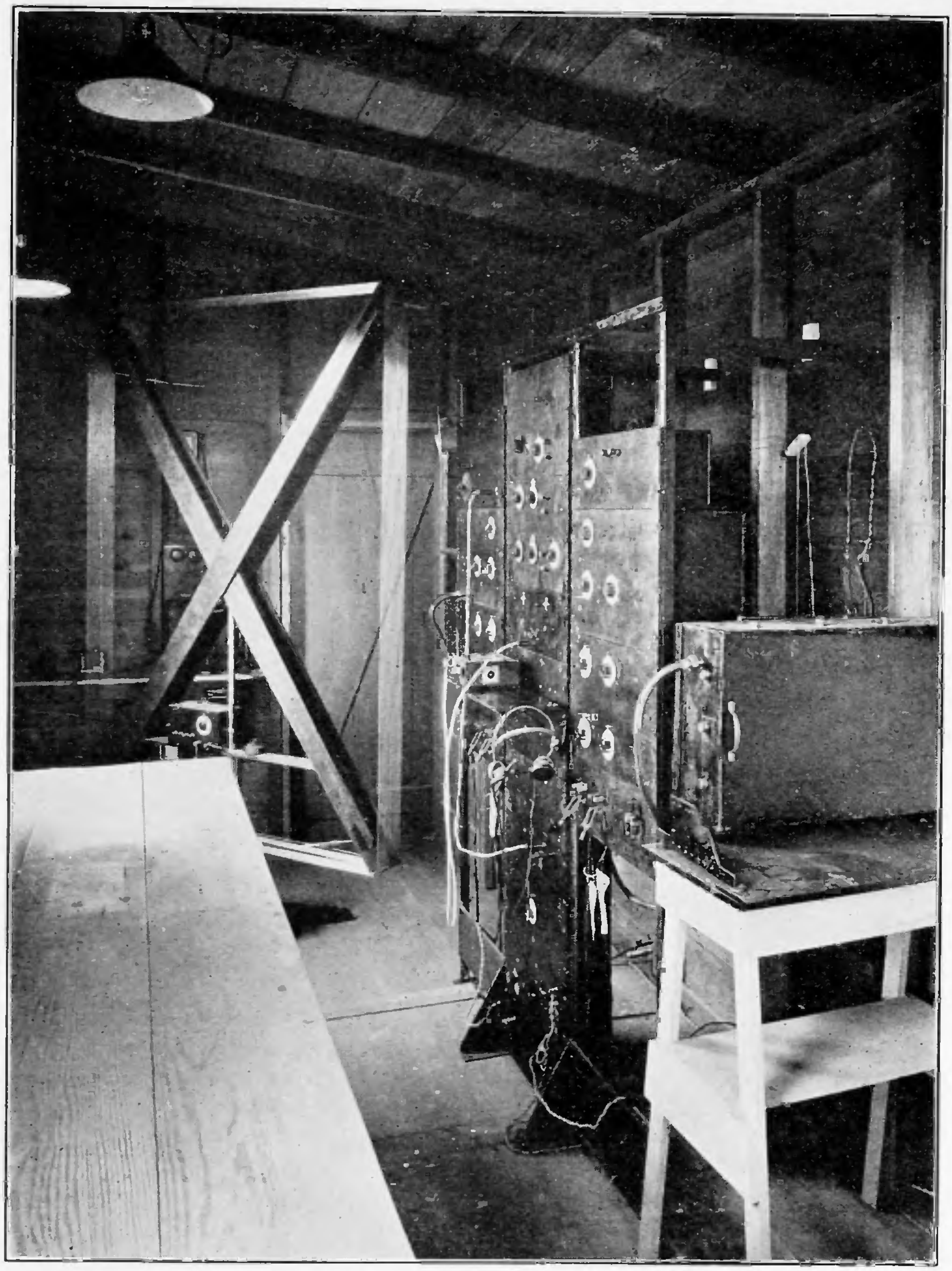


FIGURE 15—Long Wave Two-Loop System. Front View of Receiving Set

shown in Figure 1 if the set is to be used only to produce the cardioid directional characteristic. Figure 16 shows how simple it is to change an ordinary double detection receiver for this purpose. The currents in the beating oscillator circuit 3 and the tuned circuit 4 are in quadrature and the two beating current emf's in the antenna circuits will therefore also be 90° out of phase. A small change in this phase angle may be obtained by detuning circuit 4 slightly and the relative values of emf's may be adjusted by varying the mutual coupling M. A good minimum on spark signals and static, that is, a good balance of transient currents in two antenna circuits can only be obtained when the time constants of the two circuits are alike, and it is, therefore, in general, necessary to insert some extra resistance in the loop.

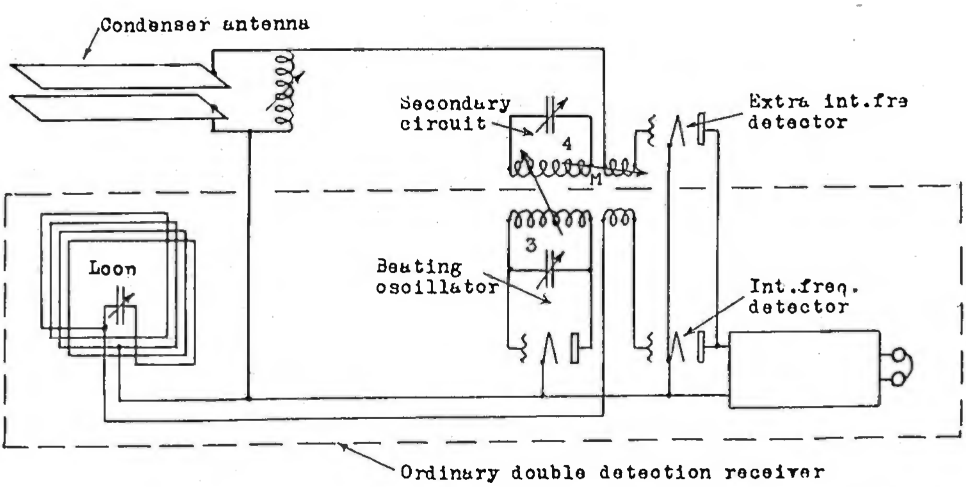


FIGURE 16—Schematic Circuit Diagram of "Condenser Antenna-Loop System"

The cardioid characteristic is inferior to the characteristic obtainable with a "two-loop system," but at long waves it has one great advantage, namely that it can be pointed towards any direction by simply turning the loop and it will give much more satisfactory results than a loop alone when it is desired to reduce interference from static or sparks, the direction of propagation of which makes an angle of at least 90° with the desired signal wave propagation.

CONCLUSIONS

A short wave directional two-loop system has been tested and shown to have the calculated directional properties. A similar long wave two-loop system has been developed and tested as far as possible under winter conditions of interference. Measured improvements in signal-to-static ratio range from unity to twenty, depending on the direction of the static. If this lies chiefly in the rear, the improvement is certainly 6 to 8 times. In fact, as far as experiments have gone, they indicate that, given a certain static distribution, the improvement could be calculated from the directional characteristic. Actual static distributions have, so far, favored the two-loop system as against the condenser antenna-loop system.

The directional antenna systems described in this paper require a double detection receiving set with a separate intermediate frequency detector for each antenna, and all necessary phase and amplitude adjustments are performed upon the beating current inputs. The system really makes use of the following general rule for double detection sets:

So far as the intermediate frequency currents are concerned, any change in phase, amplitude or frequency which it is desired to perform upon the antenna circuit current as a whole may just as well be performed upon the beating current.

The main advantages of this type of directional systems may be summarized as follows:

(1) Large reduction in static interference.

(2) Simple adjustments for interference reduction.

(3) Dimensions of not more than 1/12th of a wave length (for the two-loop system), so that only comparatively small areas are required.

(4) High efficiency antenna circuits so that excessive amplifications are not required.

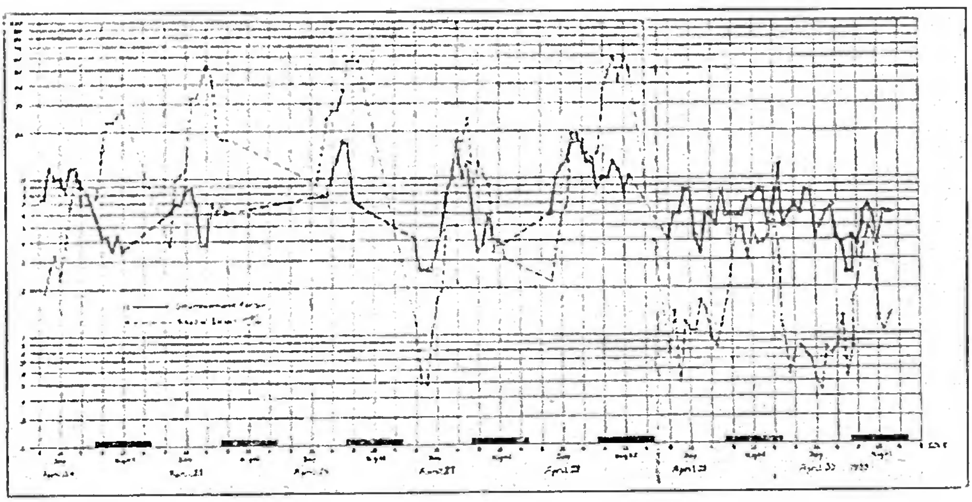
(5) Plenty of power is available in the beating current circuits, which simplifies the construction of the phase and amplitude controlling apparatus.

(6) The system can be checked quite readily.

(7) It is quite easy to house the system so as to protect it from the weather.

Appendix

Since this paper has been written it has been possible to take continuous measurements for one week on the long wave two-loop system described.



The results are given in the foregoing diagram. The full-drawn curve gives the improvement in signal-to-static ratio measured on a band approximately 2,500 cycles wide centering upon 57,000 cycles with the antenna located at Cliffwood, New Jersey, and directed for reception from England. The dotted curve gives the static level in $\mu V./m.$, as measured on one of the loops alone. The curves show that the improvement factor is approximately 10 times when the static level is larger than 5 $\mu V./m.$

SUMMARY: The paper discusses methods of combining the signal currents from the different antennas in a directional receiving system and a detailed description is given of a system by which all phase and amplitude adjustments are performed upon the beating current inputs of a double detection receiver. The theoretically derived shape of the directional characteristic of a two-loop system has been verified by experiments, and data on reduction of static for such a system are given.

AN ANALYSIS OF REGENERATIVE AMPLIFICATION*

By

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Present Conceptions of Regeneration

When Edwin H. Armstrong, in 1912, brought the phenomena of regeneration to the attention of the radio world, he undoubtedly gave rise to one of the greatest factors in the advance of radio today. Before regeneration, the electron tube served merely as a simple rectifier and amplifier. After regeneration, the same tube served as an amplifier of apparently unlimited extent. As is often the case in similar discoveries, the critical adjustments required and the enormous gain in signal strength so obtained gave the impression that the action was necessarily very complicated. Consequently at first very little attempt was made to explain the amplifying action except by stating that synchronous energy was fed from the plate circuit to the grid circuit which re-inforced the incoming signals.

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Practically all of the present discussions and analyses of regeneration can be grouped under two heads. First, that regeneration produces an equivalent reduction in the resistance of the grid circuit. Second, that regeneration is a voltage amplification due to the addition of the re-impressed grid voltage and the original applied voltage.

The first explanation of regeneration is the one more commonly accepted. This states that the equivalent resistance of the grid circuit decreases a definite (not proportional) amount as the tickler coupling is increased. A curve plotted between tickler coupling and the equivalent grid circuit resistance is a straight line and intersects the zero resistance axis at the point of critical regeneration. In this analysis an infinite amplification and an infinite response are obtained at critical regeneration regardless of the initial value of the signal voltage. This, of course, is impossible practically and so this type of analysis is always modified by the statement that the tube output is limited. This means that the characteristic curves are not straight lines in actual practice as we assumed in the theory. This variation from linear characteristics, which will be shown later to be the most important characteristic of regeneration, is absolutely omitted from almost all present mathematical analyses of these phenomena. We are, therefore, forced to reject the conclusions reached by such a method of analysis so far as they affect the conditions actually obtained in practice.

The second method of analysis shows that the final voltage applied on the grid is the sum of the initial applied voltage and the voltages introduced by the tickler. This analysis in some detail follows. When a regenerative circuit is brought near the point of oscillation, any voltage impulse on the grid will cause an oscillatory current which will die out at a rate depending on the resistance of the circuit and the value of tickler coupling. If the ratio of the amplitude of one cycle to that preceding is called "a" and if E_g is the amplitude of a given cycle, aE_g is the amplitude of the next. If an alternating voltage E_1 of resonant frequency is applied in series with the circuit, then the voltage on the grid for the first few cycles will be as follows:

$$E_1 \quad : \quad a\,E_1+E_1 \quad : \quad a\,(a\,E_1+E_1)+E_1 \quad : \quad ----$$

And for the n-th cycle will be $\dfrac{E_1(a^n-1)}{(a-1)}$

If "a" is less than "1" and "n" is infinite, then a^n is zero and

$$E_n=\frac{E_1}{(1-a)}$$

As the value of "a" approaches "1," E_n increases without limit. If "a" is larger than "1," the circuit will oscillate of its own accord. At critical regeneration $a=1$ and an infinite response is obtained regardless of the value of E_1. The criticism of the first type of analysis also holds here. That is, in practice, linear relations are not maintained, and there is no reason therefore to assume that any mathematically constructed analysis can be given as a correct explanation of regenerative phenomena.

If any further proof is needed to show the fallacy and inadequacy of the present analyses of regeneration, only a few facts are necessary. Assuming that either (or both) of the above methods of analysis are correct, it is evident that regenerative amplification is limited only by the ability of the operator to adjust the tickler to the critical value, and that with any given adjustment the final voltage will be directly proportional to the applied voltage. This means that the amplification is a constant. However, quoting from an article entitled "The Limit of Regeneration," by N. C. Little, in the August issue of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, for 1924, we find the following: "Results obtained with VT-1 and UV-201 tubes, both used with varying amounts of grid bias, show that *the relative magnitude of received signal in plate circuit to that impressed on the grid is inversely proportional to the latter*. This is called the inverse signal strength law. It states that the response of a system adjusted to critical regeneration is independent of the strength of the impressed signal, that no matter how weak the oscillating field surrounding an antenna may be, if the regeneration is pushed to its limit, a finite signal may always be obtained."

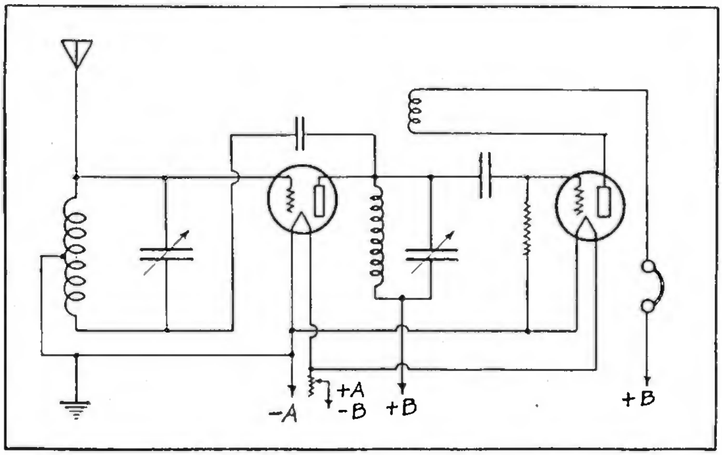
Obviously both of the above statements regarding the ratio of the final signal to the impressed signal cannot be correct. The first is quite evidently untrue, and the more experienced radio engineers will not accept the latter as an exact statement either. Something more than the previously advanced theories must be considered in the true explanation of regeneration and its limits.

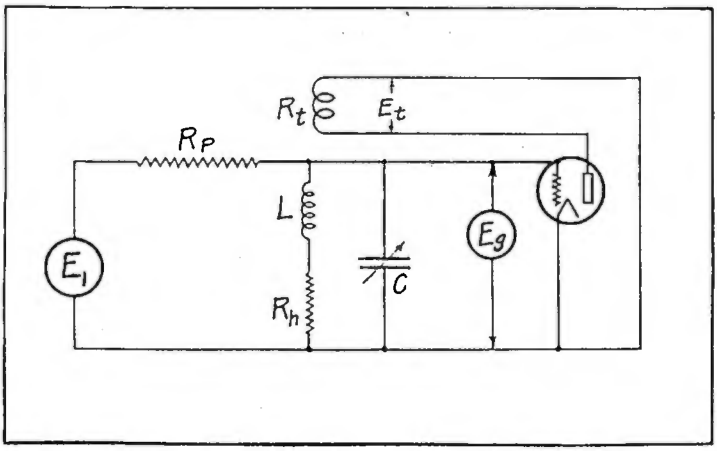
POWER BALANCE IN NON-REGENERATED CIRCUIT

To this end let us consider a few facts about the simple circuit shown in Figure 3. E_1 is an applied voltage, R_p the resistance equivalent to the tube impedance,[1] and R_h the resistance in the inductive leg of the tuned circuit of inductance L and capacity C.

[1] The action of a vacuum tube as an amplifier can be replaced by a fictitious generator of zero internal impedance, in series with a fixed resistance The voltage of this generator is μ (the amplification constant) times the voltage applied to the grid, and the resistance is equal to the plate impedance.

The resistance in the capacity branch is assumed so small as to be neglected.





FIGURES 1 and 2*

Let a voltage E_1 be impressed as shown. E_g will build up, reaching a final value which depends on the relative impedances of R_p and the tuned circuit. However, let us analyze the instantaneous conditions in the circuit. The power input at any instant is given by $E_1 \times I$. But $I = \frac{E_1 - E_g}{R_p}$ and therefore Watts Input $= E_1 \frac{(E_1 - E_g)}{R_p}$. The watts lost at any instant are given by the sum of the losses in the two resistances, R_p and R_h. The

* This figure shows the Rice method of balancing out capacity coupling between two tuned circuits. Divided plate circuit neutralization could be used with equal advantage.

loss in R_h is proportional to E_g^2. The watts lost are given by the sum of the following two terms:

Watts Lost$=\dfrac{(E_1-E_g)^2}{R_p}+\dfrac{E_g^2}{K\,R_h}$ where $\dfrac{E_g}{\sqrt{K}}$ is the effective voltage across R_h with E_g volts on the grid.

Before the final value of E_g is reached, the watts input will exceed the watts lost. This surplus is stored up in the tuned circuit as a charge on the condenser or in the magnetic field of the inductance. A stable condition is reached when the watts lost equals the watts input from E_1. Setting these two expressions equal and solving for E_g gives

$$E_g=\frac{E_1\,K\,R_h}{R_p+K\,R_h}$$

This expression is the same as that obtained by assuming that E_1 divides between R_p and the tuned circuit in the ratio of their impedances.

The instantaneous conditions can be shown to a better advantage by means of curves. These curves are shown in Figure 4, and are plotted between E_g^2 and watts. The square of E_g is used as abscissa in order to make the loss curve for R_h a straight line. The advantage of this will be more forcibly shown later. Notice that the watts input is larger than the watts lost below the point of intersection. This point determines the final voltage and is called the point of power balance.

Power Balance in a Regenerated Circuit

In the above discussion, the only power available for building up the voltage E_g came from E_1. In a regenerative electron tube, however, the plate circuit introduces power into the grid circuit. If it introduces sufficient power, the tube will oscillate. In Figure 2, let E_1 be zero. Then in order to oscillate, the plate must supply the resistance losses in both R_h and R_p. In the regener-

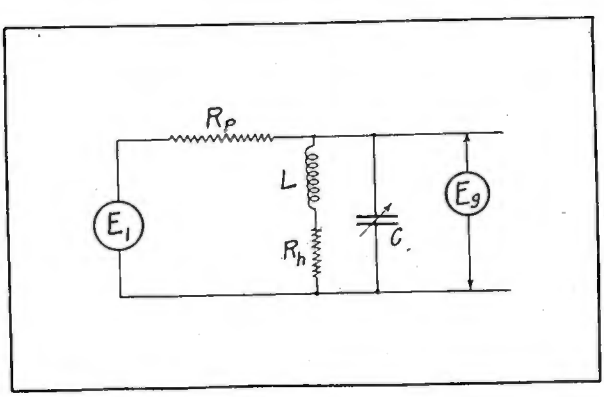


FIGURE 3

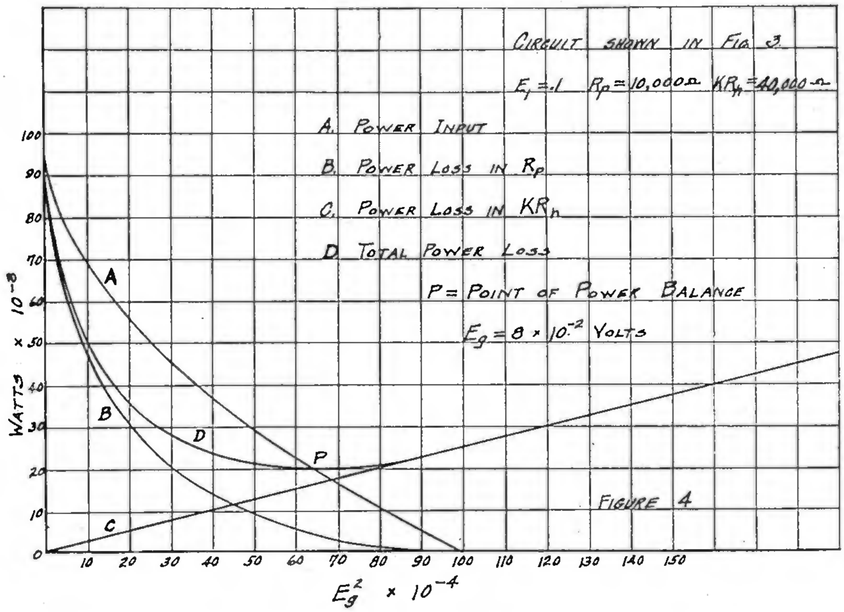


FIGURE 4

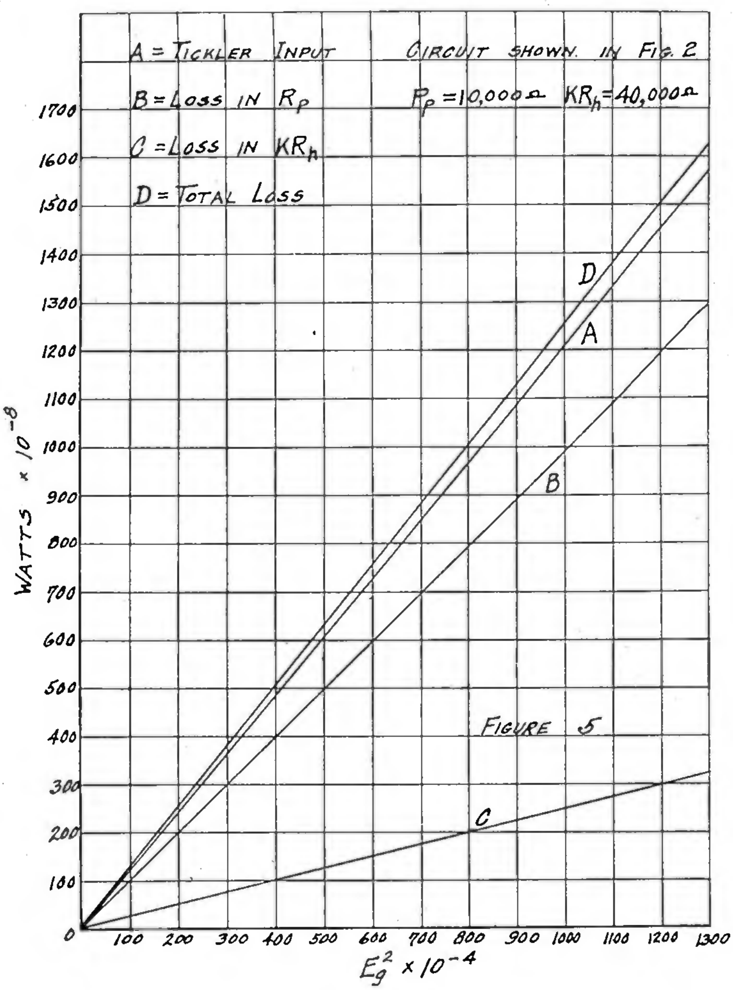


FIGURE 5

ative state, the power supply from the plate circuit for any value of voltage E_g is less than the losses in the circuit at the same value of E_g. This is shown in the curves of Figure 5. The loss in R_h is usually much less than that in R_p due to the relative impedances of the two branches and the relative resistance of the two branches. (The plate resistance branch and the inductive leg.) As long as the tube is not capable of supporting self oscillations, the sum of these losses is greater than the power input from the tickler.

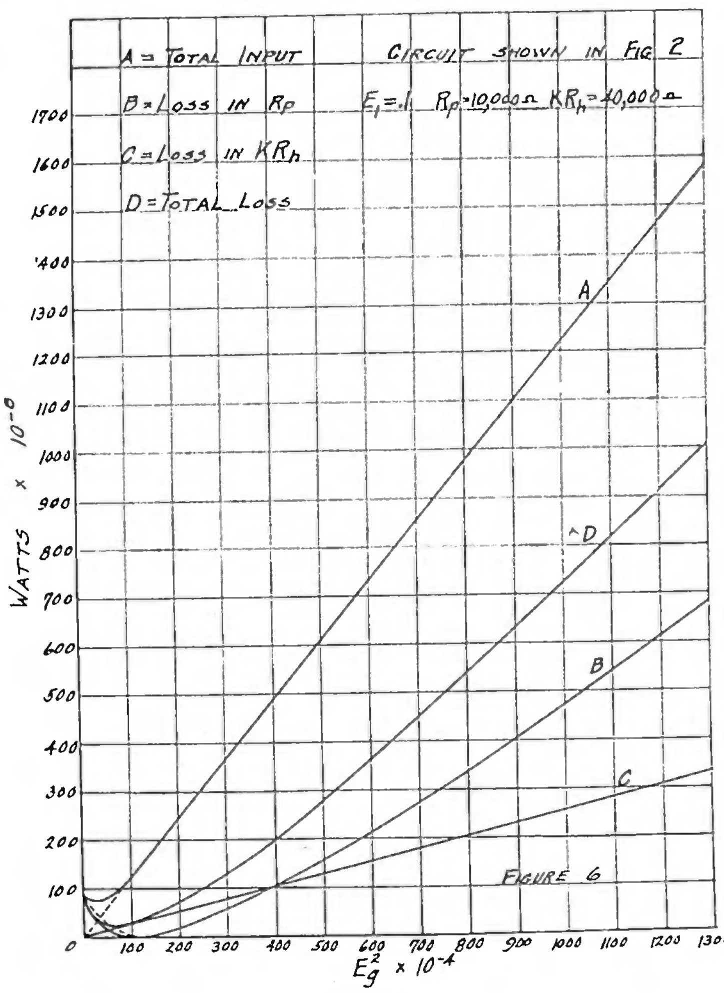


FIGURE 6

Now let E_1 be applied. At the instant of application E_1 supplies power at the rate shown in Figure 4. However, as E_g increases, the plate circuit also supplies power to the grid circuit as shown in Figure 5. The sum of these "power input" curves gives the total power input for any resulting value of E_g. This sum is shown in Figure 6. The loss in R_h is directly proportioned to E_g^2 and consequently this loss curve in Figure 6 is not different than that shown in Figure 5. However, the loss in R_p is no longer the same as before. As E_g increases, the loss in R_p decreases, due, as previously shown, to the decreased voltage across R_p. When E_g equals E_1, the loss in R_p is zero. Now as E_g further increases, the loss in R_p increases, due to a greater increase of voltage across R_p. This loss curve is shown in Figure 6. The sum of the two loss curves is shown and it may be seen that (for the range of values shown) the input is greater than the corresponding losses and the tube will "oscillate." In oscillating, the grid voltage will build up until some point of power balance is again reached, i.e., where the losses increase to equal the input As soon as E_1 is removed, the additional loss incurred in R_p wil. bring back the conditions of Figure 5, where no oscillation will be produced. Regeneration therefore simply means one thing and one only. *When a certain voltage E_1 at a frequency f_1 is introduced into a circuit tuned to f_1, as shown in Figure 2, the losses occurring in the circuit will be sufficiently lowered to allow the electron tube to oscillate at a frequency f_1.* In so oscillating, it builds up and maintains a stable value of voltage E_g on the grid. *Regeneration and oscillation are one and the same thing.*

The equation for critical regeneration is as follows:

$$\frac{E_g^2}{R_p}+\frac{E_g^2}{K\,R_h}+\frac{E_g^2}{R_g}=\frac{E_t^2}{R_t}=\frac{b\,E_g^2}{R_t} \tag{1}$$

In this equation all terms are as previously stated except R_g, E_p, E_t, b, and R_t. R_g is the grid-filament resistance of the tube used as a regenerator, E_t is the voltage across the tickler coil, b is a constant of such a value that $\frac{b\,E_g^2}{R_t}$ expresses the value of tickler input to the grid and R_t is the effective resistance of the tickler coil measured across its terminals and is due to the absorption of power in the grid circuit. When E_1 is applied, the power furnished to the first tube by E_g is

$$e_g\,I_p=E_g\frac{(E_g-E_1)^2}{R_p}=E_g^2-\frac{E_g\,E_1}{R_p} \tag{2}$$

As E_t is directly proportional to E_g (variations will be discussed later), all other terms remain constant. Subtracting (2) from the first term of (1) gives

$$\text{Watts Saved}=\frac{E_1\,E_g}{R_p} \tag{3}$$

Knowing E_1 and R_p, the total loss curve in Figure 6 with E_1 applied can be found by subtracting the above expression (3) for "watts saved" from the total loss curve of Figure 5. If the tickler feed-back is that of critical regeneration and is a straight line function of E_g^2, i.e., where (1) holds true for all values of E_g, the loss curve with E_1 applied will always be less than the tickler input and the voltage E_g will build to an infinitely large value.

Reasons for Limited Amplification

Our next and perhaps most important step is to determine what limits the amplification of a regenerative electron tube, and how these limits can be controlled in practice. It may be seen from Figure 6, that there is no power input from E_1 after E_g equals E_1. This condition will be readily reached in practice, o for the present we will limit ourselves to a discussion of the imits with respect to E_g.

In a practical example of the circuit, R_h does not vary. However, the values of the plate impedance of both tubes and the input impedance of the regenerator change with changes in amplitude of E_g. The initial values of these impedances are greatly affected by the type of tube, direct current plate and grid voltages, etc. In addition, the amount of change as E_g varies, is greatly affected by the plate and grid voltage used. Another variable introduced is that due to the fact that the tickler input does not increase as a straight line function of E_g^2.

We have seen that when E_1 is introduced, the power input exceeds the power lost. The grid voltage E_g will build up to the point where the input and the loss are again equal. This means a loss increasing faster than E_g^2, or a tickler input increasing more slowly than E_g^2, or both.

These variations from straight line characteristics serve to bring about this new point of balance. The following equation expresses the power relation at the point of balance.

$$\frac{E_g(E_g-E_1)}{R_p}+f_p(E_g-E_1)+\frac{E_g^2}{K R_h}+\frac{E_g^2}{R_g}+f_g(E_g) = \frac{bE_g^2}{R_t}-f_{bt}(E_g). \tag{4}$$

where $f_p\,(E_g-E_1)$ is a variable function of (E_g-E_1) expressing the difference between the actual power loss in the plate impedance with a finite E_g and the power loss as given by $\frac{(E_g-E_1)^2}{R_p}$: where $f_g\ (E_g)$ is a variable function of Ef, indicating a change in loss from that expressed by $\frac{(E_g^2)}{(R_g)}$: andwhere $f_{bt}\,(E_g)$ is a variable function of E_g, which includes all changes of the power supply from the tickler to the grid not in direct proportion to E_g^2.

Then

$$f_p(E_g-E_1)=\frac{E_g(E_g-E_1)}{R_p'}-\frac{(E_g-E_1)}{R_p} \tag{5}$$

$$f_g(E_g)=\frac{E_g^2}{R_p'}-\frac{E_g^2}{R_p} \tag{6}$$

$$f_{bt}(E_g)=\frac{b\,E_g^2}{R_t}-\frac{b\,E_g^2}{R_t'} \tag{7}$$

where R_p, R_g, and R_t are of values of the corresponding impedances when E_g is infinitesimal and where R_p', R_g', and R_t' are the values of impedance for finite values of E_g.

Notice that all of these functions are zero for E_g=zero, and can be positive or negative, depending upon the values of the primed numbers.

It is interesting to note that equation (4) holds good whether critical regeneration is used or not. This makes it very easy to check experimentally. A board set up of the circuit of Figure 2 was used with electron tube voltmeters to measure E_1 and E_g. With a given tickler adjustment E_1 and E_g were measured. Then E_1 was removed and the value of R_p increased by a value Y until E_g rose to its previous value due to self oscillation.

$$\frac{E_1\,E_g}{R_p} \text{ was then calculated,}$$

$$\text{and } \frac{E_g^2}{R_p}-\frac{E_g^2}{R_p+Y} \text{ was calculated.}$$

The two were found equal.

An interesting case is that at which the tickler is adjusted until $E_g=E_1$. The voltage across R_p is then zero and R_p may be varied without affecting E_g. When R_p is open circuited, the tuned circuit will be found to be oscillating with the same voltage E_g on the grid. This, too, has been experimentally verified.

If we assume critical regeneration we may subtract equation (4) from (1).

$$\frac{E_1\,E_g}{R_p}=f_p(E_g-E_1)+f_g(E_g)+f_{bt}(E_g). \tag{8}$$

The left-hand member of this equation represents the "watts saved" by introduction of E_1. The right-hand member represents the change in losses which utilize this "surplus" power at the point of balance.

Let us examine the limit of regenerative amplification as these functions are varied, keeping in mind the following:

First, in order for the tube to be in a state capable of "regenerating" but never capable of producing "self sustained" oscillations, the total loss curve must be greater than the tickler input for all values of E_g (no E_1 applied).

Second, critical regeneration is obtained where the loss curve and the tickler power input curve are tangent at $E_g = 0$.

Third, the maximum amplification is obtained when the abscissa of the point of intersection of the loss curve and the power input curve is a maximum with a given value of E_1.

Case I. When each of the above functions are equal to zero, or if the sum is equal to zero, or if the right-hand side is less than the left-hand side for all values of E_g, then E_g will increase without limit regardless of the value of E_1. This is true because under the terms of Case 1, the power input is always greater than the loss and E_g will increase indefinitely. Such a condition, although impossible practically, is shown in Figure 6.

Case II. Where the sum of the functions is finite and positive, and increases faster than the left-hand member of (8); then E_g will increase to a finite limit. This value depends on the value of E_1, but is not necessarily in direct proportion. This is shown in Figure 7. Upon removing E_1, E_g will drop to zero and the conditions given in equation (1) will be resumed.

Case III. Where the sum of all functions is negative or equal to zero for small values of E_g and become positive for larger values of E_g. In this case an infinitesimal value of E_1 will be sufficient to upset the balance of power and E_g will build up until the conditions of equation (8) are again obtained. This will occur when the sum of the functions reaches a sufficient positive value. This case is shown in Figure 8. Upon removing E_1 the voltage E_g will not fall to zero, but will remain at the point P', where the sum of the functions equals zero. To get the condition of critical regeneration again, it is necessary to reduce the tickler coupling and readjust it. This condition is known as "snapping" or "rubber-band effect" or "floppiness."

Limiting Effects Without Grid Leak and Condenser

The variations of the above functions with respect to E_g can be shown very clearly by means of curves. Figure 9 shows average curves of the variation of R_p with the variations of E_g. This grid voltage is superimposed on the direct current plate voltage of the amplifier tube.

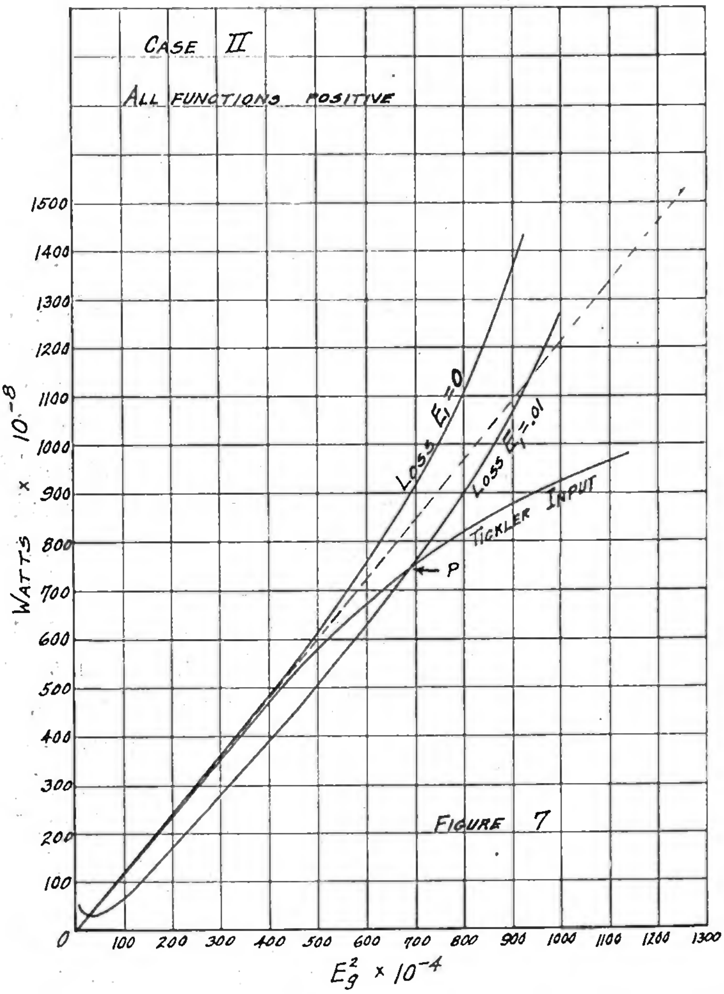


FIGURE 7

Figure 10 shows average curves of grid resistance with variations of amplitude of E_g. Various loss curves are plotted in Figure 11. These actual losses in watts are plotted against $E_g{}^2$, to produce as near a straight line curve as possible. However, in order to show the variations in loss curves by changing the circuit constants, extreme values have been chosen and the graphs are necessarily curved.

Figure 12 shows the variation in plate current with grid voltage. This is a direct current dynamic curve with resistance load in the plate circuit. From this curve are obtained the curves on Figure 13, which are alternating current dynamic curves. These curves indicate the alternating current plate current with alternating current grid volts. Notice that the grid bias very greatly affects the shape of these curves.

The power input to the grid circuit from the tickler is equal to I^2 (alternating current) $\times R_t$, where R_t is the previously men-

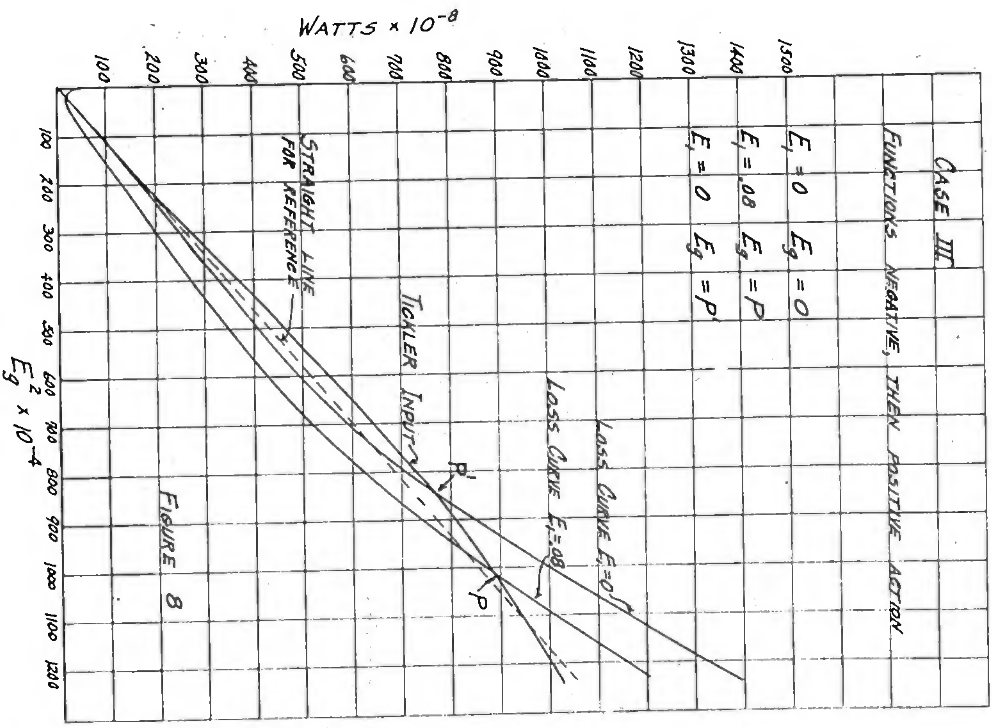


FIGURE 8

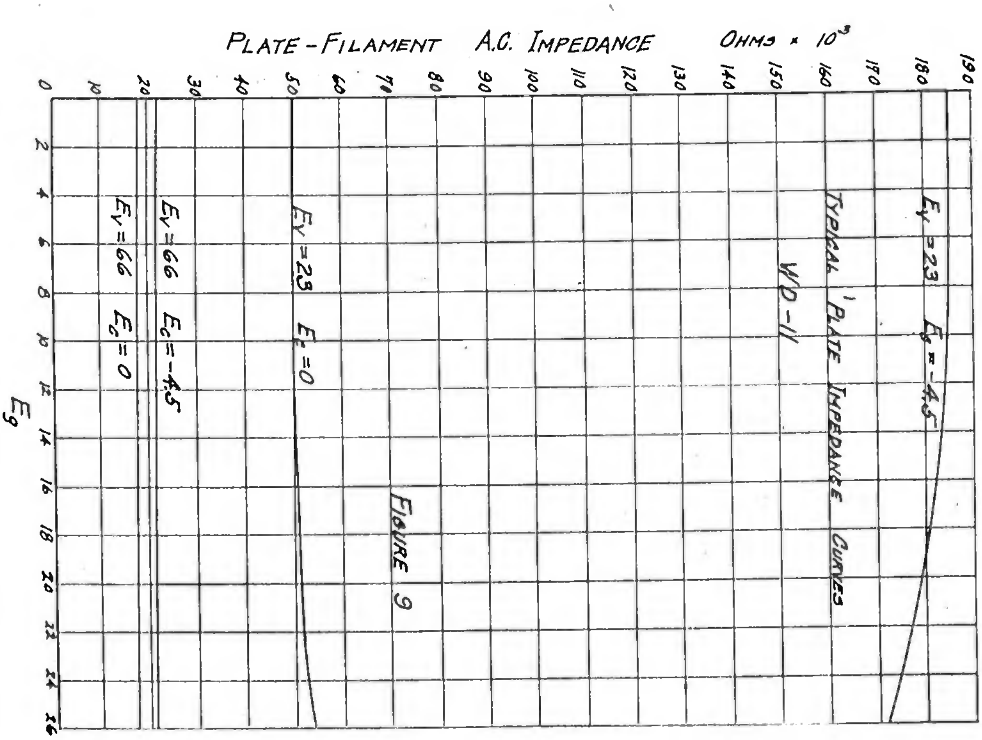


FIGURE 9

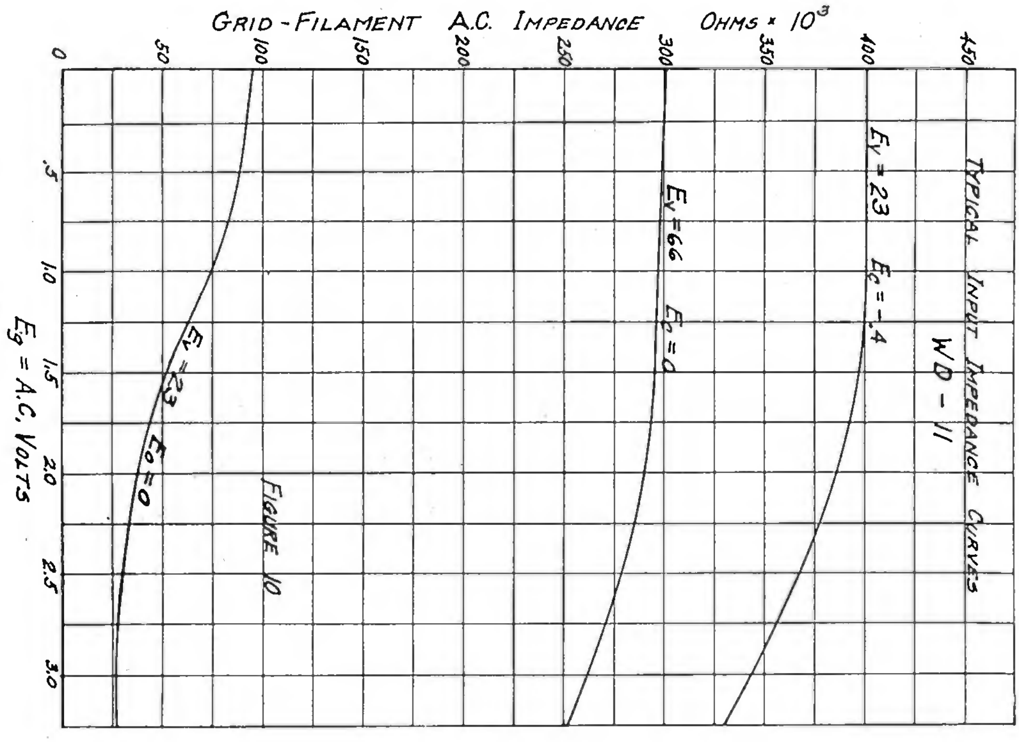


FIGURE 10

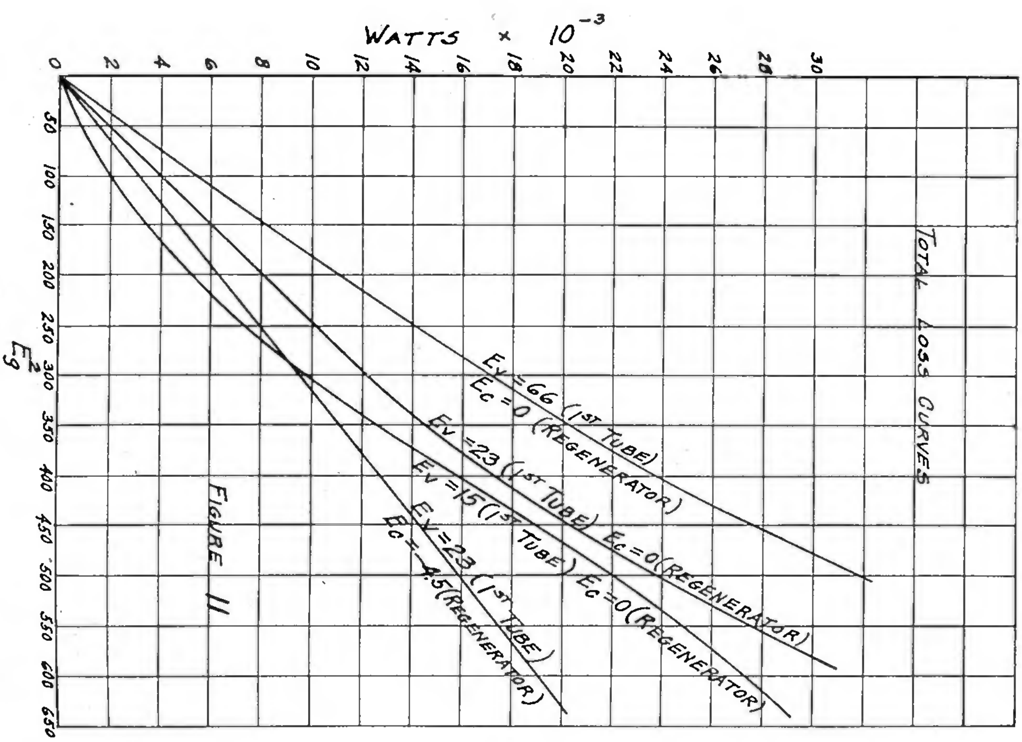


FIGURE 11

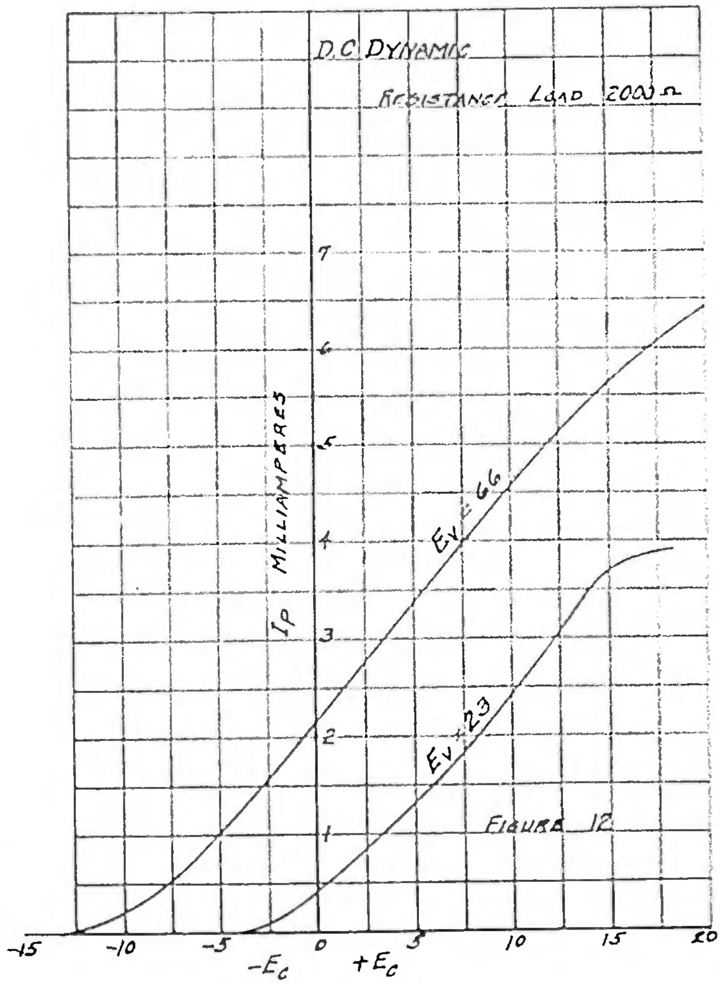


FIGURE 12

tioned resistance of the tickler coil, due to its coupling to the tuned circuit. The value of R_t increases directly as the tickler coupling is increased. I_p decreases slightly as R_t is increased, but as the value of R_t is low compared to the total impedance of the plate circuit, I_p does not change a large amount. Figure 14 shows several power dynamic curves. E_g^2 is used as abscissa to conform with the abscissa used on the loss curves; also to maintain the same shape of dynamic curves, (due to ordinates, $I_p^2 \times R_t$).

From these loss curves and input curves it is readily seen that practical conditions may vary greatly, and the limit of amplification varies with every change in operating conditions.

Limiting Effects With Grid Leak and Condensers

In the regenerative action considered so far, the tube was used simply as an amplifier and not as a detector. It is common

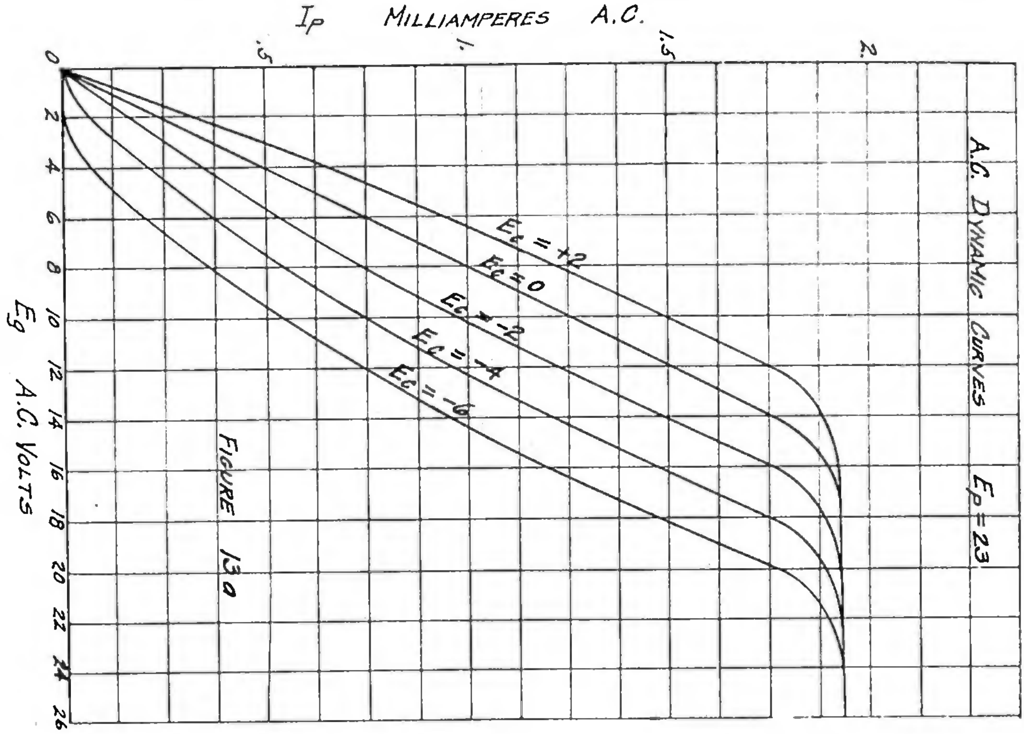


FIGURE 13a

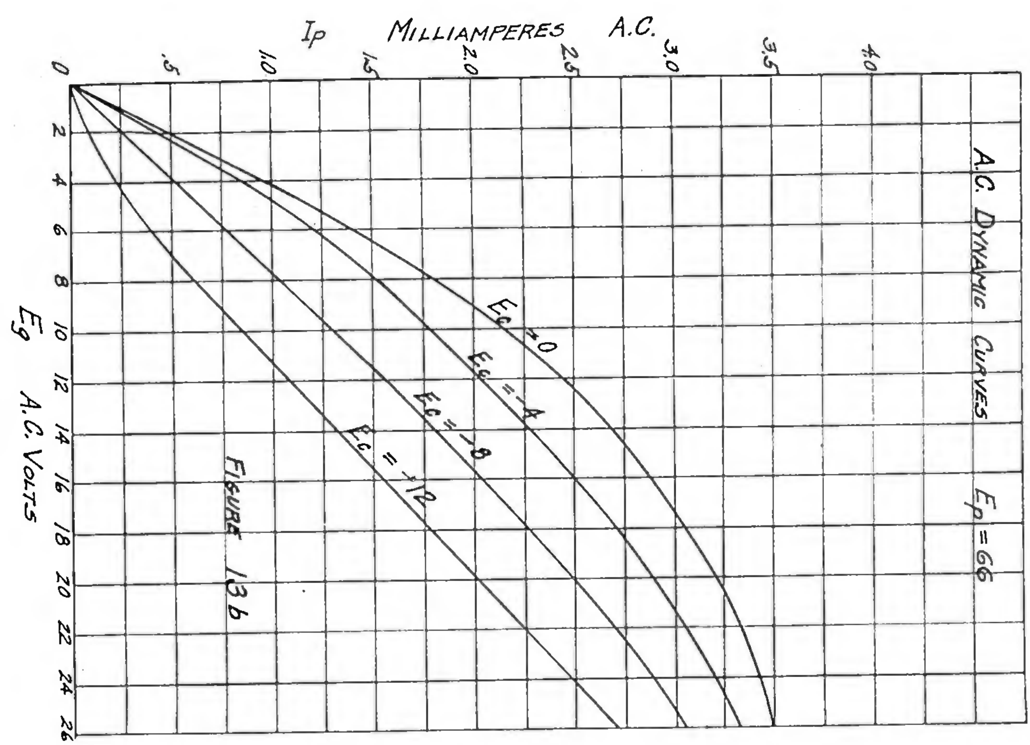


FIGURE 13b

practice, however, to regenerate the detector when using a grid leak and condenser. The component effects in this case are somewhat different than previously indicated.

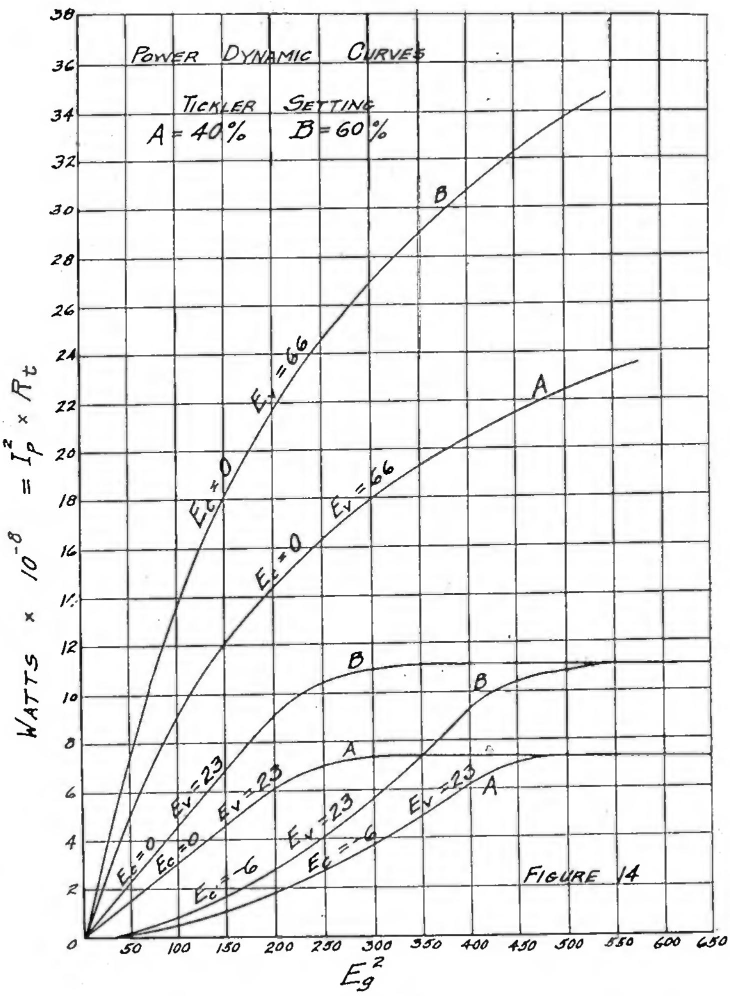


FIGURE 14

The first change is in the effective resistance of the grid leak and condenser plus the grid filament resistance. If the grid leak is small, the condition is not greatly different from no grid leak at all. That is, the current increases faster than E_g and the resistance loss increases faster than in direct proportion to E_g^2 If the grid leak is high, the grid condenser accumulates a negative change, tho the grid current must always increase as E_g increases, it will not increase as swiftly as in direct proportion. In this manner the effective resistance of the grid filament circuit increases and the loss does not increase as fast as E_g^2. These loss curves in the grid filament impedance, plotted against E_g^2, are shown in Figure 15, using two values of grid leak resistance.

The alternating current dynamic characteristic of power input is also different with grid leak and condenser. As E_g is applied, the average grid potential moves down on the static curve, thus increasing the impedance of the tube. Such a dynamic curve for $22\frac{1}{2}$ volts (E_v) with grid leak and condenser is shown in direct comparison with a similar curve without grid leak and condenser in Figure 16. The reason for this maximum power

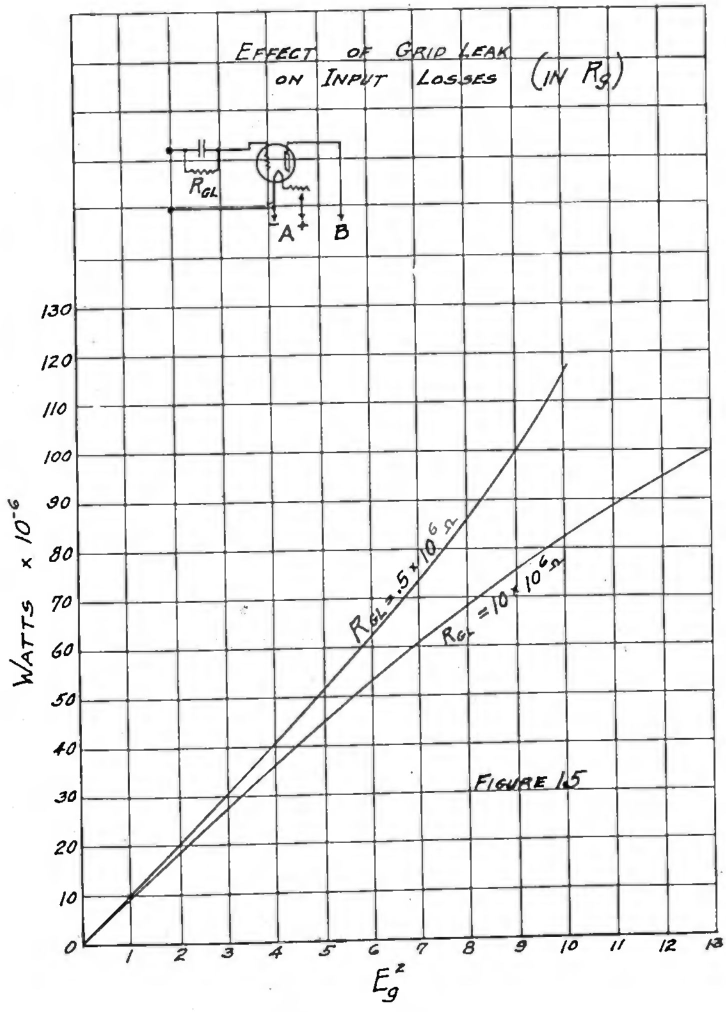


FIGURE 15

input change is seen in the static curve of Figure 17. When the grid draws current on the positive half cycle of E_g, the grid accumulates the negative charge and the average charge becomes increasingly negative. The grid becoming negative increases the tube impedance and so decreases the plate current.

With large values of plate voltage the change in plate impedance is less with a given change in grid bias. The result in the dynamic curve with grid leak and condenser is shown in Figure

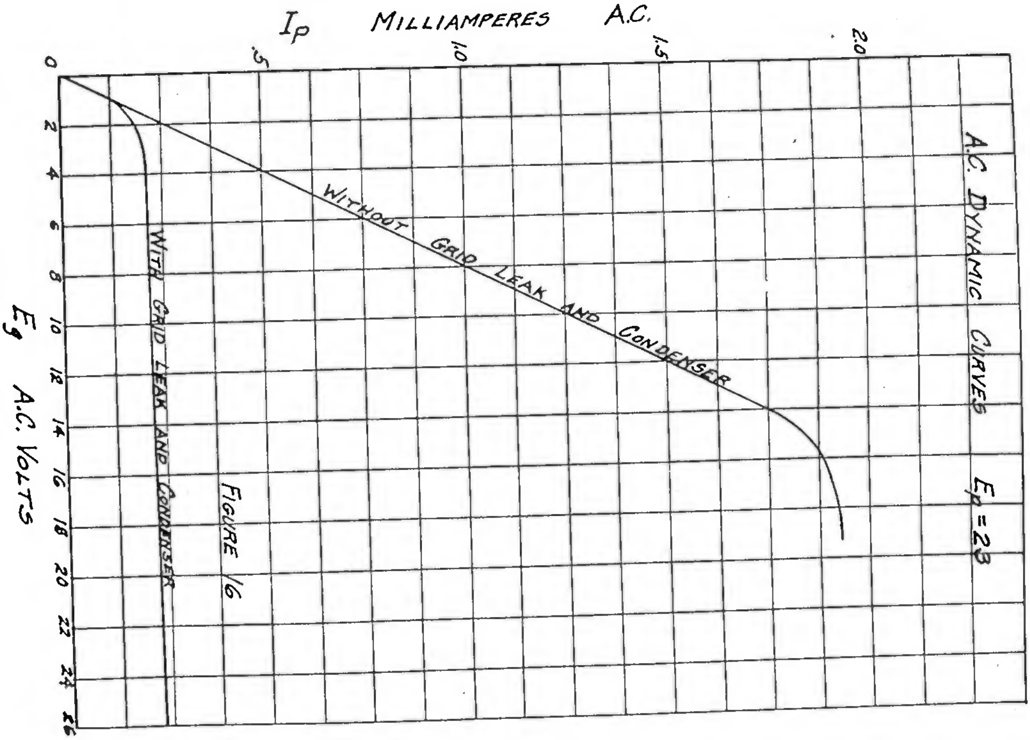


FIGURE 16

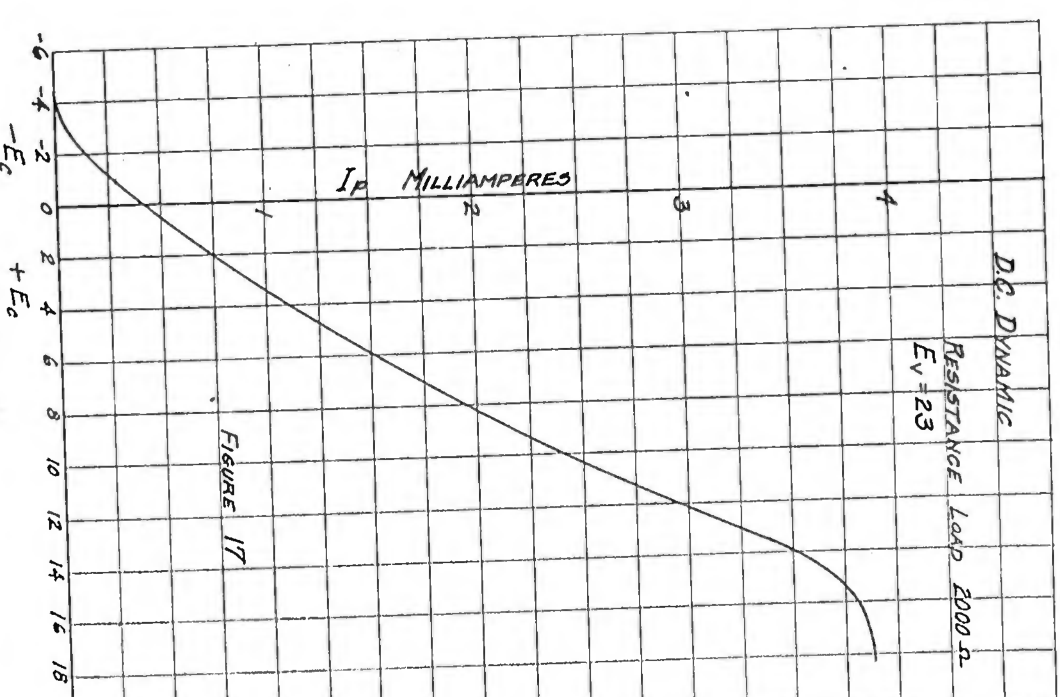


FIGURE 17

18. Comparing the static curve of Figure 19 with that of Figure 17 shows the reason for the relative changes in impedance.

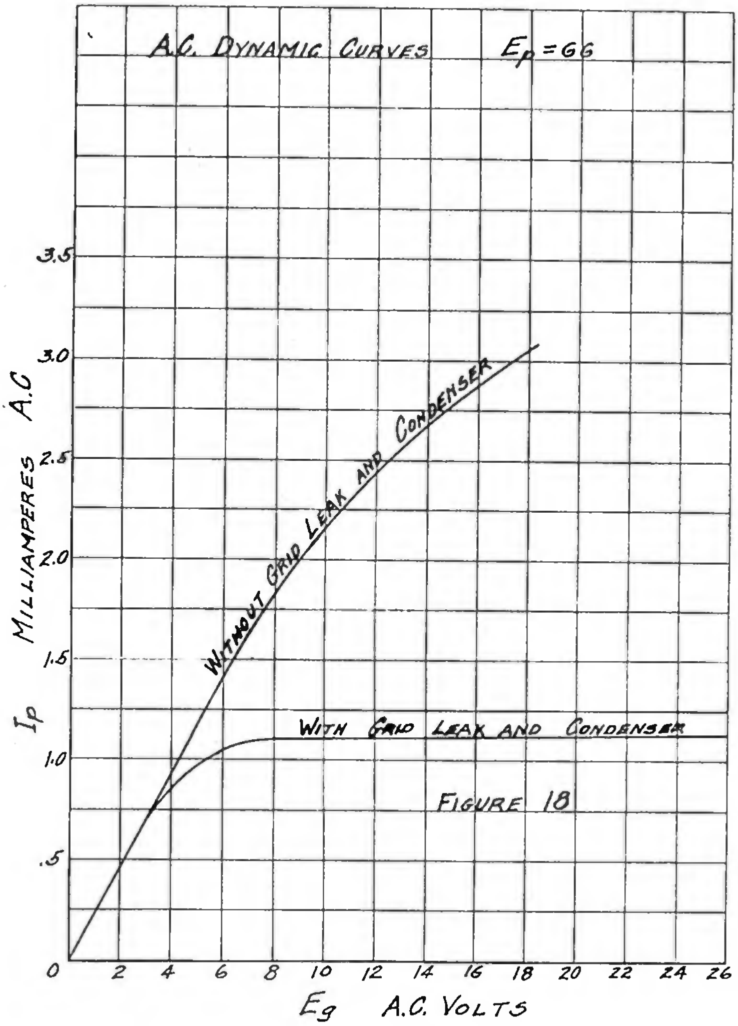


FIGURE 18

As previously pointed out, the maximum amplification will be obtained when the loss and input curves follow each other for the greatest distance. The dynamic curves for grid leak and condenser used under ordinary conditions (22½ volts on the plate), are always concave down. As seen from the loss curves we can adjust the grid leak to change the loss curve in R_g. If we continually adjust the tickler to maintain critical regeneration, decreasing the grid leak resistance increases the total loss curve, increases the tickler feed-back, and also decreases the rate of downward curvature of both curves. However, the curvature*

*In the following discussion the word curvature must be understood to mean curvature downwards. Thus if the curve under discussion is concave up, its curvature must be thought of as negative, and the phrase "decreasing its curvature" means increasing its rate of upward curvature.

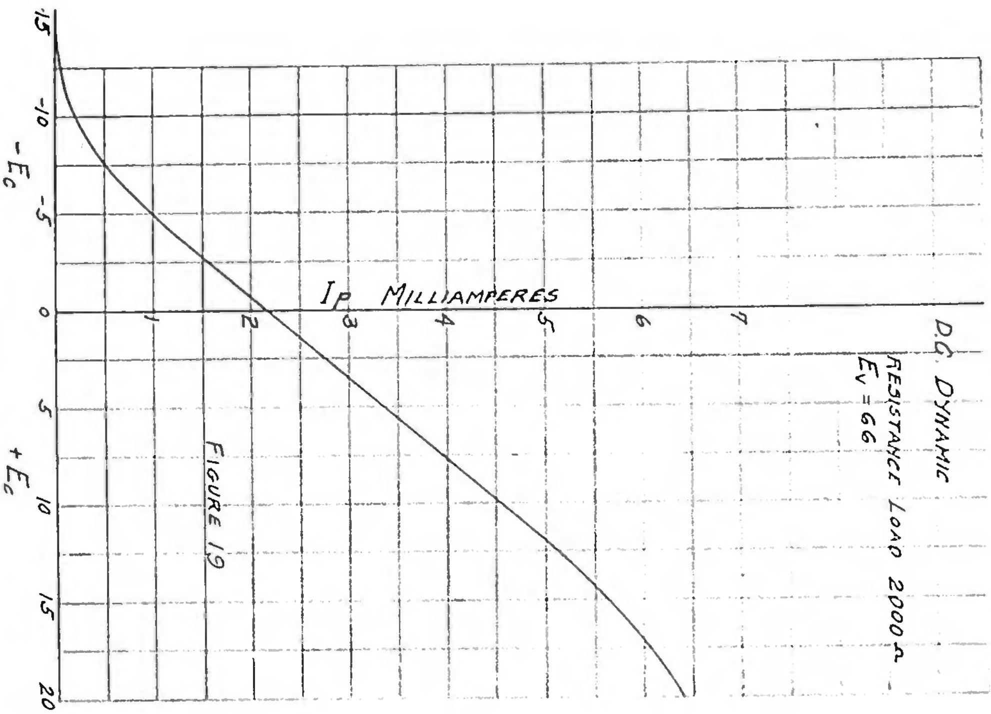


FIGURE 19

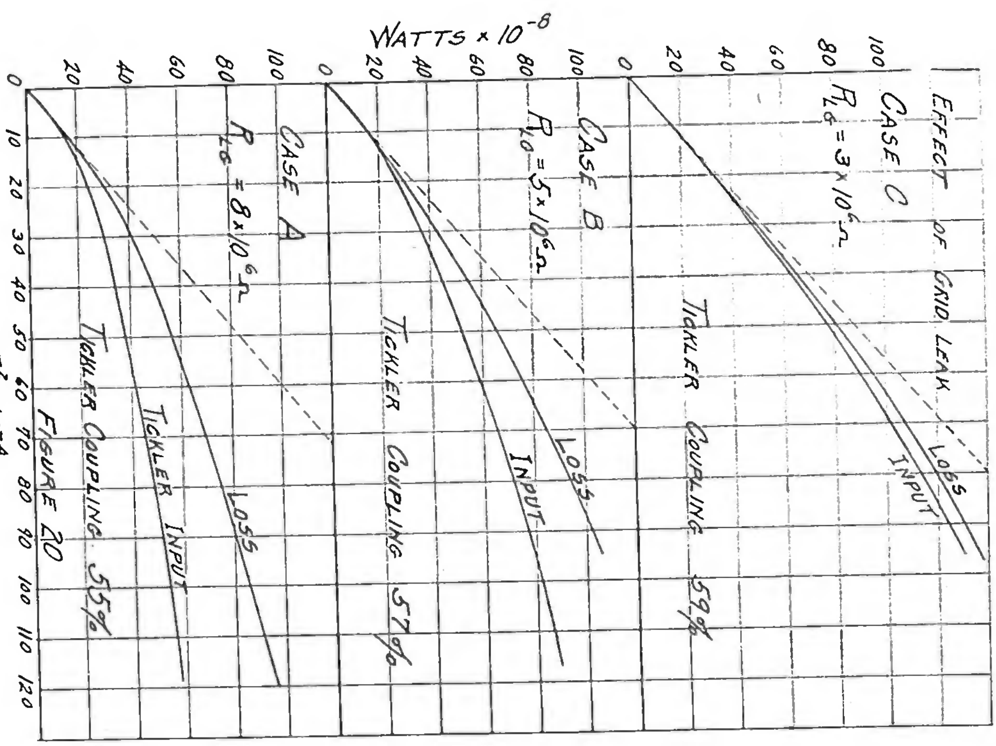


FIGURE 20

of the tickler input curve decreases faster than that of the loss curve. Thus the two curves continually approach each other. This condition is shown in the curves of Figure 20.

It is readily conceivable that with certain values of grid leak, the curvature of the tickler input may be less than that of the total loss curve. Now if the point of critical regeneration is approached, that is, where the loss curve and input curve are tangent at zero, the tube will "snap" into oscillation. Such a condition is shown in Figure 21 and in Figure 22. The input curvature is less than that of the loss in the range shown. However, with large values of E_g the input curve becomes horizontal. The loss curve always increases so that at some point these two will be equal. The tube (in Case C, both Figures) will oscillate at this value of E_g, not shown on the curve.

The relation between losses and input with high grid leak

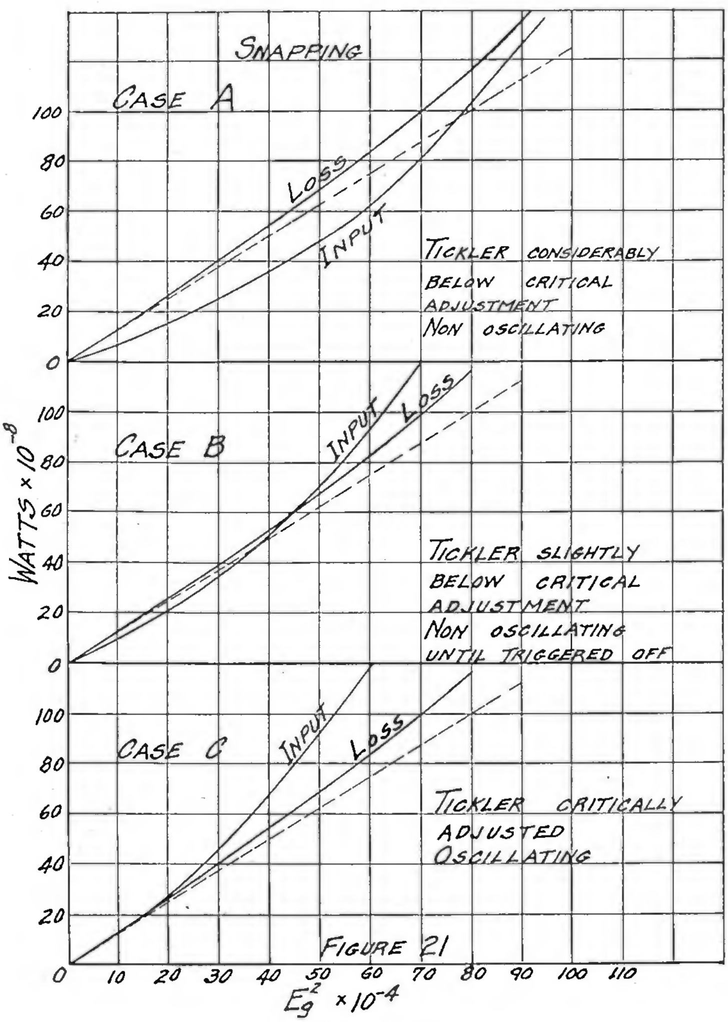


FIGURE 21

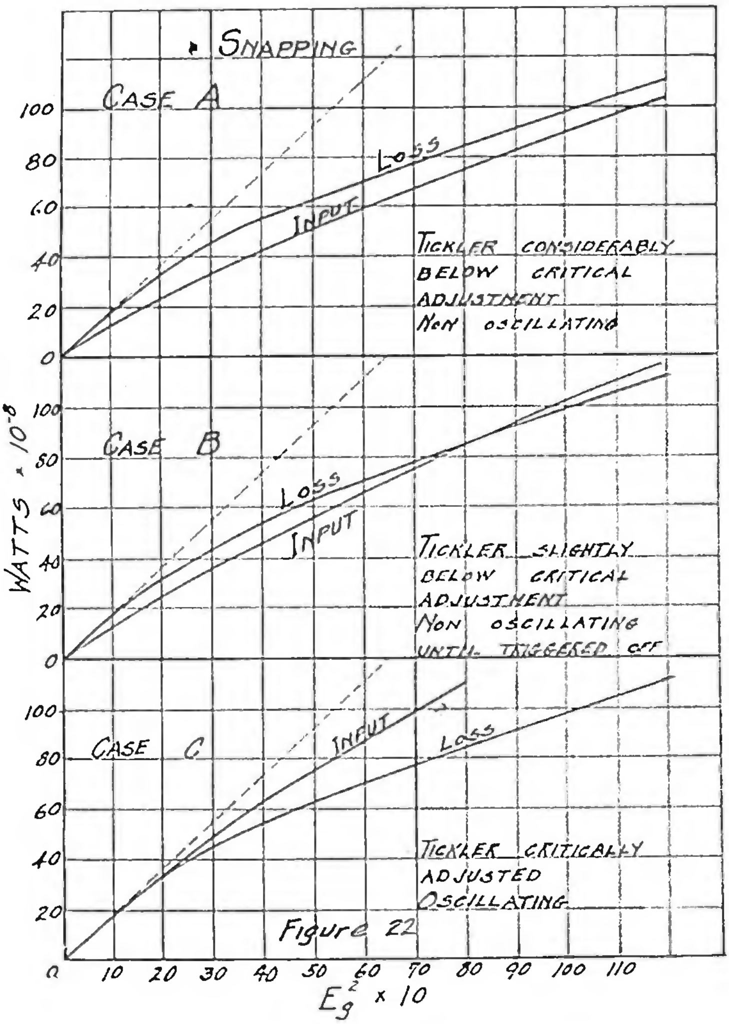


FIGURE 22

resistance shows that the two curves do not follow each other for a great distance and, therefore, the resultant amplification is low. Satisfactory operation cannot be obtained with snapping, as shown in Figure 22, so a low leak resistance will not produce any better results than a high resistance. Obviously there is some best value of grid leak to use with a regenerative detector. The results obtained when using this correct value explain in a large measure the growing use of a variable grid leak in sets where maximum sensitivity is desired on very weak signals, regardless of a multiplicity of adjustments.

Amplification With and Without Grid and Condenser

Figure 23 shows how the form of dynamic curve with grid leak and condenser effects the voltage amplification. The curves are a self-evident demonstration of the higher amplification without grid leak and condenser. These curves were taken at 22½ volts on the plate of the regenerator.

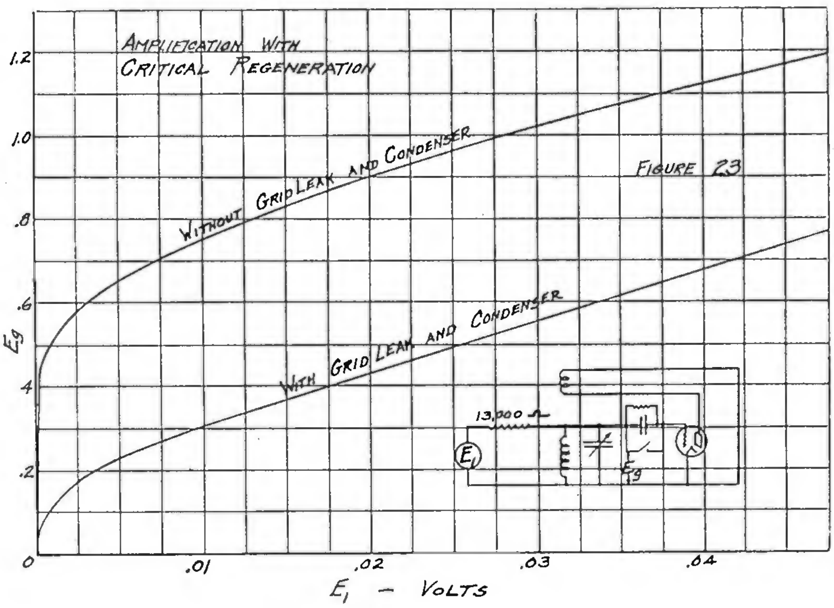


Figure 23

The decreased voltage amplification on the grid due to the introduction of the grid leak and condenser may or may not be compensated for by the increased detecting efficiency of the system. Thus the audibility of the signal as measured by an audibility meter in the plate circuit might be more with grid leak and condenser even tho a lower maximum voltage is obtained on the grid.

Effect of Resistance in the Grid Circuit

An important factor in the voltage obtained by regenerative amplification is the magnitude of the resistance in the grid circuit. So long as the tube can be made to oscillate, it is popularly supposed that the same voltage may be obtained at critical regeneration, regardless of the grid circuit resistance. Let us examine the actual results obtained by changing the resistance.

Assume the circuit of Figure 2 to be adjusted to critical regeneration and that E_1 is applied. Equation (4) will then express the power balance,

$$\frac{E_g(E_g-E_1)^2}{R_p}+f_p(E_g-E_1)+\frac{E_g^2}{K R_h}+\frac{E_g^2}{R_g}+f_g(E_g) = \frac{b\,E_g^2}{R_t}-f_{bt}(E_g) \qquad (4)$$

and a stable value of E_g will be obtained.

Now assume that E_1 is removed, R_h is increased and the tickler readjusted to critical regeneration. Assume that E_1 is now reapplied and also that the same value of E_g is momentarily existant. If the power loss under these conditions is less than the power supplied by the tickler, the voltage will build higher until a new condition of balance is obtained. If the power loss is greater than that supplied by the tickler, then we have chosen too large a value, and E_g will be smaller in a condition of balance than we have assumed to be momentarily existant.

An examination of equation (4) shows that the only terms affected by the changed conditions are:

$$\frac{E_g^2}{K R_h}; \quad \frac{b\,E_g^2}{R_t}; \quad \text{and } f_{bt}(E_g)$$

But since the tickler has been readjusted for critical regeneration with the increased resistance, the change in $\frac{E_g^2}{K R_h}$ has been exactly counterbalanced by the change in $\frac{b\,E_g^2}{R_t}$. Hence the only term tending to upset the equality is $f_{bt}(E_g)$. With a given E_g if the tickler input represented by $\frac{b\,E_g^2}{R_t}$ is increased by increasing the tickler coupling, then $f_{bt}(E_g)$ will be increased.

If the function is positive (that is, if the tickler curve is concave downward as in Figure 24), then the power loss exceeds the power input under the second set of conditions and the true balance will be obtained with a smaller value of E_g than that given by the first set of conditions. *This means that increasing the value of R_h will decrease the value of E_g even tho c.itical regeneration is obtained.*

Conversely, if this function is negative, E_g will build to a larger value in the second case than in the first. This case would seldom occur in practice. Notice, however, that the function being negative means that the tickler curve is concave upward. To obtain critical regeneration without self oscillation, under such conditions, the loss curve must have a larger rate of curvature upward than the tickler input curve. *Under these conditions, an increase in R_h would produce greater voltage amplification.*

If the function is zero (that is, if the tickler input curve is a straight line), the value of R_h will have no effect. This condition may be approximated in practice. In most cases, however, an increase in circuit resistance decreases the available amplification.

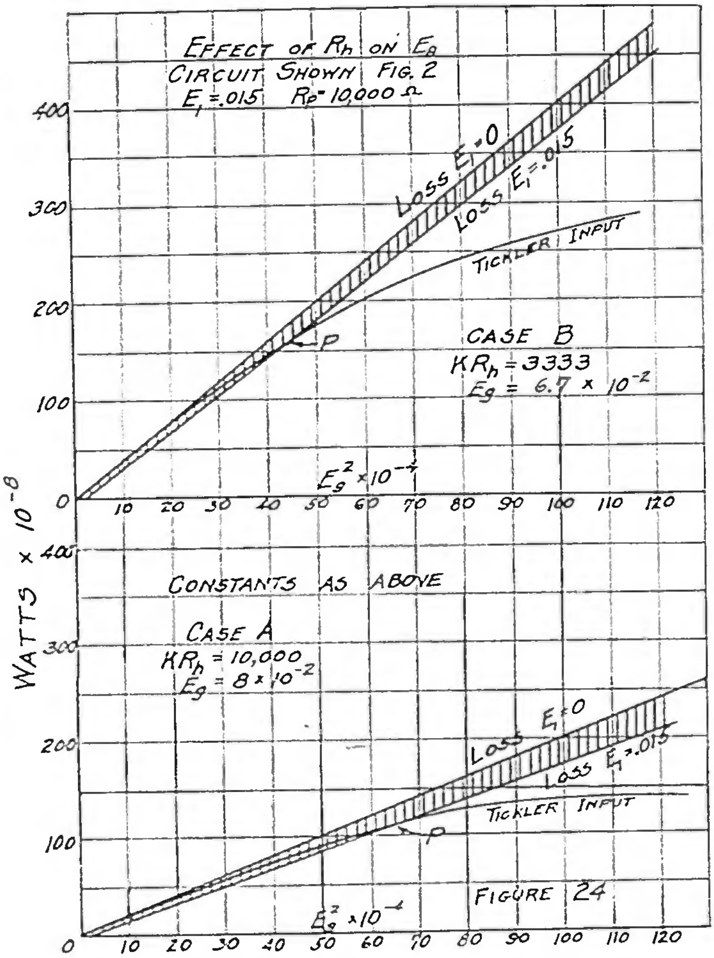


FIGURE 24

An illustration with curves will make the point clearer. Assume the circuit to be so adjusted that the loss curves are straight lines and the dynamic curve increases as shown in Case A, of Figure 24. Let the resistance loss be doubled by increasing R_h, as shown in Case B. Now double the value of tickler coupling. The effective resistance R_t is doubled, but the power input is not, due to a decrease in I_p, with increased plate circuit impedance. Thus in order to obtain the point of critical regeneration with the increased resistance losses, the tickler coupling must be more than doubled.

Let a certain value of E_1 be applied to the low loss circuit, in Case A. The new total loss curve is as shown, the area between the loss curves being slightly shaded. The lower loss curve can be obtained for any value of E_g with the given value of E_1, by subtracting the value given by (3) from the upper loss curve. The point of power balance, or the final value of E_g, is indicated.

Now consider the result when the same E_1 is applied to the higher loss circuit. The new loss curve lies the same distance below this upper loss curve as it did in the previous case. The value of "watts saved" depends only on E_1 and E_g, and is independent of R_h. This new loss curve will intersect the input curve at a very much lower value of E_g. Thus in this case increasing the resistance will decrease the voltage amplification (E_g/E_1), even tho the point of critical regeneration is obtained for the larger resistance.

Dynamic curves such as those shown in Figure 24, are obtained under normal conditions when using a grid leak and condenser. This indicates that under operating conditions, the resistance may very greatly affect the voltage amplification. Figure 25 shows the circuit which was set up to test this statement and the curve resulting. It can be readily seen that the grid voltage at critical regeneration decreases as the resistance losses in the oscillatory circuit are increased.

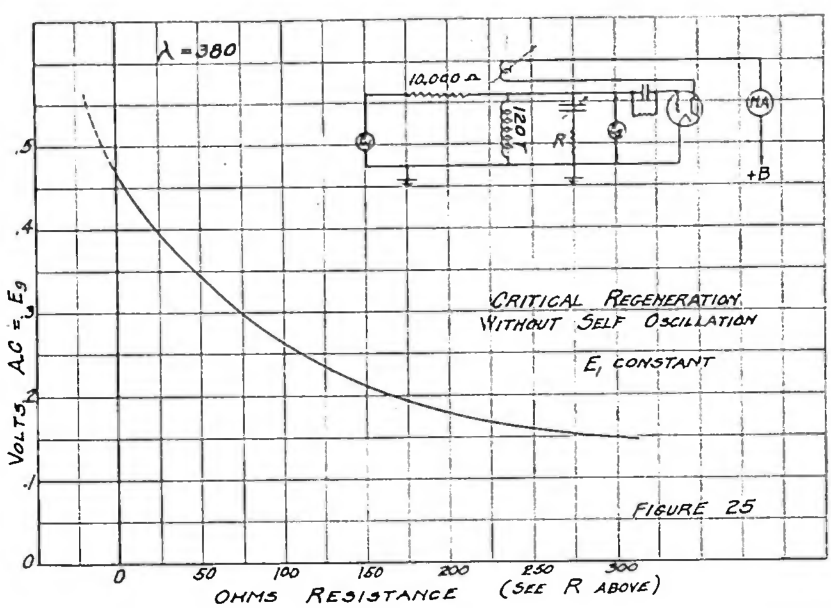


FIGURE 25

TURN RATIO FOR MAXIMUM AMPLIFICATION

Regenerated transformers often use a step-up ratio. As in the case of unregenerated transformers, there is a certain turn ratio which will give maximum amplification.

Consider the action in the circuit of Figure 2 if a step-up ratio is used. The equation for a power balance with critical regeneration and E_1 applied is

$$\frac{NE_g(N\,E_g-E_1)}{R_p}+f_p(N\,E_g-E_1)+\frac{E_g^{\,2}}{K\,R_h}+\frac{E_g^{\,2}}{R_g}+f_g(E_g) \\ =\frac{b\,E_g^{\,2}}{R_t}-f_{bt}(E_g), \tag{9}$$

where $N\,E_g$ is the voltage that would appear across the primary coil if the circuit were oscillating with E_g volts on the grid. Then N is the effective turn ratio of the primary to the secondary of the transformer.

$$\text{Watts saved } \frac{N\,E_g\,E}{R_p}=f_p(N\,E_g-E_1)+f_g(E_g)+f_{bt}(E_g).$$

For simplicity let us assume $f_p\,(N\,E_g-E_1)+f_g\,(E_g)$ is zero. In other words, the plate impedance of the first tube and the grid input impedance of the second tube do not vary as E_g changes. The effect of these on the result will be discussed later

$$\text{Then } f_{bt}\ (E_g)=\text{watts saved}=\frac{N\,E_g\,E_1}{R_p} \tag{10}$$

If we subtract "watts saved" from both sides of equation (9) and combine the terms

$$\frac{E_g^{\,2}}{K\,R_h} \text{ and } \frac{E_g^{\,2}}{R_g} \text{ into } \frac{E_g^{\,2}}{Z}$$

(where Z is the impedance of the tuned circuit across the terminals of the condenser when R_p is removed), we then have

$$\frac{N^2\,E_g^{\,2}}{R_p}+\frac{E_g^{\,2}}{Z}=\frac{b\,E_g^{\,2}}{R_t} \tag{11}$$

If we vary the ratio N and continuously readjust for critical regeneration, the rate of change of E_g with respect to N is zero at the point of maximum amplification. If we derive the above equation, letting $\dfrac{d\,E_g}{d\,N}=0$, and solve for N, we get a formula for the ratio giving maximum amplification. We must keep in mind, however, that the rate of change of $\dfrac{b\,E_g^{\,2}}{R_t}$ is not zero at this point, since the tickler is continuously readjusted. If R_t is small, compared to the tube impedance, $\dfrac{b\,E_g^{\,2}}{R_t}$ is directly proportional to $f_{bt}\,(E_g)$, hence their derivatives are proportional in the same ratio. That is

$$\frac{\frac{d}{dN}\frac{b\,E_g^2}{R_t}}{\frac{d}{dN}f_{bt}(E_g)} = \frac{\frac{b\,E_g^2}{R_t}}{f_{bt}(E_g)} \tag{12}$$

$$\frac{d}{dN}\frac{bE_g^2}{R_t} = \frac{b\,E_g}{R_t}\frac{\frac{d}{dN}f_{bt}(E_g)}{f_{bt}(E_g)} \tag{13}$$

The derivation of (10) gives

$$\frac{d}{dN}f_{bt}(E_g) = \frac{d}{dN}\frac{N\,E_g\,E_1}{R_p} = \frac{E_g\,E_1}{R_p} \tag{14}$$

Substituting in equation (13), the values given by equation (11), by equation (14) and equation (10), gives

$$\begin{aligned}\frac{d}{dN}\frac{b\,E_g^2}{R_t} &= \left[\frac{N^2\,E_g^2}{R_p} + \frac{E_g^2}{Z}\right]\frac{E_1\,E_g}{R_p}\,\frac{R_p}{N\,E_g\,E_1} \\ &= \left[\frac{N^2 E_g^3}{R_p} + \frac{E_g^3}{Z}\right]\frac{1}{N}\end{aligned} \tag{15}$$

Hence deriving equation (11), we get

$$\frac{2\,N\,E_g^2}{R_p} = \left[\frac{N^2\,E_g^3}{R_p} + \frac{E_g^2}{Z}\right]\frac{1}{N} \tag{16}$$

Simplifying (16).

$$\frac{2\,N^2\,E_g^2}{R_p} = \frac{N^2\,E_g^2}{R_p} + \frac{E_g^2}{Z} \tag{17}$$

Solving (17) for N, gives

$$\frac{N^2}{R_p} = \frac{1}{2} \tag{18}$$

$$N = \sqrt{\frac{R_p}{Z}} \tag{19}$$

Hence the best ratio is approximately $\sqrt{\frac{R_p}{Z}}$. It can be proved that this ratio is one which will give a primary impedance equal to R_p. This is the same ratio as that which would be chosen for an unregenerated transformer.

It was assumed in the above that $f_p(N E_g - E_1)$ and $f_g(E_g)$ were equal to zero. If $f_g(E_g)$ has a positive value, however, the result is changed toward a one-to-one ratio. $\frac{d}{dN} f_{bl}(E_g)$ will not be affected since $\frac{d}{dN} f_g(E_g)$ is zero at the point of maximum amplification.

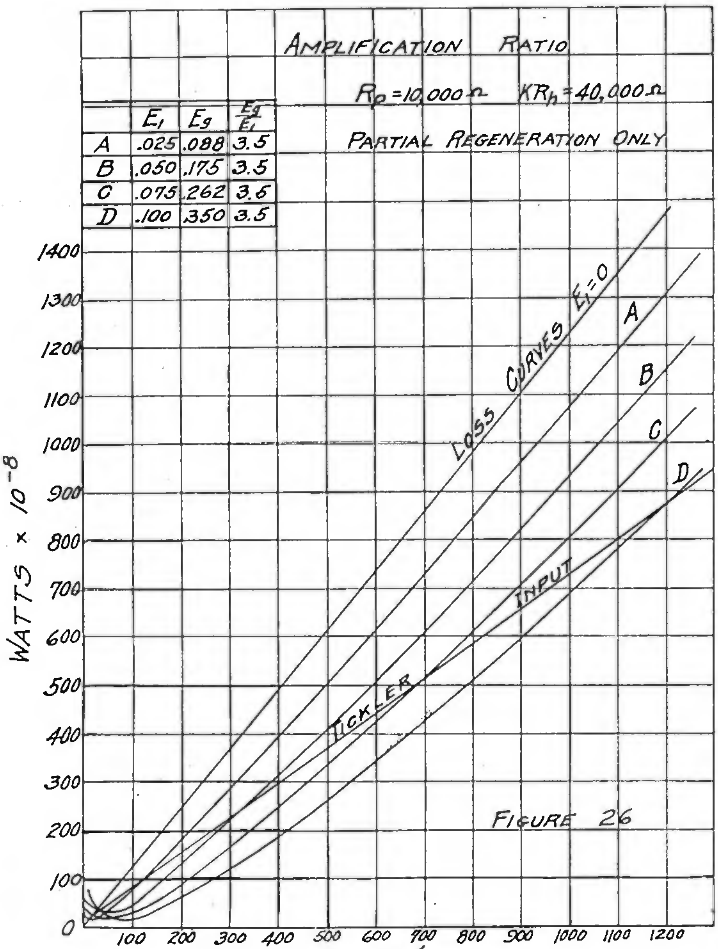


FIGURE 26

Normally $f_p(N E_g - E_1)$ will be so small as to affect the result only slightly.

It is evident from equation (19) and the above discussion, that for maximum amplification in a regenerated transformer, the correct ratio between secondary and primary turns $\left(\text{that is, } \frac{1}{N}\right)$ is equal to the best ratio of the transformer when unregenerated.

VARIATION IN AMPLIFICATION WITH VARYING INPUT

A subject of much discussion concerning regeneration and one which muct be included here, is that of the ratio between the initial and the final signal, that is, the ratio between the grid voltage on the regenerating tube and the voltage on the grid of the radio frequency amplifier.

Two conditions are possible in this consideration, first, when the tube is capable of regenerating, and second, when it is adjusted to critical regeneration. Let us consider the action when the tube is capable of regenerating but not adjusted critically. The loss curve and the tickler input curve are assumed to be straight lines. That is, all functional terms in equation (4) are zero. If we let

$$\left(\frac{1}{R_p}+\frac{1}{K R_h}+\frac{1}{R_g}-\frac{b}{R_t}\right)=+S$$

the solution for E_g is

$$E_g=E_1\frac{1}{S R_p} \tag{20}$$

This means that E_g is *directly proportional* to E_1 for all values of E_1. Or in other words, the amplification ratio is independent of the strength of the applied signal. This is one of the conclusions reached in the "Bureau of Standards Sci. Papers," Number 487. Figure 26 shows loss curves and the corresponding values of E_g with a given tickler adjustment. Note that the amplification ratio is a constant.

However, there is one fallacy in this proof. The tickler input curve is seldom a straight line, as was assumed. It is usually concave downward. It is evident that under these conditions doubling the value of E_1 will not double the value of E_g. The actual ratio for any value of E_1 will depend on $f_{bt}(E_g)$, or on the difference between the actual input curve and a straight line. Thus under normal conditions a greater amplification ratio will be obtained when using a small value of E_1. A close inspection of the observed points in Figure 3 of the above mentioned "Bureau Sic. Paper" *shows this actually to be the case in their results.* This apparent discrepancy was evidently attributed to experimental error and no attempt made at explanation.

While this variation from a constant amplification ratio is almost immeasureably small when low values of "feed back" are used, this effect becomes of supreme importance when the circuit is adjusted to critical regeneration. Such a condition is shown in Figure 27. The sharp bend in the tickler input is somewhat exaggerated, altho this can be approximated under some conditions. Due to the enormous amplification obtained at critical regeneration, smaller values of input voltage were used for these curves. The loss curves for various values of applied signal are shown and the resulting values of E_g are noted. In this case it may be seen that the amplification ratio is almost *inversely* proportional to the strength of applied signal. This conclusion was previously advanced by N. C. Little, in his article on "The Limit of Regeneration."

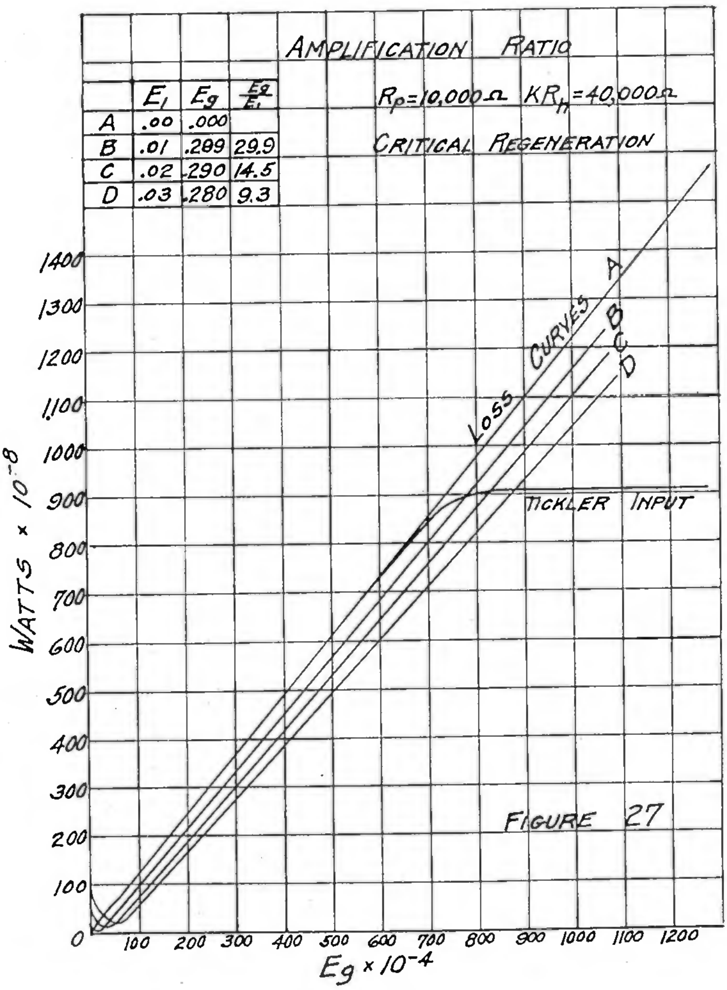


FIGURE 27

This effect accounts for the possibility of receiving distant stations when the amount of received energy is extremely small. The amplification ratio with such a weak signal may be as much as ten thousand or more at critical regeneration.

Figure 28 shows curves actually obtained by measurement between amplification ratio and the strength of applied signal. Notice how rapidly the amplification ratio increases as the strength of applied signal decreases. The curves also show the difference in amplification ratio using a grid leak and condenser. This result could have been predicted by considering the change in the shape of the tickler input curve when using a grid leak and condenser.

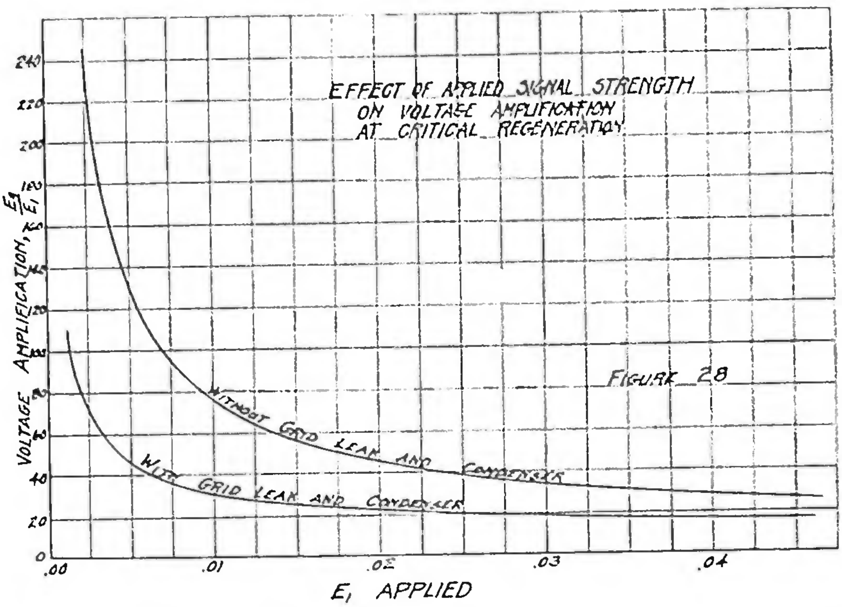


FIGURE 28

One result of this discussion is to show how greatly inferior radio frequency amplification is with respect to regeneration, especially when receiving distant and weak signals. With moderate signal strength a radio frequency amplifier may give a stronger output than a regenerative receiver, but if the signal input is reduced continuously a point will eventually be reached where the regenerative set will excel the radio frequency amplifier. The very nature of regeneration in amplifying weak signals more than strong ones is an effect which is more than we would honestly dare to require..

CIRCUIT CONDITIONS AS AFFECTING DISTORTION

In all of the above discussions we have assumed that the voltage E_1 was applied for an unlimited time and we have discussed the value of E_g after it had reached a stable condition. Under this consideration the L/C ratio (if $K\ R_h$ remains the same) will not affect the above results in the least. However, if the time of application is limited or if E_g must follow the amplitude variations of a modulated continuous wave input, then the L/C ratio is important, since it affects the rate of increase of E_g.

We see from the expression (3) that with a given E_1 (where E_1 is small), the value of "watts saved" increased almost directly as E_g. We have also seen that as E_g increases, the losses increase until a point of balance is reached. Until this point is reached the watts saved are greater than the losses. In considering the time relations involved we can assume that the average available watts for building up the voltage E_g is a small part of the value given by (3). These average available watts are used to charge the condenser as E_g increases.

The bearing of the above on the effect of the L/C ratio can best be shown by concrete example, assuming plausible values for the circuit constants.

Let $E_1 = 0.1$ v. $E_g = 2.0$ v. at the point of balance.

Let $C = 100$ mmfd. $F = 1{,}000{,}000$ cycles per second.

$R_p = 10{,}000$ ohms.

$$\text{Then } \frac{E_1 E_g}{R_p} = \frac{0.29}{10{,}000} = 29 \times 10^{-6} \text{ watts}$$

This is the value of watts saved at the point of balance due to E_1. Then 10^{-5} is approximately the average value during the rise of E_g.

Assume that about 1/5 of the average watts saved are available for charging the condenser, the rest going to supply the extra losses due to the change in the functional terms as E_g increases. Then the average value of power available for charging the condenser will be about 2×10^{-6} watts. The energy stored in the oscillating circuit when $E_g = 2$ v. is given by the expression

$$1/2\, C E_g^2 = \frac{100. \times 2}{2 \times 10^{12}} = 2 \times 10^{-10} \text{ watt seconds.}$$

$\dfrac{2 \times 10^{-10}}{2 \times 10^{-6}}$ equals 1×10^{-4} seconds. This gives the time for the voltage to build up under the above assumptions. This time is equivalent to 100 cycles at the assumed frequency.

Of course it cannot be said that E_g builds to a final value in 100 cycles, since mathematically it never reaches the stable condition but approaches the final value asympototically. In other words, as the circuit approaches the stable condition, the rate of increase of E_g decreases so that it continually approaches, but never reaches its limiting value. In a small fraction of a second, however, E_g approximates its limiting value so closely that the difference would be too small to measure. This small fraction of a second necessary for any value of E_1 to charge the condenser to the value determined by E_g, determines the maximum rate at which E_g can be changed and still obtain its maximum value. The inverse of this fraction gives the approximate frequency of modulation at which a decrease in amplification will become noticeable. In the case discussed this would be about 10^4 cycles. Higher frequencies would suffer greater decrease in amplification and lower frequencies scarcely at all. This explains why in general regeneration causes but little distortion.

If E_1 is decreased, E_g will decrease. If E_1 and E_g decrease in the same proportion, the watts available for charging the condenser will decrease in proportion to E_g^2. The energy in the condenser is directly proportional to E_g^2. Consequently the time required (and so the number of cycles) to charge the condenser is the same regardless of E_1. However, if E_g does not decrease as fast as E_1 (usually the case), the watts available for charging the condenser will be less in proportion. Thus to obtain the maximum voltage (approximately), longer time will be required when E_1 is small. In this case modulation frequencies below 10,000 cycles may suffer loss of amplification. This means that with critical regeneration, weak signals will suffer more distortion, due to the disproportionate amplification of the lower frequencies than will strong signals. This fact has often been observed in receiving voice modulated continuous waves.

If the circuit is tuned to another frequency by changing L, the energy in the condenser for given balance point (E_g) is constant. In this case the time required is independent of the carrier frequency and there is no more tendency to distort the modulation on low carrier frequencies than on high frequencies. If, however, the capacity is increased to tune to a lower frequency (higher wave length), more time in proportion will be required to obtain an approximate maximum E_g, and consequently the modulation will suffer greater distortion.

Optimum Heterodyne

It is generally understood that regenerative amplification only occurs when, independent of the signal, there are no local oscillations produced. It is nevertheless a fact that when the circuit is oscillating weakly, as in heterodyne reception, very great amplification is obtained of the beat between the signal and the local oscillations. This amplification is due to regenerative action.

Consider the action taking place when the circuit of Figure 2 is oscillating at a frequency f_2. The equation (4) previously given (using $E_1=0$), will give the equation of balance. Let E_1 be applied at a frequency f_1. At the instant of balance when E_1 and E_g are 180° out of phase, the decrease in power loss is given by

$$\frac{2\,E_1\,E_g}{R_p} \tag{3}$$

and as a result E_g will tend to increase as previously explained.

Now assume the point where E_1 and E_2 are in phase. The losses in R_p increase, due to the increased voltage across it. In this case E_1 is supplying power to the circuit.

The value of the loss in R_p when E_1 and E_g are in phase is given by

$$\text{watts in } R_p=\frac{E_g{}^2+2\,E_1\,E_g+E_1{}^2}{R_p}$$

However, E_1 is supplying part of this.

$$\text{Power supplied by } E_1=\frac{E_1{}^2+E_1\,E_g}{R_p}$$

$$\text{Power supplied by } E_g=\frac{E_g{}^2+E_1\,E_g}{R_p}$$

Extra watts supplied by E_g, due to application of E_1 is $\dfrac{E_1E_g}{R_p}$ = Extra watts loss due to E_1.

It is evident that E_g will decrease and increase at the frequency (f_1-f_2), and the amplitude of this variation depends on the values of the "watts saved" and the "watts lost." It has previously been shown that for a given E_1 a low loss tuned circuit produces more amplification (a larger E_g). In addition, in heterodyne reception we want the "watts lost" to also be as large as possible. It is evident that the power available for changing E_g increases, if the amplitude of local oscillation is increased. The rate at which the power loss and tickler input curves diverge will also increase as the initial value of E_g is increased. This has a tendency to decrease the amplification.

As these two tendencies oppose, there is obviously an optimum initial value of E_g which will give the maximum response at the heterodyne frequency.

The efficiency of detection as affected by the initial value of E_g must also be considered in a complete analysis. The optimum point can only be found by trial.

The amplitude of the heterodyne beat will not be affected as f_1 and f_2 are separated, except as the time required to change the charge on the condenser becomes appreciable. When the beat frequency is high, this effect noticeably decreases the amplitude of the beat.

"Locking-in" and Zero Beat Reception

In considering the effect of local oscillations with respect to regeneration, another well-known phenomena is to be considered. This is the effect of "locking in," where "locking in" refers to the dead spot observed in the center of the heterodyne beat note obtained between the local oscillator and the incoming signal.

Consider the circuit shown at "M," Figure 29. This is equivalent to Figure 2, except that L and C are tuned to a frequency f_2, and the tickler coupling has been increased until the tube is oscillating with the stable value E_g on the grid. E_1 is applied at frequency f_1.

FIGURE 29

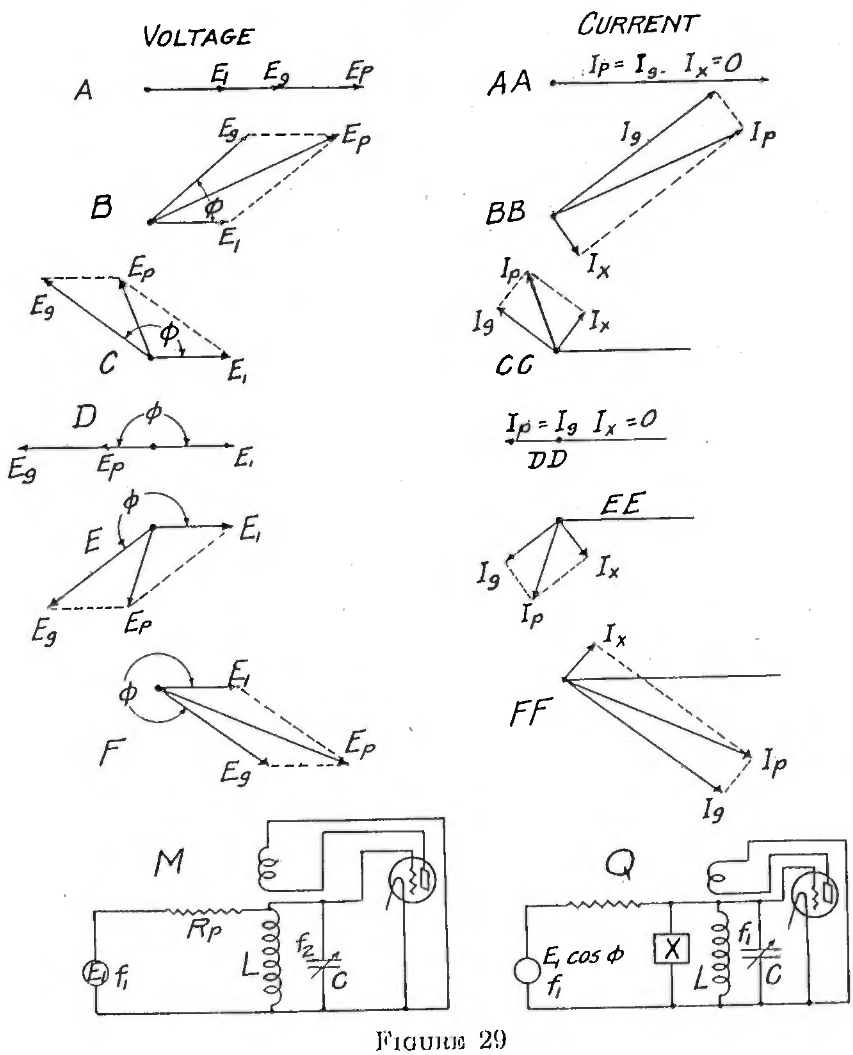


Figure 29

Let us assume an initial condition such as shown by the vectors at A. E_1 and E_g are in phase and the current in R_p is in phase with E_g (E_g is assumed slightly larger than E_1). Assume f_2 to be greater than f_1.

Then the phase relation between E_1 and E_g will change and the phase angle ϕ will increase. The phase relation for different values of ϕ are as shown at B, C, D, E, or F. The voltage E_p across the resistor and the current I_p thru it, are no longer in phase with E_g. As shown at BB, CC, DD, EE, and FF, however, the current has one component I_g, in phase with E_g and another I_x at right angles. Now

$$I_p^2 = \frac{E_g^2 + E_1^2 + 2\, E_g\, E_1 \cos\phi}{R_p^2} = I_g^2 + I_x^2$$

$$I_g = \frac{E_g + E_1 \cos\phi}{R_p}$$

$$I_x = \frac{E_1 \sin\phi}{R_p}$$

I_p is out of phase with E_g. Hence if an equivalent circuit is desired, in which the applied voltage is 180° out of phase with E_g, a reactance "X" must be shunted across R_p. Such an equivalent circuit is shown in Figure 29 at "L."

The value of "X" must be such that with E_g volts across it, a current of I_x amperes will flow thru it. Then

$$X = \frac{E_g}{I_x} = \frac{E_g\, R_p}{E_1 \sin\phi}$$

The current in R_p in this equivalent circuit is

$$I_g = \frac{E_g + E_1 \cos\phi}{R_p}$$

and the value of $I_p^2 = I_x^2 + I_g^2$ has not been changed. Hence X is an equivalent reactance introduced into the circuit by the phase relation between I_p and E_g.

Let us take a condition when f_2 is larger than f_1. Then if ϕ lies in the first or second quadrants, $\sin\phi$ is positive and X is positive (that is, an inductive reactance). Under these conditions a lagging current is flowing, as shown at BB and CC. The equivalent reactance X being inductive, the circuit is, in effect, tuned to a higher frequency and the value of f_2 is increased even farther above f_1. Hence the rate of change of ϕ is increased.

If ϕ lies in the third or fourth quadrants, then $\sin\phi$ is negative and X is negative (that is, a capacitive reactance). Hence f_2

is decreased in value toward f_1. This condition (that is, leading current) is shown at EE and FF.

If f_2-f_1 is not too large before E_1 is applied, it is evident that after E_1 is applied, X will have some value X_o at which f_2 is changed so that $f_2=f_1$.

Let α = the value of ϕ at which $X=X_o$ then

$$X_o=\frac{E_g R_p}{E_1 \sin \alpha}$$

Since f_2 was originally greater than f_1, then X_o must be negative (that is, capacitive), in order to lower the frequency of oscillation. Hence $\sin \alpha$ must be negative. There are two values that satisfy this condition, one in the third quadrant and one in the fourth. Let the third quadrant value be called α_1 and the fourth quadrant value α_2.

If ϕ lies in the first or second quadrants, X is positive and the rate of change of ϕ (that is, $2\pi(f_2-f_1)$) is large.) If ϕ lies in the third quadrant, but is less than α, $2\pi(f_2-f_1)$ is small, but is still positive; that is, ϕ increases slowly. If ϕ should happen to have a value greater than α_1 but less than α_2, then X is greater than X_o, and $2\pi(f_2-f_1)$, will be negative, that is, ϕ will decrease. If ϕ is greater than a_2, then $2\pi(f_2-f_1)$ is again positive in value until $\phi=a_1+2\pi$

From the above it is readily seen that, regardless of its initial value, ϕ will approach a_1 and remain at that value. It may be shown by means of a similar argument that if f_2 is originally less than f_1, the rate of change of ϕ is negative when E_g and E_1 are in phase. ϕ approaches a stable value a_1 in the second quadrant. The point of unstable equilibrium (where $\phi=\alpha_2$) is in the first quadrant.

If the frequency to which the circuit itself tunes is increased above f_1, the angle α_1 approaches $\frac{3\pi}{2}$ in value. If the circuit is tuned to a still higher frequency, f_2 will always be greater than f_1 and the rate of change of ϕ is always positive. In this case a beat note will be heard.

If the circuit is tuned below f_1, α_1 approaches $-\frac{3\pi}{2}$ in value and again, if the mistuning is carried farther, a beat will occur.

On the other hand, as the circuit is tuned toward f_1, the value of a_1 approaches π or 180° while a_2 approaches 0. If the circuit is tuned fairly accurately to f_1, the amplitude of the signal E_1 may be diminished to a very small value and E_g will still remain in approximately the correct phase relation, due to the stabilizing effect of X_o, which comes into play when the phase tends to shift. This is the well-known phenomena of zero beat reception. Here, as in heterodyne reception, an optimum initial value of E_g will be found by trial.

Regeneration and Sharpness of Tuning

It is a well-known fact that regeneration sharpens the tuning of a circuit by amplifying resonant frequencies to a greater extent than non-resonant frequencies. An analysis of the amplification at non-resonant frequencies will enable us to construct the resonance curve for any given tuned regenerated circuit.

First, consider a few facts concerning a non-regenerated circuit as shown in Figure 3. For convenience we assume the frequency f_1 of the applied voltage E_1 to be varied and the resonant frequency f_2 to be constant. If f_1 is higher than f_2, the tuned circuit offers a capacity reactance and the current thru R_p is leading E_1. At this time E_g is leading E_1 by some angle ϕ. As f_1 approaches f_2, E_g approaches 180° out of phase with E_1. When f_1 is less than f_2, ϕ has increased to more than 180°, or E_g is lagging with respect to E_1.

As shown under the discussion on "Locking-In and Zero Beat Reception," the current which is 90° out of phase with E_g is

$$I_x = \frac{E_1 \sin \phi}{R_p}$$

The equivalent reactance shunted across the tuned circuit thru which this current is flowing is $X = \dfrac{E_g R_p}{E_1 \sin \phi}$. It was also shown that ϕ is of such value that X tunes the circuit to resonance with the applied signal. For simplicity, we assume in this discussion that the degree of regeneration is not changed by retuning with purely reactive shunts, and also that the tube used has straight line characteristics.

From the equation giving the value of X, the value of E_g may be obtained. $E_g = \dfrac{E_1 X \sin \phi}{R_p}$. Now let us regenerate the circuit by adding a tickler, giving the circuit of Figure 2, and adjust the tickler to obtain critical regeneration. At any instant $E_1 \cos \phi$ is the effective signal applied to the circuit. Since the circuit is now regenerated, E_g increases. However, as E_g increases, the angle ϕ approaches 90°, which increases $E_1 \sin \phi$. It is evident that the limit is the point where ϕ reaches 90° in value, for at that point $\sin \phi$ is maximum. Notice also that the in-phase component of E_1 is zero. Substituting the maximum value of $sin\ \phi$ in the above equation, we get $E_g = \frac{E_1 X}{R_p}$, which is the value to which E_g will build. As f_1 approaches f_2, the value of X increases and E_g increases. The resonance curve obtained, plotted between E_g and the applied frequency f_1 is, therefore,

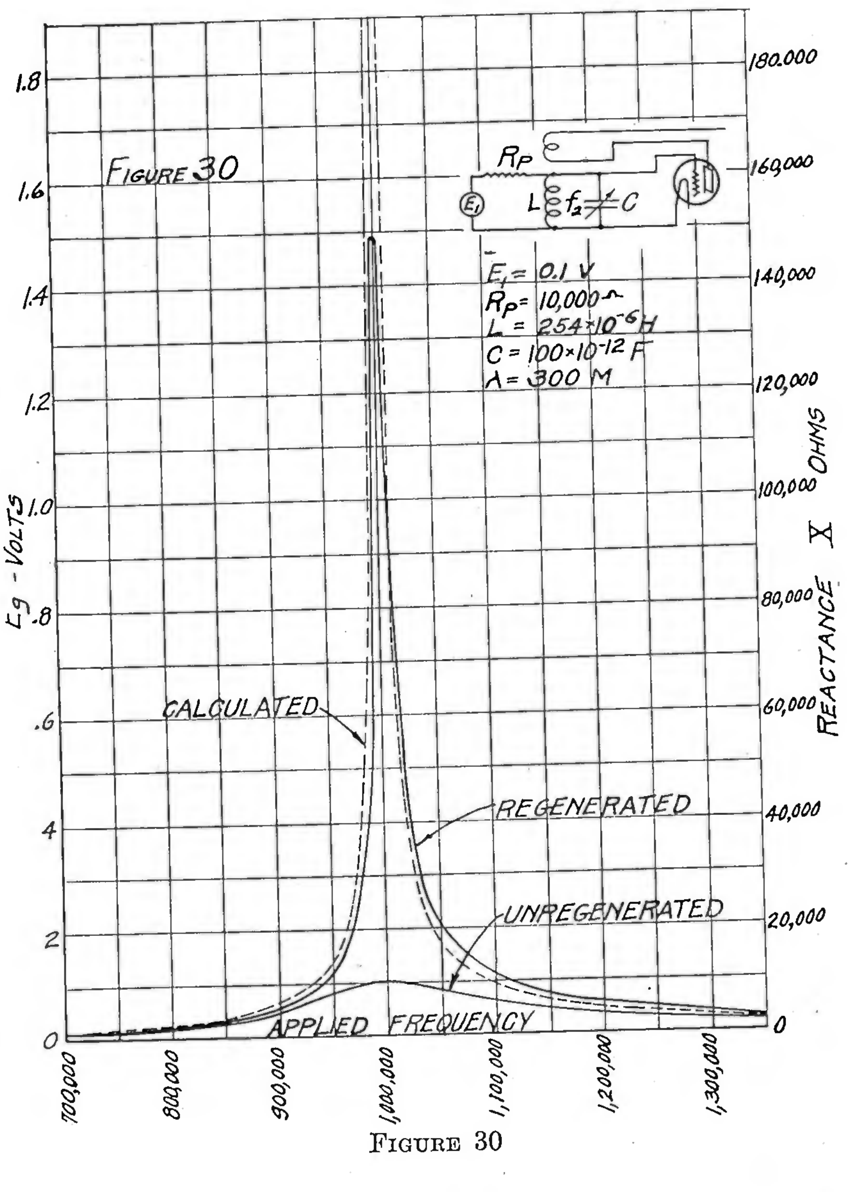


FIGURE 30

proportional to the curve plotted between f_1 and the parallel reactance necessary to tune the circuit to resonance at the frequency f_1. Such a resonance curve is shown by the dashed line in Figure 30.

The resonance curve obtained by the above method is modified by two factors. First, the limiting effect of the functional terms of equation (4) prevent E_g from reaching an infinite value, the actual maximum value being that given by equation 4.

Second, critical regeneration is not maintained when the circuit of Figure 2 is retuned by reactive shunts. When an inductive reactance is shunted across a tuned circuit, the loss in that circuit *decreases*. (E_g constant). When a capacity reactance is shunted across such a tuned circuit, the loss *increases*. If adjusted for critical regeneration at the frequency f_2, and if f_1 applied is higher than f_2, the circuit is capable of oscillating at a frequency f_2. This is due to the decreased loss when the resulting inductive reactance is shunted across the circuit.

This tends to raise the amplification above that given by the above equation. Conversely, if f_1 is lower than f_2, the loss is increased and the amplification is less than that indicated by the equation. The effect of these variables on the curve derived from the equation is shown by the regenerated resonance curve in Figure 30.

Application to an Antenna

As will be observed, all of the above discussion has applied only to a regenerative stage of radio frequency amplification. When the same type of discussion is attempted concerning a regenerated antenna circuit, it is somewhat more difficult to draw an equivalent circuit. This is because it is difficult to show, in a simple way, the manner in which the signal voltage is applied.[2]

Whether the signal voltage is introduced into an antenna in series or in any other conceivable way the following remarks must hold true for it or for any type of regenerated circuit.

The equation expressing critical regeneration must take the form

$$\frac{E_g^2}{Z} = \frac{b\,E_g^2}{R_t}$$

where Z is the impedance of the tuned circuit, and E_g is an infinitesimal voltage.

When an amount of power "W" is introduced into the circuit or if "W" watts are prevented from being lost, there is a surplus of power which tends to charge the tuning condenser to a higher

[2]The authors believe that, mathematically, the signal voltage E_1 may be considered as being applied in series with a hypothetical resistance shunted between antenna and ground binding posts of the regenerated receiver; the resistance being of such a value as to replace the actual radiation resistance in damping effect. When this assumption is made, the regenerated antenna circuit reduces to the same equivalent circuit as the regenerated transformer, and the same formulas apply thruout.

value. As E_g increases, the tickler input can no longer be represented by $\frac{b\,E_g^2}{R_t}$ but is less by an amount $f_{bt}(E_g)$.

A condition of balance[3] will be obtained such that

$$\frac{E_g^2}{Z} = W + \frac{b\,E_g^2}{R_t} - f_{bt}(E_g).$$

If the circuit resistance is increased so that Z decreases, the tickler setting must be increased to maintain critical regeneration. If the tickler coupling is greater, the term $f_{bt}(E_g)$ is larger for a given value of E_g. Hence the circuit comes to a balance with a smaller value of E_g than previously.

Since the term $f_{bt}(E_g)$ is larger when a grid leak and condenser is used, the point of balance must always be lower with a grid leak and condenser.

In a similar manner, the principal points discussed with regard to a regenerated transformer could be applied to a regenerated antenna, and the conclusions resulting must be the same.

Conclusion

The most important point in a discussion of this kind is the character and point of application of the input voltage. It may be considered as a series voltage in the secondary circuit if it is recognized that the value of this voltage will vary with any change in the current in the secondary circuit. Thus the theory of voltage summation as cited in the introduction is unsound, even if the variations of tube characteristics are considered, the reason being that E_1 (at the point considered) will vary from cycle to cycle in any practical circuit. The only way any voltage may be considered as unaffected in value by load conditions is when the impedance of the generator is considered in the discussion.

In this paper the validity of the existing theories of regenerative amplification are questioned and an explanation offered based on a power balance. The circuit conditions affecting this power balance are discussed. The resulting amplification under various circuit conditions are given, including grid leak and condenser. Some of the more commonly observed phenomena are discussed in relation to the theory advanced in this paper. The time relations involved are briefly discussed and their effect considered. The entire theory may then be transferred to an antenna.

[3] For simplicity the terms $f\ (E_g)$ and $f_g\ (E_g)$ are omitted in this general equation. Neither of these terms are very important compared to $f_{bt}(E_g)$.

In conclusion we may state that altho this paper seems rather long, many facts regarding regenerative amplification as affected by possible circuit changes are omitted. This was not due to lack of application of the present theory, but due to time limitations. Only such experimental proof as was necessary in direct confirmation of the theory was included.

It is hoped that in some measure this paper may serve to make this most useful phenomena of regeneration a little clearer and even more useful.

East Pittsburgh, Pennsylvania,
January 1, 1925.

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The authors are indebted to the following engineers of the Westinghouse Electric and Manufacturing Company for constructive criticism: Mr. L. W. Chubb, Mr. M. C. Batsel, Mr. C. R. Hanna, and Mr. Frank Falknor.

SUMMARY: This paper shows some of the defects of present theories regarding regeneration and presents a new method of analysis based on the idea of a power balance.

It is shown that a signal voltage does not supply power to a regenerated circuit but merely prevents certain losses from occurring. This upsets the balance between power input from the tickler and power lost in the circuit, so that oscillation occurs. In other words, regeneration consists of self-oscillation started and controlled by the signal voltage.

The amplification obtainable in this way has a definite limit, the limit

being caused by variations in the plate and grid impedances of the vacuum tube, as the amplitude of the grid voltage increases. The rate of variation of these impedances as the grid voltages increases, depends on the tube and on the direct voltage used.

The use of a grid leak and condenser decreases the voltage amplification, by increasing the rate of change of the plate filament impedance. In general, however, increased detecting efficiency more than makes up for the difference when audio frequency output is considered. The effect of resistance in the grid circuit is to decrease the amplification by increasing the effect of the impedance variations.

It is found that the best turn ratio to use in a regenerated transformer is the same ratio that should be used in a non-regenerated transformer. The amplification obtainable increases rapidly as the strength of an applied signal is decreased. Altho the inductance-capacity ratio does not affect the amplification obtained on an alternating current wave train, this ratio does affect the amplitude of the audio output when a modulated signal is being amplified. If a low L/C ratio is used, high notes will be lost when a weak signal is being received with full regeneration.

Regenerative amplification also occurs when a tube is in a condition of self-oscillation providing the strength of the local oscillation is weak. Under these conditions an incoming signal will be amplified whether it is beating with the local oscillation or of exactly the same frequency, as in zero beat reception. The dead spot noticed in the center of a beat note is caused by a retuning of the local circuit by the signal voltage. This voltage being out of phase with the grid voltage causes the phase of the current to shift, thus acting like a reactance.

A regenerated circuit amplifies non-resonant frequencies to a certain extent, the amount depending on the value of the reactance that would be needed to tune the circuit to the non-resonant frequency.

Altho the whole development is concerned with a regenerated transformer, the entire paper applies equally well to a regenerated antenna circuit.

DESIGNS AND EFFICIENCIES OF LARGE AIR CORE INDUCTANCES*

By

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In the operation of a radio transmitting station, two factors are of prime importance: reliability of operation and efficiency of operation. The antenna loading inductors, or tuning coils, may limit either of these two. In the early development of antenna tuning coils for use with 200-kilowatt Alexanderson alternators, reliability of operation was of first importance. With the problem of reliability solved, the question of efficiency became paramount. This paper describes, briefly, early designs which have proven to be reliable, and, in greater detail, recent designs which are reliable and more efficient.

The requirements for large air core inductances may be divided in two general classes: indoor coils and outdoor coils. In the majority of cases, outdoor coils are entirely satisfactory. In special cases in which the outdoor coils would be subjected to salt spray from the ocean, it is advantageous to protect the coils from the salt deposit. This requirement has lead to the development of an indoor coil. The designs of so-called indoor and outdoor types are essentially the same—the outdoor coil having longer creepage distances between turns and greater strength to withstand mechanical loads due to wind, and so on.

During the early development of tuning coils, wood which was used for the framework, even when protected from the weather, caused many fires and a necessary reduction in maximum antenna voltage and power. The next step forward was the substitution of porcelain for the wood, resulting in a design (Figure 1), which was superior to the wood in reliability, but which had metal supporting rings, thereby introducing additional eddy current losses. With the arrangement of the conductor on this coil, which formed a two-layer winding, the inductance and kilovolt-amperage capacity were relatively small for the physical

*Received by the Editor, April 29, 1925.

dimensions of the coil. To obtain greater inductance in the same length of coil, the multi-layer block (Figure 2) was developed. The coils using this type of block (Figures 3 and 4) not only have more turns per unit length of coil, but the spacing between turns is greater and the creepage distance many times greater.

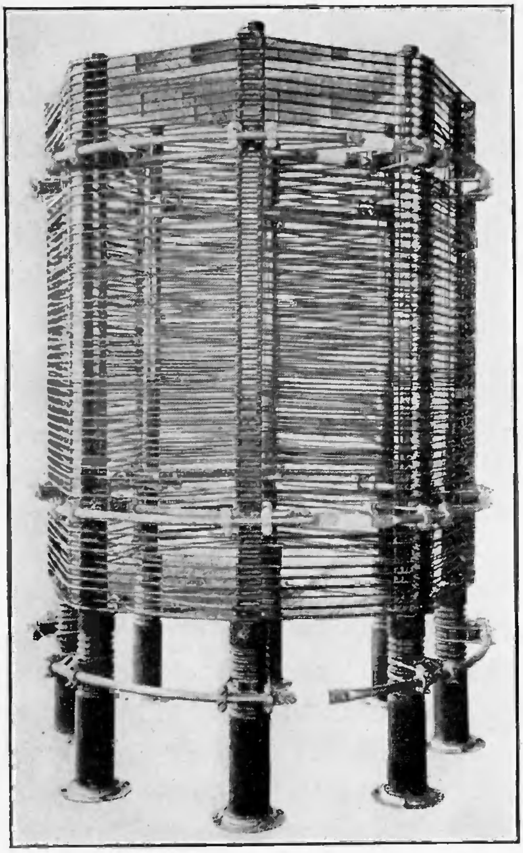

FIGURE 1—Antenna Tuning Coil

On a frame of given dimensions, it is possible to obtain with the same factor of safety approximately three times the kilovolt-amperage in the design shown in Figures 3 and 4 as in the design shown in Figure 1. By factor of safety is meant the ratio of the kilovolt-amperage at which creepage begins, to the operating kilovolt-amperage. Coils using this type of block may be wound in a variety of ways. The conductor may be wound in every groove, or as many grooves may be left idle between turns as is required for the voltage between turns. The same coil form may be wound as an inductor for low currents and high voltage or for low voltage and high currents. Small variations in inductance may be made by changing the spacing between turns. Two windings may be alternately interplaced in the grooves of the blocks to form a transformer. The great variety of possible arrangements of conductors makes it possible to build a transformer to meet practically any requirement.

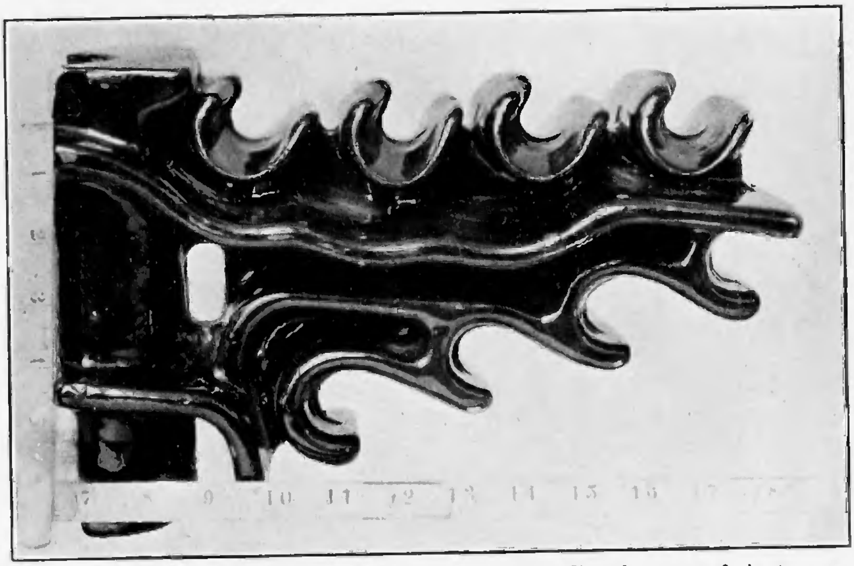

FIGURE 2—Porcelain Spacing Block for the Conductor of Antenna Tuning Coil

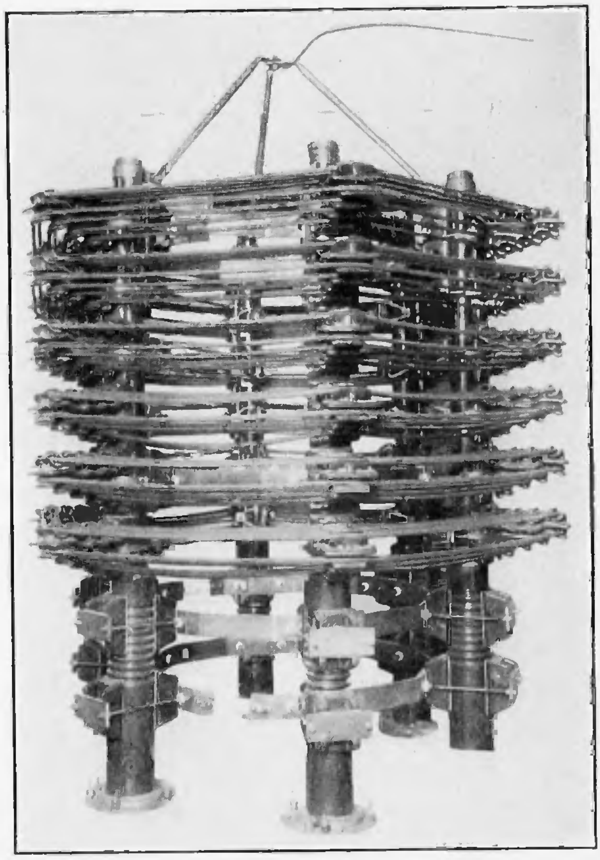

FIGURE 3—Tuning Coil for United Fruit Company, Radio Equipment

The construction of an outdoor type of coil, which has resulted from this development, is a porcelain braced coil, as shown in Figures 5, 6, and 7. This design eliminates almost all of the metal in the coil structure, and increases the rigidity of this beyond all former coils. The coil is made almost entirely of porcelain with the internal braces arranged to take advantage of the stiffness of a triangular form. During the development of this type of coil, a sample test coil was tested at the Tuckerton Station of the Radio Corporation of America. The umbrella antenna was being operated with one antenna coil in series with the alternator and in shunt with this circuit were three coils in multiple. Tests were made to determine the comparative efficiencies of the original coils, as shown in Figure 8, and the porcelain braced coil. In order to obtain data on the test coil under winter weather conditions, the conductor was wound in every other groove of the spacing blocks and the inductance

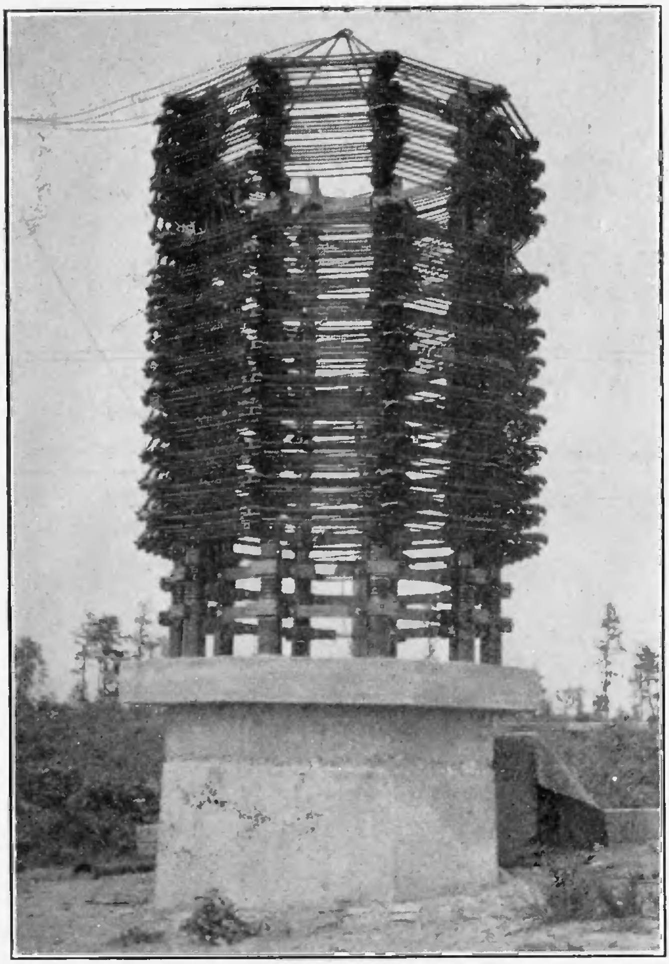

FIGURE 4

adjusted to the equivalent of the three shunt coils. The antenna resistance with this arrangement was lowered approximately sixteen percent from the value with the three-shunt coils. This coil was left in service three months, carrying approximately 18,000 kilovolt-amperes at 300 amperes during sleet, snow, rain, and fair weather conditions. As a result of these tests, the porcelain braced coil was shown to have higher efficiency at approximately four times the maximum kilovolt-amperage at which the original coils could be operated, and to be mechanically stronger than any previous types. In one case, the porcelain-braced coil was connected to replace one of the three multiple coils of the original type shown in Figure 1, and the reduction in antenna resistance was approximately 10 percent. As an extreme condition of loading, the test coil with conductor in every groove on

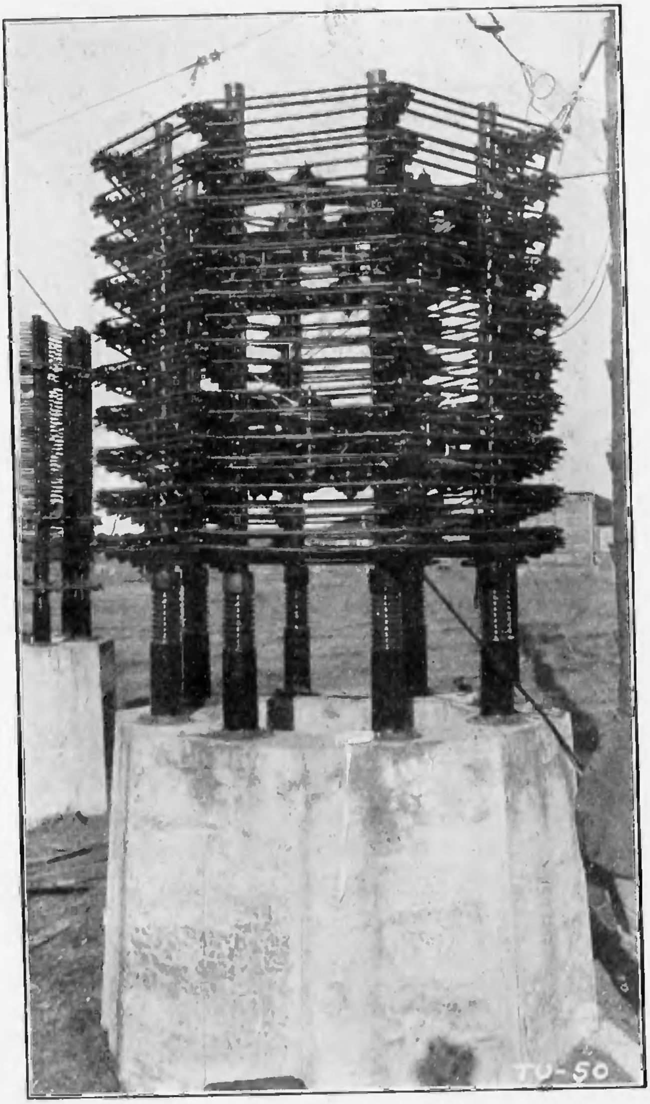

FIGURE 5

only four blocks, was loaded to more than 30,000 kilovolt-amperes at 400 amperes. This corresponded to an antenna input of approximately 200 kilowatts.

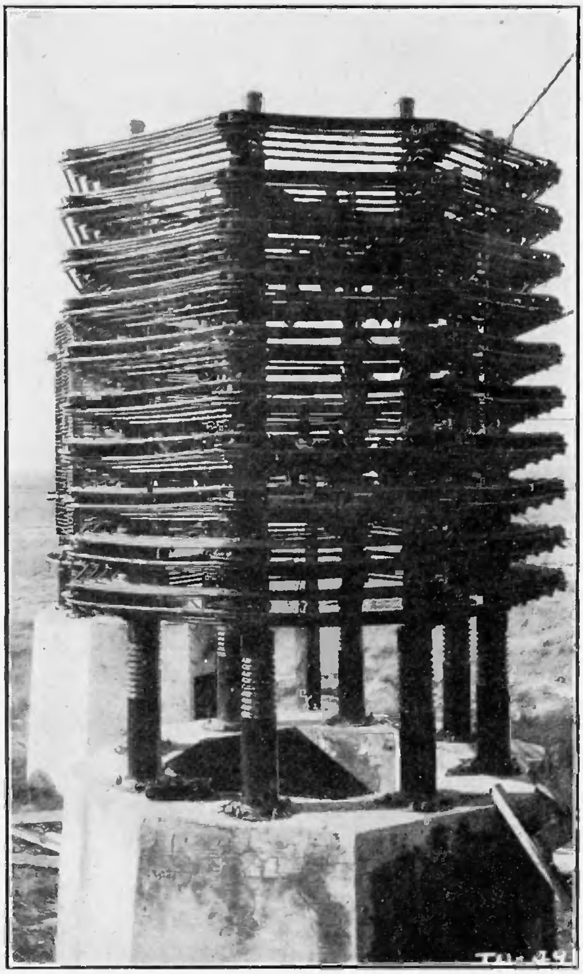

Figure 6

It is desirable to design a coil for indoor service more compactly than an outdoor coil for the following reasons:

1. Economy of space.
2. A concentrated field means less loss in surrounding objects.

The fact that shorter creepage distance for a given voltage provides the same factor of safety indoors as is obtained with longer creepage distances outdoors, makes it possible to utilize the compact design. With the smaller physical dimensions, the field, due to the coil, is concentrated, and the losses in any metal within the field would be greater than in a coil with a weaker field. For this reason, it is essential to eliminate all metal from the field except the conductor. Not only should all metal be removed, but the frame must be of good quality dielectric and be

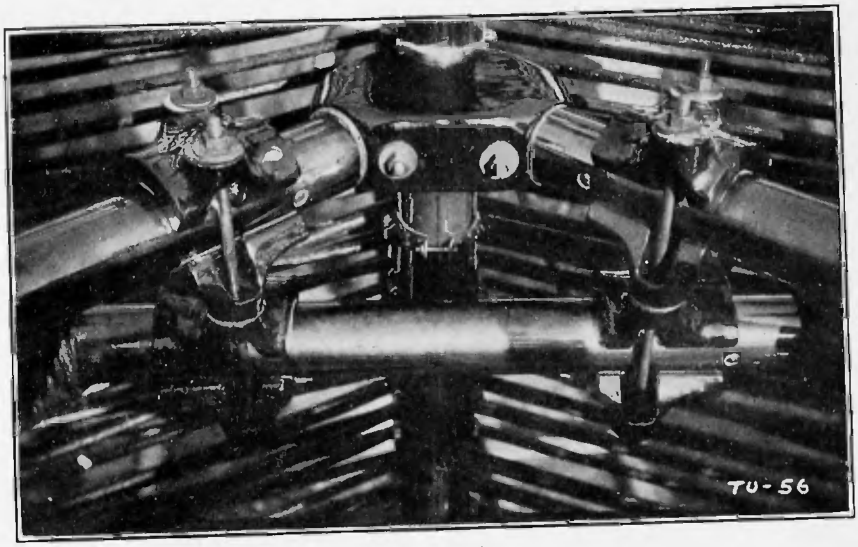

FIGURE 7

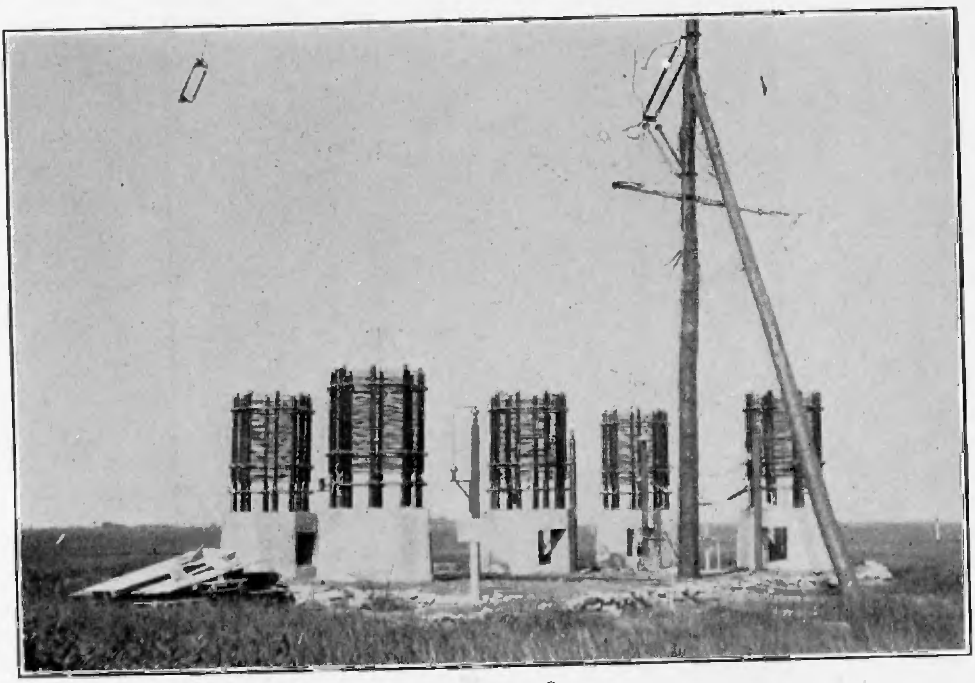

FIGURE 8

non-inflammable. The porcelain framework for such a coil is shown in Figure 9. In building up a coil of this type, three or four of the sections are stacked up, using mica sectors as cushions between the sections. This portion of the coil serves as a base and is formed of one to six sections, depending on the voltage, nearness to large metal objects, and the like. The active sections (Figure 10), may be wound before erection or may be set up as a continuation of the base and the spacing blocks and conductor applied afterward. The blocks (Figure 11) used on this coil are similar to those used on the outdoor coils, but have been modified to provide a greater number of grooves in the same height. The spacing blocks are separated from the porcelain framework by mica cushions and are held in place by the conductor. This construction is very rugged: A man may stand on the end of a block without causing the conductor to sag or the block to slip. These sections are stacked in a vertical column until the required value of inductance is obtained. Each section is nine inches high, 33 inches inside diameter, and 75 inches outside diameter and has seventeen turns if all groves in the spacing blocks are filled. This coil meets all of the indoor requirements and, in addition, has an extremely flexible design. Only two porcelain parts are required.

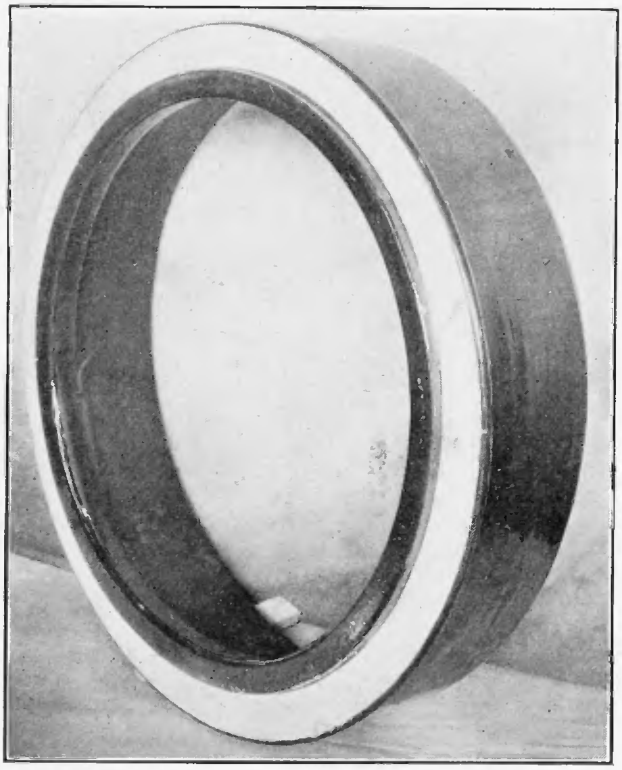

FIGURE 9—Porcelain Cylinder for Unit Type Tuning Coil 36 inches in Diameter, Bottom View

The efficiency of an antenna tuning coil is usually stated by giving the "power factor" of the coil. The power factor of an efficient coil should be below 0.002. Many of the coils in service are not within the range; however, there are some of the later designs which have power factors considerably below 0.001. Chief among the losses which go to make up the power factor of a coil are: losses in the coil frame, conductor eddy current loss, and conductor ohmic loss. In the case of the indoor coil just described, the losses in the frame have been made extremely small so that the losses in the conductor are of the greater importance. The power factor due to the ohmic drop in the conductor is easily calculated from the direct current resistance and the reactive drop; that is

$$\text{Conductor Ohmic Power Factor} = \frac{R}{X}.$$

Mr. E. F. W. Alexanderson has developed a formula for the eddy current power factor. This formula was developed with the

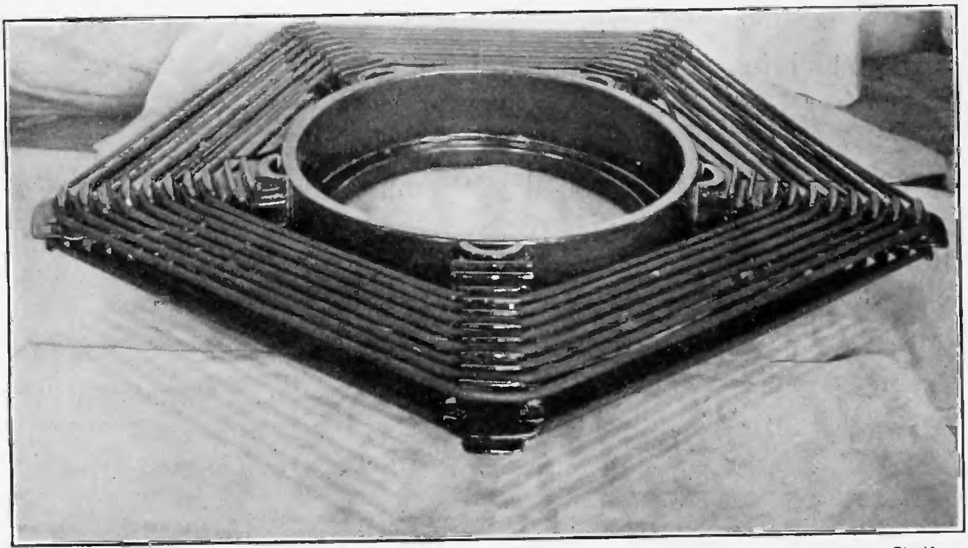

FIGURE 10—One Section of Unit Type High Power Antenna Tuning Coil

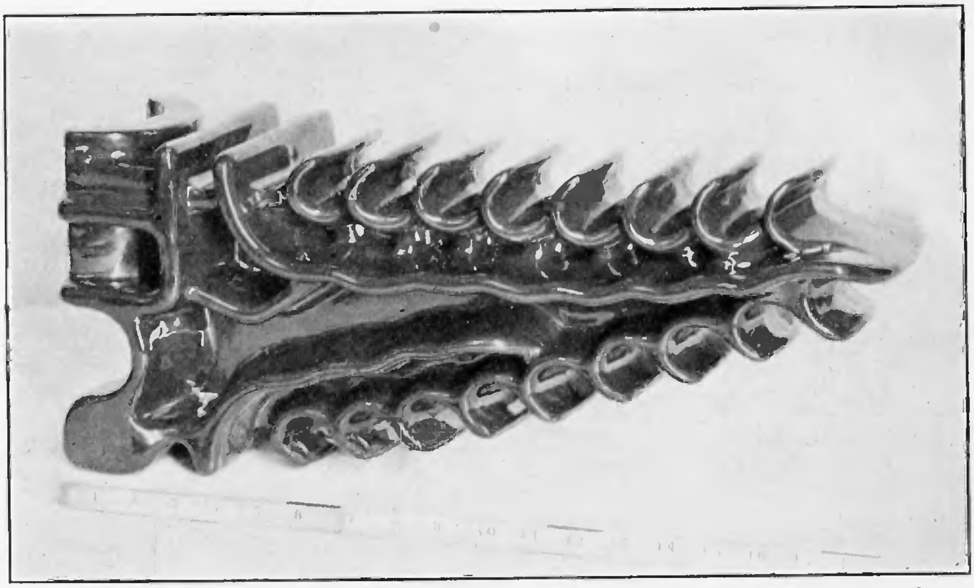

FIGURE 11—Porcelain Spacing Block with Seventeen Grooves, for Unit Type High Power Antenna Tuning Coil. Side View

assumption that the conductor was in a uniform magnetic field. This field is not absolutely obtained but is very closely approximated in the larger coils. This formula

$$\text{Conductor Eddy Current Power Factor} = \frac{X}{R}\left(\frac{d}{D}\right)^2$$

does not take into account the form of the conductor; that is, whether there is a non-conducting core or the size of such core. In the preceding formulas:

X = Reactance
R = Direct current resistance
d = Diameter of individual strands
D = Coil diameter

Tests which have been made indicate that this formula is correct to within a few percent for all conductors commonly used in high power coils. The calculated ohmic and eddy current power factors of the unit type coil with four different conductors is shown in Figure 12. The eddy current power factor increases directly with the frequency, while the ohmic decreases inversely with the frequency. The small circles at the intersection of the ohmic and eddy current power factor for each conductor indicate the optimum frequency for the coil with that particular conductor. Figure 13 shows the sum of the ohmic and eddy current power factors on the conductor number 2 with various diameters

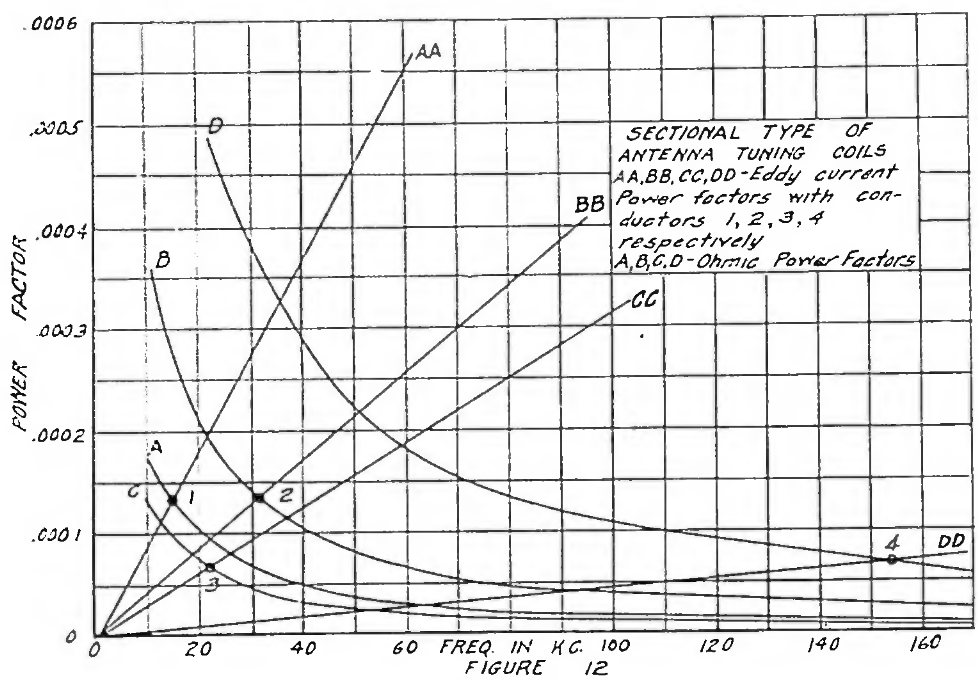

FIGURE 12

of coil. The increased efficiency of the 75-inch diameter over the 60-inch is very marked, but the increase from 75 inches to 120 inches is not so great. For this reason, 75 inches seems to be a good diameter for an indoor coil, using this conductor.

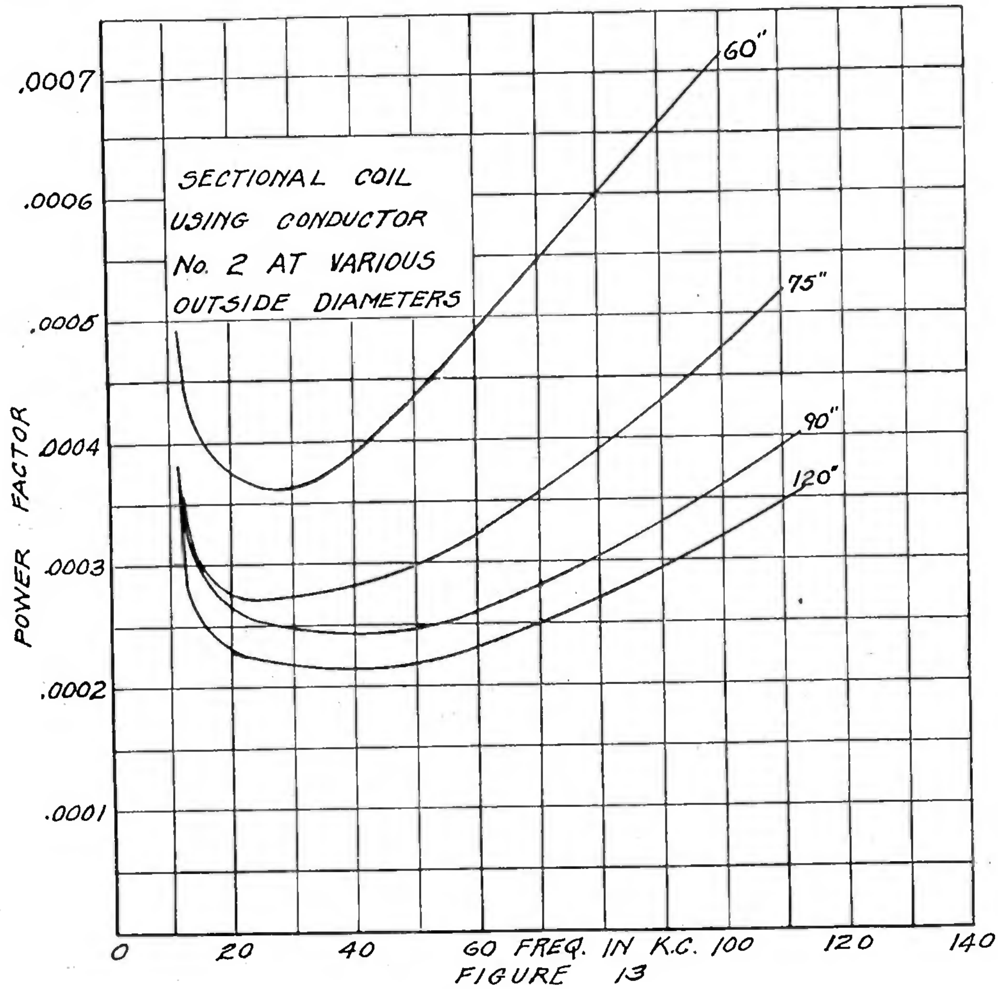

FIGURE 13

For the large air core inductors which have been described, the conductors are all of the Litzendraht or finely stranded copper conductor, each strand enameled. A large number of tests have been made on the conductors for this use. These tests show that the smaller the diameter of the strands and the larger the non-conducting core, the lower the losses are in the conductor.

March 31, 1925,
Schenectady Works, New York.

SUMMARY: Representative designs of large air core antenna tuning inductances are described, in types suitable for outdoor and indoor service. The latest improved designs are described in greater detail and compared with earlier designs on a basis of efficiency and kilovolt-ampere capacity. Formulas

are given for calculating the ohmic and eddy current conductor power factor of coils using finely stranded, separately-insulated strands.

In graphical form are shown the variations of ohmic and eddy current power factor with frequency, with four different conductors wound in a given arrangement to given dimensions; also the variation of the sum of ohmic and eddy current power factors with frequencies for a representative conductor on various diameters. These values were calculated by the formulas given, and indicate very high efficiencies for the latest types of coils.

AN EFFICIENT TUNED RADIO-FREQUENCY TRANSFORMER*

By

F. H. Drake and G. H. Browning

(Cruft Laboratory, Harvard University, Cambridge, Massachusetts

At a time when radio engineers are generally concerned with the problem of developing more sensitive methods of reception, it seems particularly desirable that exact data on the design of an efficient tuned radio-frequency transformer be made generally available. While the present paper deals specifically with the problem of radio-frequency amplification over the broadcast range, some general formulas are furnished for the design of a tuned radio transformer for any wave length band.

The figure of merit of a radio-frequency amplifier is usually considered to be its voltage amplification factor, but the ability of the amplifier to discriminate between the desired signal and signals at other frequencies is a factor of considerable importance in determining its utility. In preparing this paper, it has been our aim, first to design a transformer so that the theoretical amplification could be obtained, then to consider its selectivity, and the tendency of the circuits to break into oscillations. Values of voltage amplification predicted by theory have been checked by experiment, and from this agreement, as well as from audibility tests, the realization of the theoretical resonance curve has been concluded.

The radio-frequency transformer is represented in Figure 1 by coils L_1 and L_2, having a mutual inductance M. The primary of the transformer is untuned; the secondary is tuned by a condenser C_2. Since the primary coil L_1 is in series with the plate resistance of a vacuum tube, its resistance may be neglected in computing the current in the secondary. This fact, as will subsequently be shown, is of importance in designing the primary so that capacity coupling with the secondary will be negligible.

In this paper the over-all voltage amplification of tube and

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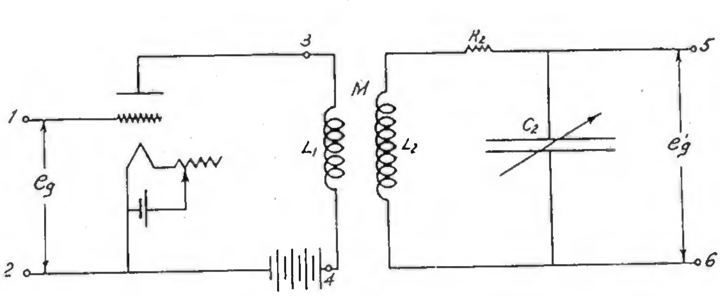


FIGURE 1

transformer is called K, and is equal to e'_g/e_g. (See Figure 1.) In comparing the values given for K with the voltage amplification obtained from a one-stage audio-amplifier, it should be remembered that the response of a vacuum tube detector is nearly proportional, for small values of e_g', to the square of the applied variation voltage so that a value $K=10$ is equivalent to an audio-amplification of 100. (A good one-stage audio-amplifier will give a voltage amplification of 30.)

It may be shown that for alternating current, Figure 1 is electrically equivalent to Figure 2 where the amplifier tube is replaced

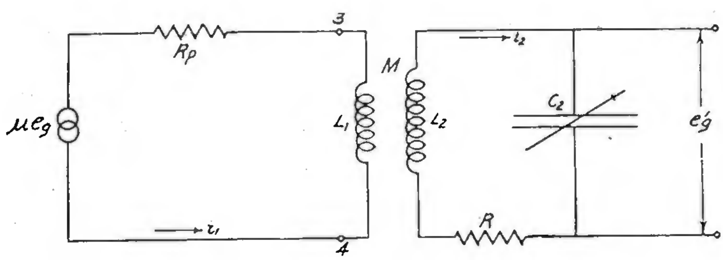


FIGURE 2

by an emf. of $\mu\, e_g$ in series with the plate resistance R_p, and the problem of determining the value of K is reduced to the problem of solving Figure 2 for i_2, since to an extremely close approximation

$$e'_g = L_2\, \omega\, i_2 \tag{1}$$

The general coupled circuit expression for secondary current is[1]

$$i_2 = \frac{M\, \omega\, \mu\, e_g}{Z_2\, Z_1'} \tag{2}$$

where
$$Z_2^2 = R_2^2 + X_2^2$$

$$(Z_1')^2 = \left(R_p + \frac{M^2\, \omega^2}{Z_2^2} R_2\right)^2 + \left(X_1 - \frac{M^2\, \omega^2}{Z_2^2} X_2\right)^2$$

The secondary current may be expressed in a form better

[1] Pierce, "Electric Oscillations and Electric Waves," Chapter XI.

adapted to computation and design by employing certain standard abbreviarions, as follows:

$$\tau = \frac{M}{\sqrt{L_1 L_2}};\ \eta_1 = \frac{R_p}{L_1 \omega};\ \eta_2 = \frac{R_2}{L_2 \omega};\ \frac{\omega}{2\pi} = f\begin{pmatrix}\text{impressed}\\ \text{frequency}\end{pmatrix}$$

$$X_1 = L_1 \omega;\ X_2 = L_2\omega - \frac{1}{C_2 \omega};\ \omega_o = \frac{1}{\sqrt{L_2 C_2}};\ \frac{\omega_o}{\omega} = \theta$$

μ = amplification factor of the tube.

With this notation and equations (1) and (2), the over-all voltage amplification factor becomes

$$K = \frac{e_a'}{e_a} = \frac{\mu \tau \sqrt{\frac{L_2}{L_1}}}{\sqrt{\eta_2{}^2 + (1-\theta^2)^2}\ \sqrt{\left\{\eta_1 + \frac{\tau^2 \eta_2}{\eta_2{}^2 + (1-\theta^2)^2}\right\}^2 + \left\{1 - \frac{\tau^2(1-\theta^2)}{\eta_2{}^2 + (1-\theta^2)^2}\right\}^2}} \tag{3}$$

In equation (3), τ is the coefficient of coupling between coils L_1 and L_2 and η_1 and η_2 are the power ratios of primary and secondary circuits. It should be noted that η_2 is the usual sort of power ratio for a coil and will be nearly constant over a considerable range away from the natural period of the coil, while η_1 contains a resistance independent of the frequency and cannot be considered constant over any range.

In tuning the transformer for maximum signal strength, condenser C_2 is varied until i_2 is a maximum, assuming, of course, that L_1 and L_2 are fixed. This partial secondary resonance relation is expressed mathematically by the relation $\frac{X_2}{X_1} = \frac{M^2 \omega^2}{Z_1{}^2}$, which,[2] using the notation above and equation (2) gives

$$K_{max} = \frac{\mu \tau \sqrt{\eta_1{}^2 + 1}\sqrt{\frac{L_2}{L_1}}}{\eta_2(\eta_1{}^2 + 1) + \tau^2 \eta_1} = (\text{approx.}) \frac{\mu \tau \sqrt{\frac{L_2}{L_1}}}{\tau^2 + \eta_1 \eta_2} \text{ if } 1 \ll \eta_1{}^2 \tag{4}$$

So far no special values of the transformer constants have been assumed except that L_1 and L_2 are fixed inductances. From equation (4) it is evident that since η_2 is approximately constant, there will exist a relation between τ and η_1 which will make K a maximum for a particular frequency. This relation may be shown to be

$$\tau^2 = \frac{\eta_2}{\eta_1}(\eta_1{}^2 + 1) = \text{approx. } \eta_1 \eta_2 \text{ as } / \ll \eta_1{}^2 \tag{5}$$

[2] Previous citation.

Equation (5) can only be satisfied for one frequency. For this one frequency the optimum K will be,

$$K_{opt} = \frac{\mu\sqrt{L_2/L_1}}{2\sqrt{\eta_1\,\eta_2}} = \frac{\mu\,L_2\,\omega}{2\sqrt{R_p\,R_2}} \tag{6}$$

In designing the transformer, it is, in the light of the (6), clearly desirable to make L_2 as large as possible consistent with ability to tune to the minimum wave length desired. The actual value of L_2 is thus a function of the distributed capacity of L_2, of the minimum capacity of C_2, and of the input capacity of the next tube. The value of frequency for which $\tau^2 = \frac{\eta_2}{\eta_1}(\eta_1{}^2+1)$ may conveniently be chosen in the middle of the wave length band for which the transformer is designed.

For plotting curves of the various equations above, the following constants have been chosen. These represent about the best design for a tuned radio-frequency transformer to include wave lengths between 300 and 600 meters with UV-199 tube. Other tubes having equally small input capacities will serve as well.

$L_1 = 0.13$ mh.	$R_p = 20{,}000$	$\mu = 6$
$L_2 = 0.400$ mh.	$\eta_2 = 0.008$	$\tau = 0.52$

With these constants $\tau^2 = \frac{\eta_2}{r_1}(\eta_1{}^2+1)$ at 400 meters.

Figure 3 shows a curve of K against τ for different values of wave length. These curves are plotted from equation (4) with n_2 assumed constant, and show that so long as τ is greater than 0.4, good amplification will be obtained over the broadcast range. The maxima of these curves are the points where $\tau^2 = \frac{\eta_2}{\eta_1}(\eta_2{}^2+1)$ and the dotted line is, therefore, the locus of that equation. Figure 4 is the value of η_2 computed from laboratory measurements of R_2. This R_2 was measured with a lighted UV-200 tube, connected across the transformer secondary. L_2 consisted of 70 turns of number 20 double cotton-covered wire wound on a tube four inches in diameter, while the secondary condenser was one of special construction designed for low losses. The values of η_2 given in Figure 4 are believed to be about the best that can be obtained with a reasonable size of L_2.

In order that the theoretical amplification could be obtained, it was early realized that capacity coupling between primary and secondary must be reduced to a value negligible compared with the inductive coupling. At the same

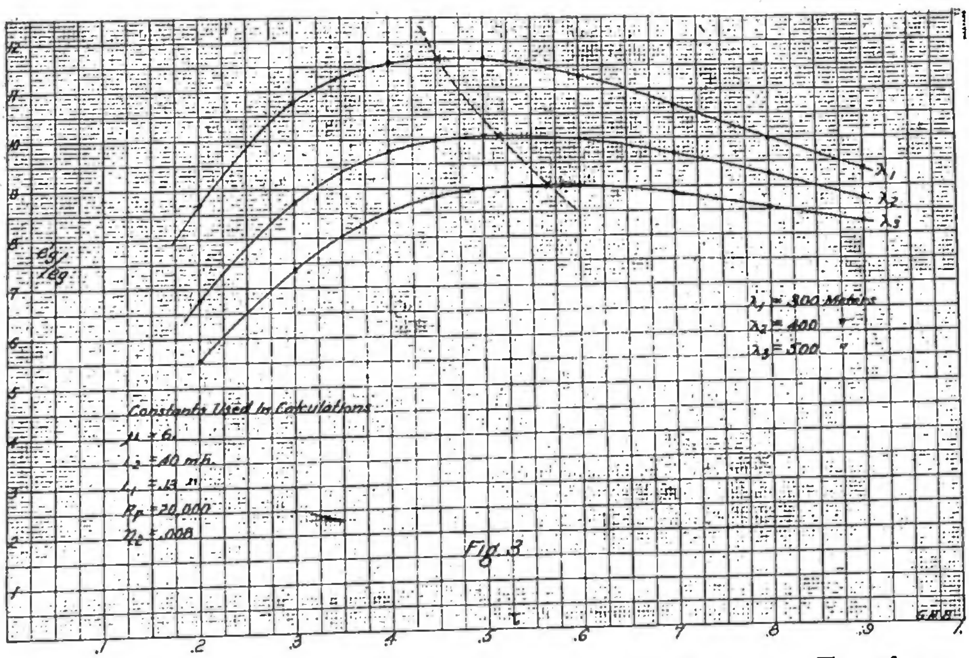


FIGURE 3—Amplification of a UV-199 Tube and Radio Frequency Transformer

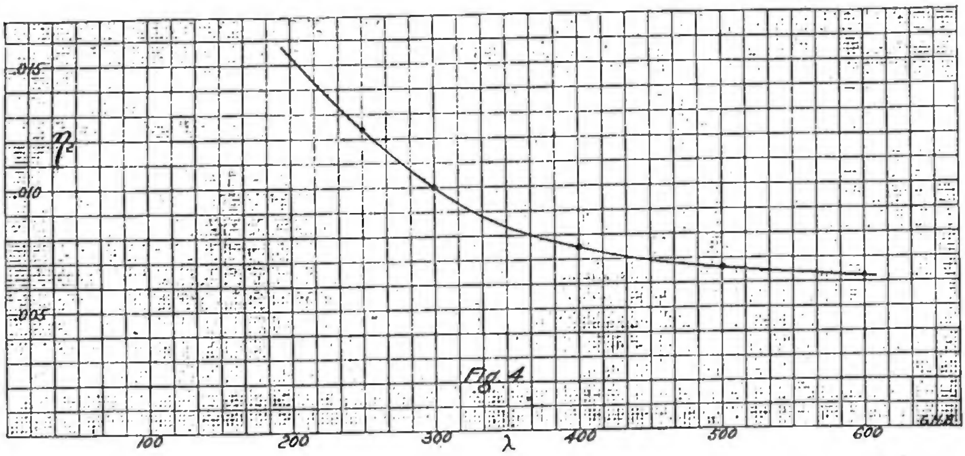


FIGURE 4—Curve of r_2 Against λ. Coil Coil Consisted of 70 Turns of Number 20 D.C.C. Wire on 4-Inch Tube. (Tube and Socket Included)

time, for reasons given later, as large a value of coefficient of coupling as possible is desirable. Capacity coupling may be assumed to be present when, in practice, a reversal of the connections to the primary of the transformer changes the value of secondary current. In general, capacity coupling will be very large if L_1 and L_2 are coaxial, single layer coils of slightly different radii. Figure 5 shows the design finally adopted for the transformer for use in the 300-600 meter range. The secondary is wound with number 20 double cotton-covered wire on a four-inch tube; the primary is wound in a small channel in a thin wooden cylinder tightly fitted inside the secondary value. Since the reactance of the primary [illegible] of the wire [illegible] should be [illegible] capacity in the primary itself. [illegible] such a transformer has very small capacity [illegible] that the primary may be placed at the center of the [illegible] the inductive coupling coefficient is a maximum. Although the calculations and measurements were made with a [illegible] the value [illegible] given in Figure [illegible] is suitable if signals from 200 to 600 meters are to be received efficiently. However, with a condenser [illegible] mf. and small minimum capacity [illegible] on the secondary will give sufficient inductance [illegible] wave lengths from 200 to 600 meters.

[illegible]

Having decided upon the transformer design, the next problem is to investigate its performance in an actual [illegible] circuit. The possibility of [illegible] sustained [illegible] in a tuned circuit connected to the grid of a three-electrode tube [illegible] adding positive resistance in the plate circuit has been [illegible] treated elsewhere.* A very rough idea of the amount of plate resistance necessary to produce such [illegible] may be formed from an examination of Figure 7 which was [illegible] by the method of Figure 5 for values of [illegible] reference. The important fact about Figure 7 is the order of magnitude of the quantities involved.

Referring to Figure [illegible] it is readily seen that [illegible] circuit is connected between points 1 and 2 [illegible] resistance may be produced due to the reaction of the plate [illegible]

*[illegible]

circuit, between points 3 and 4, of an "equivalent reactance," made up of the reactance of coil L_1 and the reactance produced by the secondary current thru M on L_1. The equivalent reactance in the plate circuit of the amplifier tube may be defined as the reactance, which, when connected between points 3 and 4 (Figure 1), would give the same value of alternating current between these points as the arrangement shown. The value of this equivalent reactance is given below equation (2), and by virtue of earlier notation is

$$X_1' = L_1 \omega \left\{ 1 - \frac{\tau^2 (1-\theta^2)}{\eta_2^2 + (1-\theta^2)^2} \right\} \tag{7}$$

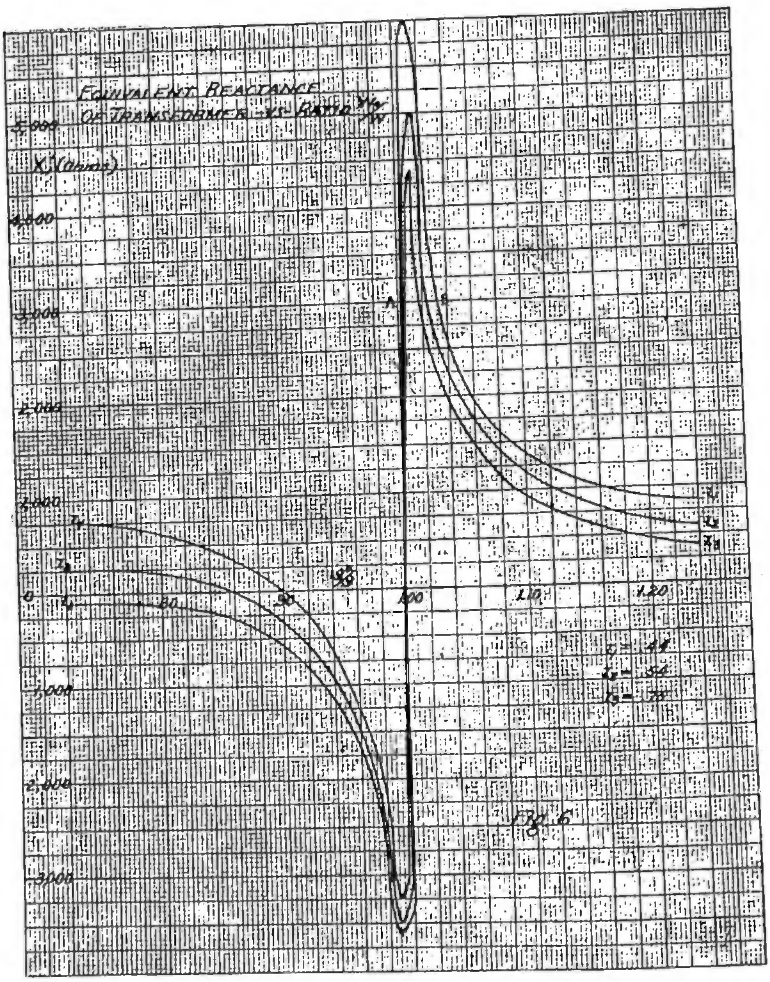


FIGURE 6

Figure 6 is plot of X_1' against θ for three values of the coefficient of coupling, for which $\tau^2 = \eta_1 \eta_2$. For large values of τ the value of L_1 is, therefore, small, and conversely. For the conditions of Figure 7 oscillations would only occur in a narrow region

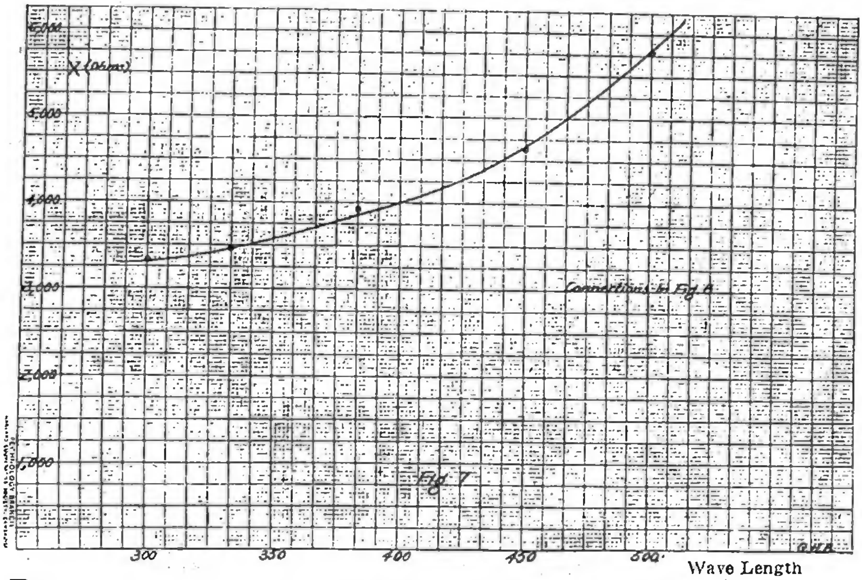


Figure 7—Reactance in Plate Circuit Necessary to Make a UV-199 Tube Oscillate

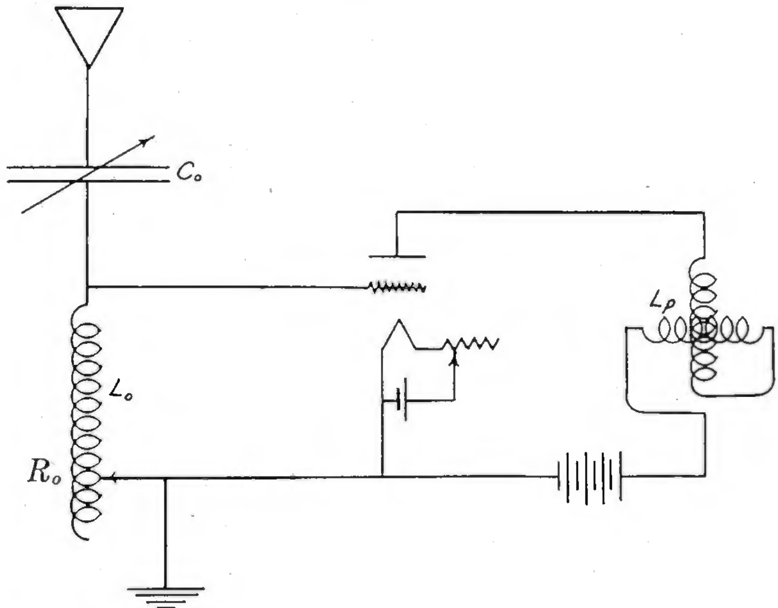


Figure 8

such as AB, and it is evident that AB will be narrower the greater the coefficient of coupling. The reason for making τ as large as possible, consistent with negligible capacity coupling, is now evident, for by doing so the region over which the amplifier will tend to oscillate is reduced to a very narrow band adjacent[4] to the setting for maximum voltage amplification, and by changing the value of $L_o/C_o R_o$ (see Figure 10) by any of the conventional

[4] The exact value of θ for maximum amplification may be shown mathematically to be very slightly less than unity. At this point the reactance is positive but less than half its maximum positive value.

methods, the narrow band may be reduced to a point, thus providing a beautiful method of determining when the antenna is in tune with the radio amplifier. Once the operator has calibrated the amplifier condenser for wave length he has only to tune the antenna circuit until the narrow band is found over which the tube oscillates, and then to reduce this band to a point, after which the whole apparatus will be almost exactly in tune with the desired wave length. This process takes longer to describe than to perform.

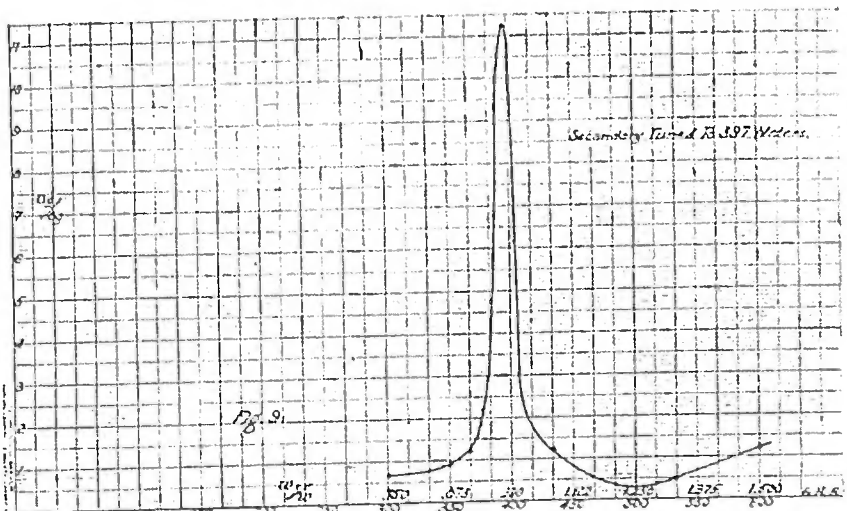


FIGURE 9—Resonance Curve of Transformer and UV-199 Tube. (Amplification Against Wave Length)

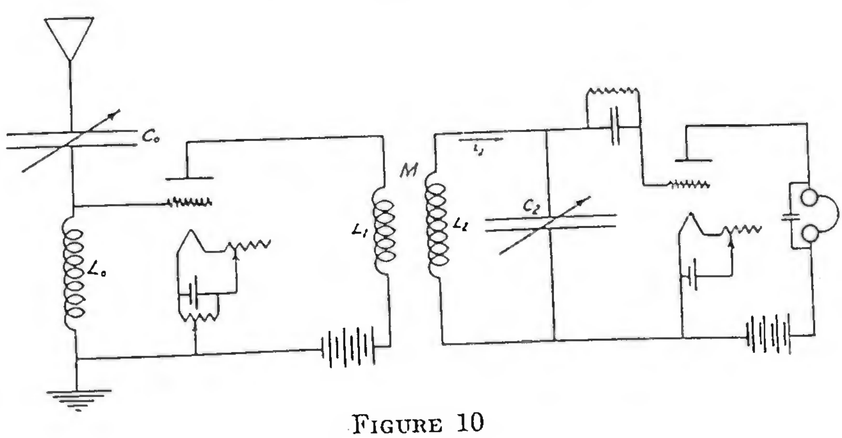


FIGURE 10

Figure 10 shows the transformer used in a single circuit one-stage radio-amplifier receiver with non-regenerative detector. A potentiometer serves to change the value of L_o/C_oR_o until the correct amount of regeneration is obtained. For the transformer above described used in conjunction with UV-199 tube, oscillations can be stopped before the grid becomes positive with respect to the filament, so that no amplification will be lost by having the grid positive. This is possible because τ is chosen large and L_1 small, thereby satisfying the condition for maximum amplification and reducing to a minimum the tendency to oscillate.

The selectivity of a one-stage radio-amplifier using a UV-199 tube and the transformer here described is *relatively* very great. Equation (2) is the analytical expression for a resonance curve of K against wave length. Figure 9 is a plot of such a resonance curve for $\theta = 1$ at 400 meters, this point also being that at which $\tau^2 = \eta_1 \eta_2$. Resonance curves at other wave lengths will be of the same general character, but will have slightly lower maxima. The selectivity is said to be "relatively" great because when the transformer is tuned to the desired signal, other signals at nearby wave lengths are amplified to a much smaller degree, yet it is to be noted that there is practically a one-to-one voltage transfer at wave lengths thruout the broadcast band. The practical importance of this is that a strong local station will be heard when the transformer is tuned to another wave just as loudly as if the radio-amplifier tube and transformer were removed and the incoming voltage applied directly to the detector, but as soon as the weak station to which the transformer is tuned is picked up, its signals will experience an absolute amplification of about ten, and so a relative amplification of ten-to-one is obtained. If absolute selectivity is desired, a coupled circuit should be used before the radio-amplifier.

Experimental Verification of Amplification

The method used to measure the voltage amplification is believed to be somewhat novel, in that it depends essentially on extrapolation from curves for different conditions than those under which the transformer is used. Figure 11 shows the connections. The voltage applied to the grid-filament of the amplifier tube is known[5] from values of I and R. The current produced by this voltage in the secondary circuit is measured by a thermo-couple and galvanometer, T and M. Since the thermo-couple happened to have a resistance (1.5 ohms) nearly equal to that introduced in circuit $L_2 C_2$ by connecting the detector tubes to points 7 and 8, and since the total secondary resistance is about 20 ohms, no correction need be applied for

[5] Great precautions against stray fields must be taken. In these measurements twisted leads carried the radio-frequency current to the resistance from an oscillator located in another room.

the resistance of the couple, and the voltage amplification is given by

$$K = \frac{L_2 \omega i_2}{R I}$$

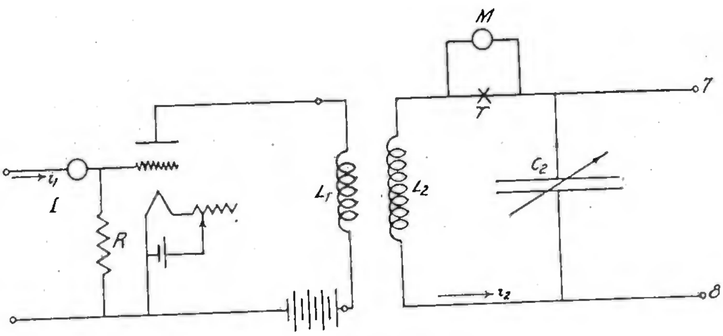


FIGURE 11

The most sensitive low resistance thermo-couple available would not give accurately measurable values of i_2 until the voltage applied to the grid of the first tube was about four volts, a value enormously greater than that produced by a strong nearby station in a tuned antenna and, therefore, available as the input voltage to the amplifier tube. Also, even with four volts on the grid, the value of i_2 could not be read with absolute accuracy, so for these reasons it was not thought desirable to rely wholly on a single value at such a large grid voltage, and the idea was conceived of obtaining several values of i_2 at grid voltages ranging from 4 to 15 volts and extrapolating to find the real i_2 for a value of grid voltage for which the excursion on the tube characteristic would not exceed the linear region. A series of such curves is shown in Figure 13. The value of plate voltage used was 50 volts. As long as μe_g is less than about 20, there is a linear relation between i_2 and e_g and a common value of four volts may conveniently be used in reading the corresponding values of i_2.

If the agreement of theoretical and experimental values for voltage amplification is taken as an indication of the validity of the experimental determination, the method outlined above is a valuable one, as Figure 12 shows. At the upper end of the broadcast band the agreement is almost perfect while divergence near the lower end is probably due to the increasing importance of small capacities in the tube and transformer at the higher frequencies and possibly due also to the fact that the value of R_p chosen to design the transformer was not that of the tube used

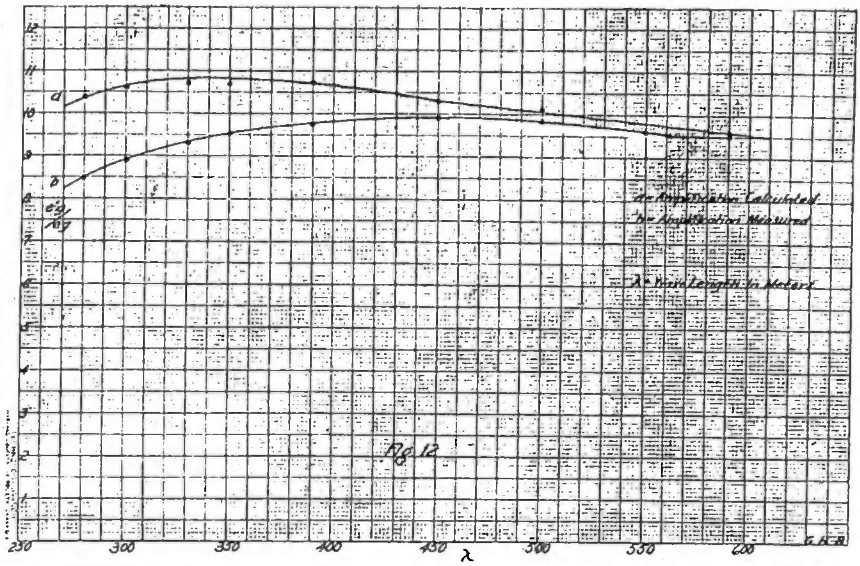


FIGURE 12—Amplification, Theoretical and Practical

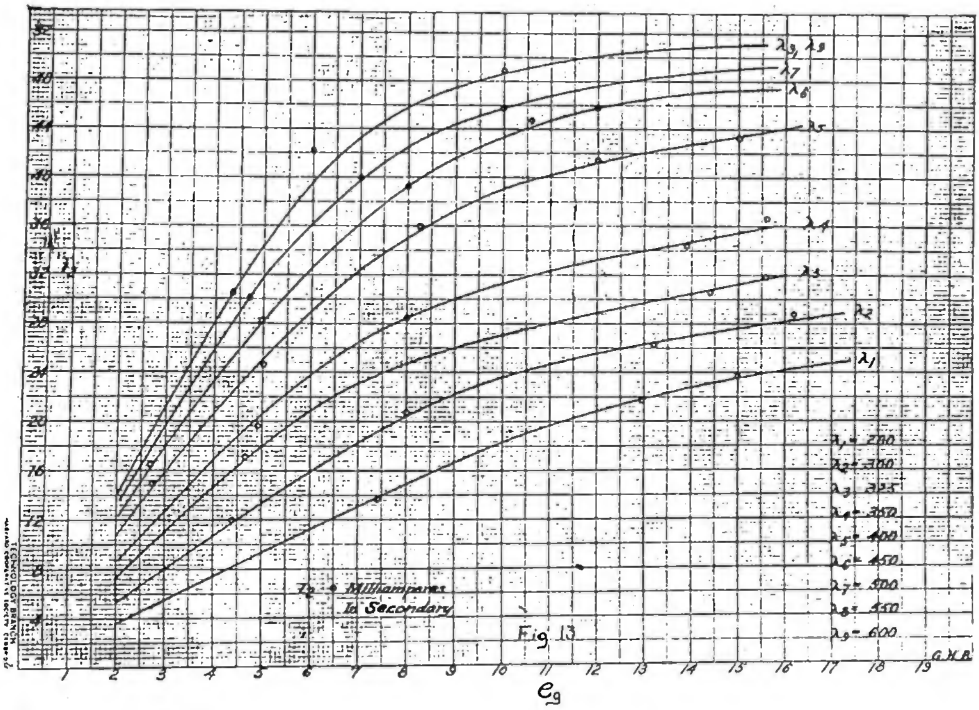


FIGURE 13—Current in the Secondary of the Radio Transformer Against Impressed Signal

in the measurements, so that the relation $\tau^2 = \eta_1 \eta_2$ was really satisfied by the transformer at a higher wave length than 400 meters. It is believed that those who have had experience with the difficulties of measurements of this sort will be lenient toward the slight discrepancies between the theoretical and experimental curves.

There are two good methods of obtaining radio amplification at short waves. One is that described in this paper and consists essentially in tuning the transformer to the signal. The other is that originated by Armstrong, the super-heterodyne, which tunes the signal to the transformer. The writers sincerely hope that the material here presented will prove of considerable value to those desirous of improving upon the sensitiveness and selectivity of the ordinary regenerative receiver by the former method. This paper should, therefore, appeal to the enthusiasts who are interested in radio-frequency amplification as well as to those who believe the utmost sensitivity is obtained by a combination of regeneration and tuned radio-frequency amplification.

Summary of Design Formulas

1. To compute L_2, let λ_o be the minimum wave length in meters to which it is desired to tune. Then L_2 in millihenrys will be given by

$$L_2 = 5 \times 10^{-6} \lambda_o^2$$

2. To satisfy the relation $\tau^2 = \eta_1 \eta_2$, assume $\tau = 0.5$ (for the construction shown in Figure 5) and substitute the value of η_2. Then knowing the plate resistance of the amplifier tube and the mean frequency to which it is desired to tune, the value of L_1 can be computed.

3. For C_2 select a low loss condenser of proper capacity to cover the wave length band desired.

DISCUSSION* ON "A NEW PHENOMENON IN SUNSET RADIO DIRECTION VARIATIONS" BY L. W. AUSTIN

By

R. L. Smith-Rose and R. H. Barfield

(Radio Research Board Station, Slough, Bucks, England)

The experimental results described by Dr. Austin in the August issue of the Proceedings are of great interest to us since we have been using tilting frame coil direction-finders for the past two years in an investigation of radio wave-fronts. As far as observations on long-wave stations at various distances are concerned, however, we have so far been unable to obtain any of the effects described by Dr. Austin, and we should like to ask one or two questions in order to obtain more definite information on this point. In a paper published earlier in the year (R. L. Smith-Rose and R. H. Barfield, "On the Determination of the Forces in Wireless Waves at the Earth's Surface"—Proc. Roy. Soc., 1925, Volume 107, pages 587-601), experiments were described in which the conductivity of the earth was measured at radio frequencies. In the same paper it was shown that this conductivity is so high in England that on long wave lengths, even were a radio wave arriving at the earth at an appreciable angle of incidence and polarized with its magnetic field in a vertical plane, the reflected wave set up at the earth's surface will be such as to tend to eliminate any vertical component of the magnetic field. The results of experiments were given supporting this theoretical prediction, and showed that the magnetic field of the arriving wave is always horizontal to within 1°, which was the limit of accuracy of the apparatus then employed. As a specific instance we may quote the observations on Leafield's transmission on a wave length of 12,400 meters at a distance of about 77 km. This station has been regularly watched now for over a year and altho all the usual effects of variations of bearings and signal intensity are observed, the resultant magnetic field has never departed from the horizontal by more than the limit above mentioned. Consequently, when the signal minimum with the coil vertical is blurred, indicating an elliptically-rotating magnetic field in the horizontal plane, no sharp signal minimum

* Received by the Editor, September 9, 1925

is found in any position except that in which the plane of the coil is swung thru the horizontal position, that is, with the vertical axis tilted at 90°.

In our own experiments somewhat elaborate precautions have been taken to eliminate any stray emf's which may be introduced into the receiving system by antenna or capacity effects or as a direct pick-up of signal on parts of the apparatus other than the frame coil. Furthermore, the site chosen for our experiments (at Slough) is situated well away from buildings, overhead or underground metalwork and so on, which might cause distortion of arriving waves. The site is known not to be ideal, but as it has been the subject of detailed investigation for the past four years, its essential defects are now fairly well understood. We should like to be assured that Dr. Austin's experiments have been carried out under similar conditions, for since the vertical component of magnetic field is at best only a residual effect, it might be possible to get comparatively large tilts by balancing the effect of this magnetic field against a stray emf. in some other part of the apparatus. On a site which is not free from metallic frame buildings or even hills and other geographical features, there is the further possibility that the arrival of down-coming waves at night will produce abnormal effects in the immediate neighborhood of such sources of disturbance. If, as we hope, the experiments are above such criticism, it is evident that the site at which they are made possesses a fairly low effective conductivity much lower than we have as yet been able to find in England. Could Dr. Austin give the results of any measurements of this conductivity at radio frequencies, either on the particular site in question or in any other parts of America? Having failed to find a place in this country (England) where the conductivity has a favorable low value, we have been compelled to resort to the alternative of using shorter waves in order to obtain measurable departures of the electric and magnetic fields at night from their normal day-time directions. Some measurements which have been carried out during the past three months on wave lengths between 300 and 500 meters have given very interesting results, and it is hoped to publish these shortly.

August 25, 1925.

L. W. Austin (by letter):* Of course I am very glad to find that Dr. Smith-Rose and Mr. Barfield have been interesting themselves in the phenomena described in my paper in the August

* Received by the Editor, October 29, 1925.

PROCEEDINGS, even though their observations do not agree with mine.

The conditions of experiment here were as follows: The double-axis radio compass coil and the well-shielded receiver were housed in a portable wooden garage about two hundred feet from the one-story radio building and a little nearer one of the 125-ft. steel towers which support the main laboratory antenna. The soil around the garage is a coarse gravel mixed with red clay and there is an average slope of the ground toward the southeast amounting to 8 or 10 degrees. Not far away there are a number of small trees. The radio-frequency resistance of the ground has never been measured with any accuracy, but from antenna resistance measurements it is believed to be high.

These conditions are not ideal, but special experiments show that the steel tower, which is by far the most dangerous source of error, plays no part in the phenomena. In these experiments the main antenna, supported by the tower in question, was tuned to WII, the station being observed for direction variation or blurred minimum, and alternately connected to the tuning inductances and left open. Within the accuracy of observation this produced no change in the deviation of the radio compass or in the vertical angle for which the minimum became sharp. The accuracy of setting was certainly better than 1 degree for the deviation and 5 degrees for the vertical angle.

Another experiment, in which the horizontal compass axis was pointed toward the transmitting station, showed a minimum of signal with the coil depressed about 10 degrees toward the southeast, i. e., bringing the coil roughly parallel with the sloping ground.

Still other experinents were made in which the antenna effect of the compass coil was varied by using a compensating capacity to earth and also by coupling a small untuned antenna to the coil. None of these, however, produced any certain changes in the phenomena.

Bureau of Standards, Washington, D. C.
October 26, 1925.

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

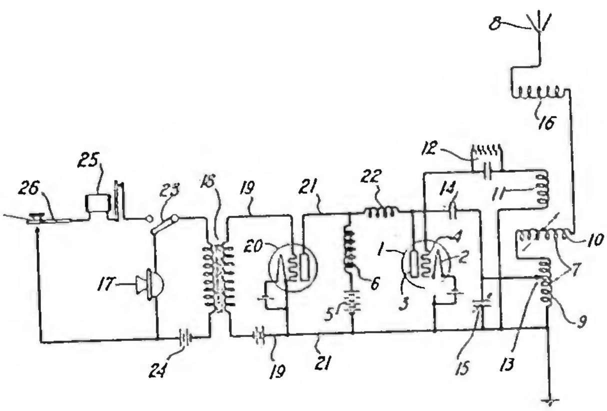
ISSUED SEPTEMBER 1, 1925—OCTOBER 27, 1925

BY

JOHN B. BRADY

(PATENT LAWYER, OURAY BUILDING, WASHINGTON, D. C.)

1,551,624—J. C. Schelleng, filed June 30, 1924, issued September 1, 1925. Assigned to Western Electric Company, Incorporated.



NUMBER 1,551,624—Circuits for Wave Transmission

CIRCUITS FOR WAVE TRANSMISSION where an electron tube oscillator is coupled with an antenna ground system through a variometer. The variometer may be readily shifted for changing the frequency of the generated oscillations and automatically adjusting the coupling with the antenna system for controlling the production of oscillations.

1,551,661—C. G. Hill, filed May 20, 1922, issued September 1, 1925.

VARIABLE CONDENSER where the plates are arranged in spiral formation in such manner that they may be varied in spaced relationship by rotative movement of a central shaft to vary the capacity of the condenser.

* Received by the Editor, November 15, 1925.

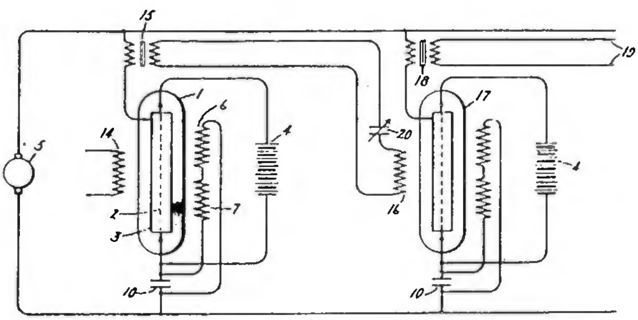
1,551,845—E. L. Popper, filed November 9, 1923, issued September 1, 1925.

CRYSTAL DETECTOR in the form of a cup-shaped unit which may be secured to a panel with a sensitive crystal secured within the cup-shaped unit between two electrodes. The detector is fixed and is intended for permanent mounting in such manner that it will not vary in its adjustment.

1,552,186—R. S. Alcox, filed July 29, 1923, issued September 1, 1925.

CONDENSER where the stator and rotor plates are formed of semi-circular concentric plates and moved to interleave with each other in an axial direction for varying the capacity of the condenser.

1,552,219—G. G. Mercer, filed April 23, 1920, issued September 1, 1925. Assigned to General Electric Company.



NUMBER 1,552,219—Vacuum Tube Circuits

VACUUM TUBE CIRCUITS in which an electron current is controlled by means of a magnetic field. The current which flows between a cathode and anode within an evacuated vessel also passes through a coil which surrounds the electrodes for producing a constant magnetic field in the space between the electrodes. The variable component of the current in the control circuit is by-passed to the cathode and does not affect the constant polarizarion supplied by the magnetic field.

1,552,266, F. C. Bradley, filed February 19, 1925, issued September 1, 1925.

A TUNING UNIT FOR RADIO RECEIVING SETS, including a condenser with a pair of hingedly mounted inductance units which may be rotated by a cam which is controlled from the shaft which carries the rotatable condenser plate. The cam varies the space relationship between the inductances in accordance with the movement of the plates of the condenser, and in this way the capacity is varied in proportion to the change in the inductances.

1,522,310—C. J. Kayko, filed July 24, 1923, issued September 1, 1925. Assigned to General Electric Company.

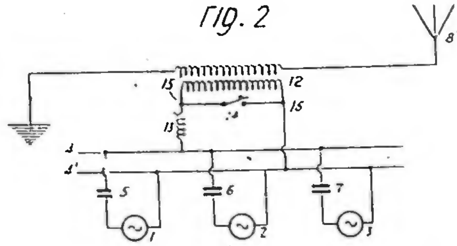
AN ELECTRODE FOR DISCHARGE TUBES which has an operating temperature which is relatively low for a high electron emission. The electron emitting cathode consists largely of nickel containing intimately incorporated therewith a compound of a highly positive metal which is sufficiently refractory to be retained in the cathode at a temperature at which the electron emission is substantial. The cathode obtains a temperature of about 1,200 deg. cent. at 0.1 to 1.0 amperes per square centimeter of surface.

1,552,606—G. Holst and E. Oosterhuis, filed November 4, 1921, issued September 8, 1925. Assigned to Naamlooze Vennootschap Philips' Glohilampen-Farriken, Eindhoven, Netherlands.

AN ELECTRON DISCHARGE DEVICE, in which the tube is filled with a gas consisting exclusively of pure argon of a pressure not exceeding one-tenth mm. of mercury. The argon filling is found to have the advantage of low tension on the anode for a maximum strength of signals, and the fact that a grid leak is unnecessary between the grid electrode and the filament when using the gas-filled tube.

1,552,670—G. Belfils, filed August 29, 1921, issued September 8, 1925.

FIG. 2

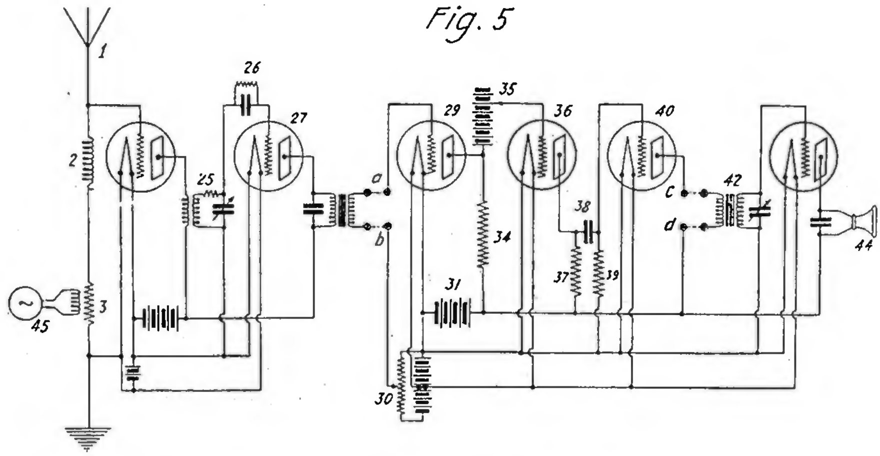


NUMBER 1,552,670—A Radio Transmitting System

A RADIO TRANSMITTING SYSTEM, in which a plurality of alternators are used in connection with the same antenna system. The alternators may be parallel connected to a radiating system and synchronized for operation with respect to each other.

1,552,829—H. J. J. M. De R. De Bellescize, filed August 29, 1921, issued September 8, 1925.

Radio Receiving System for the reduction of atmospheric disturbances in the receiving apparatus. A damped receiving circuit is provided where a current-limiting device for receiving the output of the receiving circuit is employed. The current-limiting device comprises a highly damped resonator for eliminating "grinders." A highly resonant circuit for receiving the output of the current-limiting device is provided, and the received signaling energy then impressed upon a responsive device.



Number 1,552,829—Radio Receiving System

1,552,882—F. C. Rowe, filed April 29, 1924, issued September 8, 1925.

Tuning Coil and Condenser Mounting for Radio Receivers in which an inductance coil and variable condenser are operated on the same rotatable shaft. The connections for one of the windings of the inductance system are brought out to rings which contact with brushes which are positioned with respect to the rotatable coil in such manner as to make sweeping connection with contacts carried by the rotatable coil for completing connection with an electrical circuit.

1,552,919—Gauthier, Crosby R., filed November 22, 1919, issued September 8, 1925. Assigned to Wired Radio, Inc., New York, New York.

Electrical Communication System where broadcasting stations are operated in multiple and modulated over the same interconnecting line wire system. A plurality of broadcasting stations are linked by the same line wire system in this case and modulated from the same broadcasting studio for simultaneous transmission from the several stations.

1,553,152—R. A. Fessenden, filed November 10, 1920, issued September 8, 1925. Assigned to Submarine Signal Company.

METHOD AND APPARATUS FOR GENERATING ELECTRICAL OSCILLATIONS utilizing two-electrode electron tubes by variation of electron emission of the tubes. A plurality of two-electrode electron tubes are arranged in the form of a Wheatstone bridge circuit and oscillatory variation of the electron emission secured by oscillatory surface heating of the surface of the hot cathodes.

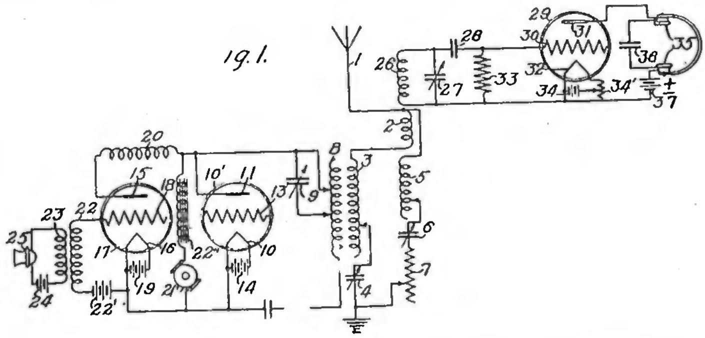
1,553,244—C. F. Jacobs, filed July 10, 1923, issued September 8, 1925.

AN ANTENNA SPREADER for an antenna where the wires are arranged in the form of a cage. The spreader consists of a flat metallic ring having notches at its periphery in which the antenna wires fit and in which they are secured in spaced relationship forming the antenna cage.

1,553,315—L. E. Gould, filed February 28, 1925, issued September 15, 1925.

LOOP FOR RADIORECEPTION in which a circular supporting turn of relatively heavy wire is provided which surrounds a plurality of turns of finer wire which are supported in spaced relationship by insulated members secure upon the heavier turn. The loop is compact in construction and may be inexpensively manufactured on a quantity production basis.

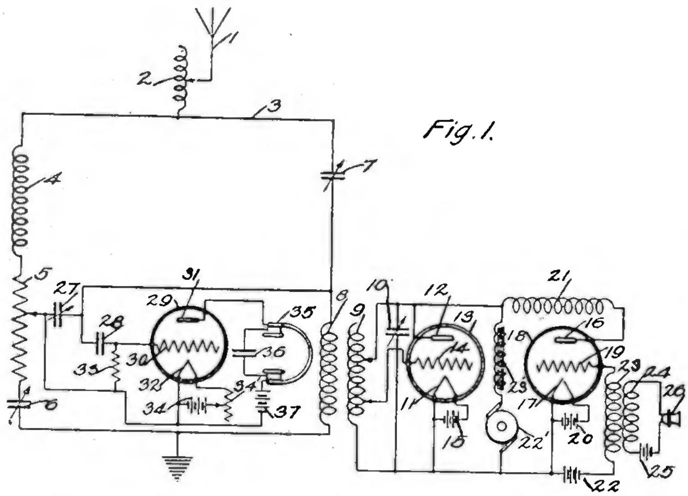
1,553,391—A. Nyman, filed July 15, 1920, issued September 15, 1925. Assigned to Westinghouse Electric and Manufacturing Company.



NUMBER 1,553,391—Combined Radio Sending and Receiving System

Combined Radio Sending and Receiving System where the effect of the local transmitting energy is neautralized by an auxiliary neutralizing reactance device. The antenna circuit is divided into two branches for connection to the transmitting and receiving apparatus, and undesired currents in the receiving branch are neutralized for preventing interference with the receiving circuit.

1,553,390—A. Nyman, filed July 15, 1920, issued September 15, 1925. Assigned to Westinghouse Electric and Manufacturing Company.



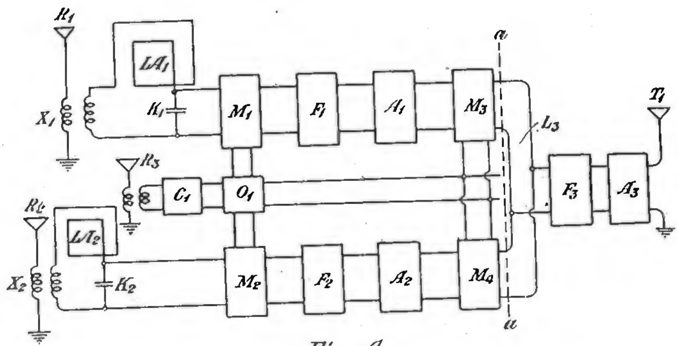
Number 1,553,390—Combined Radio Sending and Receiving System

Combined Radio Sending and Receiving System where the transmitting and receiving circuits are connected to the same antenna with the receiving apparatus connected at a selected point in the circuit at which there is a zero difference of potential arising from energy derived from the local transmitting station. By selecting a zero potential point interference from side tones is eliminated.

1,553,549—W. H. Priess, filed April 3, 1920, issued September 15, 1925. Assigned to Wireless Specialty Apparatus Company.

Condenser which is housed within a relatively heavy casing with a pressure clamp at one side of the casing for subjecting the stack to pressure, while permitting variation in length of the stack for condensers of different capacities.

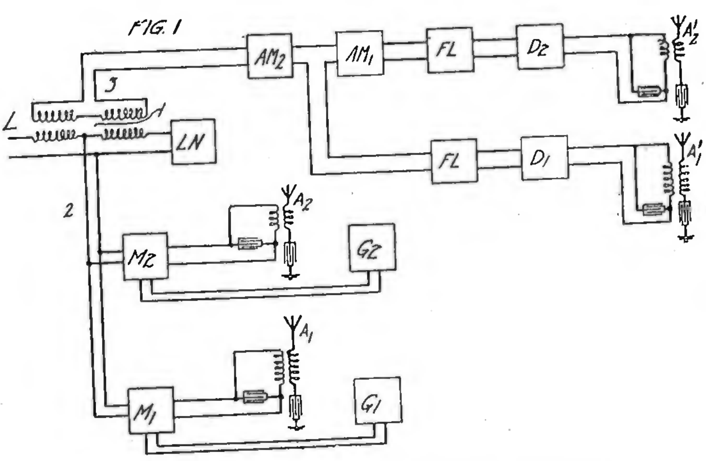
1,553,454—De Loss K. Martin, filed January 19, 1923, issued September 15, 1925. Assigned to American Telegraph and Telephone Company.



NUMBER 1,553,454—Radio Repeating System

RADIO REPEATING SYSTEM in which the number of frequencies per channel is limited to two per channel of transmission regardless of the number of repeaters employed. A directly controlled receiving antenna system is arranged to receive from different directions waves having the same frequency or band of frequencies. A frequency translating circuit is provided having means to change the frequency of each of the received waves by the same amount. A directly controlled transmitting system is provided to transmit the waves of changed frequency in different predetermined directions.

1,553,625—J. Mills, filed December 24, 1920, issued September 15, 1925. Assigned to Western Electric Company, Incorporated.



NUMBER 1,553,625—Duplex Radio System

Duplex Radio System where side tone interference at the receiving station due to simultaneous reception of the locally transmitted energy with the desired signaling energy is eliminated by the generation of an auxiliary wave by the local transmitter which opposes current in the receiving circuit induced therein by the local transmitter.

1,553,720—Cyril P. Ryan, filed February 27, 1924, issued September 15, 1925. Assigned to Vickers Limited, Westminster, England.

Radio Signaling and Control system where radio signals are emanated in regularly timed succession at a predetermined acoustic frequency together with the signaling impulses of other acoustic frequencies. A tuned receiving apparatus is arranged to respond only to the timed succession of signals for impressing said signals on a receiving apparatus which has an oscillation period corresponding to the timed succession of impulses for operating a recorder or other apparatus.

1,553,971—S. D. Apostol, filed March 10, 1925, issued September 15, 1925.

Condenser in which a cylindrical roll carries a curved metallic plate on one side thereof which may be moved toward or away from a similarly curved stator plate which is substantially U-shaped for varying the capacity of the condenser.

1,554,231—A. Press, Washington, D. C., filed February 18, 1921, issued September 22, 1925.

Hysteretic Generation of Electromagnetic Waves in a coil antenna system is employed where the antenna is immersed in a mixture of oils. The oils are selected and adjusted to provide a solution having hysteretic and purely dielectric affecting qualities.

1,554,232—A. Press, Washington, D. C., filed January 11, 1922, issued September 22, 1925.

Uniquely Resonant Coil comprising an open-ended helix with a shield extending substantially a one-half stationary wave length along the helix of the frequency of the electromotive force which is applied across the helix. The shield surrounds the helix along a selected portion thereof. The invention may be applied to wave coil systems in which standing waves are set up in an extended coil.

1,554,328—I. M. Brenner, New York, filed November 2, 1923, issued September 22, 1925.

RADIO ANTENNA of the loop type in which the conducting wires forming the loop are carried upon a frame and taps taken from various points in the loop to a selector switch. The selector switch controls a number of effective turns in the loop which may be connected to the radio receiving system.

1,558,436—I. Langmuir, Schenectady, New York, filed October 16, 1913, renewed March 14, 1916, isued October 20, 1925. Asigned to General Electric Company.

ELECTRICAL DISCHARGE APPARATUS AND PROCESS OF PREPARING AND USING THE SAME in which the tube is evacuated to such a degree that no positive ionization occurs when the impressed voltage is as high as 60 volts and the current over a working range of voltage up to 60 volts varies with the 3/2 power of the impressed voltage. This is the Langmuir patent based on the application which was involved in interference with Arnold on the subject matter of the hard tube.

1,554,598—P. J. Rudy, Centralia, Pennsylvania, filed March 14, 1925, isued September 22, 1925.

RADIO RECEIVER comprising a compact assembly of apparatus which is intended for portable use without the necessity of an antenna ground system. The circuit includes a rectifier sustem, a tuning coil, battery, and telephone receivers, the coil itself serving as a pick-up area for the signaling energy. The rectifier system consists of a plurality of steel, soft iron and lead balls making contact with each other.

1,554,501—L. C. F. Horle, Newark, New Jersey, and N. Heyman, Buffalo, N. Y., filed November 5, 1924, issued September 22, 1925. Assigned to Federal Telephone Manufacturing Corporation.

SHOCK PROOFING MOUNTING MEANS FOR VACUUM TUBES where the electron tubes are carried by a panel which is resiliently suspended on helical springs positioned between the stationary frame of the apparatus and the panel.

1,554,640—V. T. Miller, Kansas City, Mo., filed March 10, 1923, issued September 22, 1925.

CRYSTAL DETECTOR FOR RADIO RECEIVING SETS where the crystal is secured in a metallic mounting having a plurality of parallel grooves across the face thereof. The parallel grooves or corrugations have their ends overlapping the edges of the crystal, thereby firmly securing the crystal in the holder.

1,554,713—R. C. Clinker, Inwoods Rugby, England, filed July 10, 1920, issued September 22, 1925. Assigned to General Electric Company.

ELECTRICAL APPARATUS having a rotatable shaft which may be driven through coarse and fine adjustments by moving the control element of the shaft longitudinally with respect to the elements driven by the shaft, so that vernier or micrometer adjustment may be made in the apparatus unit driven by the shaft. The patent shows the invention as applied to a variable condenser.

1,554,720—C. V. Ferguson, Schenectady, New York, filed December 3, 1918, issued September 22, 1925. Assigned to General Electric Company.

ELECTRIC ARCH DEVICE, in which a thermionic discharge tube is provided with an anode consisting of an alloy of alkali metal which is vaporized during the discharge. The alkali vapor permits the operation of an arc-like discharge therein with a voltage of the order of one volt for current of about one ampere.

1,554,795—L. De Forest, New York. Filed May 10, 1915, issued September 22, 1925. Assigned to De Forest Radio Telephone and Telegraph Company.

RADIO SIGNALING SYSTEM, where an electron tube is employed for controlling the output of an arc transmitter. The electron tube has its output circuit connected in shunt with the antenna system and is used as a modulating device for the output of the arc.

1,555,251—W. H. Priess, Belmont, Massachusetts. Filed June 3, 1921, issued September 30, 1925. Assigned to Wireless Specialty Apparatus Company.

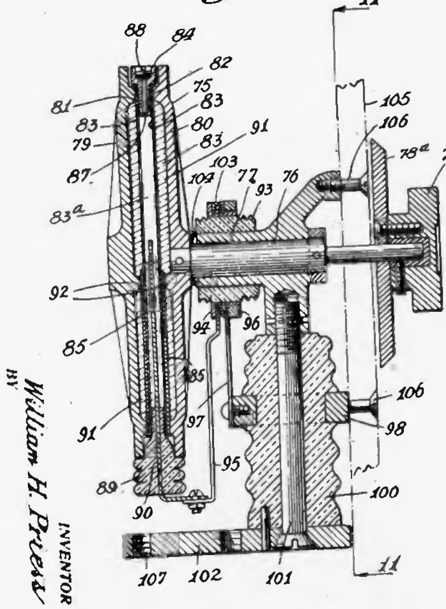
ELECTRICAL CONDENSER design for use in pack sets for the Government where the stack is arranged laterally of a casing and placed under pressure by clamps positioned on opposite sides of the casing.

1,555,252—W. H. Priess, Belmont, Massachusetts, filed June 7, 1921, issued September 30, 1925. Assigned to Wireless Specialty Apparatus Company.

Electrical Condenser for high-potential operation where the plates are built up in stacked formation where a stiff plate is interposed in the stack between the ends thereof and a clamp external to the stack for engaging the plate to brace the plate and the stack. The clamp may also be used to compress the stack from end to end.

1,555,253—W. H. Priess, Belmont, Massachusetts, filed January 6, 1922, issued September 30, 1925. Assigned to Wireless Specialty Apparatus Company.

Fig. 10.

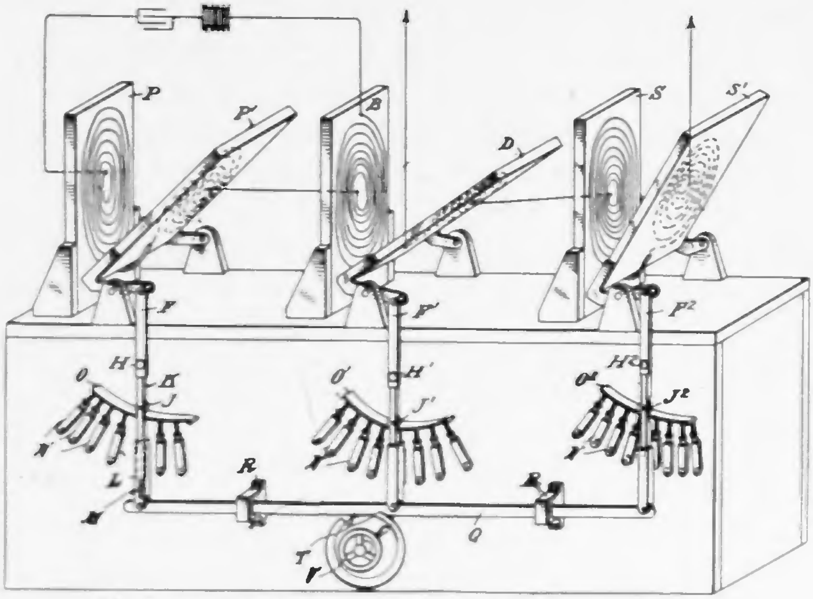


Number 1,553,253—Variable Electrical Condenser

Variable Electrical Condenser, where a body of mercury is enclosed within a rotatable reservoir which contains conductive side plates on opposite sides thereof. The reservoir may be moved to different positions to displace the mercury from the the reservoir for providing different overlapping relationships between the mercury and the side plates for varying the capacity of the condenser.

1,555,254—J. A. Proctor, Lexington, Massachusetts. Filed September 4, 1920, issued September 29, 1925. Assigned to Wireless Specialty Apparatus Company.

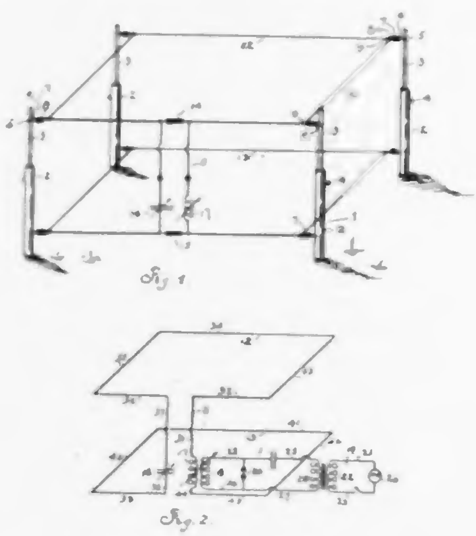
Fig. 8.



NUMBER 1,555,254—Electrical Tuning Apparatus

ELECTRICAL TUNING APPARATUS comprising a plurality of separate tuning elements which are connected by a common adjuster and the several elements simultaneously controlled from a single hand actuator. Suitable primary adjustments may be made in each of the tuning elements so that simultaneous effect of all of the tuning elements will be to efficiently control associated electrical circuits.

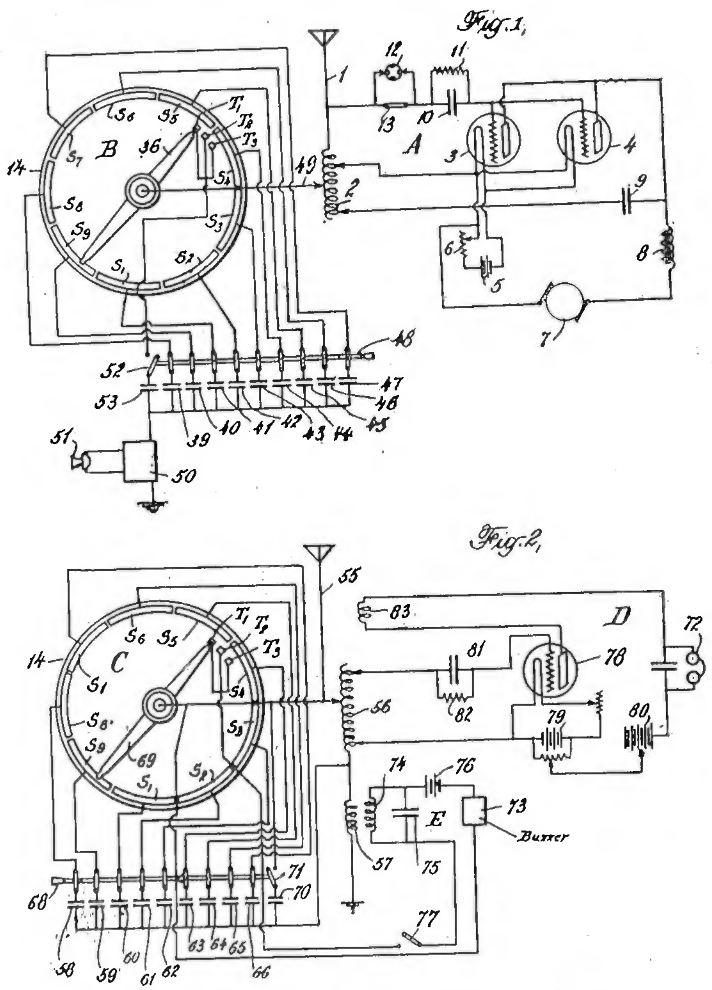
1,555,345—J. A. Willoughby, Washington, D. C., filed June 26, 1919, issued September 30, 1925.



NUMBER 1,555,345—Loop Antenna

LOOP ANTENNA for polarized wave transmission. A plurality of oscillatory loops are provided with substantially closed metallic circuits. A transmitter is arranged to excite the loops in opposition to each other. The loops are spaced substantially parallel and produce by such opposition a field of magnetic disturbance which emanates in a controllable direction.

1,555,633—L. Burch, Belleville, New Jersey. and M. K. Parkhurst, New York City, filed December 9, 1922, issued September 29, 1925.



NUMBER 1,555,633—Secret Signaling System

SECRET SIGNALING SYSTEM, in which the frequency of the transmitting station is varied over a range synchronously with the variation in frequency of the receiving station. A gyroscope is provided at each terminal and arranged to drive a switch which varies the frequency of the transmitting and receiving circuits simultaneously for maintaining the circuits in synchronism.

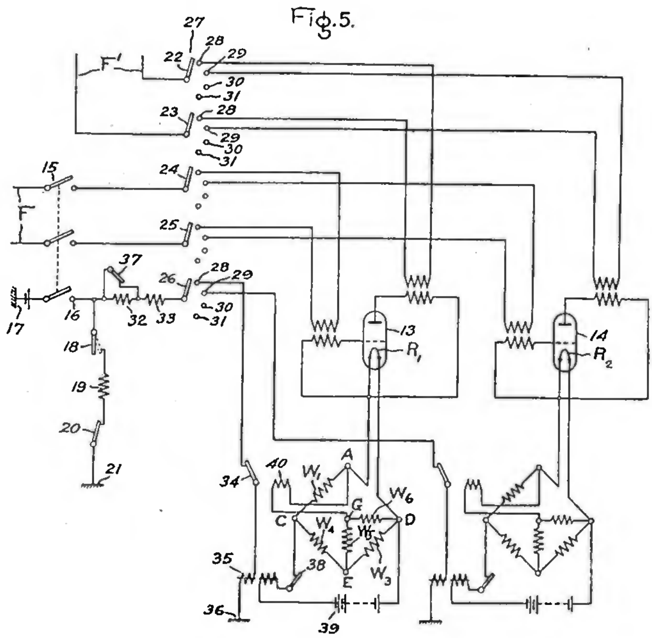
1,555,634—S. Cohen, Brooklyn, New York filed November 8, 1924, issued September 29, 1925. Assigned to General Instrument Corporation.

VARIABLE CONDENSER, in which a set of stator plates are mounted in spaced relation within a condenser frame and secured therein by insulators positioned between the frame and the stator plates. The patent illustrates the use of pedestal insulators for supporting the stator plates within the condenser frame with the rotor plates carried by a shaft which is journaled in the frame in such manner that the rotor plates may be inter-leaved with the stator plates. The condenser is designed to have minimum dielectric losses.

1,555,677—C. L. A. M. LeBlanc, Paris, France, filed March 22, 1921, issued September 29, 1925. Assigned to Societe Anonyme pour L'Exploitation des Procedes Maurice-Vickers, Paris.

ELECTRON TUBES for high-power operation where the cathode is in the form of a hollow cylinder and is heated by radiation from a resistance disposed in the interior of the cylinder and raised to incandescence. The grid and anode are also in the form of cylinders co-axial with the cathode. The cathode is heated by thermo-radiation.

1,555,757—G. Respondek, Berlin, Germany, filed November 30, 1925, issued September 29, 1925. Assigned to General Electric Company.

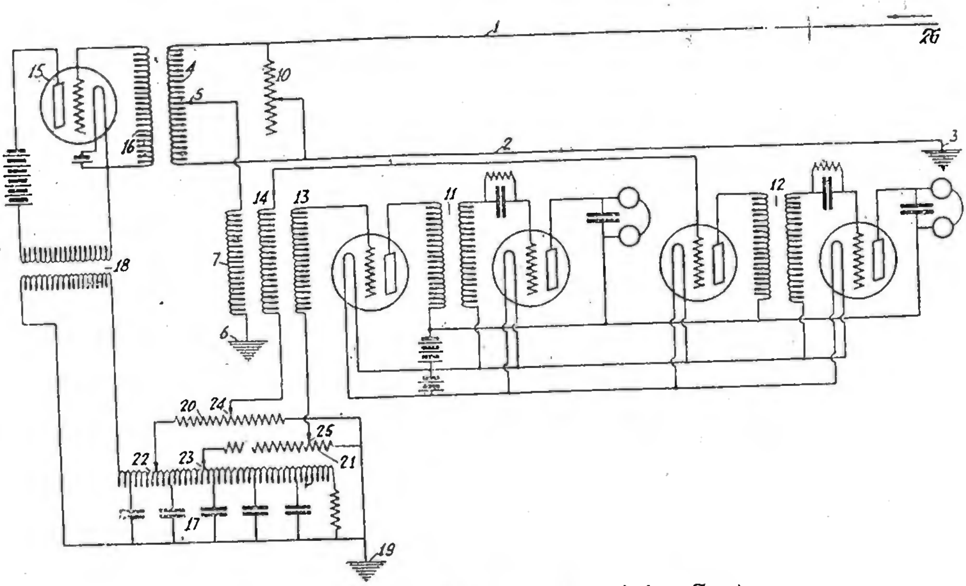


NUMBER 1,555,757—Connection for Vacuum Tubes

CONNECTION FOR VACUUM TUBES, where a spare tube may be automatically replaced for a tube which has been burned out in the course of normal operation. A Wheatstone bridge circuit is provided where one arm of the Wheatstone bridge is formed by the cathode of the electron tube. When the cathode is destroyed, the bridge is unbalanced, and a relay is operated for connecting the spare tube in the circuit.

1,556,122—A. B. Moulton, New York, filed November 1, 1922, issued October 6, 1925. Assigned to Radio Corporation of America.

Fig. 1

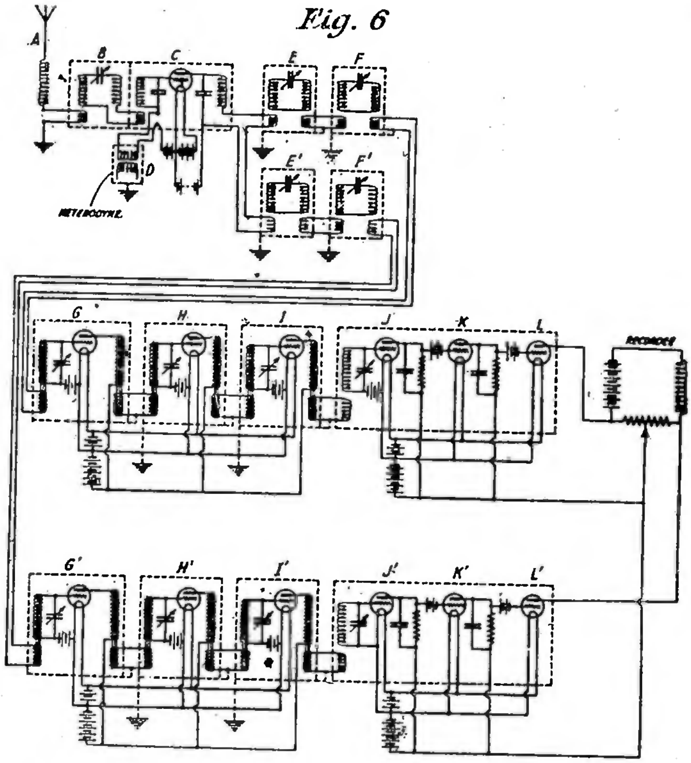


NUMBER 1,556,122—Radio Receiving System

RADIO RECEIVING SYSTEM, using a Beverage antenna system extending in the general direction of transmission. The antenna system comprises two parallel conductors with a damping circuit connected across a pair of adjoining ends comprising an inductance and a resistance in parallel with a capacity and an inductance in series with the resistance. The circuit is intended for uni-directional reception and for the elimination of interfering signals and atmospheric disturbances. The receiving system is connected to the line wire system at the end opposite the connection of the impedance circuit to the line conductors.

1,556,129—H. J. Round, London, England, filed July 9, 1921, issued October 6, 1925. Assigned to Radio Corporation of America.

Fig. 6



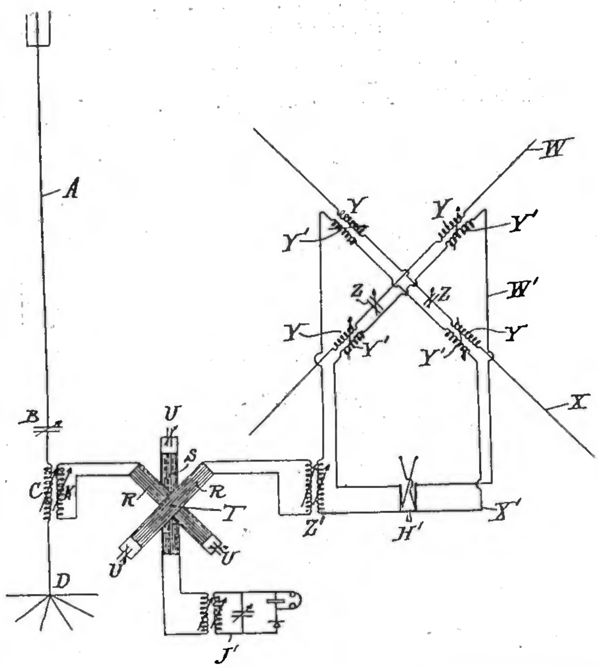
NUMBER 1,556,129—Reception of Radio Signals

RECEPTION OF RADIO SIGNALS by a circuit arrangement designed to be substantially free of atmospheric disturbances. A heterodyne is provided at the receiving circuit to produce beat tones, the currents of which are passed through resonance circuits where undesired disturbances are damped, while the desired signal is carried forward through a system of electron tubes for operating a recorder.

1,556,130—O. Schriever, Berlin, Germany, filed December 27, 1922, issued October 6, 1925. Assigned to Gesellschaft für Drahtlose Telegraphie.

CIRCUIT ARRANGEMENTS FOR RADIO SIGNALING, where transmitting stations are operated at close proximity to each other. A coupling is provided for neutralizing undesired reactive effects between separate transmitting stations which are operated simultaneously.

1,556,137—R. A. Weagant, Douglas Manor, New York, filed February 7, 1919, issued October 6, 1925. Assigned to Radio Corporation of America.



Number 1,556,137—Method and Apparatus for Radio Signaling

Method and Apparatus for Radio Signaling, whereby static interference may be reduced to a minimum. Two pick-up circuits are provided one of which efficiently receives horizontally propagated signal waves, while the other efficiently receives static impulses as currents substantially in opposite phase to the signaling currents. By this arrangement the static is balanced out while retaining the signal currents.

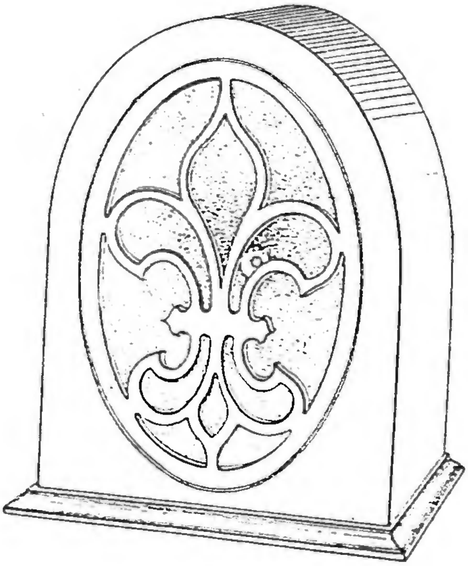
1,556,435—A. S. Gorayeb, New York, filed September 15, 1923, issued October 6, 1925.

Portable Antenna, which consists of a casing which may be fitted in the ordinary window sill with an antenna wire carried by a roll within the casing. The antenna wire may be released to extend downwardly from the window frame when the receiving set is placed in operation.

1,556,725—D. H. Shallcross, Claredon, Virginia, filed January 31, 1922, issued October 13, 1925.

Support for Radio Antennas of the loop type for direction finder work. The support is in the form of a collapsible coil frame having a plurality of members hingedly connected together. In open position the support carries all of the turns of the loop, while in closed position the support may be readily transported from place to place.

Design 68,493—M. C. Rypinski, filed August 6, 1925, issued October 13, 1925. Assigned to Brandes Laboratories, Incorporated, of Newark, New Jersey.



Design 68,493—Radio Reproducer

Radio Reproducer—This patent covers the Brandes cone speaker where an elliptical cone of relatively small size is housed within an ornamental cabinet, shaped to conform with the general contour of the diaphragm.

1,556,633—S. Ruben, New York City, filed September 13, 1924, issued October 13, 1925.

Electrical Control Method for trains and other moving vehicles, where transmitting and receiving apparatus are located upon separate trains which may be approaching each other on the same track. The transmitting apparatus may electrostatically transfer its energy to an overhead line system from the moving train, which energy is in turn transferred again to the other moving train for actuating the receiving apparatus on said train for controlling brakes or other signal when the signal strength has reached a proper degree by reason of the close approach of the trains.

1,556,740 — R. B. Woolverton, Washington, D. C., filed (original) July 25, 1921. Divisional, December 10, 1923. Issued October 13, 1925.

Transmission of Radio Signals Employing Undamped Waves developed by an arc generator, where a tuned oscillatory circuit is shunted around the arc and an antenna circuit inductively coupled with the tuned oscillatory circuit. A variable inductance is connected in series in the antenna circuit. A solenoid is arranged in a keying circuit and connected so that the inductance of the variable inductance device may be varied for destroying the resonance between the oscillatory and antenna circuits for forming signals in the antenna circuit.

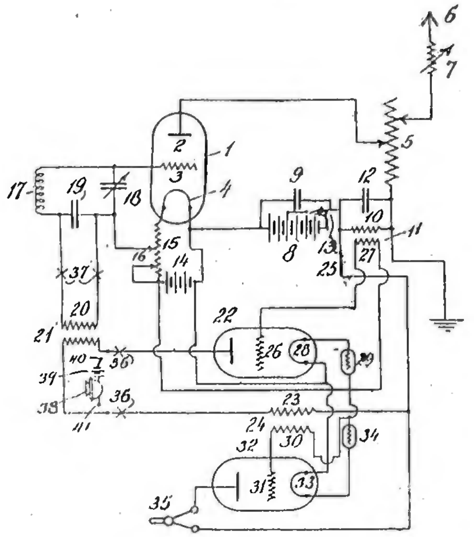
1,556,750—L. B. Bender, Washington, D. C., filed August 29, 1923, issued October 13, 1925.

ELECTRICAL SIGNALING in which the dots and dashes of the Morse code are transmitted and received at different frequencies and combined to actuate a siphon recorder for recording the received signals on a tape.

1,557,049—J. H. Hammond, Jr., Gloucester, Massachusetts, filed May 10, 1918, issued October 13, 1925.

ELECTRICAL ANTENNA for ship use, and particularly submarines, where the antenna is carried in an elongated buoy tube arranged to float on the surface of the water and trail the ship. The antenna wire is carried within the tube and connections established with the apparatus aboard the moving vessel.

1,557,067—L. Kuhn, Berlin-Charlottenburg, Germany, filed August 26, 1921, issued October 13, 1925. Assigned to Westinghouse Electric and Manufacturing Company.



NUMBER 1,557,067—Combined Transmitting and Receiving Arrangement

COMBINED TRANSMITTING AND RECEIVING ARRANGEMENT where an electron tube is arranged to function both as an oscillator and a detector and is coupled to an antenna system. Switching means are provided whereby the same tube circuits may be employed for reception or transmission.

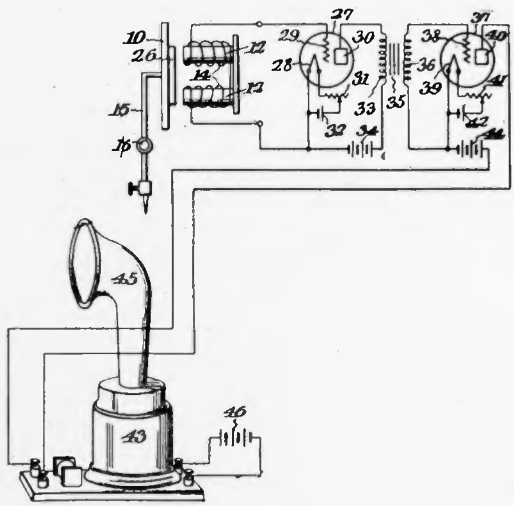
1,557,316—George H. Nobbs, Watertown, Massachusetts, filed December 16, 1924, issued October 13, 1925.

VARIABLE CONDENSER, in which a selected number of stator plates may be secured in an electrical circuit in which the condenser is connected. The condenser is of the rotatable plate variety and a switch is mounted adjacent the stator plates for establishing peripheral contact with selected plates in order to include a desired number of plates in the circuit.

1,557,389—E. N. Todd, Crisfield, Maryland, filed March 16, 1925, issued October 13, 1925.

MEANS FOR ASCERTAINING ELEVATIONS OF AIRCRAFTS comprising a radio transmitting unit which may be dropped from an aircraft and the circuits thereof automatically completed by impact of the apparatus with ground. The aviator may pick up the signals which are radiated from the transmitting apparatus which he has released and by suitable calculation determine his distance above the earth.

1,557,529—E. T. Jones, filed December 3, 1921, issued September 15, 1925.



NUMBER 1,557,529—Electrical Reproducer for Phonographs

ELECTRICAL REPRODUCER FOR PHONOGRAPHS where a diaphragm is actuated to vary the magnetic reluctance of a telephone circuit for generating electrical energy in a pair of associated windings. The energy is amplified and reproduced in accordance with the vibrations of the diaphragm.

1,557,724—W. H. Priess, Belmont, Massachusetts, filed August 2, 1921, issued October 20, 1925. Assigned to Wireless Specialty Apparatus Company.

MACHINE AND METHOD FOR BUILDING ELECTRICAL CONDENSER STACKS by building up alternately a dielectric sheet and a metallic foil sheet and flattening the foil in place on the dielectric sheet by applying a gas under pressure to the said foil. The fixed condensers may be manufactured inexpensively on a quantity production basis.

1,557,725—J. A. Proctor, Lexington, Massachusetts, filed February 1, 1921. Assigned to Wireless Specialty Apparatus Company.

VARIABLE ELECTRICAL CONDENSER where the condenser is housed in a vacuum container, and the moving condenser plates rotated in varying degrees by means of a magnet which is moved exterior of the casing which houses the condenser.

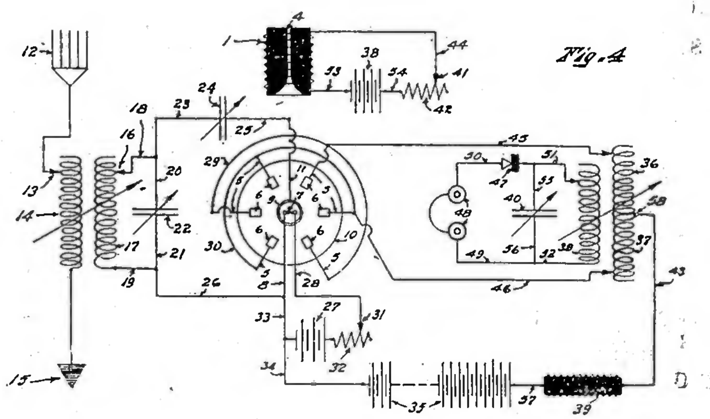
1,558,043—W. H. Priess, Belmont, Massachusetts, filed April 28, 1921, issued October 20, 1925. Assigned to Wireless Specialty Apparatus Company.

ELECTRICAL CONDENSER for high-power operation in which concentric metallic armatures are insulated one from another and embedded in sulphur which provides high insulation resistance with low dielectric loss.

1,558,111—H. E. Metcalf, San Leandro, California, filed March 23, 1925, issued October 20, 1925. Assigned to The Magnavox Company.

VACUUM TUBE in which the grid electrode is formed from a flat plate having a plurality of arms on each edge thereof. The arms are bent in relatively opposite directions to form a trough along each edge of the plate member between the arms, in which trough the filament electrode is stretched. This construction of tube if desirable from the viewpoint of manufacture and assembly for the filament is not centered betwen the electrodes.

1,558,120—F. G. Simpson, Seattle, Washington, filed April 3, 1921, issued October 20, 1925.



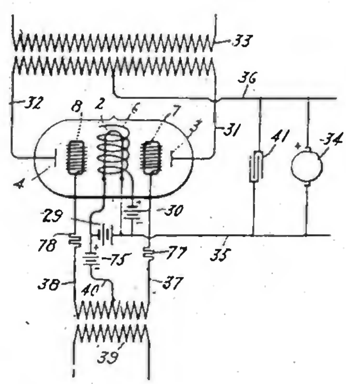
NUMBER 1,558,120—Radio Receiving System

RADIO RECEIVING SYSTEM, in which an alternating current generator consisting of an electron tube system is provided at the receiver and a magnetic field established transverse to the electron stream for varying the velocity of the electron stream in accordance with incoming signaling energy for correspondingly varying the frequency of the alternating current generator and operating a suitable observing circuit.

1,558,144—H. Chireix, Paris, France, filed August 29, 1921, issued October 20, 1925.

ELECTRIC RELAY, comprising an oscillation generator having two oscillating circuits connected thereto and tuned to different frequencies. The generator may be caused to oscillate at either of the frequencies separately. The relay may be used in various circuit arrangements.

1,558,437—I. Langmuir, Schenectady, New York, filed October 29, 1913, renewed February 29, 1924, issued October 20, 1924. Assigned to General Electric Company.



NUMBER 1,558,437—Electrical Discharge Apparatus

Electrical Discharge Apparatus in which an auxiliary conductor is provided within a three-electrode electron tube adjacent to the cathode and maintained at a substantially uniform positive potential with respect to the cathode. By this arrangement the effect of space charge in an electron discharge device is reduced, the effect of negatively charged bodies in the proximity of the cathode is eliminated, the discharge current with a given applied voltage is increased and electrons having a relatively uniform velocity are developed.

1,558,535—P. D. Delany, South Orange, New Jerssy, filed February 1, 1922, issued October 27, 1925. Assigned to International Telepost Company, Incorporated.

Secret Radio System, in which fragmentary parts of the signaling energy which make up telegraphic characters or vocal or instrumental sounds are transmitted at separate frequencies and pieced together at the receiver on a visual recorder for combining the fragmentary parts into an intelligible signal.

1,558,830—Wm. R. Brough, East Orange, New Jersey, filed May 13, 1922, issued October 27, 1925. Assigned to Western Electric Company, Incorporated.

Electron Discharge Device of high power size, where the plate electrodes are cooled by means of a water jacket which is secured by means of a screw collar around the exterior of the tube.

1,558,883—Wm. G. Housekeeper, New York, New York, filed April 20, 1921, issued October 27, 1925. Assigned to Western Electric Company, Incorporated.

A Vacuum Tube, in which the electrodes are constructed and supported away from the center of the electron tube. A support for the tube electrodes is provided where the electrodes are substantially removed from the center of the tube and supported more nearly adjacent the cylindrical sides of the vessel which houses the tube electrodes.

1,558,961—Wm. R. Bullimore, London, England, filed November 24, 1924, issued October 27, 1925.

Manufacture of Filament or Cathodes for Electric Lamps, Thermionic Tubes and the Like, in which a core of relatively high specific resistance and melting point is coated with a noble metal and an active coating of one or more compounds of the alkaline earth metals.

1,559,116—W. A. Marrison, East Orange, New Jersey, filed October 16, 1924, issued October 27, 1925. Assigned to Western Electric Company, Incorporated.

Wave Generating and Modulating System Employing Quartz Piezo Electric Crystal Circuits in which a plurality of frequencies are generated by piezo electric crystals which frequencies react to produce an audio frequency current under control of the piezo electric crystals.

1,559,193—Maurice W. Stavrum, Robert L. Olson and Wallace H. Berry, Chicago, Illinois, filed August 25, 1924, issued October 27, 1925.

Folding Loop Antenna, in which the frame members of the loop are pivotally mounted on a central clock with slotted end supports at the extremities of the frame members for spacing the turns of the loop with respect to each other. The loop frame may be folded into a small compass.

1,559,280—V. L. Ronci, Brooklyn, New York, filed March 23, 1923, issued October 27, 1925. Assigned to Western Electric Company.

Electron Discharge Device in which the electrodes are supported on threaded metallic members which extend through apertures in a flat insulated disk. The electrodes are secured upon the threaded members by means of wires which are formed into a helix and threaded upon the screw threaded members.

1,559,404—Paul Bunet, Paris, France, filed April 14, 1921, issued October 27, 1925.

High Tension Electrical Condenser in which tubular rod members are provided as the condenser electrodes. The rods making up each side of the condenser are positioned in the form of rings permanently electrically connected together so as to prevent sparking between parts. The distance between successive rods gradually increases throughout the area of the condenser.

1,559,460—S. Ruben, New York, New York, filed June 30, 1920, issued October 27, 1925.

Electron Tube in which a pair of anodes are provided with one of said anodes having an opening therethrough for the passage of electrons from an electron source within the tube. The electron stream is controlled by an external control member which is arranged adjacent the anodes. The tube is intended for high-power transmission.

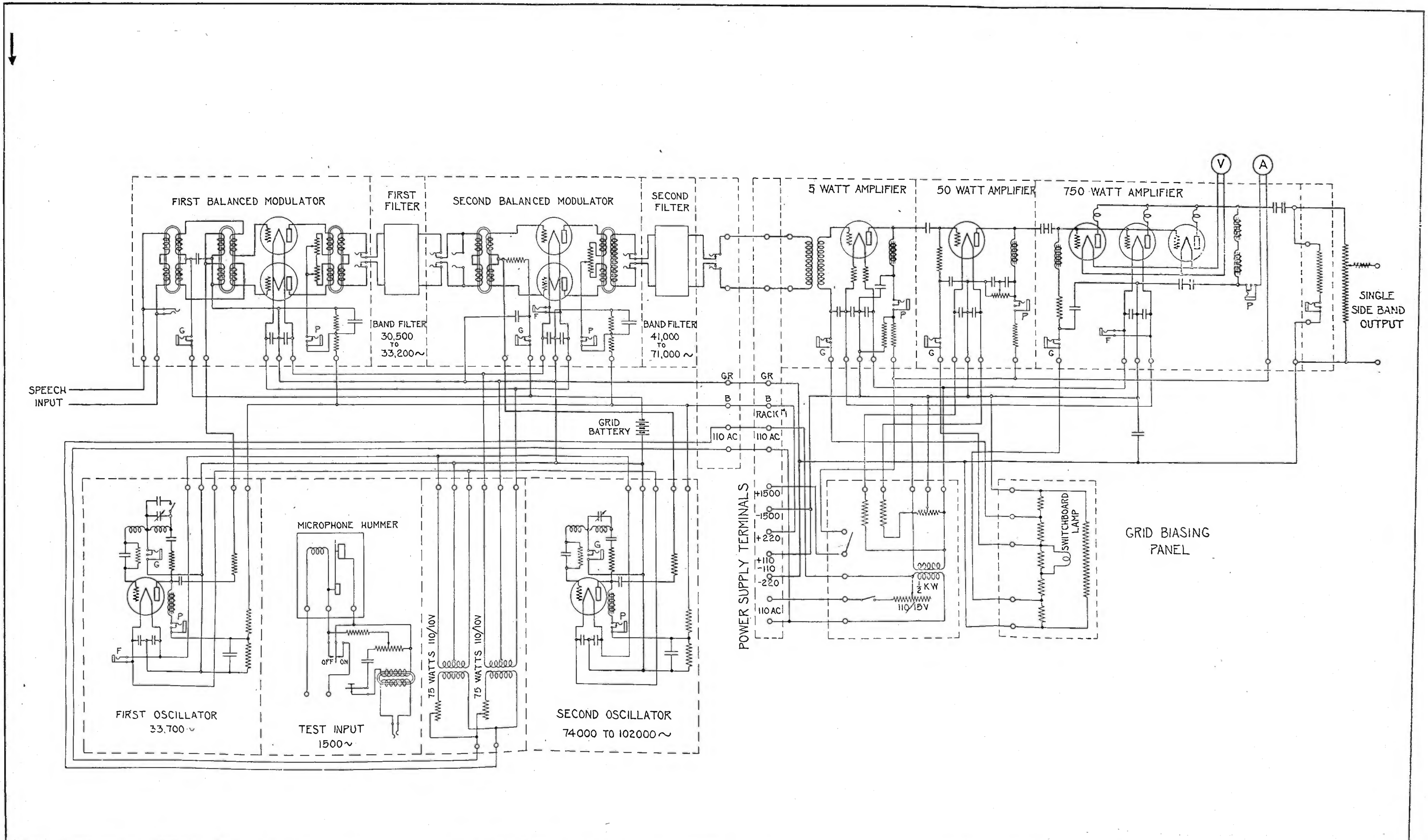


FIGURE 9—Schematic Diagram of the Circuit

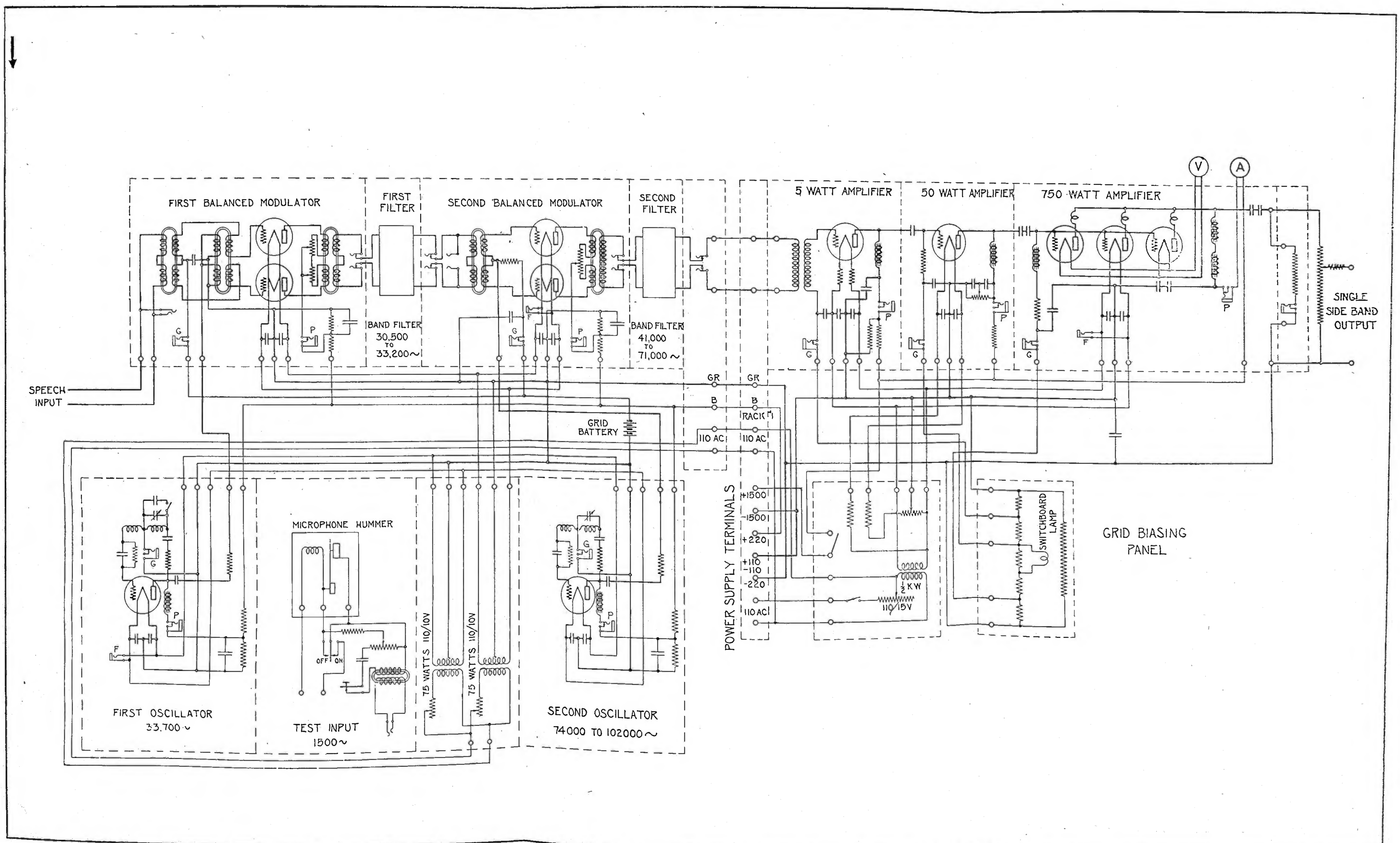


FIGURE 9—Schematic Diagram of the Circuit