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PROCEEDINGS
of
The Institute of Radio Engineers

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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 20 | July, 1932 | Number 7

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The Institute of Radio Engineers

GENERAL INFORMATION

INSTITUTE. The Institute of Radio Engineers was formed in 1912 through the amalgamation of the Society of Wireless Telegraph Engineers of Boston, Massachusetts, and the Wireless Institute of America of New York City. Its headquarters were established in New York City and the membership has grown from less than fifty members at the start to almost seven thousand by the end of 1931.

AIMS AND OBJECTS. The Institute functions solely to advance the theory and practice of radio and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members. Among the methods of accomplishing this need is the publication of papers, discussions, and communications of interest to the membership.

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	Paris, Consulat Chinois, 5 Ave. Daniel Lesueur	Wang, T. S.
Hawaii	Pearl Harbor, c/o Radio Material Office, Navy Yard	Zimmerman, P. R.

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Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before August 1, 1932. Final action on these applications will be taken on August 3, 1932.

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Louis R. Krumm
Treasurer of the Institute, 1917

Louis R. Krumm was born in Columbus, Ohio, on June 15, 1877. For six years after graduating from Ohio State University in 1899 with an electrical engineering degree, he was engaged in telephone engineering work. He was then appointed electrical engineer in the War Department in charge of signal installations for coast defenses and other military areas of the United States.

He became chief radio inspector in the Department of Commerce in 1911, continuing until 1917 when at the outbreak of the war he was appointed Captain in the Signal Corps.

He served in France for eighteen months during the World War as a member of the staff of the Chief Signal Officer, and returned to the United States in April, 1919, with the rank of Lieutenant Colonel. He is the recipient of the Legion of Honor Medal from France and the Distinguished Service Medal from the United States Government.

Resigning from military service in 1919, he served again for a year as chief radio inspector, and then became associated with the Westinghouse Electric and Manufacturing Company in New York City. Later he was transferred to the Pittsburgh general offices as superintendent of radio operations, in which capacity he was responsible for the construction and operation of KDKA and other broadcast and radiotelegraph stations. In 1922 he became associated with the largest radio distributing organization in Columbus, Ohio, and five years later left to conduct his own business as manufacturer's representative.

He became an Associate member of the Institute in 1913, and transferred to the grade of Member in 1915.

INSTITUTE NOTES

June Meeting of the Board of Directors

The June meeting of the Board of Directors was held on the 1st at the Institute office, and those present were: W. G. Cady, president; Melville Eastham, treasurer; Alfred N. Goldsmith, editor; O. H. Caldwell, J. V. L. Hogan, H. W. Houck, L. M. Hull, C. M. Jansky, Jr., R. H. Marriott, E. L. Nelson, A. F. Van Dyck, William Wilson, and H. P. Westman, secretary.

Eighty Associates, one Junior, and eight Students were elected to membership.

The financial status of the Institute was carefully scrutinized and a revision made in the budget. The revised budget calls for a reduction in the amount of money expended for PROCEEDINGS and salaries among other items.

It was agreed that the Institute office be operated on a five-day week basis, although a small force will be maintained in the Institute office on Saturday mornings to take care of visitors or such telephoned business which may eventuate.

The Emergency Employment Committee report indicated receipts of approximately $3,800 to date with expenditures approximating $4,500. During last month, three members have definitely been placed in either temporary or permanent employment. A number have been put in contact with possible employers, and between thirty and forty men have been employed on broadcast reception survey work. The original drawing account for $1,000 was increased to $1,800 to permit the committee to carry on during the summer months until the September meeting of the Board.

Radio Transmissions of Standard Frequency

The Bureau of Standards transmits standard frequencies from its station WWV, Washington, D. C., every Tuesday. The transmissions are on 5000 kilocycles, and are given continuously from 2:00 to 4:00 P.M., and from 10:00 P.M. to 12:00 midnight, Eastern Standard Time. (From October, 1931, to March, 1932, inclusive, the evening schedule was two hours earlier.) This service may be used by transmitting stations in adjusting their transmitters to exact frequency, and by the public in calibrating frequency standards and transmitting and receiving apparatus. The transmissions can be heard and utilized by stations equipped for continuous-wave reception throughout the United States be ad-

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The committee discussed reports made by the chairmen of the various technical committees and established a general program looking forward to the preparation of the preliminary draft of the final report.

Technical Committee on Electro-Acoustic Devices—IRE

On June 3 a meeting of the Technical Committee on Electro-Acoustic Devices of the Institute was held with E. D. Cook, chairman; Benjamin Olney, L. J. Sivian, and B. Dudley, secretary, in attendance.

The committee finished its report which is now being set up in type for final consideration by the main Standards Committee.

Technical Committee on Fundamental Units and Measurements

H. M. Turner, chairman; C. R. Englund, R. F. Field, G. C. Southworth, and B. Dudley, secretary, attended the May 27 meeting of the Technical Committee on Fundamental Units and Measurements which was held at the Institute office.

At this meeting the committee practically concluded its report, and a few remaining items of minor nature will be taken care of in subsequent correspondence.

Technical Committee on Radio Receivers

The Technical Committee on Radio Receivers held a meeting on May 12 in the office of the Institute and those present were H. A. Wheeler, chairman; R. D. Brown (representing W. E. Holland), E. T. Dickey, Virgil M. Graham, David Grimes, F. A. Hinners, F. A. Polkinghorn, A. E. Thiessen, L. P. Tuckerman (representing C. E. Brigham), Lincoln Walsh, W. T. Wintringham, and B. Dudley, secretary.

At this meeting the committee finished the major portion of its work, and its report will be available for inclusion with the others in the preliminary report to be considered by the main Standards Committee in September.

Institute Meetings

Atlanta Section

A meeting of the Atlanta Section was held on March 31 at the Monitoring Station of the Department of Commerce at Marietta, Ga. Members and their guests met at the Atlanta Athletic Club and drove to the Monitoring Station. The various antenna systems of the station were inspected and their methods of operation explained by W. J. Holey, Assistant Radio Inspector, who was responsible for their design.

The paper was discussed by Messrs. Kilgour, Glessner, Osterbrok, and Wilson of the twenty-nine members and guests in attendance. Twenty-one of these attended the informal dinner which preceded the meeting.

Detroit Section

H. L. Byerlay, chairman, presided at the May 20 meeting of the Detroit Section held at the Edison Boat Club.

The paper of the evening on "Television" was presented by J. E. Brown and R. L. Osborne. Mr. Brown described his experimental television transmitter, pointing out many of the difficulties encountered in its design and construction, and methods employed to overcome them. He then described some of the results that have been obtained using a scanning disk for dividing the image into its elementary areas. Some results obtained with cathode ray tubes were then outlined. The paper was followed by a demonstration of television employing two Jenkins receivers to pick up signals broadcast from Mr. Brown's station which was located about five miles away. The demonstration was considered very successful and the paper was discussed by a number of the 153 members and guests in attendance. The informal dinner which preceded the meeting was attended by sixty.

New York Meeting

Two papers were presented at the June 1 New York meeting held in the auditorium of the Engineering Societies Building and presided over by R. H. Marriott in the absence of Dr. Cady.

The first of these papers "Transoceanic Reception of High-Frequency Telephone Signals" by R. M. Morris and W. A. R. Brown of the National Broadcasting Company was presented by Mr. Morris, and treated the application of high-frequency telephone transmission to international rebroadcasting. Methods used in rating the suitability of reception for rebroadcasting were outlined and the effects of magnetic disturbances upon transmission, the correlation of magnetic activity with transmission, and the forecasting of magnetic disturbances and resultant transmission conditions were discussed.

The second paper, presented by Clifford N. Anderson of the American Telephone and Telegraph Company, was on "North Atlantic Ship-Shore Radiotelephone Transmission during 1930 and 1931." The paper was based upon the considerable data on radio transmission collected during the years 1930 and 1931 incidental to the operation of a ship-shore radiotelephone service with several passenger ships operating in the North Atlantic. The results of an analysis of these data were discussed. Contour diagrams were given which showed the variation of signal fields with distance and time of day for the various seasons of the year on frequencies of approximately 4, 9, 13, and 18 megacycles. Similar diagrams were employed to show the distribution of commercial circuits and curves were also given to enable the data to be applied more generally under other conditions of noise and radiated power.

The meeting was attended by 100 members and guests, several of whom participated in the discussion which followed the presentation of the papers.

San Francisco Section

The San Francisco Section held its May meeting on the 18th at the Bellevue Hotel, Chairman R. M. Heintz presiding.

A paper on "The Communication System of the Mackay Radio and Telegraph Company" was presented by G. T. Royden, division engineer of that organization.

The meeting was attended by seventeen members and guests.

Seattle Section

L. C. Austin, chairman, presided at the April 28 meeting of the Seattle Section held at the University of Washington.

A paper by J. R. Tolmie of the Pacific Telephone and Telegraph Company was presented on "The Application of Line Transmission Theory to Radio Transmission."

At the conclusion of the paper, a practical demonstration was give by T. M. Libby of the theory and conclusions covered in it. Part of this demonstration employed an antenna made up of special neon tubes a half wave in length.

The meeting was attended by thirty-eight members and guests, a number of whom participated in the discussion of the paper.

Toronto Section

The annual meeting of the Toronto Section was held on May 11 at the University of Toronto with F. K. Dalton, chairman, presiding. The chairman opened the meeting by reviewing the activities of the section during the past season, then calling upon the Secretary-Treasurer for the financial report and the chairmen of the various committees for reports on their activities. The Nominating Committee then submitted its report and as no further nominations were made from the floor the single nominee for each office whose name was submitted by the Nominations Committee was declared elected. The new officers are: Chairman, R. A. Hackbusch; Vice Chairman, W. F. Choat; Secretary-Treasurer, G. E. Pipe.

The paper of the evening on "The Development and Work of the Approvals Laboratories" was presented by W. C. Cale, Approval Engineer of the Hydro-Electric Power Commission of Ontario.

The paper was preceded by the showing of a motion picture film of the various sections of the laboratory and some of the equipment in operation.

The author then proceeded with his paper discussing the foundation for the present system as laid down a number of years ago. It was found that the work of the Inspection Department could be facilitated if all equipment was inspected and approved before being offered for sale. A laboratory for inspection and approval work and the preparing of special specifications was then set up, and after some preliminary work had been done a Canadian electrical code was established and published. The scope of the code and some of its requirements were outlined. The speaker then explained the organization that administers the specifications, telling in detail how the work is carried out. He also discussed label service, field inspection reëxamination, service in plants, inspection and approval work as done in other provinces.

Messrs. Hackbusch, Nesbitt, Oxley, Patience, Pipe, and Price of the sixty-five members and guests in attendance participated in the discussion.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 7 *July, 1932*

TECHNICAL PAPERS

APPLICATION OF THE CLASS B AUDIO AMPLIFIER TO A-C OPERATED RECEIVERS*

By

Loy E. Barton

(RCA Victor Company, Inc., Camden, N. J.)

Summary—*The demand for increased power output from a radio receiver has reached a point where it is no longer economical to increase the output of the present class A output systems. The class B audio output system is a somewhat radical departure from the present system, and for a given cost permits an output of two to three times the power output of the present class A amplifier. This paper discusses the special circuit requirements of an a-c receiver to use the new RCA-46 class B tube successfully in a class B audio output system.*

Introduction

THE general trend in receiver design has been toward higher output power until at the present time two pentodes (247-type) in push-pull are used in a majority of receivers. The maximum power output of these sets is 6 to 7 watts with appreciable distortion as compared to 4 to 5 watts for the push-pull 245 output system previously used. The relatively small increase in power output of the pentodes represents only 10 to 15 per cent increase in sound output, which is not easily noticed by the average person. The result has been a receiver that has fallen short of the expectations the public had for the pentode tube.

The pentode operates as a class A amplifier and consumes approximately the maximum plate power that can be economically supplied by the average rectifier tube and filter system. The question of heat dissipation is another important factor as well as the power consumed from the house lighting circuit if class A amplifier systems are used for higher power outputs.

Although it is probable that the average listener does not use an average power greater than one-half watt, there are times when more power is desired and the peak power requirements for an average of one-half watt is probably above 6 watts. Therefore, a higher power

* Decimal classification: R363.2. Original manuscript received by the Institute, March 18, 1932. Presented before Twentieth Anniversary Convention, Pittsburgh, Pa., April 8, 1932.

output system is to be desired because of its ability to handle peak powers. A higher output power also permits refinements in the acoustic system for better fidelity at some expense of power, while an output system of the present type does not permit the waste of power. This effect is similar to a high power motor in an automobile which permits refinements in riding qualities and safety, but at greater expense. However, the class B audio amplifier has the advantages of a higher audio output at little increase in cost of construction, a lower power consumption from the lighting circuit, and less heat to be dissipated from the tubes and chassis parts.

The class B audio amplifier was first used commercially in a battery-operated receiver because independent sources of voltage could be easily obtained for bias and plate voltage. However, if the average three-element tubes were used in an a-c operated receiver for class B operation, the plate and grid would require a separate rectifier source because self-bias, as usually defined, cannot be used to reduce the plate current essentially to cut-off for this type of amplifier. This bias difficulty prevented an early application of the class B audio output system to an average receiver. Since a fixed bias cannot be supplied from a single rectified a-c supply, special tubes requiring no bias or very low bias must be used for the class B output amplifier. The satisfactory development of such a tube was a necessary step in the design of an a-c receiver having the class B output system.

The purpose of this paper is to present means for using the class B audio amplifier as the output system for an a-c operated receiver. A brief review of the use of the class B audio system in battery-operated receivers is given first to indicate the desirable features of the system and to indicate certain features that must be considered for a-c receiver application. The fundamentals of the class B amplifier were discussed in a previous paper[1] and will not be repeated here, except as related to the special tube for a-c operation.

Battery-Operated Class B Audio Amplifiers

The class B audio amplifier was first made available to the public in a battery-operated superheterodyne receiver by the company with which the writer is associated. This receiver was introduced in July, 1931. It is interesting to note that the above receiver has an output of about 1.25 watts and uses RCA-230 tubes in the output system. A set of heavy-duty B batteries and an air cell filament supply will give approximately one year of service for about 2 to 2½ hours daily

[1] Loy E. Barton, "High audio power from relatively small tubes," Proc. I.R.E., July, (1931).

use. It is also interesting to note that there is no pilot lamp on this receiver, because it would consume as much power as the filaments of the eight tubes in the receiver. The cost of battery power for the receiver is slightly less than two times the cost of power for the average a-c operated receiver, while the cost of power for a battery receiver having the same power output from a class A system would be about 8 to 10 times the cost of power to operate an average a-c receiver. The convenience of replacing batteries and long intervals between battery replacements are other very desirable features resulting from the class B audio output system.

The class B audio amplifier has also been successfully used in the automobile receiver and portable receiver recently announced by the above company and possesses the inherent advantage of the class B audio output system with respect to conservation of battery power and increased power output.

The performance of the class B amplifier in the battery-operated receivers indicates the two principal advantages of this type of amplifier, which are high output power from small tubes, and the conservation of plate power. To obtain the above results, from the ordinary tubes, it is necessary to have a well regulated plate supply voltage and a separate and well-regulated bias voltage. A battery source for the above voltages is ideal and presents no difficulty if the impedance of the batteries is low. To supply the voltages from a single rectifier to the average tube is impractical for normal operation and, in general, unsatisfactory.

Zero-Bias Class B Audio Tube

The simplest and most practical way of eliminating the bias difficulty is to use a tube which requires no bias for the class B audio output system. The first attempt to build a tube for the class B amplifier was more or less successful in limiting the plate current to a low value for class B operation, with zero bias on the grid. This was accomplished by using a grid with very fine mesh. The tube was successfully operated as a class B output tube, but the grid took a comparatively high current on positive grid swings and the peak plate current was comparatively low. This type of construction for a zero-bias tube required considerable power to drive the grids and the output was not as high as was desired. The experience with these samples of the first zero-bias tubes indicated that the commercial application of a class B output system to an a-c receiver depended to a very large extent upon the development of a satisfactory zero-bias tube because it was apparent that satisfactory plate voltage regulation could be obtained from rectified a-c supply.

The power pentode (247) was studied in an attempt to find a possible solution for a zero-bias class B tube. Since one of the requirements of a class B tube was to have essentially zero plate current for no signal and normal plate voltage, different combinations of screen-grid and control-grid voltages were tried. As the screen-grid voltage was decreased, a lower grid voltage was necessary to obtain the desired minimum plate current until at zero screen-grid voltage the control-grid voltage for the desired plate current was practically zero. This general characteristic was expected, but the desired plate current at zero bias

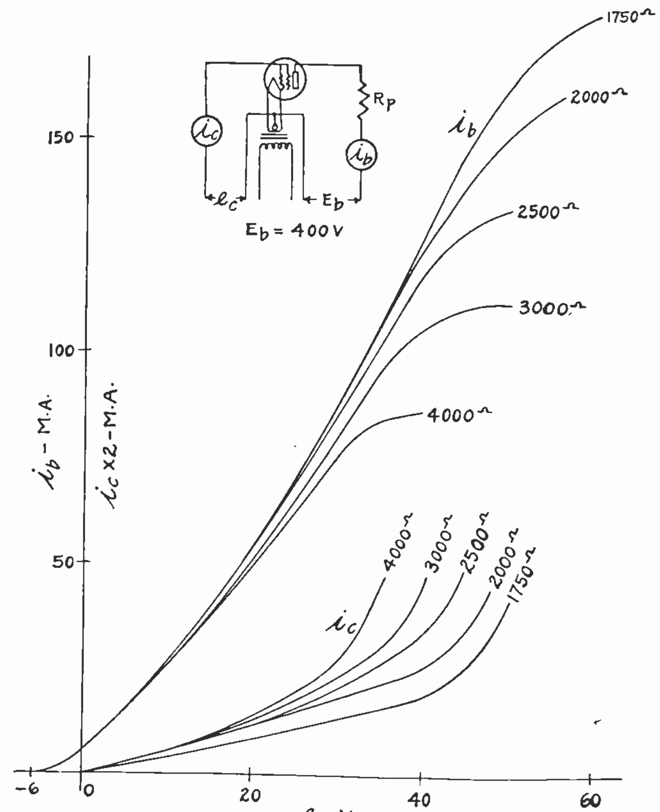

Fig. 1—Load characteristics of an RCA-46 for class B operation.

on the grid was more or less fortunate. The grid current to the control grid on positive swings was still high for good operation. The next step was to connect the screen grid and control grid together as a control grid. The combined grid currents for a given plate-current control was reduced materially. The power output of two pentodes (247) as class B audio amplifiers when the two grids are connected as a control grid is about 10 watts and is limited by comparatively low peak plate currents and by high grid currents on positive swings on the grid. The tube laboratories were advised of the results of the special operation of the pentode as a class B zero-bias tube as an aid in developing a satisfactory tube for class B operation. Two of the tubes that were developed easily deliver 10 watts or more of audio power and may be made to deliver more than 20 watts for auditorium amplifiers or other service where high audio output is desired. These tubes have two grids connected together as a control grid, which results in a three-element tube with a comparatively high amplification factor and high plate resistance, and are operated at zero bias. The tube is the new RCA-46 recently announced to the radio industry.

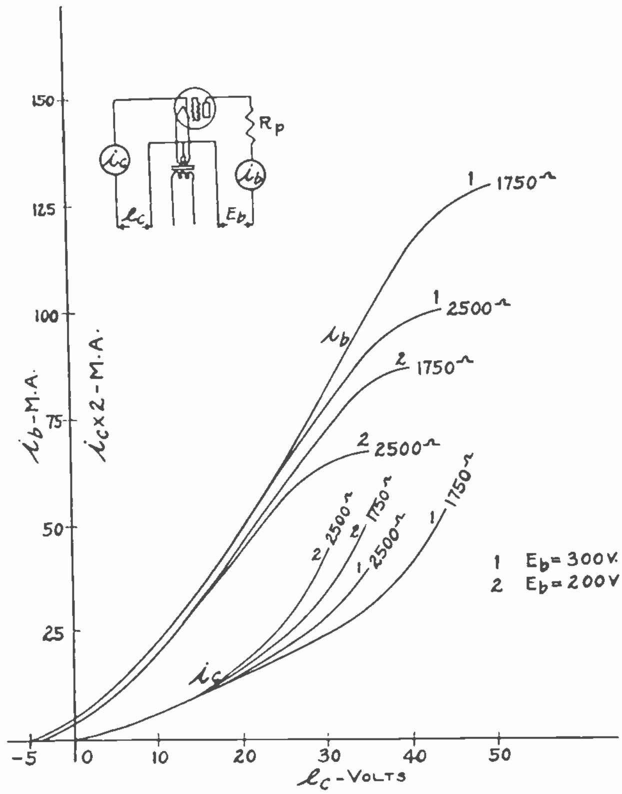

Fig. 2—Load characteristics of an RCA-46 for class B audio amplifier.

The characteristics of the RCA-46 that are particularly useful for the application of these tubes to a class B audio amplifier are given in Figs. 1, 2, and 3. The measurements for the curves in Fig. 1 were made under static conditions, but represent the dynamic characteristics of the RCA-46 that are particularly useful in calculations of the class B audio output for various load resistances and also indicate the grid input power requirements. The circuit indicated in Fig. 1 is the circuit used to obtain the data for the curves. A load resistance R_p is in series with the plate with the desired plate voltage applied at the end of the resistor. Since the normal plate current is zero, or essentially so, no correction is necessary for supply voltage. The grid voltage was varied and the corresponding plate and grid currents were read. The plate supply voltage for the curves in Fig. 1 was 400 volts, and the supply voltages for the curves in Fig. 2 were 200 and 300 volts.

The plate- and grid-current curves for the 2000-ohm load at 400 volts are plotted in Fig. 3. The upper curves represent one tube in a push-pull system such as V_1 in Fig. 4 and the lower curves represent V_2 in Fig. 4. If a signal is applied to the grids of V_1 and V_2 through T_1 in Fig. 4, V_1 will be positive when V_2 is negative. Referring to Fig. 3, the

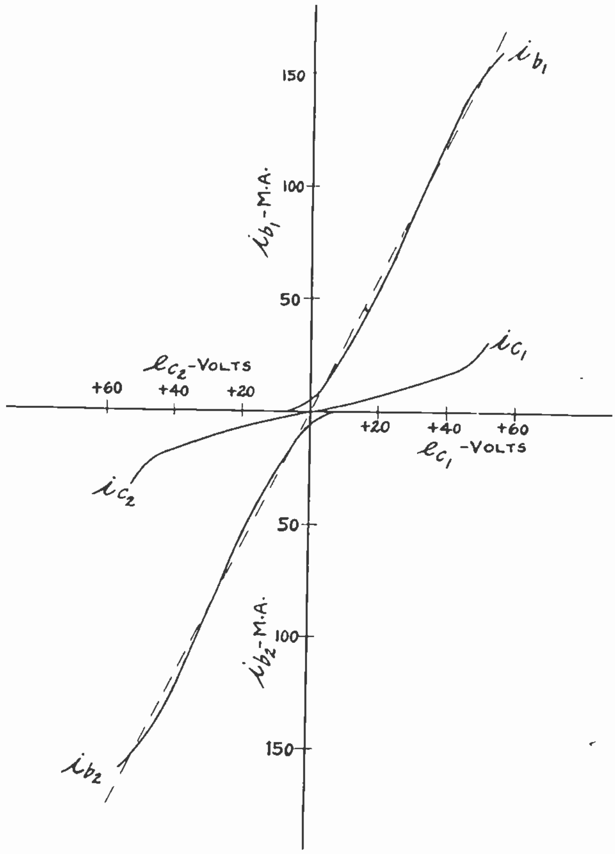

Fig. 3—Load characteristics of two RCA-46 tubes arranged for class B operation.

plate current will increase along the line i_{b_1} as the instantaneous grid voltage moves to the right along the zero plate-current axis. The instantaneous grid current will follow the grid-current curve i_{c_1} for positive swings on the grid for the tube represented by the upper curves and represented by V_1 in Fig. 4. During this period of the cycle V_2 is idle because its grid is beyond plate-current cut-off in the negative direction. As the signal through T_1 reverses for the other half cycle, the grid of V_2 swings positive and the plate and grid currents increase as was the case for V_1 in the previous half cycle but the grid current in T_1

has changed direction and the plate current for V_2 flows in the opposite direction in the other half of the output transformer primary. Therefore, the resultant output wave is similar in shape to the input voltage wave to the grids.

It will be noted that a mean straight line which conforms most closely to the upper and lower curves in Fig. 3 intersects the zero point for grid bias. A theoretically correct plate current curve is one in which all the values of plate current are on a straight line and the line intersects the bias line at zero. Another theoretically correct curve is one in which all values of plate current fall upon a straight line which if extended will pass through the zero bias point and the tailing-off portion of the curve near the zero plate-current axis conforms to a

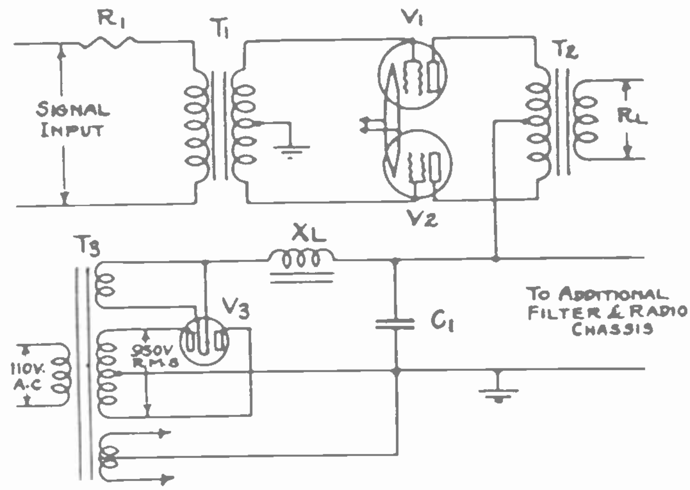

Fig. 4—Circuit for plate voltage supply and for class B audio amplifier.

square-law curve, which becomes tangent to the straight line at a point on the positive grid swing equal to the point of plate-current cut-off on the negative grid swing. Another way of expressing the latter condition is that the plate-current curves of the two tubes should conform to a square law over the portion of the grid swing during which both tubes contribute to the output because of the small plate current. If the lower portion of the plate-current curves conforms to a square law the sum of the currents contributed by the two tubes over this portion of the curve is a direct function of signal voltage. However, after one tube reaches plate-current cut-off on its negative grid swing, it is necessary that the plate current of the other tube be linear with respect to grid voltage from this point.

The plate-current curves in Fig. 3 are sufficiently close to the theoretical condition for practical applications. The question of distortion will be discussed later and actual distortion curves given to indicate the percentage of harmonics. It should be noted that the grid-current curves form an almost straight line to about 45 volts swing. This essentially straight line represents an essentially constant input resistance, which simplifies the problem of selecting the proper means for driving the grids of the class B audio amplifier to the point at which the curves rise sharply.

The instantaneous voltages and currents as related to load resistance for the amplifier may best be understood by referring to Fig. 5.

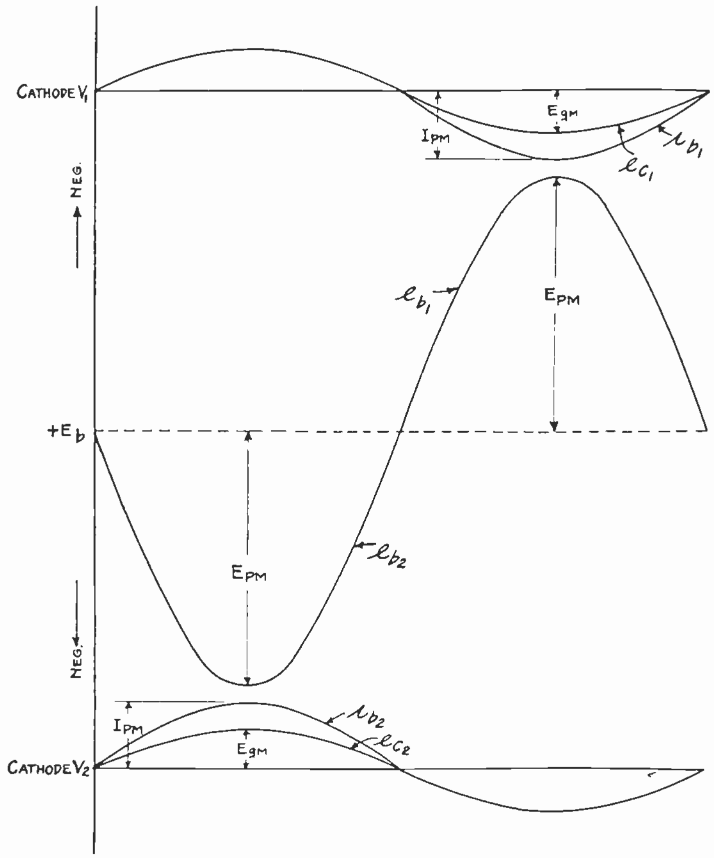

Fig. 5—Diagram of instantaneous voltages and currents for a class B audio amplifier.

The plate-voltage line is used as a reference in order that the plate currents for the tubes V_1 and V_2 as represented in Fig. 4 may be shown reversed in phase. The reversal of phase for the plate currents takes place in the primary of the output transformer. If the tube V_2 in Fig. 5 is considered first, it will be noted that the grid voltage swings in a positive direction and that the signal or grid voltage for the other half of the input transformer swings in a negative direction on the grid of tube V_1. The plate current for V_2 is i_{b_2} and the plate current for V_1 is zero for the first half cycle. The plate current i_{b_2} causes the plate potential to decrease on V_2 as indicated by e_{b_2} because this current passes through the equivalent load resistance. The curve also indicates the plate voltage increase on the plate of V_1. The instantaneous plate voltage below the supply voltage for V_2 and above the supply voltage for V_1 is given by the product of the instantaneous i_{b_2} and the equivalent load resistance in series with the plate of V_2.

During the next half cycle, the grid of V_1 swings positive and plate current flows according to i_{b_1} while the plate current for V_2 is zero because its grid is negative. Note that the instantaneous plate potential for V_1 now decreases with increase of i_{b_1} while the instantaneous plate potential for V_2 increases. If the two tubes have essentially linear and similar characteristics and have the same equivalent load resistance in series with each plate, the two parts of plate voltage curves e_{b_1} and e_{b_2} will combine so that the resultant wave will be similar to the grid input voltage curve and will appear across one side of the output transformer. Since the value of E_{pm} is a function of I_{pm} and load resistance, the output power may be easily calculated for sinusoidal waves by the relation:

$$\frac{I_{pm}E_{pm}}{2} = \text{power output} \tag{1}$$

or,

$$\frac{I_{pm}^2R_p}{2} = \text{power output} \tag{2}$$

in which,

I_{pm} = maximum I_p

E_{pm} = maximum decrease in plate volts

R_p = load resistance in series with one plate.

The plate power input is equal to the product of the average plate current and plate supply voltage which is given by the following expression:

$$0.637\ I_{pm}E_b = \text{power input} \tag{3}$$

in which,

E_b = d-c plate supply voltage

I_{pm} = maximum plate current.

The above relation is true as long as a linear relation exists between grid voltage and plate current. Referring to Fig. 1, it is easy to determine when the plate-current curve deviates appreciably from a straight line at the upper end. If the upper limit of plate current is taken as indicated in the table below, for the various load resistances, the power output may be calculated by using (2), and power input by using (3). The plate dissipation per tube is the difference between power input and power output divided by two. The above power relations in watts are approximately as shown in Table I.

TABLE I

R_L	I_{pm}	Power Output	Power Input	Plate Loss per Tube	Plate Eff.
1750	170	25.2	43.2	9.0	58.0
2000	150	22.5	38.3	7.9	58.5
2500	125	19.5	31.8	6.1	61.2
3000	104	16.2	26.5	5.1	61.5
4000	80	12.8	20.4	3.8	62.5

The plate loss in the above cases is the plate dissipation at maximum output and in general does not exceed this value at lower outputs for the same load resistances. However, if the no-signal plate current is somewhat high, the plate dissipation may be somewhat higher at lower outputs. The no-signal plate dissipation for the curves in Fig. 1 is approximately 2.0 watts per tube. The lower input power to the output tubes and low plate dissipation for a given power output is easily appreciated from the table above.

To obtain the above power outputs, it is necessary to supply essentially an undistorted input voltage swing on the grids to a point at which the plate-current curve begins to deviate appreciably from a straight line at the peak values of plate current. Referring to Fig. 1, if maximum plate current for the 2000-ohm load curves is to be obtained, a grid swing of about 48 volts is required. The grid-current curve for this swing has deviated considerably from a straight line and at the upper end represents a resistance of about 600 ohms. If the input voltage is to have little distortion, the effective resistance in series with each grid should be about 100 ohms or less. This is obtained by a step-down impedance ratio in T_1, Fig. 4, from the effective resistance R_1 in series with the primary to one side of the secondary. Two 245's in push-pull with a plate voltage of 225 volts and biased to about 20 ma plate current will successfully drive the class B tubes to plate current limitations if a properly designed coupling transformer is used having a step-down turn ratio of 6- or 7-to-one or an impedance ratio of about 40-to-1 from total primary to each side of the secondary. If greater distortion is permitted, the plate voltage on the 245's may be reduced and a lower step-down input transformer may be used.

If full output is desired as indicated in the above table, special precautions must be taken to prevent serious distortion and in general, such high powers are not desired from a receiver. The portion of the grid-current curves below the point at which the grid current takes a sharp rise is fairly straight, but the slope varies for different load resistances. Therefore, it is necessary still to have good regulation of the input voltages because of varying loads due to changes in frequencies, but a driver system for about 10 watts output can be obtained very economically.

A peak power output of about 10 watts for a 2000-ohm load requires a grid swing of about 35 volts and the approximate minimum input resistance to the grid of the output tube is about 1500 ohms. The equivalent resistance in series with the grid in this case may be about 300 ohms. If two 227 tubes are used in push-pull as drivers with maximum plate voltage and plate current, a step-down transformer may be used having a primary to each side of secondary turn ratio of about 8-to-1 step-down. It will be seen that the proper design of the input system depends upon the particular driver tubes and upon the desired power output. If a limited output of 4 to 5 watts is desired, a lower plate voltage may be used and the driver system designed accordingly. The curves for lower plate voltages are given in Fig. 2. However, in general, power outputs under 5 watts are more economically obtained from a class A amplifier system unless it is desired to take advantage of the high efficiency of the class B system to reduce heating and size of parts.

The Class B Audio Amplifier Plate-Voltage Supply

A circuit for the rectifier and filter system for plate-voltage supply to the class B audio amplifier is given in Fig. 4. The rectifier tube was developed primarily to supply plate voltage to two RCA-46 tubes in push-pull as class B audio amplifiers and is known as the RCA-82 indicated as V_3. The mercury rectifier tube is used because of its inherent characteristic of about 15 volts internal drop for all load currents. The resistance of the high voltage winding of T_3 should be as low as possible, but in general, a commercial design to prevent overheating on full load is satisfactory. The filter reactor X_L should have a low resistance to aid in plate-voltage regulation and a saturating characteristic is desired which further aids the plate-voltage regulation when plate currents to the output tubes vary in response to the input signal. The condenser C_1 is 4 to 10 microfarads, which, in general, is sufficiently large to filter the pulsating currents to the output tubes effectively, and also to reduce the ripple in the plate supply. Considerable ripple can be tolerated to the plates of the output tubes because of the balanced relation of the tubes and also because of the high plate impedance of the tubes. These characteristics of the amplifier permit the simple filter system as indicated with a regulation of the output voltage of 10 per cent or better. In general, it is desirable to put the reactor in the high potential side of the filter system to help reduce the radio-frequency disturbance caused by the mercury rectifier.

The above rectifier circuit may be used for a receiver having 10 to 12 watts output and additional filtering from C_1 may consist of a 200-volt loud speaker field for a filter reactor and another condenser similar to C_1. The plate supply for the radio chassis and driver stage would then be about 200 volts. A bleeder resistance across the chassis supply voltage should be used to increase the current through the speaker field to the desired value. A higher voltage may be had for the chassis by decreasing the loud speaker field resistance with a corresponding increase in field current. It is not advisable to attempt to supply the plate current to the output tubes through a speaker field because of the resultant poor regulation of the plate voltage to the output tubes and to the radio chassis. If it is desired to use a common filament winding, the loud speaker field should be placed in the positive lead.

A poorly regulated plate supply does not affect the output tubes appreciably except that the output power will be limited by the plate voltage at maximum current drain from the plate-voltage supply. Another very important reason for having good regulation of the plate supply voltage is that it may be necessary to supply essentially constant voltage to the rest of the radio receiver. Certain types of circuits for the receiver tend to flutter if the plate voltage varies with signal to the small degree experienced with the above rectifier system so that it is necessary to design the receiver in such a manner that small variations in plate voltage will not affect the receiver performance.

It is evident from the above discussion of the class B audio output system that the associated circuits should be designed especially to meet the more or less severe requirements of the input signal voltage to the output tubes and the plate-current variations with signal.

Distortion in the Class B Audio Amplifier

The calculated power outputs of the class B audio amplifier from data in Fig. 1 and tabulated in Table I, does not consider the efficiency of the output transformer, but, in general, actual net outputs for approximately 5 per cent third harmonic distortion is very close to the outputs calculated from the curves in Figs. 1 and 2 for coupling transformers of average efficiency. As the upper limit of the plate-current curves in Figs. 1 and 2 are approached, the third harmonic increases very rapdily if the grid swing causes the plate-current curve to flatten appreciably. This flattening of the plate-current curve is the limit of the power capability of the tube for the particular plate voltage and loads used in obtaining data for the curves. The variation in slope of the plate-current curve through the 10- to 40-volt grid swing for the 2000-ohm curve in Fig. 1 causes relatively little distortion and, in general, is not objectionable. However, the lower end of the plate-current curves cause distortion in the class B audio amplifier, which, in some cases, becomes objectionable. At some low point for the positive grid swing, on one tube, the plate current for the opposite tube begins to flow and the power output is the sum of the powers delivered by the two tubes from this point to a point beyond the zero bias line at which the plate current for the first tube becomes zero. As pointed out above, if the plate-current curve on each tube follows a square law over this portion of grid swing, there will be no distortion. However, the lower end of the plate-current curve is not usually ideal so that some distortion results.

If the maximum grid swing for a sine wave input is limited to the region in which both tubes supply power, the harmonics generated will be of the lower order. If the grid swing increases beyond this point, one tube will be idle over a portion of the cycle while the other tube supplies the output power, which results in a shorter interval of time in the transferring of signal from one tube to the other when the grid signal swings through the zero axis. This shorter period of transfer from one tube to the other over the distortion area results in harmonics or transients of the higher order. When the grid swing is at a maximum, this period becomes a minimum and the harmonics are of a still higher order. These higher order harmonics or transients may tend to produce a sound similar to a buzz, which, in general, is not noticeable, except on sustained pure tones and may be eliminated entirely. The measurements of harmonics that indicate the above condition are given in Fig. 6. These measurements were made on an amplifier especially designed for approximately maximum output, in which two 245's were used to drive two RCA-46 output tubes through a step-down transformer having a ratio of 6 to 1 from primary to one side of secondary. The circuit for the rectifier system was similar to the circuit in Fig. 4. The load on the output tubes was approximately 1900 ohms per tube. It should be noted that the load on each tube during the time it functions is one-fourth of the equivalent load from plate to plate of the output tubes.

The curves in Fig. 6, indicate clearly that the disturbance at the lower end of the plate-current curve causes a peak in third harmonic distortion at low grid swings and that as the amplitude of grid swing increases, the higher order harmonics peak at somewhat higher output levels as anticipated. It should be noted that the second harmonic is quite low and is the result of unbalancing in some part of the output system. The harmonic distortion is a minimum at about 4 watts output and the third harmonic increases gradually until the limit of output is reached at which it rises quite rapidly. This is also true of the other order harmonics except that their amplitudes are much lower.

The curves shown in Fig. 6 indicate how accurately the power output may be calculated from the curves in Fig. 1 if the values of plate currents are properly chosen. It is probable that the distortion in the above amplifier is not objectionable at any power level below about 25 watts. It should be noted that the above sample amplifier requires a maximum plate current to the output tubes of about 125 ma at 400

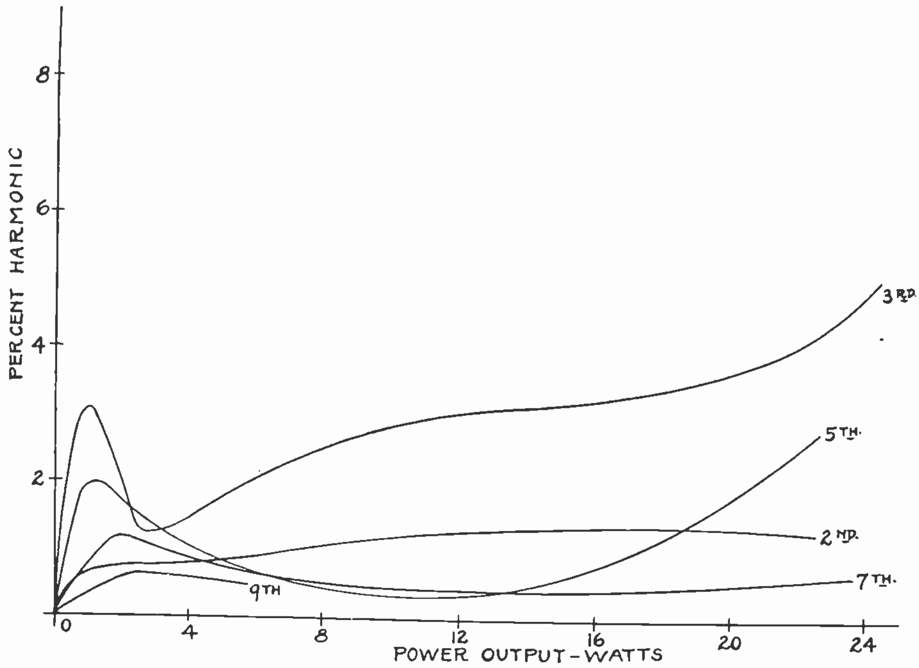

Fig. 6—Distortion curves for an a-c operated amplifier using two RCA-46 tubes and designed for approximately maximum output.

volts and that high power driver tubes are necessary. Therefore, the increased cost to meet the special requirements to obtain maximum output is probably not justified except for an amplifier in which the maximum power is needed, such as auditorium amplifiers and amplifiers for modulation purposes. For lower outputs of approximately 10 watts, the special requirements of the class B amplifier can be economically met for an a-c receiver. Such a system would use 227's or equivalent for drivers, and the maximum plate current to the output tubes is 70 ma or less.

The curves in Fig. 6 were taken with a resistance load and will be altered considerably if a loud speaker is substituted for the resistance load. If an average loud speaker is used, in which the impedance increases with frequency and no precautions are taken to limit the impedance, the harmonics as indicated in Fig. 6 will increase considerably. In general, the increase in percentage harmonics will be a function of the impedance of the speaker at the fundamental frequency and the frequency of the harmonics. For example, the average loud speaker has a minimum impedance at approximately 200 cycles. If harmonic measurements are made with a fundamental frequency of 200 cycles and using the speaker as the load, the third harmonic will be increased because of the increase in speaker impedance at 600 cycles and the fifth seventh, ninth, and higher order harmonics will be increased in proportion to the speaker impedance at the particular frequency of the harmonic. The result of this higher transient or harmonic voltage across the speaker coil is a buzz effect, which is usually noticeable if constant tones are used and in some cases, is noticeable on speech or music. This effect can be reduced materially if some filter scheme is used; for example, keeping the load impedance nearly constant or decreasing the impedance as the frequency increases. If a decrease in impedance characteristic is used, a rising frequency characteristic may be used in the driver stage to compensate for the change in frequency characteristic of the output system.

The above frequency characteristic of the class B output amplifier for a load impedance that varies with frequency is due to the high internal plate resistance of the tubes. This impedance is of the order of 50,000 to 100,000 ohms, depending upon the positive grid voltage at which the plate resistance is measured. The high plate resistance of the output tubes results in practically constant current to the load impedance so that the voltage across the speaker is approximately proportional to its impedance in combination with any system that may be used to maintain a constant impedance load for the output amplifier. If no means is used to maintain a constant load impedance, the high-frequency response of the average speaker is usually higher than is desired. Considerable work has been done to develop a suitable speaker for the class B system. It is probable that a continued use of high impedance output systems such as discussed, and which also includes the pentode for output systems, will result in the development of loud speakers having a constant impedance or requiring a constant current throughout the frequency range for best fidelity. No compensation for loud speaker impedance changes would be necessary for such loud speakers.

Conclusions

It is apparent from the above precautions and special requirements for the class B output system for a radio receiver that the RCA-46 tube cannot be used to an advantage unless the receiver is designed to work with such a system.

The voltage regulation of the plate supply is important because it affects the maximum output of the class B amplifier and affects the driver stage and the radio-frequency system. The regulation of the plate supply is improved to a satisfactory degree by using a mercury rectifier and a special filter system. The rest of the receiver can be designed so that slight fluctuations of plate supply will not affect its operation.

The driver stage for the output tubes must have a low effective resistance in series with the grids of the class B tubes, so that varying grid currents may be supplied without appreciable variation of the signal wave. This is accomplished by combining the proper driver tube or tubes and the proper transformer to couple to the grid of the output tubes.

The output transformer is such that the tubes work into the proper load resistance for the particular power desired and for a given driver system and plate voltage. Precautions are also necessary to compensate for the speaker impedance changes due to frequency variation so that the load impedance is more nearly constant or as needed for best fidelity from the particular speaker used.

The class B audio output system has the advantage of high power output from relatively small tubes with the associated reduction in size of parts and decreased heat to be dissipated. These advantages make the system very attractive and offset the special requirements in receiver design. By using this output system at little or no increase in cost for a receiver the public can be offered a receiver with greater output power capability, smaller space requirements and greater fidelity because audio power can be consumed by compensating means to improve the fidelity. These desirable features can be had with a lower power consumption from the house circuit than is required for the average receiver having considerably lower audio output power.

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DESIGN OF RESISTORS FOR PRECISE HIGH-FREQUENCY MEASUREMENTS*

By

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Summary—*New shielded and unshielded resistance boxes and fixed standards of resistance for use in precise a-c measurements, are described in detail and numerical values are given for the residual inductance or capacitance of the individual coils and of the boxes at various settings, and for the resistance error at 1 and 50 kc.*

A new coil construction and two new types of decades are used. In one of the resistance boxes, for any setting of the dials, only one coil of each decade is in the circuit, while the idle coils are completely disconnected, and in addition the configuration of the circuit inside the box remains constant for all settings of the dials.

Introduction

Recent developments in the design of resistance boxes, including both the arrangement of the decades and the construction of the individual resistance units, has resulted in instruments whose time constants for the complete resistance box, over a wide range of settings, are less than 10^{-7} second.

A resistance box for use with alternating current must satisfy all the requirements for d-c boxes in regard to temperature coefficient, accuracy of adjustment and the like and in addition, (1) the resistance to alternating current should be the same as to direct current, and (2) the reactance component should be zero.

The method of stating the performance in these two respects requires a brief explanation. The reactance characteristic of a resistor is commonly, and conveniently, expressed in terms of its time constant. If the resistor is inductive and is thought of as an inductance in series with a resistance, the time constant is L/R; if the coil is capacitive and is considered to be a capacitance in shunt with a resistance, the time constant is $-CR$.[1] The time constant (T) is simply related to the phase angle for $2\pi fT$ is equal to the tangent of the angle between the voltage applied to the resistor and the current through it. If the current is considered as the reference vector, a positive value for the tangent indicates that the voltage leads the current.

It is also common to describe a resistor by the numerical values of the shunting capacitance or series inductance mentioned above. A

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[1] This convention in regard to the negative sign for CR is not universal.

10,000-ohm coil, for example, may be described as having a residual capacitance of 0.8 $\mu\mu$f or a time constant of $-(0.8\times10^{-12})$ (10,000) $=-0.8\times10^{-8}$ second. It is often convenient to refer to inductance as a negative capacitance and vice versa, the numerical values being so chosen as to result in the correct value for the time constant. Thus the 0.8 $\mu\mu$f of the 10,000-ohm coil is equivalent to -80 μh for $-(0.8\times10^{-12})$ $(10{,}000)=(-80\times10^{-6})/10{,}000$.

The resistance error is expressed in per cent and is equal to 100 $(R_{dc}-R_{ac})/R_{dc}$ where R_{dc} is the resistance as measured with direct current and R_{ac} is the resistance component as measured with alternating current of the frequency under consideration.

The effective inductance and capacitance of a shielded resistance box depend upon how the box is used. Two sets of characteristics are given below corresponding to the two most common ways of using a resistance box. The curves labeled "box grounded" apply when the shield of the box is connected directly to that terminal of the box at which the lower decades terminate. The curves marked "direct resistance characteristic" correspond to the case when the shield is maintained at the same potential as one of the terminals of the resistor but is not actually connected to either of the terminals.[2] The former method of connecting the shield is a convenient one in most uses of a resistance box, the latter connection is of most common occurrence in a bridge with a Wagner earth connection where the shield is connected to ground, while a terminal of the box is also at ground potential but is not connected to ground.

Unshielded Resistance Box

A four-dial unshielded resistance box incorporating a new decade and coil construction is shown in Fig. 1. Fig. 2 is a plan view of the decades with each decade at a different setting between 0 and 3. There are six equal coils in each decade and the resistance of each coil is twice the decade step. Five of these coils are mounted on the rotor and are joined in series, with the junction points of the coils and the terminals of the first and last coils connected to switch segments on the rotor. One of the terminals is also connected to the central stud on the rotor. The sixth coil is stationary and is connected to two brushes which bear on the rotor switch segments. For even-valued sett ngs of the dial, both brushes contact with the same switch segment; for odd settings the brushes are in contact with different but adjacent segments so that the

[2] For an interesting and brief discussion of the difference between these two methods of using a resistance box, *see* Hartshorn and Wilmotte, *Jour. Sci. Inst.*, vol. 4, p. 33, (1926). For the general concept of direct resistance and direct impedance *see also* G. A. Campbell, *Bell Sys. Tech. Jour.*, vol. 1, p. 18, (1922).

stationary coil is in shunt with the rotor coil connected to those two segments. Thus the operation is such that the even-valued resistance settings are obtained by including in the circuit an appropriate num-

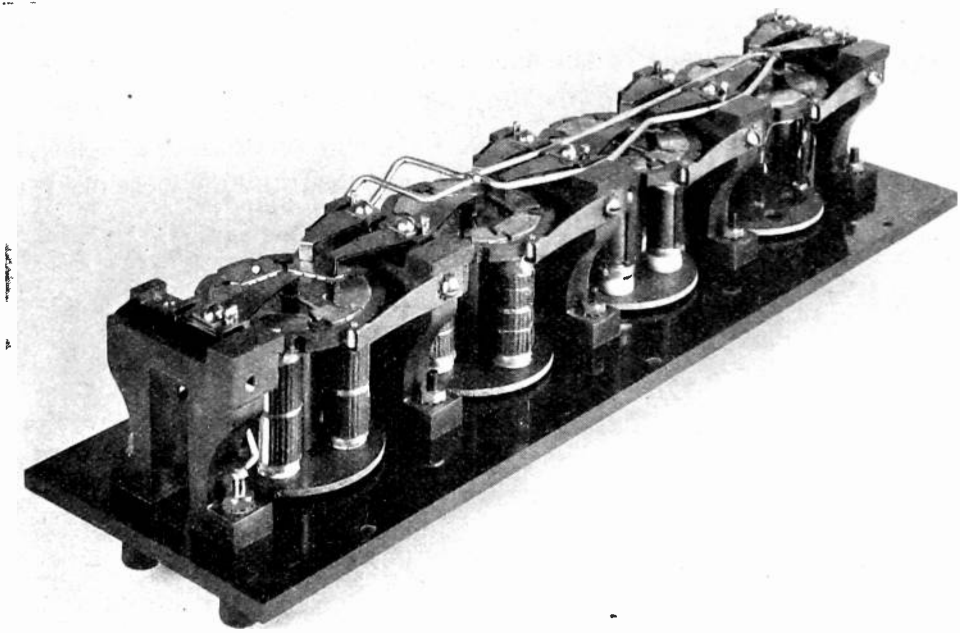


Fig. 1—Four-dial resistance box.

ber of the rotor coils in series; for the odd values the stationary coil is in parallel with one of the coils on the rotor.

The decade can be set in unit steps from 0 to 10 and, unlike previous decades with fewer coils than steps, operation of the dial changes

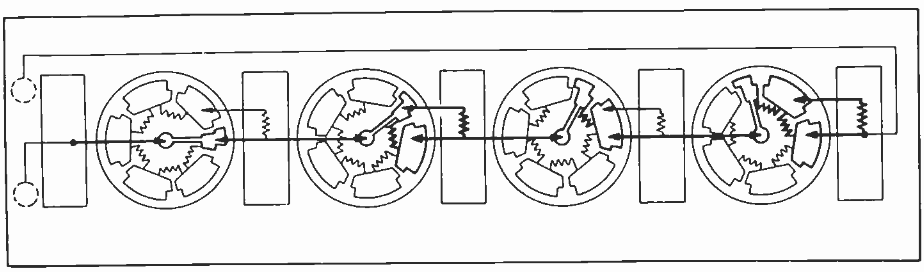


Fig. 2—Four-dial resistance box. Box set for 123. ohms 1000-, 100-, 10-, and 1-ohm decades. 1000-ohm decade at left.

the resistance directly from one value to the next, at the even-valued settings only one of the two brushes connected to the sixth coil is effectively in the circuit, for the odd settings the brushes are in parallel branches so that contact resistance effects due to these two brushes are at most equal to the contact resistance of one brush.

A compact construction is possible with a minimum of connections. The length of the brushes is not limited by the size of the circle on which the switch segments are located. The rotor shaft is the only part of the decade which projects out of the resistance box and this shaft is not in contact with any of the coils of the decade.

The a-c performace of the box is given by the curves of Fig. 3. Each curve applies to one decade and gives the time constants, as measured at the terminals of the box, for every setting of that decade and with all the other decades at zero. The measurements were made at

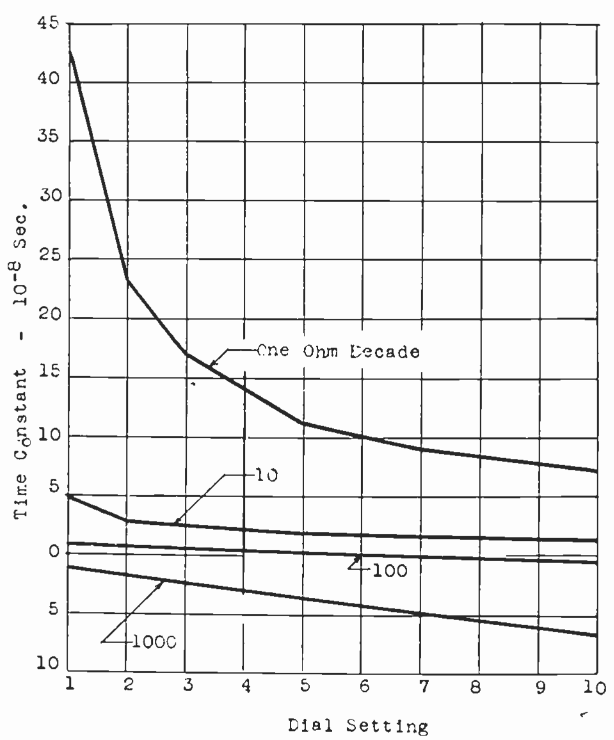


Fig. 3—Characteristics of four-dial unshielded box.

50 kc on six boxes. For any setting, the curve corresponds to the box having the largest time constant for that setting. The differences between the boxes are indicated by the results for the 1000-ohm dial; at the 1000-ohm setting the time constants of the six boxes are included in the limits -0.9_9 to $-1.1_4 \times 10^{-8}$ second and at the 10,000-ohm setting the range is -6.6 to -6.9×10^{-8} second. The change of time constant with frequency is small. For a particular box, typical of them all, the time constants for 10,000 ohms are -7.0_5 and $-6.7_5 \times 10^{-8}$ at 1 and 50 kc, respectively, and the corresponding values for 1000 ohms are -1.0_6 and $-1.0_3 \times 10^{-8}$.

The resistance error as defined above,[3] at 50 kc is $+0.02_7$ per cent for 10,000 ohms, $+0.00_4$ per cent for 1000 ohms and $+0.00_7$ per cent for 10×100 ohms. Measurements on four boxes agreed to one unit in the last place. At 1 kc the resistance error is too small to measure with accuracy; it is certainly much less than 0.01 per cent.

The curve for the 1-ohm decade applies when that decade is used by itself. When any other decade is used with the 1-ohm decade, the

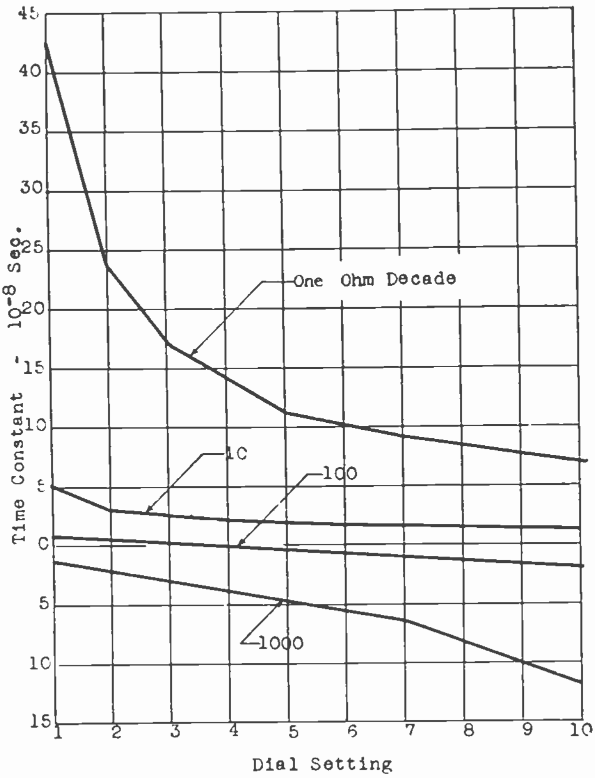


Fig. 4—Characteristics of four-dial shielded box. Shield connected to low terminal.

inductance of the circuit, as computed from the time constant curve for the higher decade, is increased by 0.03_5 microhenry per ohm of the 1-ohm decade.

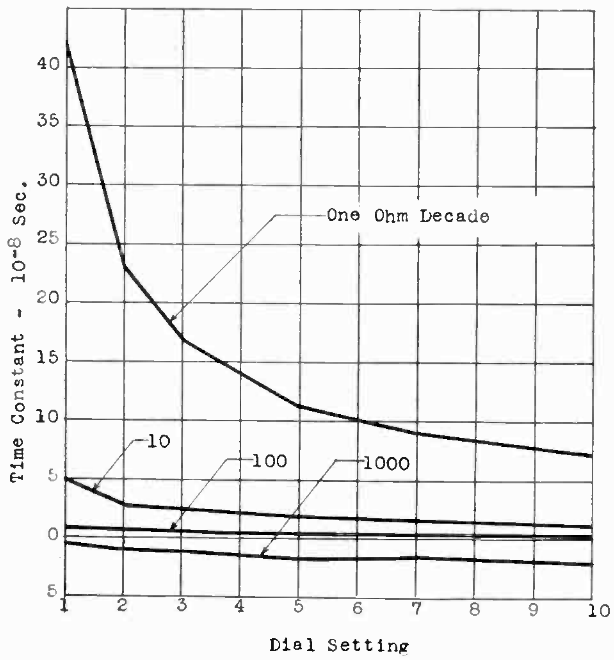
Four-Dial Shielded Box

The unit decade construction described for the unshielded box can also be used in a shielded box. Such a box is generally similar to the unshielded one except that the top panel and the case are made of metal and a third binding post is provided for making connection to the shield.

As previously mentioned, the reactance characteristics of such a box

[3] This refers to the difference between the a-c and d-c resistance values and should not be confused with the tolerance in the adjustment of the d-c resistance values.

depend on how the box is used and two sets of curves are given which correspond to the most common methods of use. Fig. 4 applies when the shield is joined to the terminal which is connected to the 1-ohm decade. Fig. 5 gives the "direct resistance" characteristic and refers to the use of the box with the shield at the same potential as either one of the terminals of the resistor but not actually connected to the resistor. This is the condition that exists in the use of a bridge with a Wagner earth connection.



[Fig. 5—Direct resistance characteristics of four-dial shielded box.

The resistance error, change of time constant with frequency, and uniformity among different boxes are substantially the same for the shielded as for the unshielded box.

Six-Decade Shielded Box

The resistance boxes described are capable of improvement in several respects:

(1) The unused coils, particularly of the higher decades, remain connected to the box circuit and add to the capacitance.

(2) Rotation of the dials changes the geometrical configuration of the loop joining the resistance coils and thus changes the inductance with change in setting.

(3) In general, for any setting of a decade, more than one resistance coil is in the circuit and this usually increases the inductance or capacitance.

The six-decade resistance box illustrated in Fig. 6 eliminates these objections as shown schematically for two decades in Fig. 7. Each decade has two stationary brushes and a rotor which carries ten coils and

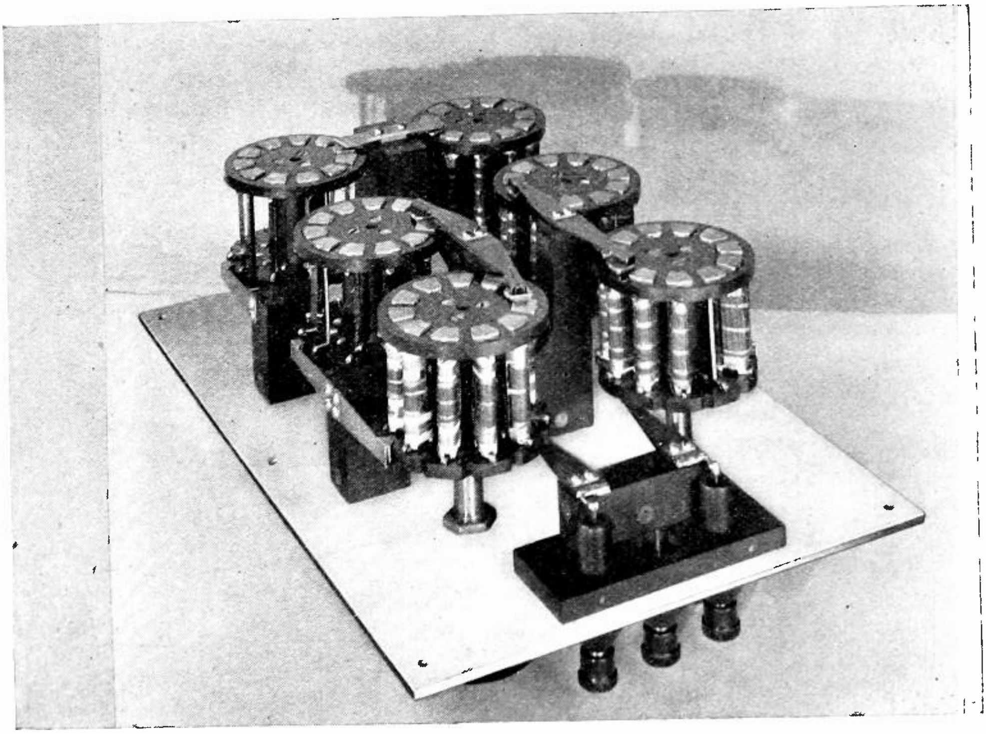


Fig. 6—Six-dial shielded box.

a short-circuiting strip. The coils all have different values, there being one of one unit, one of two units, and so on up to ten. Each coil and the short-circuiting strip terminates in two switch segments but otherwise

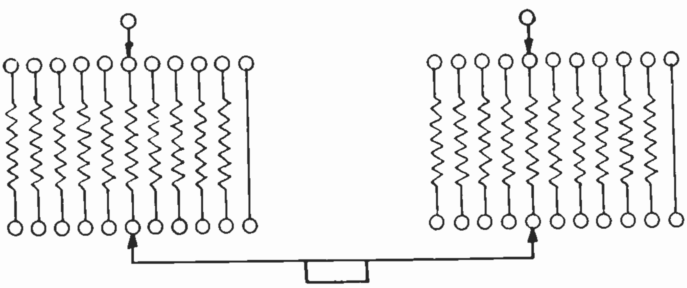


Fig. 7—Decade construction six-dial shielded box.

has no connection to the circuit. Operation of the dial turns the rotor so as to bring the appropriate two switch segments in contact with the brushes and include in the circuit any *one* of the coils or the short-circuiting strip. For any dial setting therefore, only one of the coils of a decade is used; the active coil always occupies the same position and each coil is used for only one setting of the decade.

This box is subject to an inconvenience, which, in practice, has proved to be a minor one. When a dial is being changed from one setting

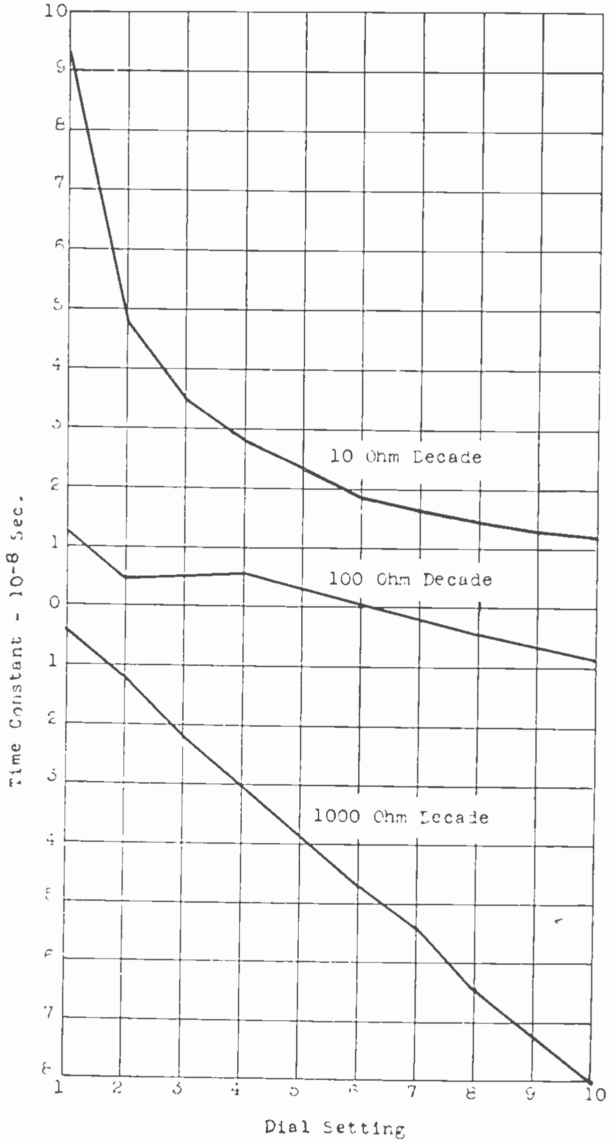


Fig. 8—Characteristics of six-dial shielded box. Shield connected to low terminal.

to the next, the two neighboring coils are momentarily connected in parallel. Thus, while rotating the 1-ohm dial from 3 to 4, the two coils are in parallel for a short time and the decade resistance is about 1.7 ohms.

Fig. 8 and Fig. 9 give the time constants for the two connections of the shield as discussed above. Each curve gives the time constants for various settings of the indicated decade, as measured at the termi-

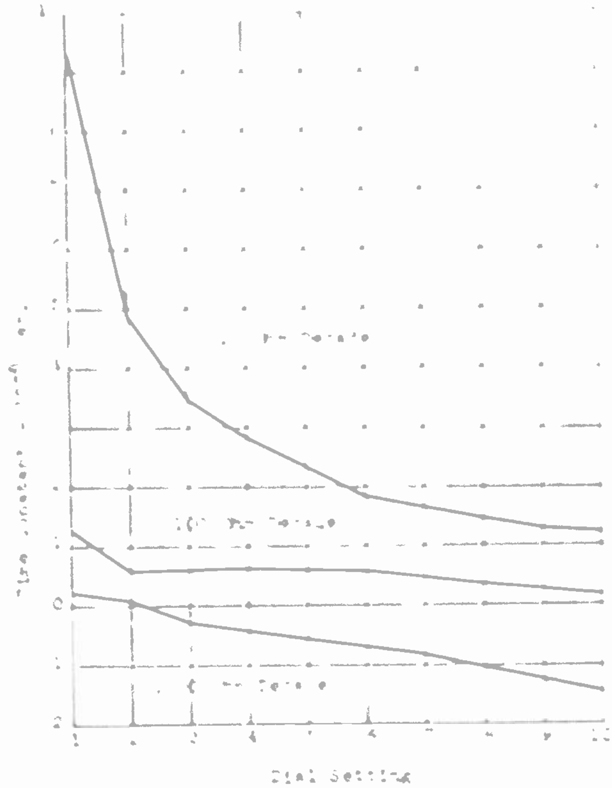


Fig. 9—Direct resistance characteristics of six-dial shielded box.

nals of the resistance box and with the other decades set at zero. Only the time constant curves for the 1000-, 100-, and 10-ohm decades appear there; the characteristics of the lower decades are presented in

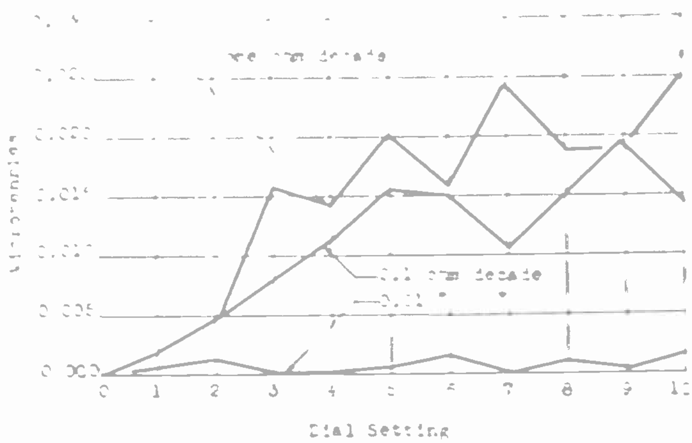


Fig. 10—Change of inductance with dial setting for six-dial shielded box.

pear there; the characteristics of the lower decades are presented in Fig. 10, where inductance is plotted against dial setting. The inductance referred to is $L-L_0$, where L is the inductance as measured at the terminals of the resistance box for the appropriate setting of the decade under consideration and with the other decades at zero, and L_0 is the inductance between the terminals of the box with all the decades at zero. These inductance curves apply to either method of shielding. Where the three lower decades are used with any one of the three upper decades, the total inductance of the box is the inductance of the upper decade, as computed from its time constant of Fig. 8 or Fig. 9, plus the sum of the inductances of the three lower decades as read from Fig. 10. The inductance of the upper decade may be positive or negative as already discussed.

With the box connected to the shield, the time constant for the 10,000-ohm setting is $-8.0_2\times10^{-8}$ at 1 kc and $-7.9_4\times10^{-8}$ at 50 kc. For 1000 ohms the measured time constants were identical for the two frequencies, -4.3×10^{-9}. The resistance error[4], at 50 kc, is $+0.01_0$ per cent for 10,000 ohms and less than $+0.00_1$ per cent for 1000 ohms.

Resistance Standards

The characteristics of single-valued a-c resistance standards of values between 1000 and 20,000 ohms are listed in Table I. These standards consist of resistors, woven as described below, mounted in an insulating case and provided with petticoated rubber terminals. Their

TABLE I
Time Constants of Resistance Standards

Resistance	*Time Constant*
1,000 ohms	$+0.06\times10^{-8}$
2,000	-0.08
3,000	-0.3
4,000	-0.6
5,000	-0.8
6,000	-1.0
7,000	-1.2
8,000	-1.4
9,000	-1.5
10,000	-1.9
20,000	-4.0

change in time constant between 1 and 50 kc is less than one unit in the last given figure. At 50 kc the resistance error for the 10,000-ohm standard is $+0.00_3$ per cent and for the 1000-ohm unit is less than $+0.00_1$ per cent.

Resistance Units

Both the inductance and capacitance of coils as used in the above instruments, are kept small by the use of a minimum of wire. Wire of

[4] This refers to the difference between the a-c and d-c resistance values and should not be confused with the tolerance in the adjustment of the d-c resistance values.

high resistivity, small cross section, and correspondingly short length for a given resistance, is therefore desirable, but what can be done along these lines is fairly definitely limited by the resistivity of the material available, surface needed for heat dissipation, desired stability of resistance with time, convenience of manipulation, etc. The character and size of the wire being thus determined there remain as controllable factors the shape of the cross section and the arrangement of the wire.

The inductance of a coil is kept low by so winding the coil that adjacent lengths of wire carry equal and opposite currents. The ideal condition is to have the go and return currents superimposed, and this condition is approached by reducing to a minimum the mean distance between the filaments of current flow in the two directions. For low and medium resistance coils, the mean distance may be made small by using two or more wires in parallel and interleaving the go and return paths, and by using wire in the form of a flat tape. The capacitance of the coil as a whole is minimized by making the capacitance between its parts small and so arranging this distributed capacitance that it is smallest between parts of the coil which have the largest difference of potential.

The requirements for minimum inductance and for minimum capacitance are in general conflicting. Their relative importance depends upon the resistance of the coil. When both inductance and capacitance are present, the time constant is approximately $L/R - CR$.[5] In the case of low resistances, the inductance term, with R in the denominator, is the dominating factor, while for high resistances the capacitance term, with R as a multiplier, is the larger.

A consideration of the numerical values is helpful. We are concerned with coils having time constants of the order of $1.\times 10^{-8}$. A capacitance of 1000 $\mu\mu$f shunted around a 10-ohm coil will produce a time constant of the assumed value while for a 10,000-ohm coil the capacitance need be only 1 $\mu\mu$f. In the case of inductance, a 10-ohm coil with 0.1 μh and a 10,000-ohm coil with 100 μh each have a time constant of $1.\times 10^{-8}$.

These considerations suggest the possibility of making coils with a low time constant by the use of a winding in which the inductance and capacitance are so proportioned as to satisfy the relation $L/R - CR = 0$. This procedure is open to the objection that uncontrollable variation in the resistivity and diameter of the wire make uniform results difficult. Capacitance thus purposely introduced usually has a comparatively poor dielectric so that both the time constant and resistance of

[5] G. A. Campbell, *Elec. World*, vol. 4, p. 728, (1904).

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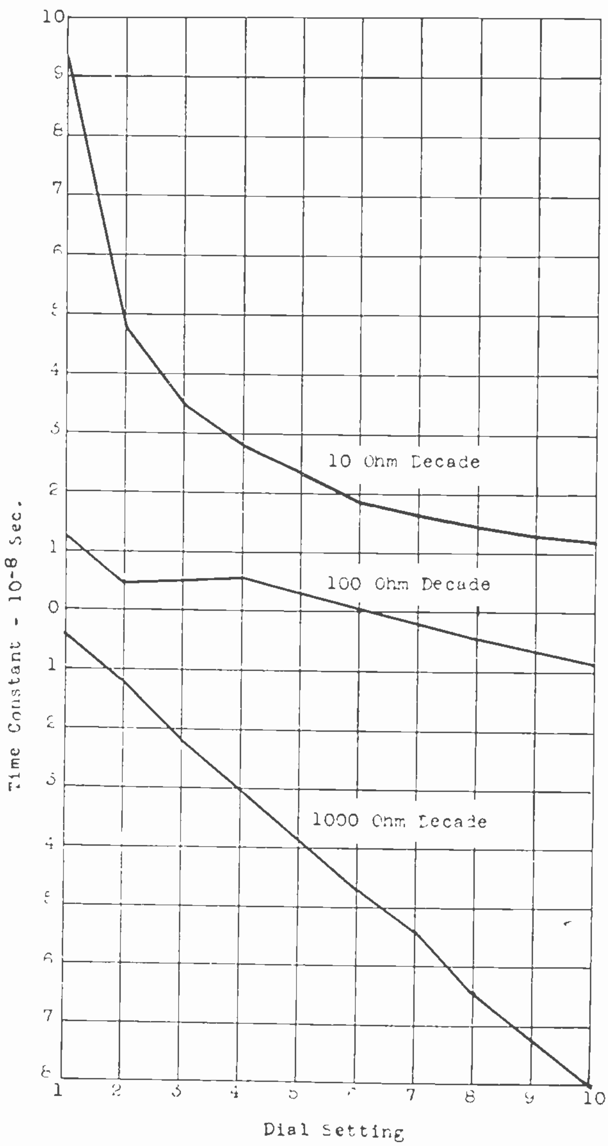

Fig. 8—Characteristics of six-dial shielded box. Shield connected to low terminal.

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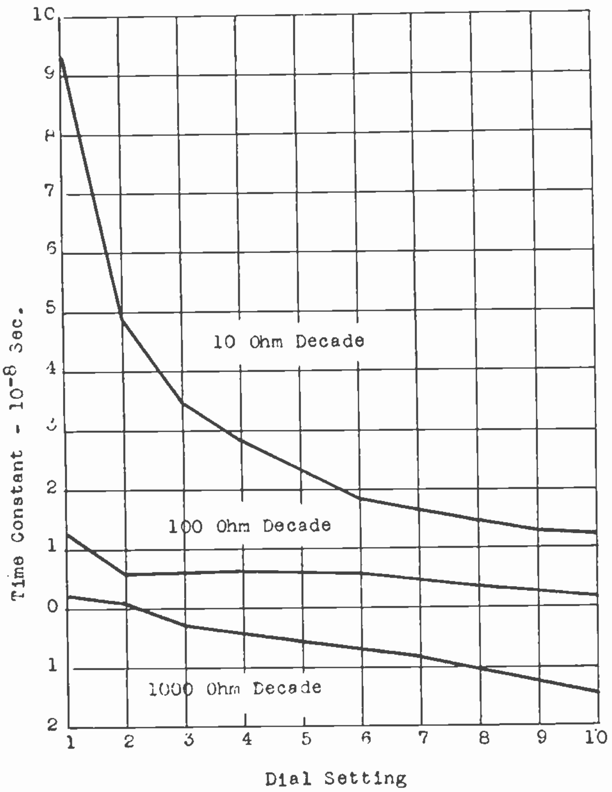

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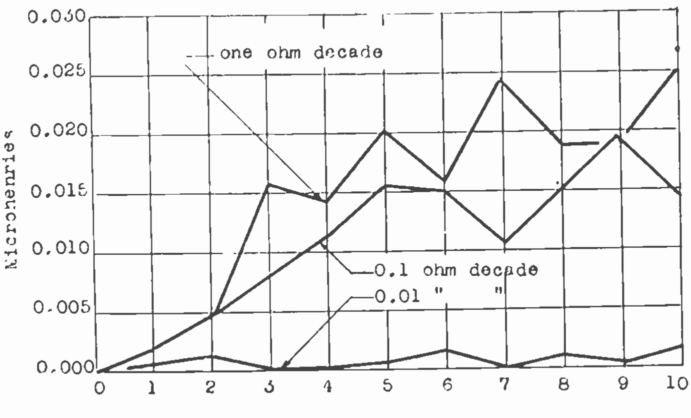

Fig. 10—Change of inductance with dial setting for six-dial shielded box.

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The inductance of a coil is kept low by so winding the coil that adjacent lengths of wire carry equal and opposite currents. The ideal condition is to have the go and return currents superimposed, and this condition is approached by reducing to a minimum the mean distance between the filaments of current flow in the two directions. For low and medium resistance coils, the mean distance may be made small by using two or more wires in parallel and interleaving the go and return paths, and by using wire in the form of a flat tape. The capacitance of the coil as a whole is minimized by making the capacitance between its parts small and so arranging this distributed capacitance that it is smallest between parts of the coil which have the largest difference of potential.

The requirements for minimum inductance and for minimum capacitance are in general conflicting. Their relative importance depends upon the resistance of the coil. When both inductance and capacitance are present, the time constant is approximately $L/R-CR$.[5] In the case of low resistances, the inductance term, with R in the denominator, is the dominating factor, while for high resistances the capacitance term, with R as a multiplier, is the larger.

A consideration of the numerical values is helpful. We are concerned with coils having time constants of the order of $1.\times 10^{-8}$. A capacitance of 1000 $\mu\mu$f shunted around a 10-ohm coil will produce a time constant of the assumed value while for a 10,000-ohm coil the capacitance need be only 1 $\mu\mu$f. In the case of inductance, a 10-ohm coil with 0.1 μh and a 10,000-ohm coil with 100 μh each have a time constant of $1.\times 10^{-8}$.

These considerations suggest the possibility of making coils with a low time constant by the use of a winding in which the inductance and capacitance are so proportioned as to satisfy the relation $L/R-CR=0$. This procedure is open to the objection that uncontrollable variation in the resistivity and diameter of the wire make uniform results difficult. Capacitance thus purposely introduced usually has a comparatively poor dielectric so that both the time constant and resistance of

[5] G. A. Campbell, *Elec. World*, vol. 4, p. 728, (1904).

the coil change with frequency. More satisfactory results are generally obtained by minimizing both the inductance and the capacitance. These remarks do not apply to all cases, and where compensation is resorted to in order to reduce the unavoidable residual, the compensating means is conveniently placed external to the coil and is adjusted after the coil is finished and as a result of residual measurements on the coil as actually made.

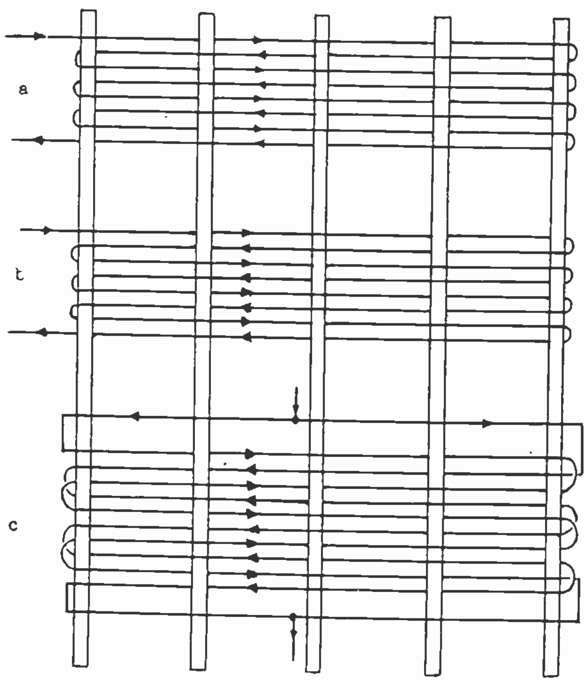

Fig. 11—Woven resistors.

Weaving has long been employed[6] in the manufacture of resistance coils, particularly those of higher values. Such construction, in effect, arranges the wire in a large number of bifilar sections. The bifilar arrangement results in a low inductance while the subdivision into a large number of sections fixes the potential difference between adjacent portions of wire at a small fraction of the total voltage. The method of weave hitherto used is shown in Fig. 11a and it is evident that the wire lengths are not as close together as they can be, for consecutive lengths pass on opposite sides of the warp. Two methods of weaving are now being used which avoid this objection. In Fig. 11b a single wire is used and the "go" and "return" lengths lie next to each other and on the same side of the warp. In Fig. 11c, the coil is made of two wires in parallel which are always side by side and through which current passes in opposite directions. This weave is particularly useful for the lower-valued woven resistors.

[6] A. C. Hague, "Alternating Current Bridge Methods," p. 70.

The webbing made as above is conveniently mounted by wrapping it about a cylindrical form of suitable insulating material. The wrapping is done with the warp parallel to the axis of the cylinder, as this maintains the maximum separation between those parts of the webbing between which there is a large difference of potential.

Table II gives the characteristics of several typical woven coils. In the column headed "weave" the letters refer to Fig. 11 above, illustrating the types of weave.

TABLE II
CHARACTERISTICS OF INDIVIDUAL RESISTANCE COILS

Resistance	*Wire Size*	*Weave*	*Time Constant*
20,000 ohms	No. 44 B and S	b	-1.4×10^{-8}
10,000	44	b	-0.8
1,000	44	b	$+0.11$
1,000	41	b	$+0.20$
1,000	41	a	$+0.80$
1,000	38	b	$+0.37$
100	41	b	$+0.36$
100	38	b	$+0.52$
100	36	b	$+0.92$
100	39	c	$+0.33$
100	38	c	$+0.43$

Resistances between about 5 and 50 ohms are generally wound of wire, lower resistances are usually of strip. Several different types of winding may be used; their performance, however, does not differ very much from the simple bifilar. The principal deciding factors are the size and shape of the cross section and whether there are several resistance wires or strips in parallel.

It is intended to describe in a future paper the methods of measurement used in obtaining the results presented here.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 7 *July, 1932*

DISCUSSION ON "DESIGN OF RESISTORS FOR PRECISE HIGH-FREQUENCY MEASUREMENTS"*

L. Behr and R. E. Tarpley

R. F. Field:[1] The choice as to whether the reactance of a resistor should be considered inductive or capacitive depends on whether this reactance is thought of as being in series or in parallel with the resistance. The logical choice is that which yields a value essentially constant at low frequencies. A resistor has inductance in series with its resistance and capacitance in parallel. The capacitance is distributed in the coil, but whenever the resistor is mounted on a switch or has terminals, lumped capacitance is added which is usually much larger than the

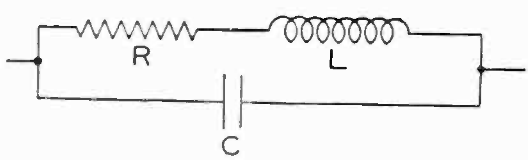

Fig. 1

distributed capacitance. The resistor is then equivalent to the circuit of Fig. 1. The equivalent series reactance of the circuit at frequencies low enough so that the terms containing the square of the frequency are negligible compared to unity is

$$\hat{X} = \omega(L - R^2C) \tag{1}$$

$$\therefore \hat{L} = L - R^2C. \tag{2}$$

Hence the resistor should be considered as inductive. When in the case of high resistances the term R^2C becomes larger than L it should be considered as having a negative inductance.

The parallel capacitance C may be found by transposing (2)

$$C = \frac{L - \hat{L}}{R^2} \tag{3}$$

if the series inductance is known. For resistances above 1 kilohm $L \angle\angle \hat{L}$ so that it may be neglected in expressing C. The value of this capacitance for any particular decade depends considerably upon its position with respect to the other decades. Roughly this capacitance is doubled over that for an isolated decade when connected as the end dial of a 4-dial box and set at maximum, increasing to three times at zero setting. When connected as an inner dial, its value is approximately constant at three times its isolated value. This indicates a large effect due to mutual capacitance. These values for the 4-dial Leeds & Northrup box are about 4, 8, and 12 $\mu\mu f$, respectively.

The addition of a shield connected to the lowest decade increases these considerably. When this shield is not connected to either terminal but kept at the potential of the lower decade, all capacitances to the shield are eliminated together with a considerable part of the capacitance between terminals, the mutual

* Proc. I.R.E., this issue, pp. 1101–1113.
[1] Engineer, General Radio Co., Cambridge, Mass.

capacitance, and the capacitance of the switches themselves. The resultant capacitance is between $\frac{1}{2}$ and $\frac{2}{3}$ that of a single isolated decade, having a value of about 2 $\mu\mu$f.

When more than one dial reads, the equivalent inductances add approximately, from which the time constant can be calculated if desired. But since resistance comes into (2) as a squared term, there is a cross term, $2R_1R_2C$, by which the inductance of the two in series is less than their sum. For high resistances which have large negative inductances, this term is important, and it becomes necessary to determine by a series of measurements the portion of the parallel capacitance common to each pair of decades.

The resistance and inductance at zero setting are very important quantities which can be neglected only when substitution methods are used so that the box is permanently in circuit. Time constant plots do not give these values. For the 4-dial Leeds & Northrup box they are 0.018Ω and 0.35 μh, respectively.

Skin effect depends upon the size of the wire used, its specific resistivity and permeability, and the inductance of the winding. It is not appreciable at 50 kc, but at 1 megacycle it may easily be 1 per cent. The skin effect of the zero loop is large, increasing the zero resistance to 0.01Ω, a factor of 5.

The 6-dial box is a very interesting attempt to produce a box with minimum residual errors. Of the three respects in which improvement may be sought, only the first, the capacitance of the connected but unused decades, has been completely realized and that is of least importance. The change in inductance from one coil to the next is usually much larger than that due to the coil placing, at least in the higher decades. The inductance is not appreciably reduced below that of the 4-dial box, because of the large zero inductance, 0.9 μh, which is due to the irregular shape of the zero loop.

The inductance and capacitance of a resistor depend upon the length of wire used in its construction and hence on the physical size of the finished resistor. The heat radiation of the resistor also depends directly on its size so that beyond a certain point small values of inductance and capacitance can be obtained only at the expense of current-carrying capacity unless extra radiating surface is added. The limit to which this process may be carried is illustrated by the straight fine wire resistors used as high-frequency standards up to a value of 1000 ohms.

L. Behr and R. E. Tarpley:[2] Equations of the form given by Mr. Field are of assistance to the resistance box designer in estimating the relative importance of changes in construction. Attention is called to the fact that the numerical values for the reactance characteristics, as presented in the time-constant curves of the paper, are the result of measurements made at the terminals of the resistance box. The numerical values are therefore independent of any assumptions as to the location and nature of the inductance and capacitance in the box, such as are involved in the derivation of the above equations.

In regard to the remarks on the changes introduced by the addition of a shield connected to the terminal which is joined to the lowest decade, Figs. 3 and 4 of the paper give quantitative results on this point. The 1- and 10-ohm dials are entirely unaffected and the change in the 100-ohm dial is just barely noticeable. There is a sensible increase in the capacitance of the 1000-ohm dial but in view of the amount of metal introduced by the shield, the increase in capacitance is really surprisingly small. The greatest change is at the 10,000-ohm setting, where the capacitance is increased from about 7 $\mu\mu$f to 11.

[2] Research Department, Leeds and Northrup Co., Philadelphia, Pa.

In the case of the direct resistance characteristics, where the shield is not connected to a terminal, it should perhaps be emphasized that the curves apply when the shield is maintained at the same potential as the terminal connected to either the highest or lowest decade.

The time-constant curves do not give the inductance of the box directly but they do give all the information needed to determine the inductance. For example, from the time-constant curve for the one-ohm decade in Fig. 3 and from

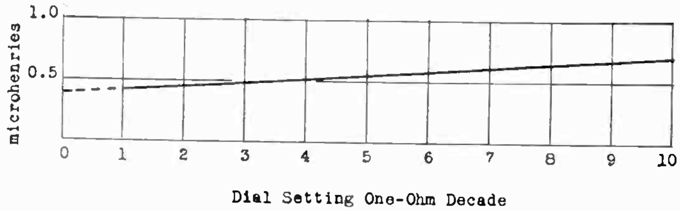

Fig. 12

the definition of the time constant as L/R, the inductance for each setting of the one-ohm decade may be computed, and plotted as in Fig. 12. The intercept on the axis of zero resistance is the zero inductance.

We do not agree with the general remarks on the single-coil-decade, six-dial resistance box. The change of inductance with dial setting is the important limiting factor in the performance of the lower decades of a resistance box, while for the higher decades it is the change of capacitance which is significant.

It is evident that, in a decade design in which the resistance is increased by the addition of elements to the circuit, while all previously included circuit ele-

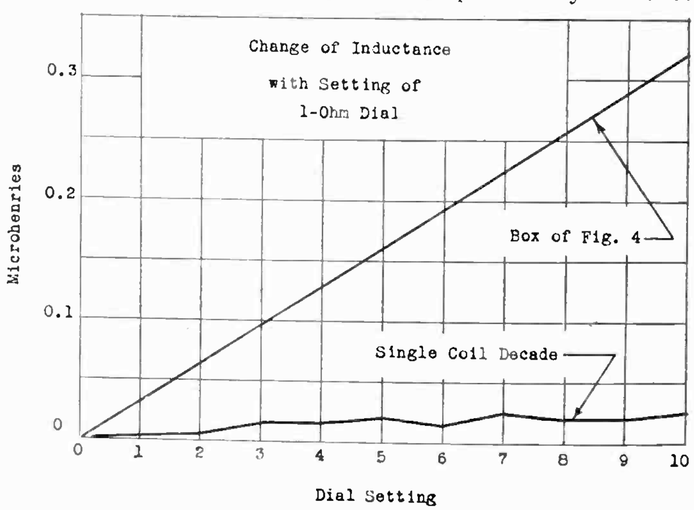

Fig. 13

ments are retained, the inductance must increase with increase of resistance. The substitution of one resistance unit for another, theoretically makes it possible to maintain the inductance of the circuit absolutely constant. In practice, coils have been designed and are regularly manufactured for the single-coil-decade box which closely approach this condition. Fig. 13 gives the change of inductance with dial setting for the 1-ohm decades of the single-coil decade box, and of the box shown in the photograph of Fig. 4, and shows that the change of inductance with dial setting for the former box is less than one-tenth of the change for the latter. The box of Fig. 4 was chosen as a basis for comparison because its rate of change of inductance is smaller than that of any available box we know of, using ten equal coils per decade.

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TWO-WAY RADIOTELEPHONE CIRCUITS*

By

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(American Telephone and Telegraph Company, New York City)

Summary—*This paper deals with the problems of joining long-distance radio-telephone transmission paths to the ordinary telephone plant. It gives the possibilities and limitations of various methods of two-way operation of such circuits wher the radio channels employ either long or short waves. It also describes the special terminal apparatus for switching the transmission paths under control of voice currents and lists the advantages of using voice-operated devices.*

Radiotelephone circuits are now in regular use between New York and London, New York and Buenos Aires, San Francisco and Honolulu, and many other points. At each end of such a circuit there are a transmitting radio station and a receiving radio station, usually geographically separated. The radio provides two one-way transmission paths. The circuit is completed by means of one-way wire lines which connect the radio sending and receiving stations to a common point. At this common point some rather intricate apparatus is called for in order to permit switching of the circuit to the wire telephone plants at the two terminals. This paper explains why this intricate apparatus is necessary even for the comparatively simple case of short-wave radio circuits which use different frequency bands to transmit in opposite directions. It also describes the latest form of this terminal apparatus in which provision is made for certain switching of privacy apparatus by means of which an important saving is made in the amount of privacy apparatus required. The original form of this apparatus is described in an earlier paper.[1]

Transmission Paths

Radiotelephone circuits may employ the same frequency band for transmission in the two directions or they may employ separate bands. The present long-wave telephone circuit between New York and London is of the first type, while most existing short-wave circuits are of the second.

A short-wave circuit, using separate frequency bands, is shown in its simplest form in Fig. 1. It is formed of two sets of terminal appara-

* Decimal classification: R450. Original manuscript received by the Institute, March 18, 1932. Presented before Twentieth Anniversary Convention, Pittsburgh, Pa., April 8, 1932.

[1] S. B. Wright and H. C. Silent, "New York-London telephone circuit," *Bell Sys. Tech. Jour.*, vol. 6, pp. 736–749; October, (1927).

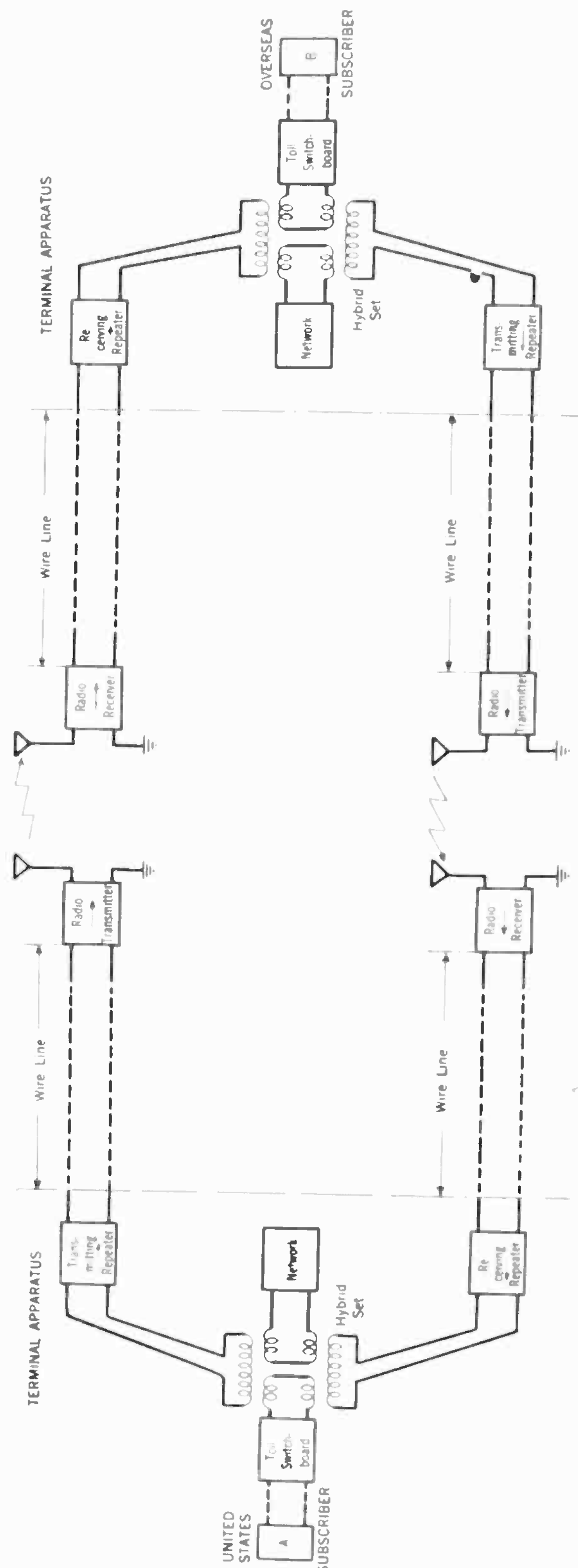

Fig. 1—Two separate radio links connected as a four-wire circuit.

tus connected by two one-way channels, each of which consists of a transmitting wire line, a radio link, and a receiving wire line. The function of the terminal or "combining" apparatus is to tie together these two one-way paths in such a manner that they may be connected at the switching centers to various telephone subscribers via the usual telephone circuits.

When the United States subscriber, designated as A in Fig. 1, talks, electrical waves set up by his voice pass over a wire line to a toll office. They then divide in a hybrid set. Part of the energy is dissipated in the output of a receiving repeater, and part is amplified by a transmitting repeater and passes over a wire line to a radio transmitter, as indicated in the upper transmission path of Fig. 1.

The waves are then amplified and transformed into radio-frequency energy and radiated. Some energy is picked up by a distant radio receiver, amplified, and transformed back into voice-frequency energy which passes over a wire line to the overseas terminal. The receiving repeater at this point makes up for the loss of the receiving wire line. From its output the waves pass into a hybrid set, part being dissipated in the network and the other part going through the toll office to the overseas subscriber B. Due to the imperfect balance between the subscriber's line and the network, a portion of this energy will be returned over the lower transmission path to the United States subscriber A as echo.

The action when the overseas subscriber B talks is substantially the same as that described above except that the useful speech waves pass over the lower transmission path.

In long-wave radio circuits the scarcity of suitable radio channels makes it highly desirable to use the same frequency band for transmission in both directions. This results in two additional radio paths becoming important, namely, those between the radio transmitter and the radio receiver at each end of the circuit. By using specially directive antenna arrangements, tranmission over these paths may be partly balanced out. In practice, this balance cannot be made very effective in reducing the relative importance of these paths without sacrificing materially the receiving directivity against natural radio noise. The effect of these added transmission paths is to make the transmission problem more difficult, as will be explained.

Transmission Characteristics

Returning to consideration of the simple four-wire set-up involved in short-wave operation, the transmission characteristics of the circuit evidently depend on the sum of the effects of the radio and wire line portions. In view of the relatively higher cost and greater length of the radio links, the highest grade wire circuits available are usually justified so that, in general, they should not be allowed to contribute much transmission impairment. In general, the added delay introduced by these wire lines is their most important effect.

If the radio links are quite stable and fairly free from atmospheric disturbance, the circuit may be operated like four-wire land telephone circuits. That is, the total amplification in the circuit may be kept at such a value that it never exceeds the total attenuation, and over-all singing will not occur. Four-wire terminating sets or hybrid sets placed at the ends of the circuit, as shown in Fig. 1, are adequate to prevent singing and minimize echo effects in four-wire circuits which have over-all transmission times less than about 0.02 second, provided the net loss from switchboard to switchboard does not become lower than about 5 db.

If the radio links or wire lines are long, the circuit will produce annoying echoes exactly as a four-wire cable circuit will, due to delay or instability, or both. Also, as in the case of four-wire cable circuits, echo effects may be reduced appreciably by voice-operated echo suppressors which block the path of the delayed echoes while the other path is transmitting speech. The possibilities and limitations of this type of device are discussed elsewhere.[7]

When the radio channels are more noisy and/or less stable, the transmission may be greatly improved through more efficient use of the radio links. The noise may be minimized by bringing the speech waves of all talkers, strong or weak, to the same "electrical volume" or strength at the input to the radio transmitter. Thus, practically full modulation may be maintained on the radio transmitter at all times and the ratio of the desired signal to the radio noise kept a maximum. Large changes in gain between the two-wire line and the radio transmitter are necessary to accomplish this result. These changes are made by technical operators who make the adjustments with variable loss devices. An indication of the volume is obtained through the aid of electrical meters called "volume indicators." In practice, the over-all transmission of a long radio circuit may be varied by the technical operators from a 30-db loss to a 30-db gain within a few minutes.

In short-wave circuits, the phenomenon known as "fading" introduces an effect which is of great importance in two-way operation. Where fading results in variations of the entire transmitted band of frequencies, automatic gain control at the radio receiver is effective in

[7] A. B. Clark and R. C. Mathes, "Echo suppressors for long telephone circuits," *Jour. A.I.E.E.*, vol. 44, pp. 618–626, June, (1925).

maintaining the received volume at a substantially constant value. The gain control is operated by the incoming carrier. When the fading is of the type in which the different frequencies in the transmitted band do not fade simultaneously, the automatic gain control is not so effective and considerable variations in volume out of the receiver may occur in a short time.

In the long-wave circuit, the variations are too slow to be classed as fading, and occasional manual adjustments of receiver gain result in keeping the volume at the receiving end within about ± 5 db.

Because of the gain adjustments to reduce noise, combined with changes in radio receiver gain to compensate for fading or for variations in radio attenuation, "singing" would occur if the hybrid coils and echo suppressor were not augmented by additional means of singing prevention. One way of preventing singing would be to reduce gain in the receiving leg whenever gain was introduced in the transmitting leg of the circuit. Volume penalties to the listener as great as 25 db would frequently be encountered if this were done, and, in addition, considerable agility would be required on the part of the technical operators to keep the circuit adjusted. However, this method would not compensate for gain changes in the radio receivers, so that singing might still occur under unfavorable conditions.

Also, in the case of a long-wave transatlantic circuit, singing could occur over transmission paths between the local radio transmitter and receiver. The volume received from the local transmitter may occasionally be as much as 40 db stronger than that from the distant station if the transmitter and receiver are about 90 miles apart, even though antenna directivity were used at both the transmitters and the receivers. In general, if the receiver gain is adjusted to give the proper volume on the distant station, the amplification in the local radio path is entirely out of reason.

It is, therefore, necessary to provide other means of preventing singing to maintain optimum transmission conditions.

Vodas

There has been developed for meeting these difficulties an antisinging voice-operated device known as a "vodas."[3] Fig. 6 shows a radiotelephone circuit arranged with a vodas in its simplest form at each end of the circuit. The vodas consists of a transmitting delay circuit, detector, and certain relays, and a receiving delay circuit, detector, and relay. These devices are operated by the voice currents in the circuit so as to keep all singing paths blocked at all times.

[3] Taken from initials of the words "Voice-Operated Device Antisinging."

The vodases in Fig. 6 are shown for the condition when no speech is being transmitted. Relay 1 keeps the transmitting circuit blocked so that singing cannot occur around the complete circuit or through a local radio path and terminal. Transmission is free to pass the contacts of relay 2. When the United States subscriber speaks, voice currents go into the transmitting detector and delay circuit. While they are traversing the delay circuit, relays 1 and 2 become operated provided relay 3 has not been operated previously. The operation of relay 1 permits the voice currents to travel on to the radio transmitter. The operation of relay 2 blocks the receiving path and prevents echoes and singing that might otherwise occur when relay 1 is operated.

Upon being received at the distant end, the voice currents operate relay 3 from the receiving detector, thus protecting the transmitting detector and relays against operation by echoes of received speech currents. These echoes are returned from unbalances in the two-wire portion of the connection beyond the terminal. The receiving delay circuit delays the speech long enough to insure complete operation of relay 3 before the echoes return. When the subscriber stops speaking, the relays return to normal.

By adding two more relays to the transmitting side of the vodas, it is possible to save part of the apparatus which is used to increase privacy on the circuit. This saving is made by using the same privacy device for both transmitting and receiving. This is possible provided the action of the privacy device is the same for distorting the voice waves at the transmitting end as it is for restoring them at the receiving end of the circuit. An arrangement of the vodas having this feature, which is now in general use, is illustrated in Fig. 2. The apparatus additional to the simple vodas consists of relays 4 and 5, the privacy device, a hybrid set, and two one-way repeaters. In Fig. 2 this apparatus is labeled "Privacy Switching Circuit."

The action of the device shown in Fig. 2 on transmitting speech waves is as follows: Useful waves from subscriber A are impressed on a potentiometer ahead of the transmitting repeater, which is kept adjusted by the technical operator to maintain constant volume at the output of the transmitting repeater. The waves then pass into the vodas where they first reach the transmitting delay circuit and are stored for a short interval. A small part of these waves enters the transmitting detector and operates relays 1, 2, 4, and 5, provided relay 3 has not been operated by the receiving detector previously. The interval of the transmitting delay circuit is several times as great as the operating time of relays 1, 2, 4, and 5 so that initial weak parts of speech syllables may be stored in this delay circuit until stronger parts arriving later have had a chance to operate the relays.

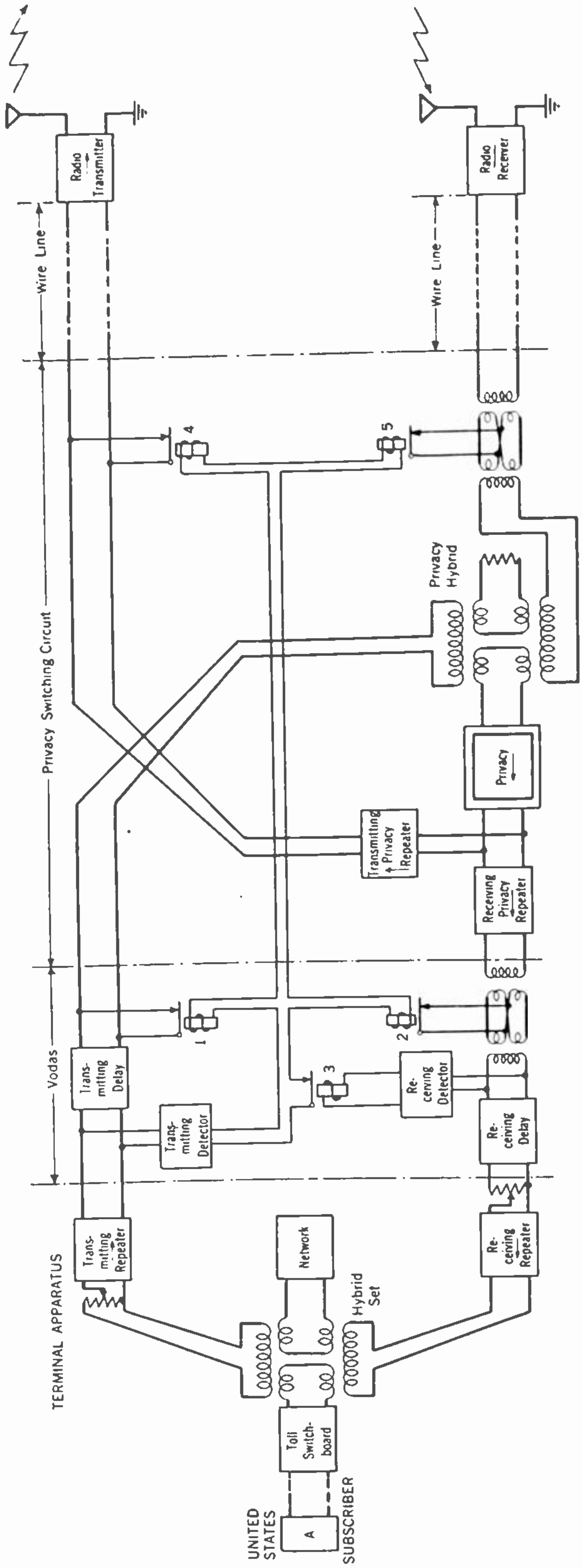

Fig. 2—Voice-operated device used on New York-Buenos Aires and other radiotelephone circuits.

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leased.

In practice it is necessary to protect the vodas against operation from noise. This is done by frequency discrimination as well as by amplitude discrimination and the use of artificial delay. The detectors have their input circuits arranged to keep out frequencies that are not essential for speech operation. In addition, their sensitivity is made adjustable. The transmitting detector is generally worked at a value which results in no perceptible loss of intelligibility due to failure of the transmitting relays to operate, at the same time allowing a maximum amount of line noise to be applied without operating the relays falsely. The receiving detector is adjusted frequently by the technical operator so as to obtain the best operation on incoming speech without false operation from radio noise. When the incoming noise is low, relay 3 may be made very sensitive. Any incoming speech which does not

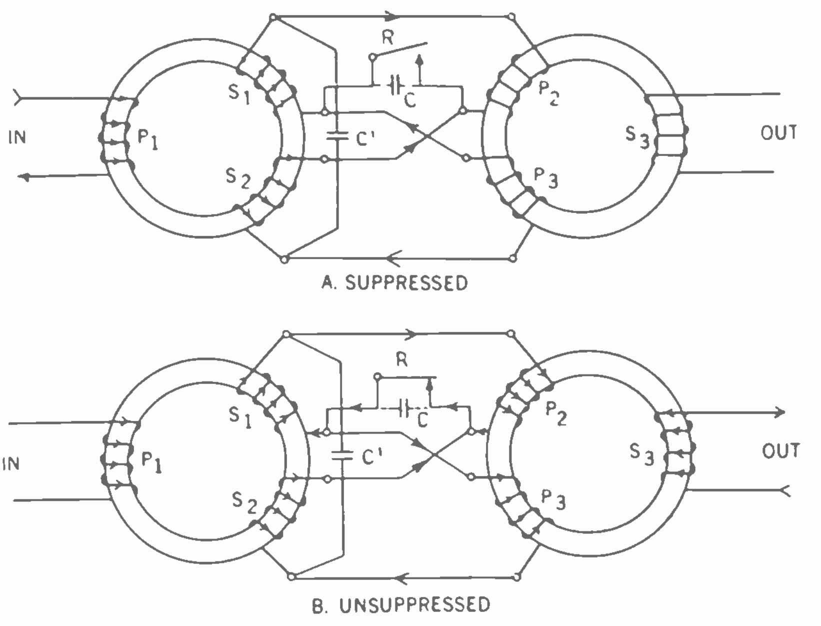

Fig. 3—Repeating coil arrangement for suppressing echoes.

operate relay 3 is thus weak, and the receiving volume may be kept high without danger of echoes operating the transmitting relays. When the noise is high, relay 3 is made insensitive, requiring more loss in the echo path and, consequently, lower volume to the listener.

The method of suppressing transmission by opening a single relay contact is illustrated in detail by Fig. 3. In A of this figure, the relay (*R* on the figure) is assumed to be operated so that transmission is suppressed. The voltages induced in windings S_1 and S_2 of the first coil are opposed to each other in a circuit including P_2 and P_3 of the second coil, the resulting flux in the core of this coil being very small. Losses as high as 75 db are produced by this arrangement by proper design of coils and adding a small condenser (C') to balance the capacity (C) of the leads to the relay contacts. In practice, a pair of wires is cut off to

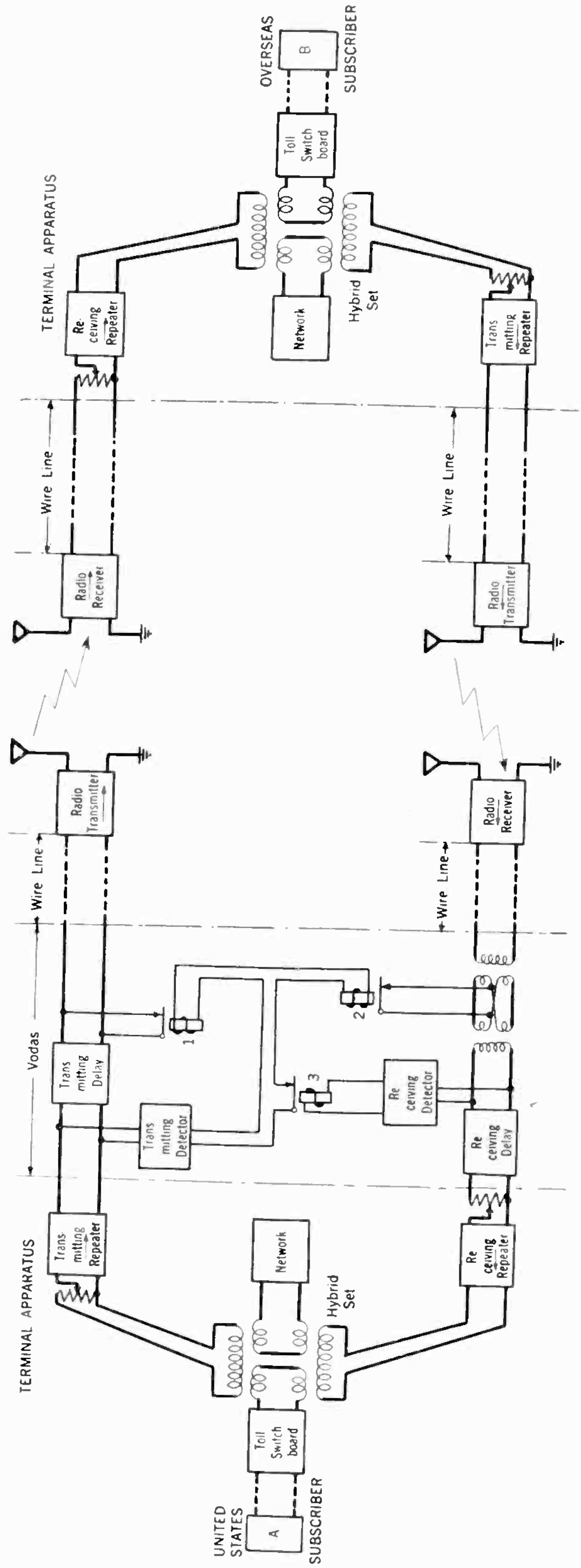

Fig. 4—Radiotelephone circuit with vodas at one end only.

give the right capacity, and then laced into the cable form. In B of the figure, the circuit is shown in the normal condition. Transmission through the coils is affected only by their normal transition loss. While the windings S_1 and S_2 in A are effectively in series opposing, in B they are connected by separate circuits to the corresponding windings P_2 and P_3, due to the extra path through the relay contact.

Operation of a Radiotelephone Circuit

Having in the vodas a means for suppressing echoes and singing under extreme conditions (with the additional advantages of suppressing intermediate "cross-transmission" paths), it is important to consider the broader application of such a device to radiotelephone circuits. Three cases of operation with antisinging devices are of interest:

1. Vodas at one end, plain hybrid set at the other.
2. Vodas at one end, echo suppressor (without antisinging relay) at the other.
3. Vodases at both ends.

1. Vodas at One End, Plain Hybrid Set at the Other

This arrangement is shown in Fig. 4. In this and the next figures, the privacy switching circuit has been omitted for simplicity. The vodas prevents singing and echo effects from unbalances at the *A* end and also prevents the *A* subscriber from hearing echoes.

The disadvantages of these arrangements may be understood by considering the transmission received at the *B* end which has no voice-operated relays. Speech received over the circuit would be returned to the local radio transmitter as an echo or echoes. If a weak talker were connected at the *B* end, the volume control device would amplify these echoes to an appreciable extent. In addition to overloading the radio transmitter, such echo would permit both sides of the conversations to be broadcast from the same station, thus reducing privacy. Radio noise might also be received at *B* and transmitted as echoes to the *A* end of the circuit. In addition, line noise from a two-wire circuit at the *B* end would be freely transmitted to the *A* end causing a limitation of the sensitivity of relay 3 and consequently a reduction of volume at the *A* end.

2. Vodas at One End, Echo Suppressor at the Other

The above possibilities of transmitter overloading and speech reradiation due to echoes might be prevented by adding a voice-operated echo suppressor at the *B* end as shown in Fig. 5. This device would be operated by received speech so as to disable the transmitting branch of the circuit. Its sensitivity would be limited as would also the received volume at one or both ends. It should be noted that if no cross trans-

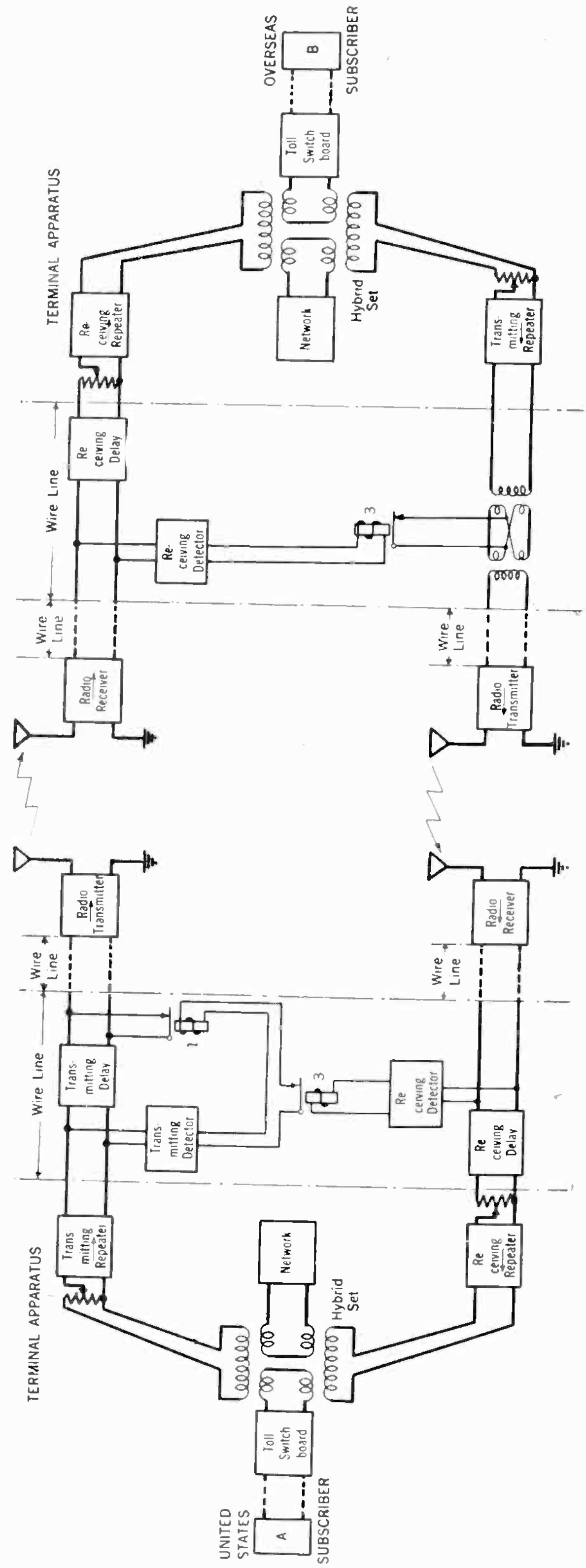

Fig. 5—Radiotelephone circuit with vodas at one end and echo suppressor at opposite end.

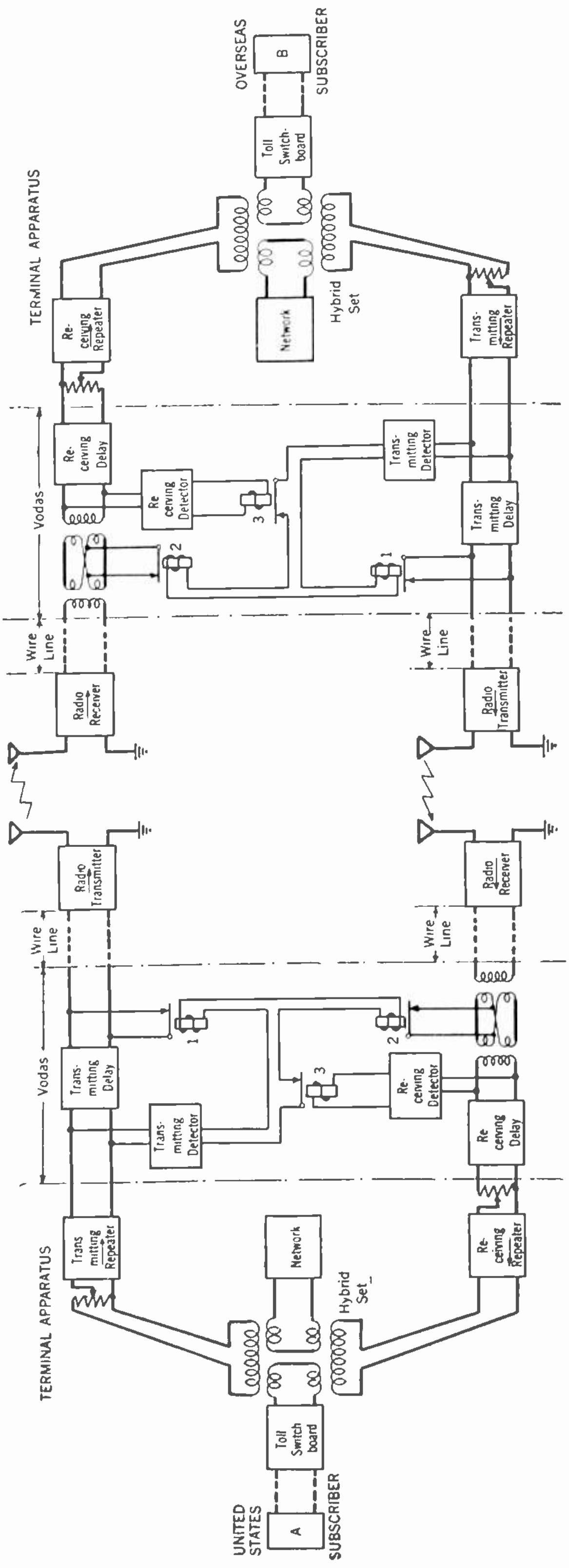

Fig. 6—Radiotelephone circuit with vodases at both ends.

mission paths existed, relay 2 of the vodas at the A end could be omitted when this device is used.

The echo suppressor would not suppress echoes of radio noise or direct transmission of line noise from the B end to the receiving relay 3 at the A end. Relay 3 would, therefore, need to have its sensitivity reduced so as not to be operated by these noises. This gives rise to an additional limitation of received volume at the A end. The amount of the penalty would, of course, depend on the noise conditions, the talker volume and the echoes in two-wire circuits at the B end. Under extreme conditions, the necessary reduction in receiving volume at the A end might be as much as 25 db. This is considered to be an important disadvantage inasmuch as conditions at the two ends are not independent and lack of an antisinging device at one end penalizes the received volume at the other.

Another solution would be to limit the transmission gain at the B end so that the noise transmitted past the echo suppressor would never limit the sensitivity of relay 3 at the A end. This would mean that the received signal-to-noise ratio at the A end would be reduced, particularly if the talker at the B end were weak.

3. Vodases at Both Ends

To summarize, it may be said that anything short of antisinging devices at both ends does not make the two ends of the circuit independent and may penalize the transmission at the vodas end when there is radio noise or line noise at the end without an antisinging device. The preferred arrangement is shown in Fig. 6.

Results of Vodas Operation

In general, the results of operation with the vodas have been good and, when radio conditions are favorable, the circuits are not appreciably different from land circuits of comparable length.

Occasionally, the vodases introduce minor difficulties. False operation by noise and simultaneous talking by the two subscribers both tend to cause speech mutilation. The large transmission advantages afforded by their use greatly outweigh any such troubles. These advantages may be summarized as follows:

1. Suppress echoes and singing which would otherwise be heard due to adjustments to reduce radio noise, instability and cross-transmission paths.

2. Prevent retransmission of echoes which would cause overloading and two-way broadcasting at the radio transmitters.

3. Save privacy apparatus.

4. Permit the telephone listeners to hear louder speech waves.

5. Afford independent technical operation of the two ends of the circuit.

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KENNELLY-HEAVISIDE LAYER STUDIES EMPLOYING A RAPID METHOD OF VIRTUAL-HEIGHT DETERMINATION*

By

J. P. Schafer and W. M. Goodall

(Bell Telephone Laboratories, Deal, N. J.)

Summary—*This paper describes a new method of determining the virtual height of the ionized regions by visual observations of the received pulse pattern on a cathode ray oscillograph tube, both for single frequencies and for two frequencies simultaneously. A résumé of the data obtained during observations of some three hundred hours is given. The frequencies used for these tests were 1604 kc, 2398 kc, 3256 kc, 4795 kc, and 6425 kc. A number of the tests included measurements made upon two frequencies in rapid rotation. The more important results may be summarized as follows:*

(1) On a large number of occasions during the night, a phenomenon has been observed apparently indicating an increase in the density of ionization in the lower layer. This is important because the ionization is usually assumed to decrease during the night hours.

(2) Reflections are often observed simultaneously from both ionized layers. An explanation of this phenomenon is given.

(3) The virtual heights of the reflecting layers are rarely duplicated from day to day for a given time and frequency.

(4) Large numbers of multiple reflections are frequently obtained representing a path distance of over 5000 km for the last reflection. This fact indicates that the multiple-hop mode of propagation is probable for long-distance transmission.

Introduction

In May, 1931, experimental studies of the virtual heights of the Kennelly-Heaviside layer were resumed at Deal. The general pulse transmission system described by Breit and Tuve was again used. Previous work in this field[1] had indicated the need for a means of observation which would give immediate values of virtual heights, rather than the lengthy and cumbersome photographic process formerly employed. A new method has accordingly been devised for obtaining these heights in which visual observations of the pattern on a cathode ray

* Decimal classification: R113.61. Original manuscript received by the Institute, March 3, 1932. Presented before Twentieth Anniversary Convention, Pittsburgh, Pa., April 7, 1932.

[1] Schafer and Goodall, Proc. I.R.E., August, (1931).

oscillograph[2] make it possible to plot the data showing time variations in the virtual heights while the test is still in progress.

By employing this new method considerable data have been obtained in the range between 1.5 and 6.5 megacycles. Many of the tests included measurements made upon two frequencies in rapid rotation. In this case two complete transmitters and receivers were used, and oscillographic observations were made first on one and then on the other frequency.

A method of observing two frequencies simultaneously on the same oscillograph tube has also been developed and preliminary tests indicate that it will function satisfactorily.

Most previous methods for obtaining virtual heights using pulse transmissions required the use of an ordinary string oscillograph. Since

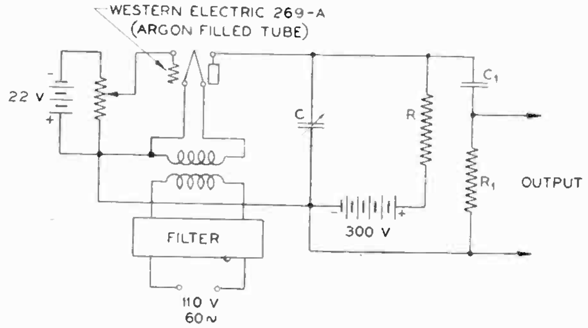

Fig. 1—Circuit of the pulse generator. The pulse generator is a relaxation oscillator in which the number of pulses generated per second is controlled by the time constant CR, while the duration of the pulses is controlled by the time constant C_1R_1. The control-frequency voltage is obtained by heating the filament of the tube with 60-cycle alternating current.

no accurate visual observations could be made with this instrument it was necessary to obtain records photographically. After the film had been developed it was necessary to make measurements of the time intervals between the pulses recorded on the film. For the most accurate results this measurement was made with a micrometer comparator and was a very lengthy and tedious process. If only one oscillogram were taken every ten minutes, it required about a week to measure the films taken during a twenty-four hour test. When the cathode ray tube is used, it is possible to obtain practically continuous values im-

[2] A Braun tube seems first to have been used by Goubau, but accurate data was obtained photographically and not visually. *See* G. Goubau, *Phys. Zeit.*, vol. 31, p. 333, (1930), and Goubau and Zenneck, *Zeit. für Hochfrequenz.*, June, (1931).

Since our method was developed other investigators have published results using cathode ray tubes for visual observations.

Ivo Ranzi, "Nuovo Cimento," Anno VII, N. 6; June, (1931); E. L. C. White, *Proc. Cambridge Phil. Soc.*, vol. XVII, part 3, July, (1931).

mediately. This feature is of special importance in the study of rapid variations.

In the present method, the transmitted pulse frequency and pulse phase are controlled by a voltage of constant frequency. In our case this is obtained from the commercial 60-cycle power supply (Fig. 1). Voltage of this same frequency is used for a timing wave on the oscillograph tube at the receiver. The pulse pattern observed at the receiver consists of the direct or ground pulse, which is stationary, and the overhead or reflected pulses, which are constantly changing in amplitude and displacement in accordance with variations in the ionized layers. By measuring the separation between the ground pulse and the reflections it is possible to obtain the difference in time between the two paths, knowing the frequency and the wave form of the voltage used for the timing wave.

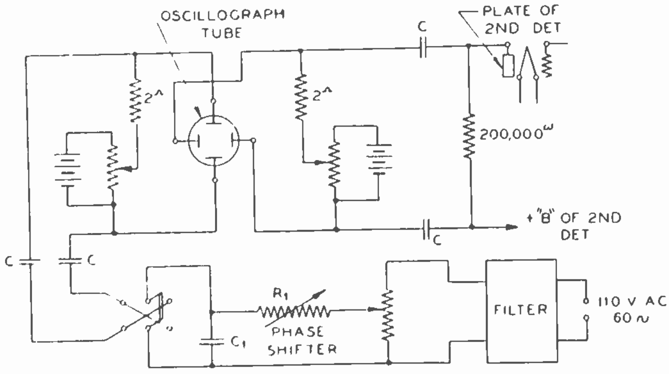

Fig. 2 Circuit of cathode ray oscillograph and output of radio receiver.

In the present series of experiments the direct or ground path between the transmitter and receiver is very short (1 km) and, therefore, the ground-wave pulses are very much stronger than the pulses which are reflected from the upper atmosphere. Even when the horizontal dipoles which are used for transmitting and receiving are pointed end-on, the difference between the two waves is of the order of 50 db (or greater) for weak reflections. With the arrangement of receiving apparatus used, however, (see Fig. 2 for schematic circuit diagram) this large difference does not cause any particular difficulty. All that is necessary is to increase the gain of the receiver until the reflections are visible on the oscillograph screen. For this condition, the ground-wave pulse has a very large amplitude and only its beginning and end are visible. Values of time intervals between pulses are obtained from measurements of the displacement between the beginnings of the two pulses and do not depend upon the whole of the ground pulse being visible on the screen. No very serious attempt has therefore been made to reduce the ground-wave amplitude below its present value, although this may be desirable in future experiments.

Experimental Procedure

The method of procedure in making observations of the virtual heights of the reflecting regions may be stated briefly as follows:

The receiver is tuned to the desired frequency and the ground-wave pulse is observed on the screen of the oscillograph tube. The position of this pulse is shifted to the zero voltage point of the 60-cycle timing wave by means of a phase shifter in the timing-wave voltage supply (Fig. 2). The amplitude of the timing-wave voltage is adjusted to a value which gives a standard deflection on the scale of the oscillograph tube. The gain of the receiver is then increased until reflections are observed. The scale reading between the beginning of the ground pulse and the beginning of the reflected pulse is then noted. From this reading it is possible to calculate the time interval and, therefore, the virtual height, knowing that the 60-cycle timing wave is sinusoidal. For a given amplitude of timing wave a calibration chart may be drawn giving the direct relation between the scale reading and virtual height. In this calibration chart a correction is made for any nonlinearity that may be found in the characteristic of the cathode ray tube.

The observer at the receiver reads the values corresponding to the various reflections, the data being recorded at the transmitter. This is accomplished by telephone communication between the two laboratories. This method avoids the delay which would arise if it were necessary for the observer to record the results. In this manner it is possible to obtain a practically continuous record of several simultaneous reflections, even during periods of rapid variations, since the observer can watch the oscillograph tube continuously and report any change as soon as it has taken place. Another advantage of this method is that it is less tiring on the eyes of the observer since they are focused continually on the tube and are not shifting back and forth from the tube to the data sheet as would be necessary if the data were recorded at the receiver.

The zero voltage point of the timing wave is adjusted, by means of a d-c bias, so that it falls on one edge of the screen. The amplitude of the timing-wave voltage is then adjusted so that its trace extends in one direction across the whole screen. This is equivalent to about $3\frac{1}{2}$ or 4 inches. The part of the sine wave which is visible at any instant is, therefore, $\frac{1}{2}$ cycle or 1/120th second. The trace of this half-cycle consists of two lines which are superimposed, the rising voltage during the first quarter-cycle and the falling voltage during the second quarter-cycle. Actually the two lines are separated a trifle due to stray 60-cycle pick-up, which is desirable in order to distinguish between reflections which happen to occur during the different quarter-cycles. Reflections which occur during the third and fourth quarters of the cycle may be observed by the use of a polarity-reversing switch in the timing-wave circuit. Pulses which occur during the first quarter-cycle are due to reflections which have a time retardation up to 1/240th (4.17×10^{-3}) second. This corresponds to virtual-height values up to 625 km. In this manner, for one complete cycle, reflections corresponding to virtual heights of any value up to 2500 km may be observed.

Usually the reflections occurring during the first quarter-cycle are of most interest as the reflections in this range include those from the lower and from the upper ionized regions having virtual heights of approximately 100 and 300 km, respectively. The scale which has been used in our observations has been divided into 20 parts per inch and it is easy to obtain readings which are accurate to less than $\frac{1}{2}$ division. This gives an accuracy of about ± 2 km for values between 100 and 350 km. For readings which represent heights between 400 and 800 km the accuracy is less than this due to the fact that the reflections are now observed at that part of the sine wave where the voltage changes more slowly with time. At the crest, one-half division represents a difference in virtual height of about 25 km. On the whole, however, the use of the sine wave is believed to be preferable to the usual type of linear time scale. In the latter case, for a 60-cycle control frequency, all reflections corresponding to virtual heights up to 2500 km would be crowded into a single "sweep" and a height of 100 km would be represented by a reading of only 0.16 inch, whereas when the sine wave is used as described above a 100-km height would correspond to a deflection of 1.0 inch. The accuracy of either time scale could be increased by using a frequency of 4 or 5 times 60 cycles, but then reflections representing heights greater than 500 or 600 km would be superimposed upon those representing heights less than this value making observations very difficult. The accuracy of the measurements for the lower layer could be increased by increasing the amplitude of the timing-wave voltage, but this would cause the reflections from the upper layer to be deflected off the screen.

When two frequencies are to be observed simultaneously, the transmitters are modulated with thirty pulses per second, with the pulses of one transmitter beginning 1/60th of a second later than those of the other transmitter. These pulses are generated by using two relaxation oscillators (Fig. 1) with the time-constant CR adjusted to give 30 pulses per second with a phase difference of 180 degrees between the pulses of the two oscillators. Any other convenient method can be used. The output circuits of the two receivers are connected in parallel to one set of plates of the cathode ray tube. The usual 60-cycle timing-wave voltage is applied to the other set of plates. Thus on one cycle, the ground pulse and reflections are due to one radio-frequency signal, while on the next cycle they are due to the other. One of the receivers is equipped with a one-stage amplifier which causes the pulses of one frequency to be deflected upward while those of the other frequency are deflected downward. In this manner it is possible to observe two frequencies at the same time on the same oscillograph tube. A switching arrangement is provided so that it is possible to select either frequency at any time. Although only preliminary tests have been made using simultaneous observations the results indicate that the method will be very useful.

Experimental Results

Observations have been made over a period of several months (May to December, 1931). In addition to the method described for obtaining measurements on two frequencies by rapid rotation, it should be mentioned that in practically all cases, observations have been made on each of the four frequencies between 1.5 and 5 megacycles at the beginning and at the end of each test. Since the beginning of September, during the day, observations have also been made on 6.4 mc. After observations had been made on all of the frequencies, the two to be observed continuously were chosen so as to obtain observations giving both upper- and lower-layer virtual heights, whenever this was possible. A few typical data curves and a résumé of the principal phenomena which have been obtained during some three hundred hours of observation will be given below.

(1) *Base Line*

Little difficulty has been experienced in obtaining strong reflections at the receiver in spite of the fact that the distance between the transmitter and receiver was only one kilometer. It was necessary, however, to use a horizontal dipole as an antenna at either the transmitter or the receiver in order to obtain strong reflected signals. In practically all of our measurements horizontal dipoles have been used at both transmitters and receivers.

(2) *Variation of Absorption*

The reflections obtained during the middle of the day are usually weak, while toward sunrise and sunset, and during the night, they are much stronger. Often this amplitude change takes place without an appreciable change in virtual height. This indicates that in addition to the absorption in the lower reflecting layer, there may also be an absorbing region below this layer which attenuates the signal but does not materially alter the virtual height.[3] This absorbing region is probably an important factor in daytime transmission for medium frequencies and makes the use of high frequencies necessary for long-distance transmission.

(3) Variation in Layer Height

The virtual height of the lower layer shows relatively small variations, either with time of day or with frequency, nearly all measurements indicating heights between 90 and 125 km.

The virtual height of the upper layer shows large variations depending upon the time of day and upon the frequency; generally the measured height is between 240 and 350 km. At certain critical periods, however, heights up to 650 km are not unusual. No values of virtual heights greater than 700 km have been observed when only a single reflection has been present.

(4) Frequent Appearance of Lower Layer at Night

The virtual height of the lower layer frequently decreases during the night, while at the same time reflections are again obtained at a higher frequency. It is possible that this effect can be explained on the basis of a changing ionic gradient due to recombination. It seems more probable, however, that it is due to an increase in the ionic density in the region where reflection occurs. We believe that such an increase would be due to an additional ionizing agency.

A nighttime increase in the ionization of these regions cannot be caused by the agencies suggested by Chapman;[4] i.e., high velocity neutral particles from the sun for the lower layer, and ultra-violet light for the upper layer. These phenomena would only give daylight effects whereas it would be necessary to explain the night increases as well.

One assumption which might be made to explain an increase in ionization would be that during such times the changing ionization is due to charged particles from the sun which have been deflected by the earth's magnetic field and in this way reach the dark side of the earth. These charged particles might not necessarily have to be present in such quantities as to cause an appreciable magnetic disturbance.

Another assumption might be that the increased ionization is caused by collision between molecules of the atmosphere and meteoric matter which is probably present in varying quantities in the earth's path.

In any event it seems rather important to know that frequently the

[3] R. A. Heising, Proc. I.R.E., January, (1928); S. Chapman, *Proc. Roy. Soc. A*, vol. 132, p. 353; August 1, (1931). See also footnote 1.
[4] *Loco. cit.*

ionization of the lower layer is sufficient to give reflections from this layer during such a large part of the night. The effect of this phenomenon would probably be most pronounced in radio transmission for medium distances and frequencies.

(5) *Existence of Two Layers*

In practically every extended test of the present series we have obtained data which show two distinct groups of virtual heights. While it is possible that a rigorous mathematical treatment of the problem might show that the data could be explained on a single layer basis,

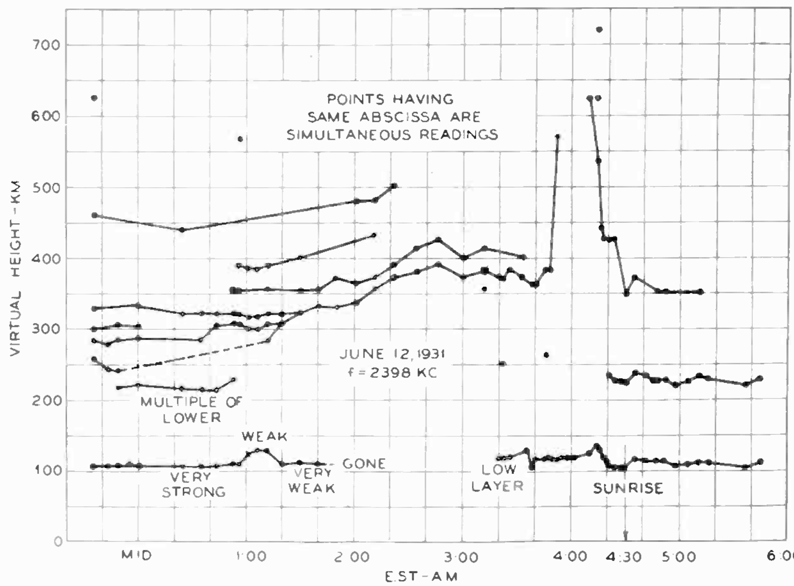

Fig. 3—Virtual-height variations during the early morning of June 12, 1931, for a frequency of 2398 kc. Note the varying intensity of the amplitudes of reflections from the lower layer and the fact that the virtual-height variations are not very great. The virtual-height variations of the upper layer are much greater during the same period. Split peaks are shown for the upper layer as well as a critical period shortly before sunrise.

we believe, as has been assumed by other investigators, notably Appleton, that these two groups of virtual heights indicate the existence of two layers or regions of maximum ionization.[5]

Fig. 3 is an example of the data obtained during the tests and shows how the virtual heights of the two layers vary independently with time. The reflection amplitudes vary independently also and the upper-layer reflections are not multiples of the lower. Multiple reflections, however, are also obtained from both layers.

[5] It is believed that the reason these two groups of virtual heights were not observed by Goubau and Zenneck (footnote 2) is mainly due to the wavelength used in their experiments (530 meters). It is probable that during the periods of their tests the conditions of ionization of the lower layer prevented upper-layer reflections from being received at this low frequency.

(6) *Simultaneous Reflection from Both Layers*

This phenomenon is found more often during the late afternoon and the night than during the middle of the day. Whenever lower-layer reflections are received, upper-layer reflections are also observed, except during periods of maximum ionization. These simultaneous reflections have been observed for at least one frequency in nearly every extended test of the present series. The fact that they had never been found during our previous tests may indicate that, due to seasonal or annual changes, conditions are now different from those prevailing in the spring of 1929. It may also be due to the fact that the angle of incidence is much smaller in these experiments than it was in those of 1929.

For all frequencies below 5 mc a transition from upper-layer reflections to lower-layer reflections usually takes place after sunrise. This change occurs as follows: Reflections are first observed from the upper layer. Then weak reflections from the lower layer are observed simultaneously with those from the upper layer. From this point on, the lower-layer reflections become stronger and those from the upper layer become weaker, until only lower-layer reflections remain. Both layers show a gradual decrease in height during this period. In the afternoon or evening the reverse of this phenomenon may take place.

When conditions of ionization are such that a reflection of appreciable amplitude would be predicted on the basis of the gradual bending of the rays due to refraction, the signal would not be expected to penetrate the layer and consequently reflections would be obtained from only one layer at a time. It is, therefore, apparent that this simple ray theory cannot satisfactorily explain simultaneous reflections. Epstein[6] has recently demonstrated that for the case when the dielectric factor of the medium attains a minimum value in a vertical direction and then increases again, partial rather than total reflection should take place. It, of course, follows that if the dielectric factor passes through a maximum and then decreases sufficiently at some greater height, the wave which penetrates the lower layer will be reflected from this upper region. The amplitude of the upper-layer reflection however, may not be sufficiently strong to be observed at the receiver. A study of this problem has been undertaken, but so far only preliminary conclusions have been reached. From this study it appears that for certain conditions of ionization, which might reasonably be expected to exist in the layers, and for normal incidence, the reflections which are returned simultaneously from both layers could have approximately equal amplitudes.

[6] P. S. Epstein, *Proc. Nat. Acad. Sci.*, vol. 16, October 15, (1930). Also *see* D. R. Hartree, *Proc. Roy. Soc.*, series A, vol. 131, no. A817, p. 428.

(7) Day-to-Day Variations

Virtual heights measured for a certain time on a given frequency are not duplicated from day to day; i.e., upper-layer reflections may be observed one day while reflections from the lower layer are found the next day, or both may be observed simultaneously. This variability is found even during periods when magnetic conditions are substantially undisturbed.

An example of this is given in Fig. 4 which shows the results of measurements taken on four successive days. Apparently the conditions of ionization were different on all four days.[7] In fact it is unusual to obtain virtual height values on any day which duplicate those obtained at the same time on the previous day.

(8) Multiple Reflections

Multiple reflections are often observed from both layers. Usually when strong multiples are obtained from one layer, reflections from the other layer are either weak or absent. For the condition of strong multiples from the lower layer, the virtual heights measured are less than the average of values obtained at other times. This indicates that the ionization is stronger than usual. At such times absorption in the lower layer probably prevents the reception of reflections from the upper layer.

In general the number of multiple reflections and the frequency of their occurrence is greater for the upper layer than for the lower layer. As many as ten or eleven upper-layer reflections (v.h. 250 km) and as many as five lower-layer reflections (v.h. 100 km) have been measured.

(9) Split Peaks

A phenomenon often observed is that the received echo is composed of several peaks. (See Fig. 3.) These "split peaks" indicate differences in virtual heights of 10 to 50 km and are very prevalent especially in

[7] (a) Magnetic disturbances—A mild disturbance was reported on May 26th by Science Service Research Bulletin. The other three days were reported as quiet. The mild disturbance noted on one day is hardly believed to be sufficient cause for the variations found over the four-day period.

(b) Earth currents—A study of earth current records shows no unusual variations for any day during the periods when virtual-height measurements were being made.

(c) Radio field strengths—A study of the field strength records of transatlantic signals between 12 and 18 megacycles shows that the fields were normal on the 25th and the 28th, and somewhat low on the 26th and the 27th, the minimum being on the 26th of May. The latter day is the one day when only lower layer reflections were obtained locally, and it may be that when conditions of ionization are such as to cause this local reflection condition, the transatlantic signal strength is reduced by absorption in the lower regions, whereas when reflections occur locally from the upper layer the attenuation to transatlantic signals is less.

reflections from the upper layer.[8] Usually there are two peaks formed on one pulse group but often three or more are present. The different peaks fade rapidly and independently. On some occasions the upper-

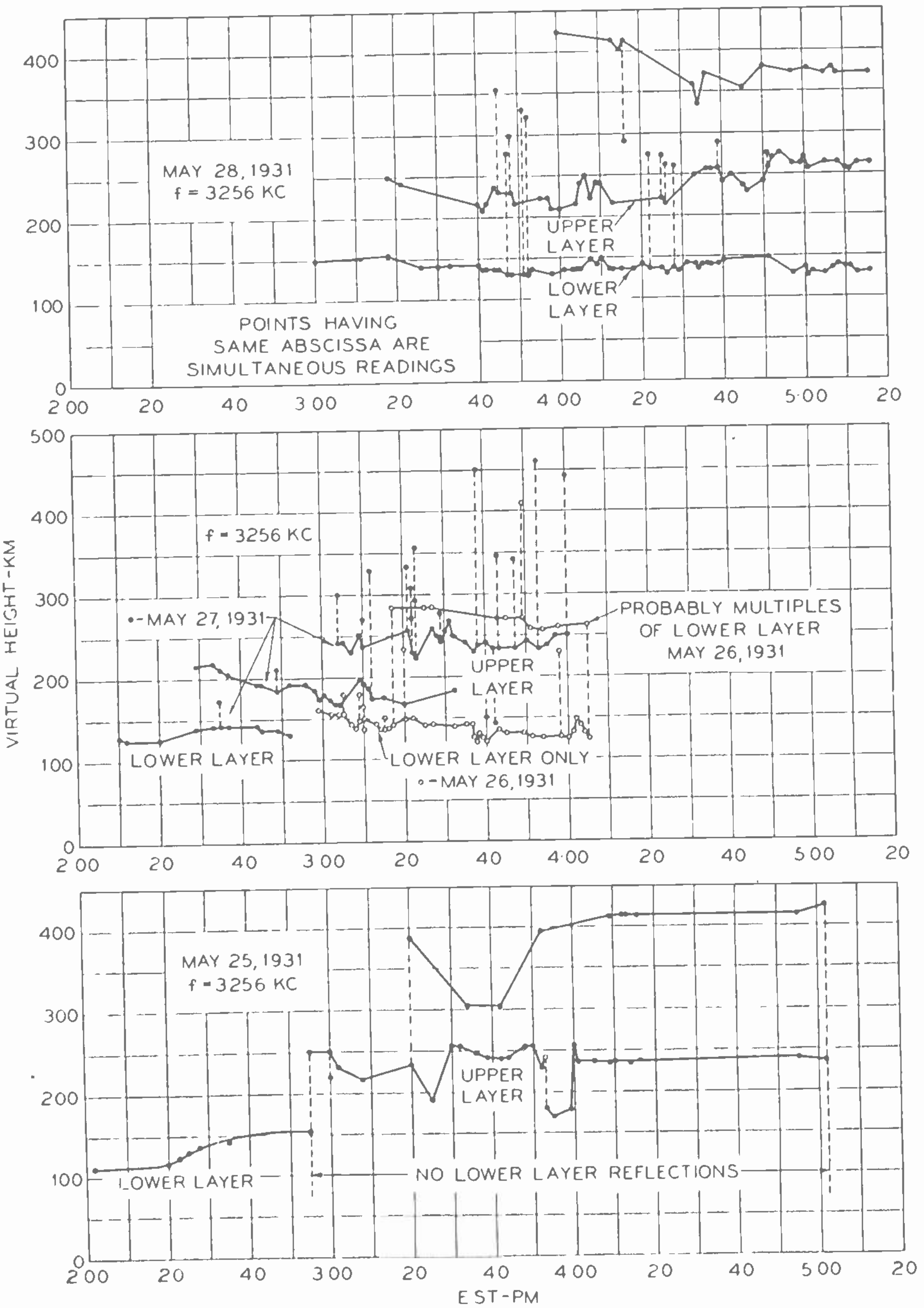

Fig. 4—Virtual-height variations during the afternoon for four successive days, May 25th to 28th, 1931, for a frequency of 3256 kc. Note the lack of duplication in the day-to-day virtual-height values.

[8] See Fig. 5 of the paper referred to in footnote 1. The present method of measurement is a vast improvement over the old method for such cases.

layer reflections may consist of a series of these split peaks having many values of virtual heights between 250 and 700 km. In general if the upper-layer reflections consist of many split peaks, the reflections having virtual heights greater than 700 km are either weak or absent. These facts suggest that when conditions are suitable for a large number of split peaks, they are not suitable for multiple reflections from the upper layer.

Another phenomenon which is observed at times in connection with split peaks is as follows: One of the peaks in a group will disappear for a time, returning with no measurable change in virtual height. At other times, first one and then another of the peaks of a group will become much stronger than the others, and when the full gain of the receivers is not used it appears as though the virtual height of one peak is changing rapidly, while in reality the virtual heights of both peaks remain constant and the apparent change is due to amplitude fading. When the amplitudes of the various peaks are changing fairly rapidly it is easy to be sure that such a change as that described above takes place. It is also believed that the same phenomenon takes place when the amplitude changes are much slower. For this reason it is important to make continuous observations in order to be certain that an actual change in virtual height is really taking place.

These split peaks are not inconsistent with the two-layer hypothesis. Thus they may be due to internal reflection within either layer, or to reflection between the two layers, or to combinations of both types of reflection. Since it is believed that total reflection should not take place in either layer, this type of reflection phenomenon might reasonably be expected. (See section (6) of experimental results.)

It is probable that some of the split peaks are due to double refraction caused by the earth's magnetic field. At times, what appears to be a split peak is really due to the coexistence of a multiple reflection from the lower together with a single reflection from the upper layer. It is also possible that these peaks may be due to turbulent conditions or to horizontal changes in the layers. Turbulence appears to be the cause of the many peaks which are observed during magnetic storms. (See section (12) of experimental results.)

(10) Critical Conditions of Ionization

As the ionization in the layer changes, a critical condition is found for which the virtual height varies rapidly with time for one frequency, while it is practically constant for other frequencies. At such times the measured virtual heights for the critical frequency may be several times the actual height of penetration due to the slow group velocity of the wave in the medium. No critical condition in the lower layer which causes any marked change in lower-layer reflections at or near sunset has been observed, except that the reflection amplitude has been found to increase as night approaches. This indicates different conditions than were found in the spring of 1929.[1] The reasons for this difference are probably the same as those given in Section (6) to explain the simultaneous reflections from two layers.

A critical condition in the upper layer has often been observed for the higher frequencies (3256, 4795, and 6425 kc). This is evidenced by a rather rapid increase in virtual-height values up to about 600 km followed by an absence of reflections for a period of several hours. When reflections are again observed the reverse of this process occurs, beginning with weak reflections involving large virtual heights followed by a rapid decrease in height to normal upper-layer values. The time of day at which these critical conditions occur depends upon the frequency which is being used and upon the state of ionization in the layer on the particular day in which tests are being made. These conditions have not been found during every test, but on certain days the virtual heights remained fairly constant at the upper-layer value of 300 km. For the lower frequencies the critical period mentioned above has not been observed for usual conditions of ionization, except for a few occasions near sunrise. Fig. 3 is an example of a critical condition for a frequency of 2398 kc.

When only a single reflection is present, and that reflection indicates large virtual heights, (600 to 700 km) the amplitude is invariably weak, indicating a critical condition. Weak reflections showing these large virtual heights have also been observed simultaneously with those indicating normal lower- and upper-layer values. This type of reflection may be due either to reflections between the two layers, or to reflections within the upper layer. These weak reflections might also be due to a condition in the lower layer which causes abnormally slow velocities for the wave which penetrates this layer. In no case, however, have strong reflections indicating virtual heights above 400 or 500 km occurred when these strong reflections were not some multiple value of a preceding echo.

(11) *Meteor Shower Observations*

Observations have been made during four meteor showers. Unusually disturbed conditions have been found with intermittent periods of increased ionization in the lower layer during the night hours (11 P.M. to 5 A.M.), but no definite conclusions as to the effect of such showers can be made at the present time, due in part to the fact that mag-

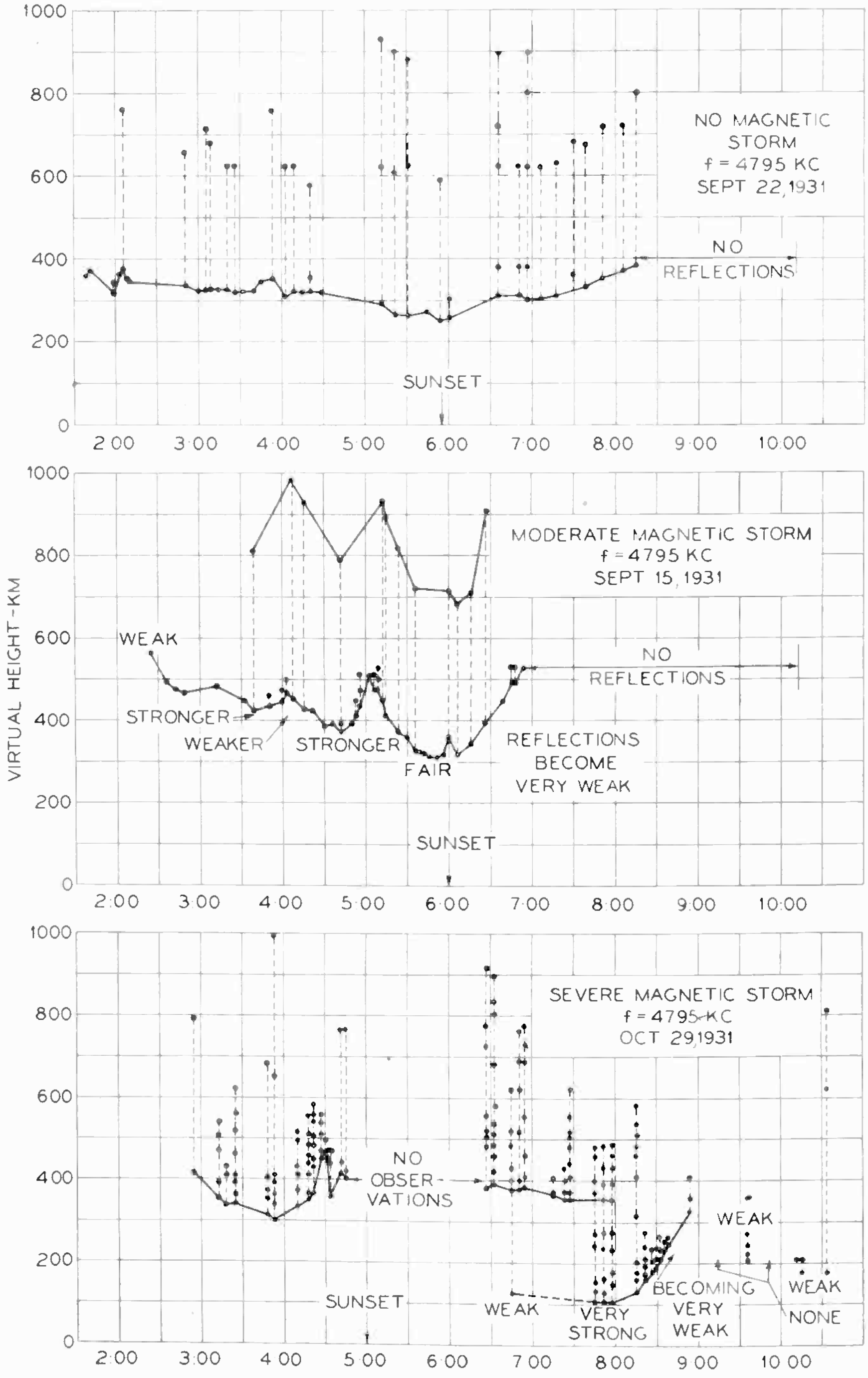

Fig. 5—Virtual-height variations for three days during which the magnetic conditions were different. Note that the virtual heights become much more irregular as the magnetic disturbance increases.

netic conditions were moderately disturbed during three of these tests. A more detailed account of experiments conducted during the Leonid shower of November 1931 will be published later.

(12) Magnetic Storms

Observations have been made during several magnetic storms. The following tendencies have been noted:

On those days when the magnetic disturbance was not very pronounced no marked peculiarities in the virtual heights were obtained.

During moderate magnetic disturbances the virtual height of the lower layer seemed to decrease while that of the upper layer seemed to increase.

During severe magnetic storms the data suggest a condition of great turbulence in the ionized regions. In marked contrast to the slow changes ordinarily found, the retardations of individual reflections were found to change in an erratic manner. Large numbers of weak reflections were sometimes obtained, representing virtual heights between 70 and 600 km with only 20 or 30 km separation between peaks. For such cases it is unusually difficult, if not impossible, to draw a line of demarcation between reflections due to the upper layer and those due to the lower layer. In fact we are not justified on the strength of these data in assuming that only two distinct reflecting regions exist under these particular conditions.

Fig. 5 shows examples of virtual height variations for different magnetic conditions.

(13) Limiting Frequency

Reflections for 6425 kc[9] have been received only during daylight hours, and then not for very long periods. Usually the virtual heights were of the order of 300 to 350 km with a rapid change from and to heights of over 700 km at the beginning and end of the period. This indicates that the ionization in the upper layer is only sufficient to cause reflections at this frequency during a limited period of time for normal incidence.

Reflections from the lower layer were obtained on only one occasion and then at a virtual height of 160 km. This indicates that the proper condition of ionization to give reflections from this layer at normal incidence is found on very rare occasions.

(14) Seasonal Effect

For the period covered by these tests there have been no striking

[9] As mentioned before, measurements using this frequency were begun in September.

changes in reflection phenomena which could definitely be attributed to seasonal variations. The following tendencies have been noted:

(1) Beginning in October there appears to have been a decrease in the amount of daytime absorption in the region of the lower layer.

(2) The time at which critical periods occur has changed to correspond to the changing time of sunrise and sunset.

(3) It is also possible that there has been some change in virtual heights, but because of the day-to-day fluctuations which are ordinarily found a more detailed study of the data will be necessary in order to give a satisfactory answer to this question.

Discussion and Conclusions

From a study of the results presented above it appears that certain phenomena have been observed, for transmission at normal incidence to the layers, which would not be expected from the simple ray treatment of the problem. For example, the occurrence of simultaneous reflections from both layers would not be expected if the layers were perfectly regular in the horizontal direction. Simultaneous reflections have actually been found, and from the frequency of occurrence and duration of this phenomenon, it seems unlikely that horizontal irregularities could have been responsible for all of the observed results. A similar statement can be made for the large number of split peaks often observed.

From the preliminary study of wave propagation previously mentioned it appears that simultaneous reflections from both layers would be expected for normal incidence. Under certain conditions, for angles of incidence different from normal and for frequencies in the range of those used in these tests, the waves which penetrate the lower layer during the period when reflections are obtained from both layers should be attenuated with increasing angles. For this reason, and others, it appears desirable to conduct simultaneous reception experiments over several base lines.

The widely varying values of virtual height and number of reflections obtained make it difficult, with the amount of data now available, to draw definite conclusions as to the probable composition of the ionized layers. The most important characteristics, however, are in agreement with the following hypotheses:

(1) There are two reflecting layers or regions. A lower layer at a height of approximately 100 km and an upper layer at a height between 200 and 300 km.

(2) There may be an absorbing region below the lower layer. This is indicated by the fact that the amplitude of reflections from the lower

layer sometimes show a gradual decrease with time and then disappear without any noticeable change in virtual height.

(3) The presence or lack of multiple reflections is governed by the amount of absorption in the ionized regions, and not by reflection loss at the ground. The ground conditions are practically constant, and since large numbers of multiple reflections are sometimes obtained, it follows that the reflection loss at the earth's surface is small if not negligible. Since as many as ten reflections from the upper layer have often been received, the last of these having traveled about 5000 km, it would appear that in long-distance transmission the earth is a smaller cause of absorption than the ionized layer, and that propagation occurs by multiple reflection rather than by the single-hop method favored by Pedersen.[10] Although these results have been obtained for normal incidence it is believed that similar results would be found for other angles of incidence and higher frequencies.

(4) The virtual height of each layer increases with frequency as would be expected from the assumption of increasing ionization with height. Critical conditions of ionization are sometimes present, causing a rapid though continuous change in virtual height. The unusually large virtual heights (500 to 700 km) found during these critical periods are probably due to a slow group velocity in the medium, rather than to an actual penetration to these heights.

(5) There is a region of maximum ionization in the upper as well as in the lower layer. Above this upper maximum point the ionization probably decreases and no experimental evidence has been obtained to indicate that the ionization increases again. As has been pointed out previously[1] there is probably a minimum point of ionization between the two layers.

A point of maximum ionization in the upper layer is indicated from the observation of unusually large virtual heights for a critical frequency.

(6) The lower layer (100 km) is present during both the day and night. The ionization often increases at night, and occasionally even attains a value comparable with that which occurs during the middle of the day.

(7) The ionized layers are greatly disturbed during severe magnetic storms and at such times there does not seem to be any distinct separation between the upper and lower layers. The ionization in the region of the lower layer increases and changes in an erratic manner. The erratic behavior and lack of strong multiple reflections together with the

[10] P. O. Pedersen, "Propagation of Radio Waves," Chap. XI, pp. 179–218.

increased absorption observed at such times probably prevent successful high-frequency long-distance transmission.

Acknowledgment

We are indebted to Mr. J. C. Schelleng for many helpful suggestions and criticisms in connection with these studies; to Mr. A. M. Skellett for suggesting the measurement of layer heights during meteor showers and for data upon magnetic and solar phenomena, and earth currents; and to Mr. E. C. French for assistance in the experimental work.

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TRIPLE-TWIN TUBES*

By

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Summary—*This paper describes a new tube and circuit which utilizes the positive, as well as the negative, region of the E_g-I_p characteristic and has a negligible amount of distortion. A high input impedance is maintained. Minimum distortion and maximum output occurs at the same load impedance. The tube comprises in one envelope two sets of special triodes. The cathode of the input set is directly connected to the grid of the output. The first section supplies the power demanded by the grid of the second section. Although this paper only treats audio amplification, the development is not limited to this field.*

The fundamental circuit is described. Analysis of the "grid current compensation" is treated from data obtained by practical application rather than a mathematical evolution. Operating notes and a push-pull circuit are also discussed.

A commercial triple-twin's output and sensitivity is compared with a pentode and a triode. All have the same plate voltage rating. The triple-twin delivers nearly twice the pentode's power and three times the triode. Its power sensitivity is many times greater than its contemporaries.

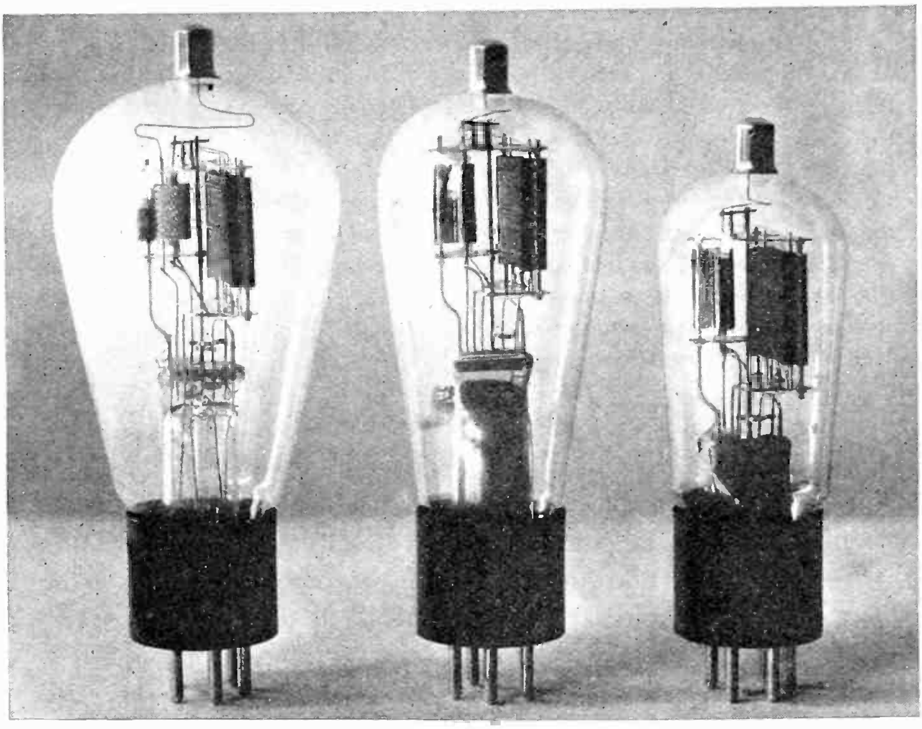
Introduction

An attempt to meet the demand for increased audio power, requiring neither higher voltages nor multistages led to the development of the pentode. Now the triple-twin tube with its circuit introduces a method for obtaining approximately twice the power and many times the sensitivity of a pentode at the same plate voltage. It has the additional advantage of operating into a load nearly equal to its own low internal impedance.

In the customary voltage or power amplification systems, the excitation signal is restricted to the negative region of the grid voltage, plate current characteristic. This restriction is evident as incursions into the positive region produce a grid current flow. If bias shifting is minimized by limiting the amount of resistance in series with the grid, the grid current still creates an undesirable condition, as the varying input impedance reflects a changing load to the driving source. This means that to prevent distortion of the signal the driver must produce a voltage which is independent of the load. Since the reflected load changes during only a part of the cycle, it is difficult to provide such a driver capable of supplying an undistorted wave form to the input of the amplifier.

* Decimal classification: R339. Original manuscript received by the Institute, March 1, 1932. Presented before Twentieth Anniversary Convention, Pittsburgh, Pa., April 8, 1932.

The problem of dealing with grid current by using two triodes and employing a push-push system with the grids normally biased to plate current cut-off ([1]class B) is becoming popular. This system, in its present form, has definite limitations and complications. An additional power tube is necessary as a driver. The amplitude of the peak signal required to load the tubes is great, consequently the sensitivity is low. The speaker load must reflect an almost constant impedance to the push-push plates. Distortion at low power level becomes objectionable



Speed triple-twin tubes. Type 295 for a-c operation; Type 291 for d-c, 110-volt operation; Type 293 for automobile receivers.

especially when the tubes are unevenly matched. Its use is confined to a plate supply whose voltage is practically constant and independent of load current.

The purpose of this paper is to introduce a tube and system, known as triple-twin, that utilizes both the positive and negative region of the E_g-I_p characteristic curve, and still maintains a high input impedance. The problem of the voltage supply system is identical to usual amplification as the total static power delivered to the plate is in excess of that demanded with maximum excitation. Although the chief objective of this development is to widen the scope of audio amplification, it is by

[1] Loy E. Barton, "High audio power from relatively small tubes," Proc. I.R.E., vol. 19, July, (1931).

no means limited to this field. Detector-amplifier action is being successfully employed. This problem will be treated separately in a subsequent paper. Its application to radio-frequency amplification merits investigation.

Combination tubes previously presented have been unfavorably received because of the coupling methods employed and the necessity for an increased number of terminals. Direct coupling and lack of spurious intercoupling makes the triple-twin a practical tube electrically. In the system to be described two triodes will operate as well as a triple-twin provided that the characteristics of the dual tube system are the same as those of the triple-twin system.

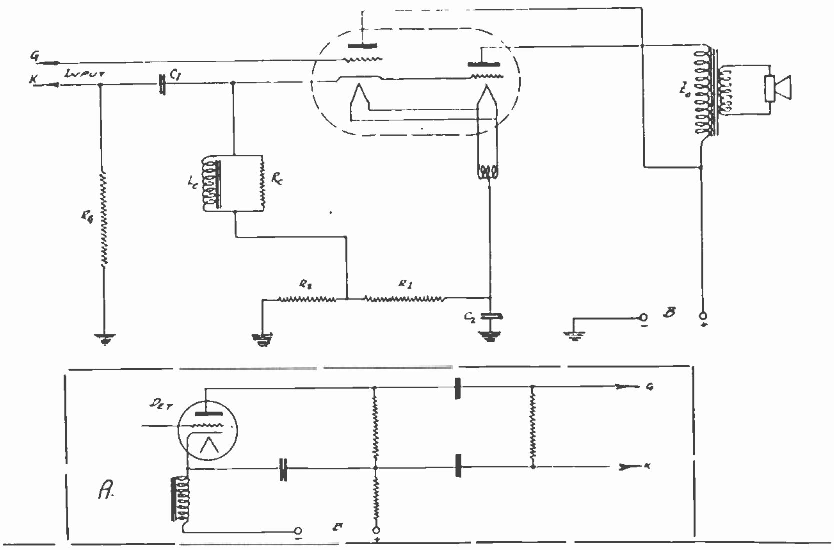


Fig. 1—Fundamental a-c triple twin amplifier circuit. Insert—suggested resistance coupled circuit to the input of the triple-twin for television.

No attempt will be made to give a mathematical treatise of this system's functions and characteristics.

The data available are empirical, and it is not felt that a mathematical treatment would be justified at the present time.

Fundamental Circuit

The triple-twin tube consists of two sets of three elements; the first set handles the input, and the second, the output. The object of the input section is to present at all times a high input impedance and output characteristics that will supply the power demanded by the output grid. This section employs an indirectly heated cathode which electrically isolates it from the output filament. This cathode is internally connected to the output grid.

Referring to Fig. 1, the fundamentals of the circuit will be discussed. The input of the first section is similar to usual operation, as this grid does not take current, but it differs in that the cathode is above ground potential. This means that the applied signal voltage must also be above ground potential. The signal reaches the cathode through a small condenser C_1, which offers a low impedance to the incoming signal. The grid receives its bias by the d-c drop in the load impedance of the first section (L_c) and the IR drop in resistance R_2. The d-c return path to this grid is through resistance R_g. It is signifi-

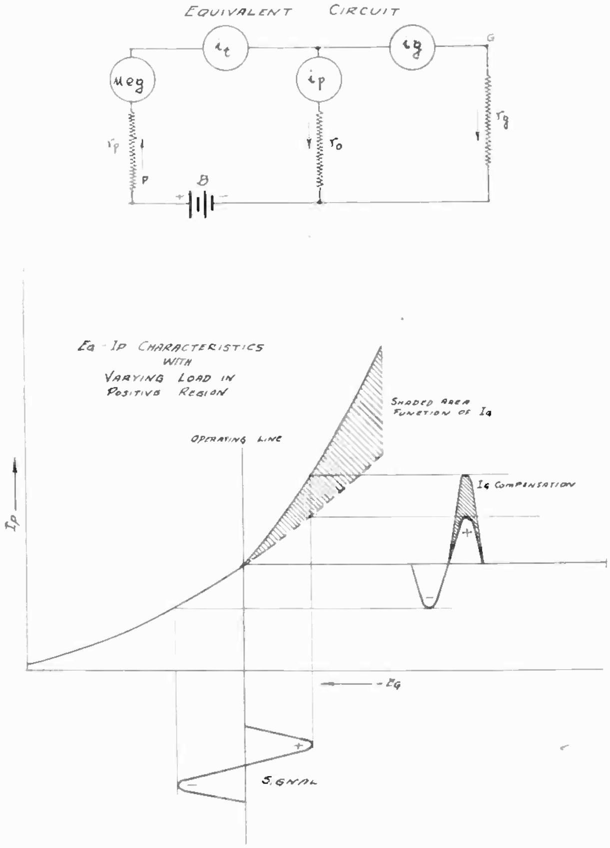


Fig. 2—Equivalent circuit and action of grid current compensation.

cant that the load impedance of the first section exists between cathode and ground and is substantially the combined parallel value of resistance R_c and the grid impedance of the second section. The inductance L_c is shunted across this combination but its impedance is high, except at low frequencies, compared to the other values, and its function is to allow a low d-c resistance path for grid and plate returns. Its d-c voltage component also augments the voltage drop in R_2 and opposes the drop in R_1, but this effect is negligible as the resistance of its winding is small. Resistance R_1 establishes the grid of the output section several volts negative and is necessary only in a-c operation to suppress hum. Condenser C_2 by-passes the audio frequency in these resistors, preventing intercoupling between plate and grid. The plate circuit of the second section is identical to triode operation.

Theory of Compensation

The simplified equivalent circuit, Fig. 2, is useful in analyzing the triple-twin. μe_g and r_p represent the generator action and internal impedance of the input section. Its grid action is eliminated as it has no bearing on the grid current compensation analysis. The effective load impedance offered by R_c and L_c is represented by r_o. For simplicity, the frequency-impedance characteristic of the combination will be neglected, and consequently, r_o will be considered constant. r_g is the grid impedance of the second section. μe_g also will be considered constant. The currents in the circuit are represented by $i_p + i_g = i_t$, where i_p is the current through the load r_o; i_g grid current of the output section; and i_t total plate current. Hereafter, we shall consider these as instantaneous peak values and disregard the steady d-c component in i_t and i_p. The impressed signal will be assumed a sinusoid.

The object of the compensation is to supply power to the grid circuit of the second section when required and to establish the voltage applied to this grid independent of varying impedance. We will assume that when the grid of the output section swings positive, it draws current. Then its impedance cannot be considered constant, but some function of the positive cycle of the voltage developed between cathode and ground. Consequently, r_g is a varying value which, in effect, when combined with r_o, offers a varying load to the plate characteristic of the input section

Now if $r_p = 0$ (neglecting the internal resistance of battery B) the addition of i_g to the load r_o would not affect the voltage across r_o. As $r_p \neq 0$ and if $i_p r_o$ is to remain constant, r_p must vary. Since $i_g = f(+i_p r_o)'$ and the positive values of i_p should equal the negative, the rate of change in r_p must always satisfy the relation $i_p = i_t - i_g$. Therefore, the compensation must operate in considerable curvature and nonparallelism of the plate characteristics in order to obtain the necessary variations in r_p.

To approximate the compensating action graphically, the two sections are treated individually. The usual family of plate characteristics for the input section is plotted, Fig. 3. Load line AOB represents r_o with O as its operating point. Its slope is obtained by the customary method, triangle DOC. Now if we assume that the grid is taking current whenever the signal is positive, its impedance will be reflected on only one side of the operating line. The compensation is represented by the shaded area, and it appears only at the left of the operating point O. The significance of this position is that the output voltage of the first section is in phase with its plate current; this is opposite to usual amplifiers where there is a phase difference of 180 degrees. This is due to the load existing between cathode and $B-$. The reference level for the pulsating current in the plate circuit is, therefore, reversed. Of course, the plate current, in both cases, is in phase with the grid signal voltage. The significance of this relation will be shown later. In determining the true phase relation, the reactance of the load circuit must be taken into account. Any point along OB such as C represents a plate

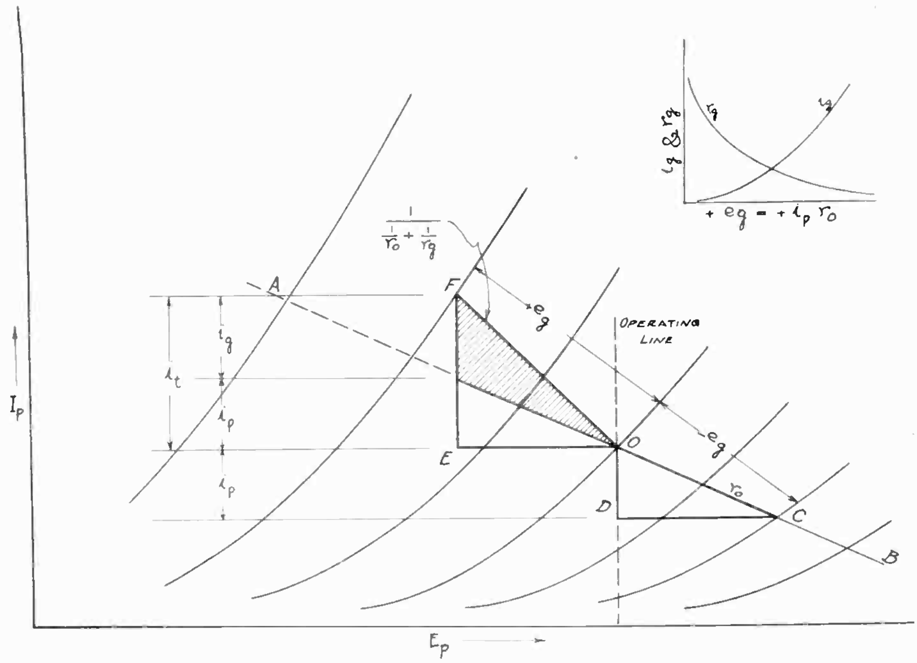


Fig. 3—Grid-current compensation theory. Plate characteristics of input section.

current whose instantaneous negative magnitude is OD or i_p; and this is produced by an input signal $-e_g$ whose magnitude swings the grid from its operating point O to the point C. Then CD represents the negative peak of the voltage across the load. As the signal is a sinusoid, the positive peak of the output voltage should equal the negative. With the voltage value CD, the instantaneous i_g and r_g can be obtained from the dynamic input impedance curve of the second section. This curve must be taken with proper load impedance in the plate circuit, as r_g is not independent of the plate voltage. Since at all times $i_p+i_g=i_t$, the compensation must produce changes in i_t when $i_g>0$. Let F be the point where the abscissa of i_t intersects a characteristic to the left of the operating point, whose grid voltage is determined by the positive signal swing $+e_g$. Now, if the compensation is to be satisfied, the slope of the new triangle EFO must represent a resistance equal to $[1/((1/r_p)+(1/r_o))]$. Thus the positive output voltage peak represented by the base of the triangle EO equals the negative, CD. Many other points, like F, are obtained by the same procedure. A line connecting these, forms a general contour which represents the dynamic condition of the load.

The result from Fig. 3 can be transposed to a dynamic e_g-i_p characteristic of the input section, as in Fig. 2. It should be understood that both Figs. 2 and 3 are arbitrarily drawn, merely for representation, and they cannot be transposed graphically. The grid current

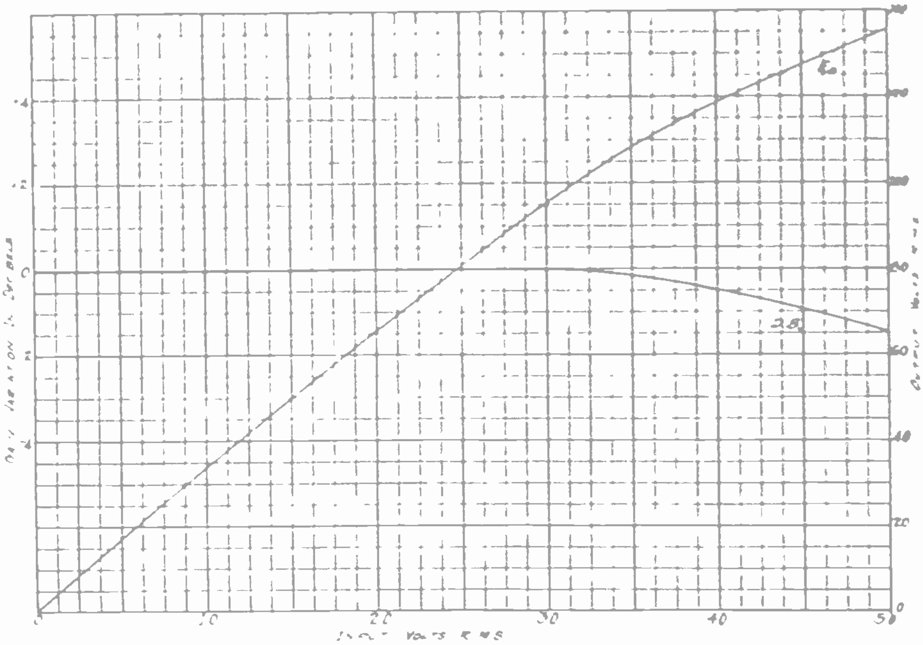


Fig. 4—Voltage output and gain variation versus input voltage. The slight loss in db represents nonlinearity of the e_o-e_i characteristic.

peak is shown with good wave form. In reality, the nonlinear shape of the e_g-i_g characteristic alters this form. The grid current is a pulsating half wave at signal frequencies. The grid current does not affect the steady static condition of the output grid, because this current's path is not through r_o.

It is evident that the compensation is possible because of the peculiar phase relation. If this relation was as usual, the grid load would be reflected on the negative section of the operating line where the "fanning" would oppose the desired action.

From the foregoing discussion, it becomes obvious that the plate impedance of the input section varies with the magnitude of the grid current. Therefore, the arrangement is not "matched" but a "self-compensating" device.

In practical application, the compensation can diverge considerably from ideal conditions without any appreciable effect in over-all results. In fact, the test spread used in production of ordinary triodes can be successfully used with triple-twins. This freedom from the necessity of exact matching is partially due to the proper design of r_p and r_o. They should be of such magnitude that small variations in r_g will have little effect on the voltage across r_o, irrespective of the compensation. The nonparallelism of the characteristics produces some

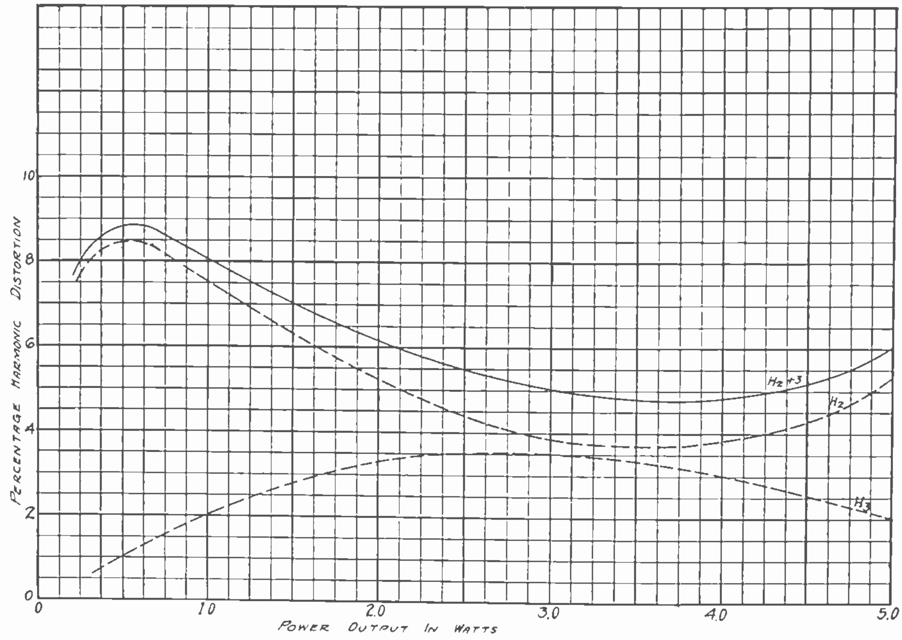


Fig. 5—Typical triple-twin harmonic distortion $\left(\frac{E_h}{E_f}\times 100\right)$ versus power output characteristic. Notice predominance of second harmonic. Speed 295.

amplitude distortion which means that the output voltage does not vary linearly with the input. However, this divergence from linearity will be at relatively high power levels where the ear sensitivity is quite low. In Fig. 4, the over-all effect of amplitude distortion for a typical triple-twin is shown.

e_o varies approximately as the first power of e_g up to 110 volts. This voltage represents an output of 3 watts. From this level to 5 watts, there is only 1.5-db loss.

Fig. 5 shows percentages of total distortion and of individual harmonics plotted as a function of power output. An increase in distortion occurs at low power levels, due to incomplete compensation. This distortion peak is below 9 per cent and, as it is chiefly second harmonic, it is probably not perceivable to the average ear. If the value of R_c is reduced, the peak may be lowered at some sacrifice in sensitivity. With a peak total distortion of 6.5 per cent, the sensitivity is down approximately 20 per cent. The second harmonic predominates throughout which is a decided advantage in push-pull applications where the even harmonics cancel.

Output Load Impedance

The output of the second section is comparable to a usual power triode. Consideration of the output load impedance naturally involves

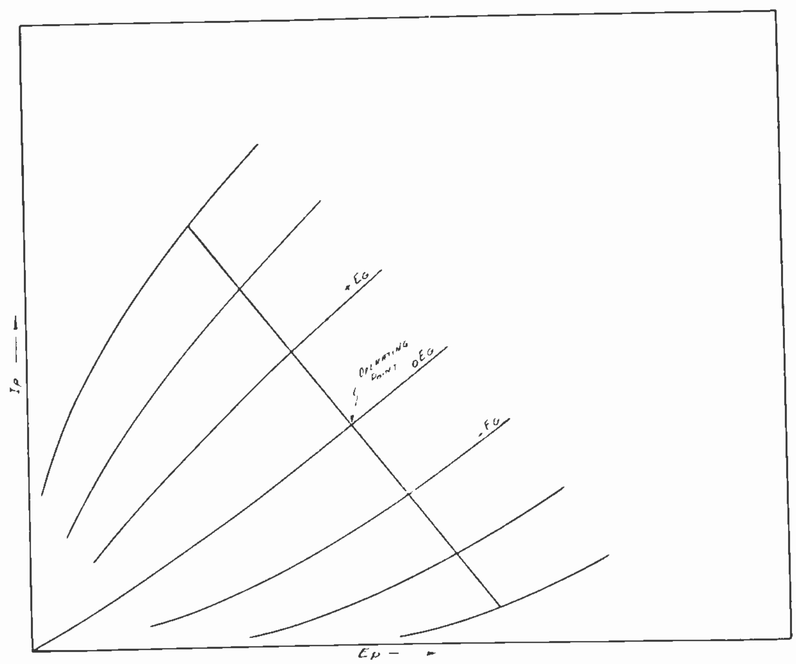


Fig. 6—Explanatory output plate characteristics. Illustrates "fanning" in both positive and negative regions of the output plate characteristics.

the curvature of the output plate characteristics. In the triple-twin, the operating point falls nearly in the middle of the widest spread. Fig. 6 graphically shows the operating point at zero bias, and this point is the center for the load line. Near the negative cut-off region, the "fanning" is, as usual, appreciable. Near the positive region the curves have the same shape and spacing but are diagonally opposite in curvature. This latter phenomenon is somewhat complicated, but is chiefly caused by the crowding effect of electrons arriving at the grid, and the space-charge produced by grid secondary electrons. This condition restricts the number of electrons arriving at the plate. Of course, emission saturation will bend the curves, but this becomes negligible when employing oxide-coated emitters that are designed to supply the demand abundantly. If both curvature and nonparallelism are equal, but one group diagonally opposite in curvature to the other, and the axis of the load line is on zero grid bias, the extremities of the load line may extend equally into both regious. Therefore, the internal impedance may be designed so that the load for minimum distortion will nearly equal this value. These conditions also permit maximum energy transfer.

Amplitude distortion resulting from characteristics of this type is not necessarily serious. The over-all divergence from linearity is given

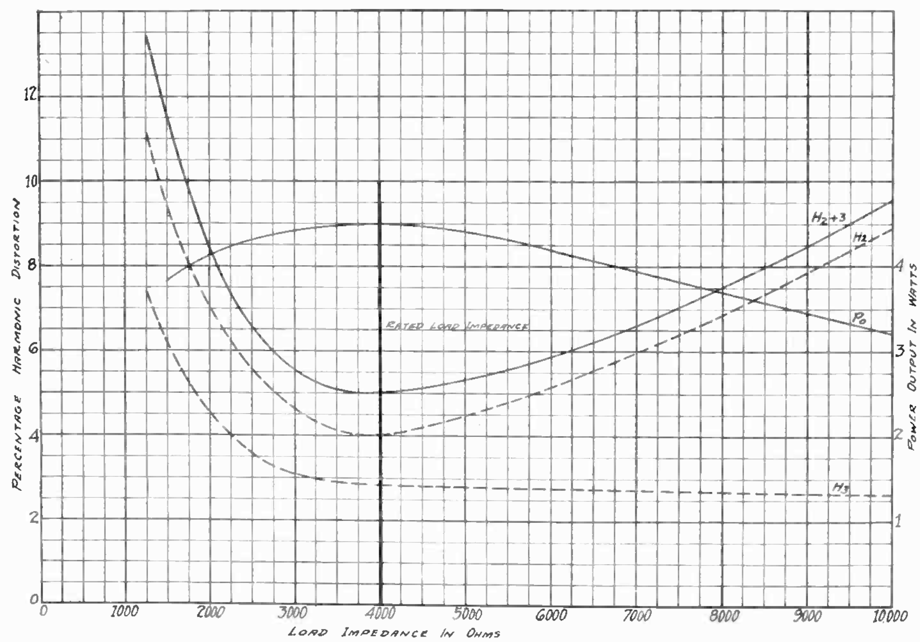


Fig. 7—Curve showing maximum power output occurring at the same load as minimum distortion $\left(\dfrac{E_h}{E_f}\times 100\right)$. Notice predominance of second harmonic. Speed 295.

in Fig. 4. The significance of this curve has already been discussed under "Theory of Compensation." It should be mentioned that care must be exercised in obtaining a family of static plate characteristics because of grid heating. This produces an unnatural condition that does not appear in dynamic operation where the heating is reduced by at least one half.

In Fig. 7, the over-all power output and harmonic distortion is plotted as a function of load impedance. A constant input signal was used that gave full power at rated load. Maximum power occurs at 4000 ohms, and at this value the distortion is at a minimum. Note that the second harmonic predominates, and the third is constant at higher load values. This relation is again ideal in push-pull applications. The output versus load characteristic is comparable with triodes, but the percentage of power loss for the new tube is even less.

Frequency Response

The frequency response of the typical triple-twin is shown in Fig. 8. The frequency-impedance characteristic of the coupling choke L_c is not important because of the relatively low shunt resistance, (R_c). The impedance of the choke would be a factor in any case only at low frequencies. Its distributed capacity is no appreciable shunt to the resistances. It should be observed that there is no appreciable power loss at high frequencies, even those used in television, to the order of 50,000

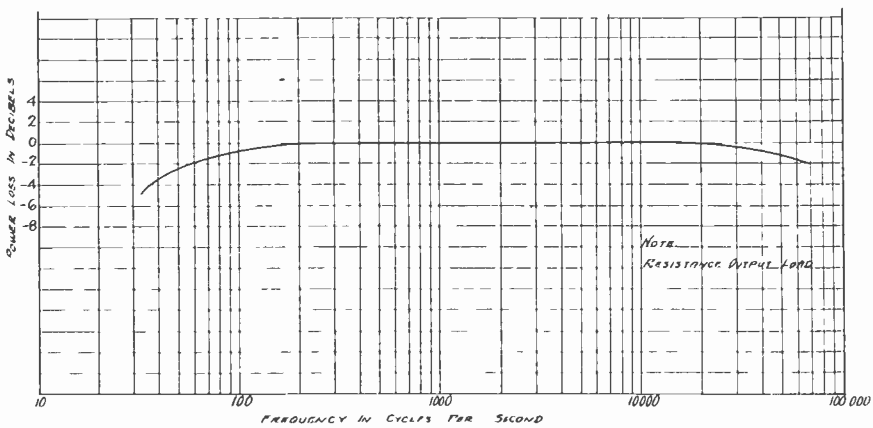


Fig. 8—Triple-twin fidelity curve. Speed 295.

cycles. The choke used in obtaining this curve had an inductance of 20 henries at 60 cycles, and a d-c resistance of 192 ohms. The direct-current saturation, of course, can be small as only the low plate current of the first section passes through this choke. The shunt resistance (R_c) was 12,500 ohms.

The low register is subject to the effectiveness of the by-pass condensers C_1 and C_2. However, the usual problem when employing high gain tubes, that of eliminating grid-to-plate coupling, is not troublesome as the high over-all gain is divided between the two sections. The value of condenser C_1 was $2\mu f$. Condenser C_2 is the common by-pass for both biasing resistors. Although the gain in the last section is somewhat greater than the usual power triode, the bias resistance value is much less. Consequently, the capacity for effective by-passing can be compared with triode operation. C_2 was a low voltage electrolytic condenser of $25\ \mu f$.

It should be realized that these capacity values may be reduced without serious loss in low response. No appreciable advantage is obtained by individually by-passing R_1 and R_2 as the ohmic value of R_1 is small.

Comparisons with Contemporary Tubes

In Table I a typical[2] Speed triple-twin type 295 is compared with the pentode 247 and triode 245. These tubes were compared as they all have the same plate voltage rating. The efficiency of the tubes is given by the ratio of the a-c power output to the d-c power input. The product of the total voltages, that is, plate voltage plus grid voltage, and the total currents, plate currents, and screen-grid current, is used for the power input. This method differs from that usually employed, but since most applications for these particular tubes use a self-biasing system, the power dissipated in the bias resistor should be regarded as part of the power input. The grid power in the triple-twin's second section is not represented because it is an instantaneous value, and the

TABLE I

		Total Drain				Per Cent Efficiency	Sensitivity
Type	E_p	$E_p - E_g$	$\times (I_p - I_{sg})$	P_o	R_o/r_p	P_o/P_i	$P_s = \sqrt{P_o}/e_g$ (r-m-s)
Triple-Twin 295	250	267	56	4.5	1.33	30	0.42
Pentode 247	250	266.5	39.5	2.5	0.2	23.7	0.15
Triode 245	250	300	34	1.6	1.95	15.7	0.046

sum of the instantaneous power demands is always less than the dissipation at no signal periods. The power sensitivity is calculated by Ballantine's[3] equation, $P_s = \sqrt{P_o}/e_g$. Even though the sensitivity comparisons are not made with equal power ratings, these values are useful. The high sensitivity of the 295 is strikingly evident. It becomes even more significant when it is realized that only a 5-volt signal is required to produce 4.5 watts output.

It is well to point out the reasons for the high efficiency and sensitivity of the 295. For a given power output, when the input signal is confined to the negative portion of the $E_g - I_p$ characteristics due to grid current limitation, as in triodes, the anode voltage to produce this output must be high to draw the electrons through the negative field produced by the heavily biased grid. In pentode operation, only the negative portion of the characteristic is used, but a positive auxiliary grid reduces space-charge effect, improving efficiency compared to a triode.

[2] Manufactured by the Cable Radio Tube Corporation.
[3] Stuart Ballantine, Proc. I.R.E., vol. 18, p. 452; March, (1930).

However, this auxiliary grid consumes energy and a cathode grid becomes necessary to reduce eccentric characteristic curvature caused by primary and secondary plate electrons. To overcome the shielding effect of this latter grid, a higher plate potential is required. Further reduction of efficiency is caused by the necessity of operating into a load approximately one-fifth of the internal impedance of the tube. The triple-twin operates at approximately zero bias, so that the signal swings equally into the positive and negative regions. This tube works into a load almost equal to its own impedance. With this arrangement, the same power output can be obtained at considerably reduced plate voltage.

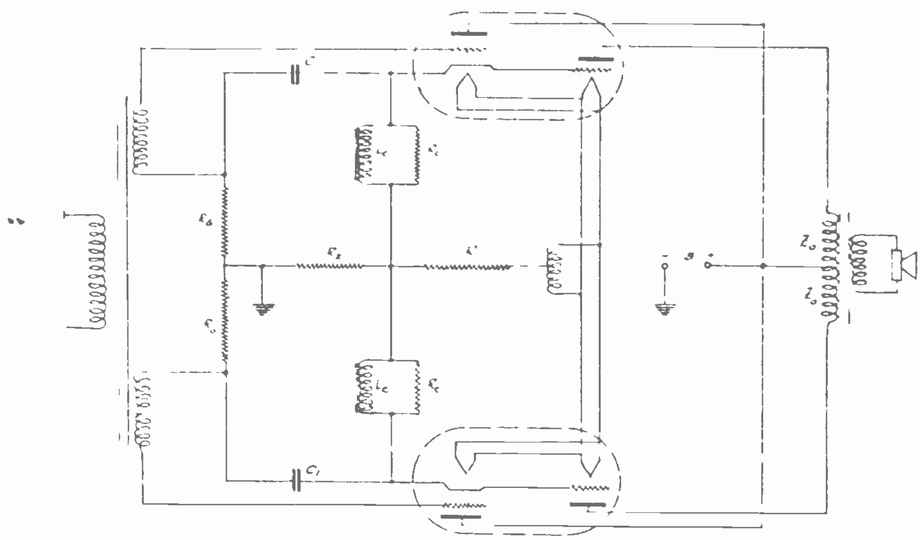


Fig. 9—Suggested push-pull circuit for triple-twin type 295. Note split secondary input transformer.

The power sensitivity of the triple-twin is high, due to the no-loss effects of direct coupling and the high gain in both the input and output sections. The effective grid area of the output section may be large, as the plate current is not limited by a strong negative field. This allows high amplification with a low plate impedance.

Operating Notes and Push-Pull

The fundamental circuit in Fig. 1 shows the input terminals above ground potential, that is, terminal K is above ground by the amplitude of the voltage developed in the combination L_c and R_c. When transformer coupling is used to feed the tube, the secondary winding is simply left above ground. In ordinary resistance coupling, the a-c voltage in the coupling resistance is developed between the high side (the end near the plate) and the B battery. In order to use this type of coupling with the triple-twin on a common B supply, without additional complications, condenser C_1 and resistor R_g are removed and R is terminated at G. Now the input to the tube is, as in usual amplifiers, between grid and ground. This arrangement is entirely satisfactory with respect to fidelity, but the sensitivity is reduced by the amount of the gain normally obtained from the input section. Even with this loss, the sensitivity is greater than the usual power triode. A special resistance coupling circuit, with its output terminals above ground, has been used with good success, particularly for television. In certain television applications, where high power is wanted over a wide frequency range, high sensitivity is most desirable. The method is given in Fig. 1-A. The coupling condensers offer a low impedance to the signals which are to be amplified. An inductance is used in the cathode circuit, so that the d-c resistance may be small.

A suggested push-pull system is shown in Fig. 9. The input transformer has a split secondary so that the individual inputs can be isolated and kept above ground. Tapped resistance R_g offers a d-c path for the respective input grid biases, which are obtained from the IR drop in L_c and R_2. The biases for the output grids are obtained by the drop in R_1 minus the respective drops in L_c.

Conclusion

The development of the triple-twin offers a new solution for many design problems, as the circuit is simple and flexible. While the subject has been considered generally with specific reference to an a-c tube, two types of d-c tubes are also virtually completed. One is designed for 110 volts d-c operation. In this field, the problem of obtaining high audio power has been difficult. The new tube delivers four times the power of a commercial pentode, operating at this voltage. The second tube is designed for a higher voltage, but with a lower current consumption. It particularly lends itself to automobile and aircraft receivers.

Tubes designed for high power used in public address and sound picture systems should effect material economy in the cost of the apparatus, as prestages and voltages can be considerably reduced.

Acknowledgment

In conclusion the writer wishes to express his appreciation for the assistance of the engineering organization and particularly that of Messrs. Paret and Bair who collaborated in the solution of many of the problems encountered.

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TRANSMISSION LINES FOR SHORT-WAVE RADIO SYSTEMS*

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Summary—*The requirements imposed on transmission lines by short-wave radio systems are discussed, and the difference in the requirements for transmitting and receiving purposes is emphasized. Various line types are discussed, particular attention being given to concentric tube lines and balanced two-wire lines. The concentric tube line is particularly valuable in receiving stations where great directional discrimination is involved and low noise and static pick-up is required.*

Excellent agreement between calculations and measurements is found for the high-frequency resistance of concentric lines, using the asymptotic skin effect formula of Russell. Other losses in correctly designed concentric tube lines are found to be negligible. Measured losses in two-wire lines are found to be greater than losses predicted by the asymptotic skin effect formula owing, in part, to losses brought about by unbalanced currents.

Practical aspects of line construction such as joints, insulation, and provision for expansion with increasing temperature are discussed.

Some difficulties encountered in transmission line practice, such as losses due to radiation, reflections from irregularities, effects of weather, and spurious couplings between antenna and line are discussed.

I. General Requirements

THE transmission line systems employed for the purpose of transferring energy between radio units and antennas are fundamentally no different from line systems used in power or telephone work. Owing, however, to the high frequencies employed in radio transmission an operating technique differing from that found economical in low-frequency practice is necessary. An important consideration in 60-cycle power practice is that the voltage at the far end of the line be maintained constant irrespective of load variations. At radio frequencies a transmission line may be many wavelengths long and the reflections from a load other than one equal to the characteristic impedance of the line produce standing waves. Transmission losses in radio-frequency lines are appreciably augmented when the currents and voltages on the line appear in the form of standing waves. The operation of the radio unit connected to the line is sometimes affected by the presence of standing waves.

Induction and cross-talk problems familiar to every telephone engineer are increasingly important as line operation approaches radio

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concentric tube lines capable of transmitting several kilowatts of modulated radio-frequency power require special consideration.

The lines commonly employed in radio stations may be divided into four classes: single wire lines, balanced open wire lines, multiple wire lines, and concentric tube lines.

Single wire lines are of limited utility owing to the low efficiencies arising from the marked radiation characteristics of such wires. The power radiated by a single wire line several wavelengths long may be equal to that radiated by the antenna to which it is connected.[1] In fact single wire lines, particularly when terminated, are for certain services desirable radiating elements. Diamond-shaped arrays of such elements are employed in some of the radio facilities of the Bell System.[2]

It is generally appreciated that the power losses due to radiation may be reduced by employing two conductors in a go-and-return circuit, the wires being separated a small fraction of a wavelength. A necessary requirement is that the two wires carry equal currents exactly opposite in phase. Otherwise, there will appear current components which employ the two conductors in parallel. In the latter event the radiation losses ascribed to single wire conductors occur.

Although there is a very great reduction in radiated power in balanced two-wire lines as compared with single wire lines there are many practical cases where the radiation from two-wire lines produces cross-talk and loss of signal discrimination. Multiple wire lines comprising several pairs of conductors in go-and-return circuits may be employed to reduce the undesired radiation couplings. As in the two-wire case care must be exercised in maintaining the required current amplitudes and phases since otherwise the radiation losses ascribed to single wire lines may destroy the utility of the multiple wire system. Multiple wire lines, of course, reduce static and noise interference.

From the standpoint of isolation an ideal electrical connection between antennas and radio apparatus is approached when one conductor completely encloses the other conductor. A concentric tube line comprising an outer sheath and an inner conductor is the practical form of this construction. Long transmission lines often pick up a large amount of static and other electrical disturbances. Spurious couplings may introduce these disturbances into the radio circuit. Electrical disturbances so introduced are greatly reduced when the outer sheath of a line may be grounded at frequent intervals. In fact, concentric tube lines may be buried in the ground.

The effect of weather is a factor which in some instances may determine the type of construction to be employed in radio-frequency lines. It is generally appreciated that rain and sleet storms may materially lower the insulation of a line. The velocity of propagation and characteristic impedance also are affected by a coating of water or sleet upon the wires. Concentric tube lines may be constructed so as to be weatherproof.

[1] See calculations in the appendix.
[2] E. Bruce, Proc. I.R.E., p. 1406, August, (1931).

This paper will be confined entirely to concentric tube lines, to balanced two-wire lines, and to the apparatus associated with these two line types.

II. Concentric Tube Lines

A shielded line comprising an inner tubular conductor and an outer concentric shield is the form most commonly employed in radio practice. Owing to the circular symmetry of the line the case is capable of rather exact mathematical analysis. At radio frequencies the results are surprisingly simple. This simplicity is very evident for the two important parameters of a transmission line, the propagation constant P and the characteristic impedance Z_0.

The propagation constant of a line is defined by:

$$P = \sqrt{R + j\omega L} \cdot \sqrt{G + j\omega C} \tag{1}$$

in which

$(R+j\omega L)$ is the complex impedance, and

$(G+j\omega C)$ is the complex admittance, both per unit length. It is well known that the propagation constant is a complex number and that at radio frequencies (1) reduces to[3]:

$$P = \alpha + j\beta = \frac{R}{2Z_0} + \frac{GZ_0}{2} + \frac{j2\Pi}{\lambda}. \tag{1a}$$

In which R is the resistance and G is the leakage conductance, both per unit length and at the wavelength λ.

The characteristic impedance Z_0 is defined by the ratio:

$$Z_0 = \frac{\sqrt{R + j\omega L}}{\sqrt{G + j\omega C}}. \tag{2}$$

The characteristic impedance also is a complex quantity, but at radio frequencies it is for most practical purposes the real quantity:

$$Z_0 = \sqrt{\frac{L}{C}}. \tag{2a}$$

In the case of concentric tube lines the expression for the capacity C per unit length is the familiar relation:

$$C = \frac{1}{2 \log_e \frac{b}{a}} \text{ e.s.u.} \tag{2b}$$

[3] J. A. Fleming, "The Propagation of Electric Currents."

in which a is the outer radius of the inner conductor and b is the inner radius of the outer conductor. The inductance L per unit length may be obtained from an expression derived by Lord Rayleigh upon assuming that the two tubes comprising the line are of negligible thickness. This is permissible because at radio frequencies the conduction of currents is essentially a skin effect. Upon this basis the inductance per unit length of a concentric tube line becomes:

$$L = 2 \log_e \frac{b}{a} \text{ e.m.u.} \tag{2c}$$

Upon substituting (2b) and (2c) into (2a) with proper regard for units a simple expression for the characteristic impedance at radio frequencies is obtained:

$$Z_0 = 138 \log_{10} \frac{b}{a} \text{ ohms.} \tag{2d}$$

The high-frequency resistance of concentric tube lines has been treated by a number of investigators, notably by A. Russell.[4] The asymptotic formula for resistance as the frequency is increased without limit is:

$$R = \sqrt{\rho\mu f}\left(\frac{1}{a} + \frac{1}{b}\right) 10^{-9} \text{ ohms/cm} \tag{3}$$

in which,

- ρ is the resistivity in e.m.u. (for pure copper ρ is about 1730 e.m.u.)
- μ is the magnetic permeability
- f is the frequency, c.p.s.
- a is the outer radius of the inner conductor, and
- b is the inner radius of the outer conductor, the two latter being in centimeters.

It is of interest to note that the wall thickness of the conductor is not involved. At radio frequencies the current is confined to a very thin layer on the outside of the inner conductor and on the inside of the outer conductor.[5] The skin effect is, of course, not so pronounced at low frequencies and more complicated formulas involving wall thickness must be employed.

Some typical experimental data are submitted to show that for frequencies higher than one megacycle and for several practical line

[4] A. Russell, *Phil. Mag.*, April, (1909); and "Alternating Currents," vol. 1, p. 222, (1914), Cambridge Press.

[5] Frequency is not the sole criterion, resistivity, wall thickness, and diameter also being involved.

constructions the foregoing equation (3) holds with a very useful degree of accuracy. The physical dimensions and construction details of the lines for which the observations were made appear in Fig. 1. With the exception of one rubber insulated line all inner conductors were

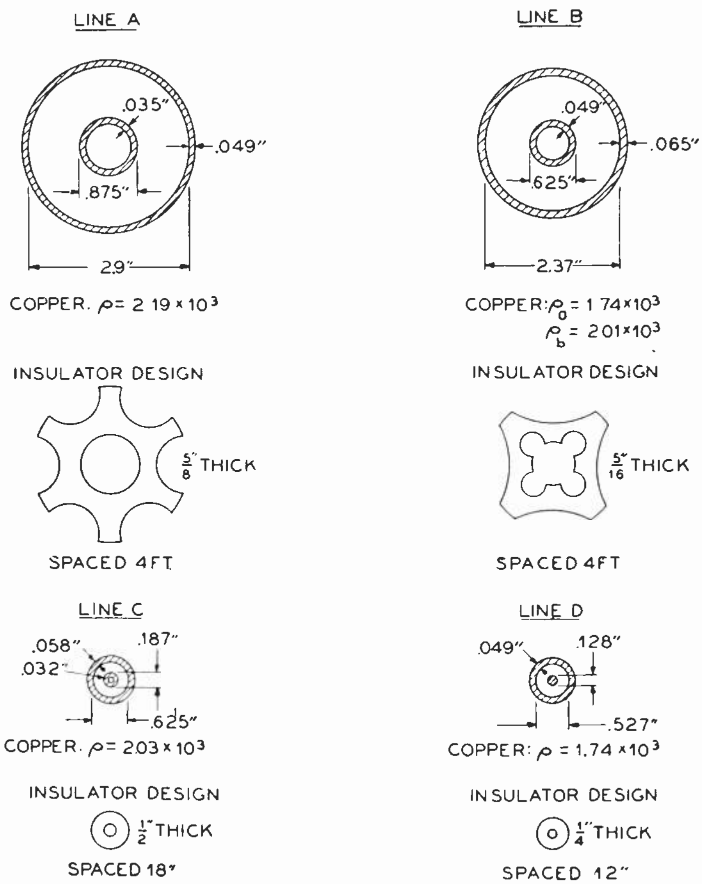

Fig. 1—Principal dimensions of the concentric tube lines upon which the experimental resistance measurements shown in Fig. 2 were made.
Line *E*—Same dimensions as "*C*." Brass ($\rho = 6.6 \times 10^3$) outer pipe.
Line *F*—Same dimensions as "*C*." All brass ($\rho = 6.6 \times 10^3$), insulators spaced 22 inches.
Line *G*—Same as "*C*" but filled with insulators.
Line *H*—Lead sheath cable, No. 18 B and S copper ($\rho = 1.7 \times 10^3$), rubber insulation, lead ($\rho = 17 \times 10^3$) $\frac{1}{8}$ inch inside diameter.
Line *I*—Same as "*C*" but insulators spaced 9 inches.
Line *J*—Same as "*F*" but insulators spaced 18 inches.

supported on porcelain insulators. The latter were attached to the inner conductor by means of spring clips, extruded metal ears, or by means of soldered rings. Some measurements were made on lines assembled with soldered joints and some on lines connected by means of pipe unions with miniature plug-and-jack connections for the inner pipe. Various line lengths were employed. Most of the observations comprised measurements of the quantity $(R/2 + GZ_0^2/2)$. The measurement procedure will be described later.

The results of these measurements appear on Fig. 2. The solid curves were computed by means of (3), neglecting the conductance term. The points are the experimental observations. Note that, excepting lines G and H, the small margin between measured and calculated values shows that the leakage losses are very small. It is believed

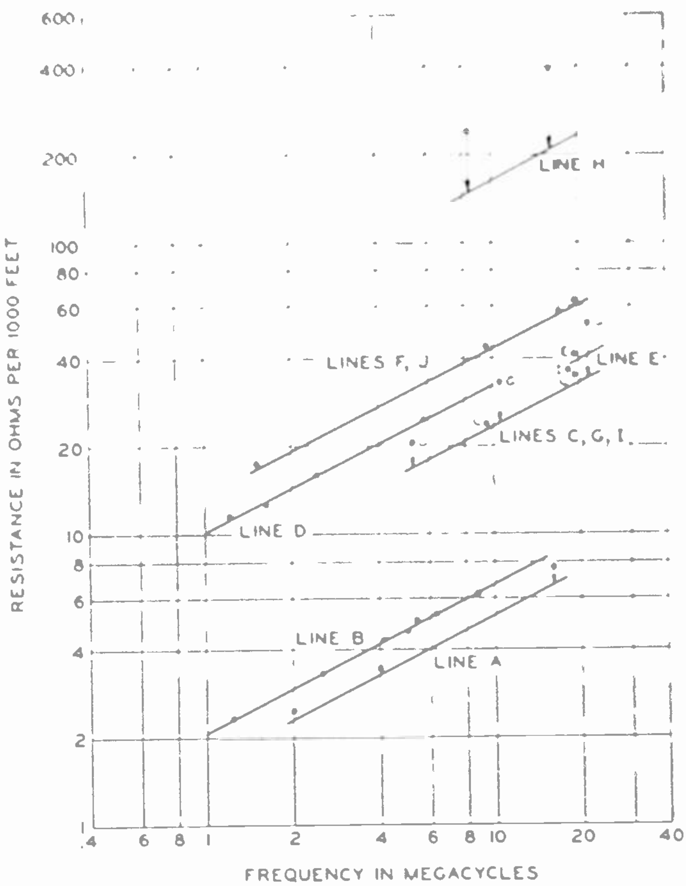

Fig. 2—Radio-frequency resistance measurements upon concentric tube lines. The curves are based upon computations and the points are experimental observations. See Fig. 1 for line details.

that the scattering of the observed points could have been reduced appreciably if corrections for variations of resistivity with temperature had been made.

In lines B, E, and H the sheath and the inner conductor comprised materials of different resistivities. Calculations for these cases were made with the assistance of a modified form of (3):

$$R = \frac{1}{a}\sqrt{\rho_a f} + \frac{1}{b}\sqrt{\rho_b f} \tag{3a}$$

in which the subscripts denote the inner and the outer conductors.

Line G was completely filled with porcelain insulators and line H was a rubber insulated, lead sheathed, cable. The curves of Fig. 3 were derived from the difference between the observed and calculated re-

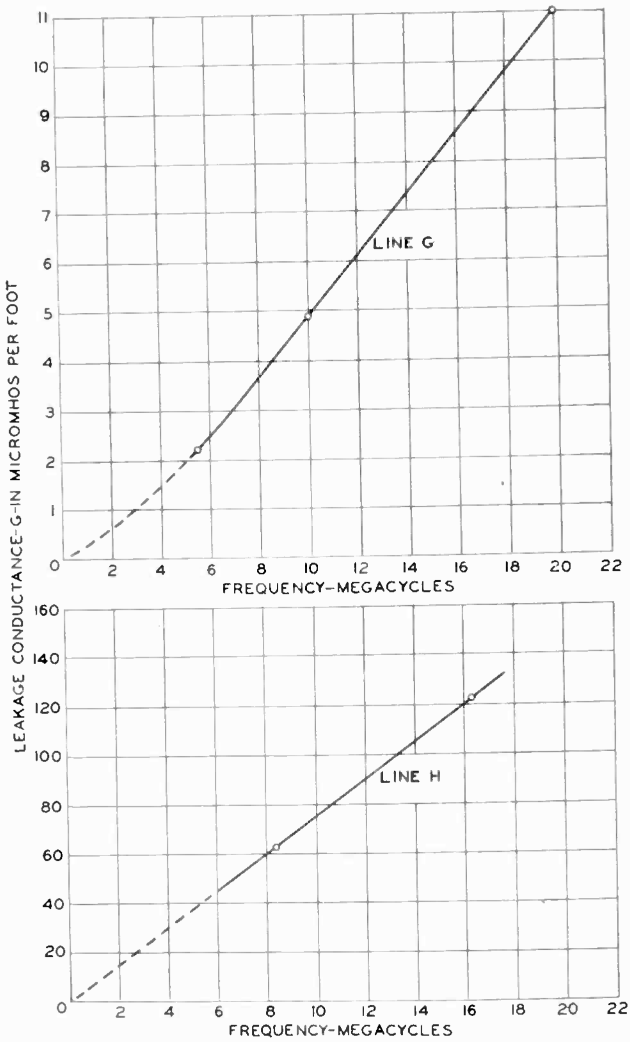

Fig. 3—Derived leakage conductance for the concentric tube lines G and H of Fig. 1. Line G was completely filled with porcelain insulators. Line H comprised a No. 18 B and S conductor with rubber insulation and lead sheath.

sistance in these two cases. The results so obtained are a fair approximation of the leakage losses. Note that the curves are nearly proportional to the frequency which is to be expected if, for a constant voltage, the dielectric absorbs a fixed amount of energy each cycle.

As may be expected the velocity of propagation for both lines G and H was reduced by a factor of approximately 1.8 which corresponds to a dielectric constant of about 3.2. Line I, which was made with insulators spaced at 9-inch intervals, was the only other line which showed a pronounced reduction in the velocity of propagation, the factor in this case being 1.18.

It is of interest to observe that if for economic reasons the diameter of the outer conductor is fixed there is an optimum inner conductor size for minimum attenuation. Employing (1a), (2d), and (3) the real part of the propagation constant may be written as:

$$\alpha = \frac{\sqrt{\rho\mu f}}{276} \frac{\left(\frac{1}{a} + \frac{1}{b}\right)}{\log_{10} \frac{b}{a}} \times 10^{-9}. \tag{4}$$

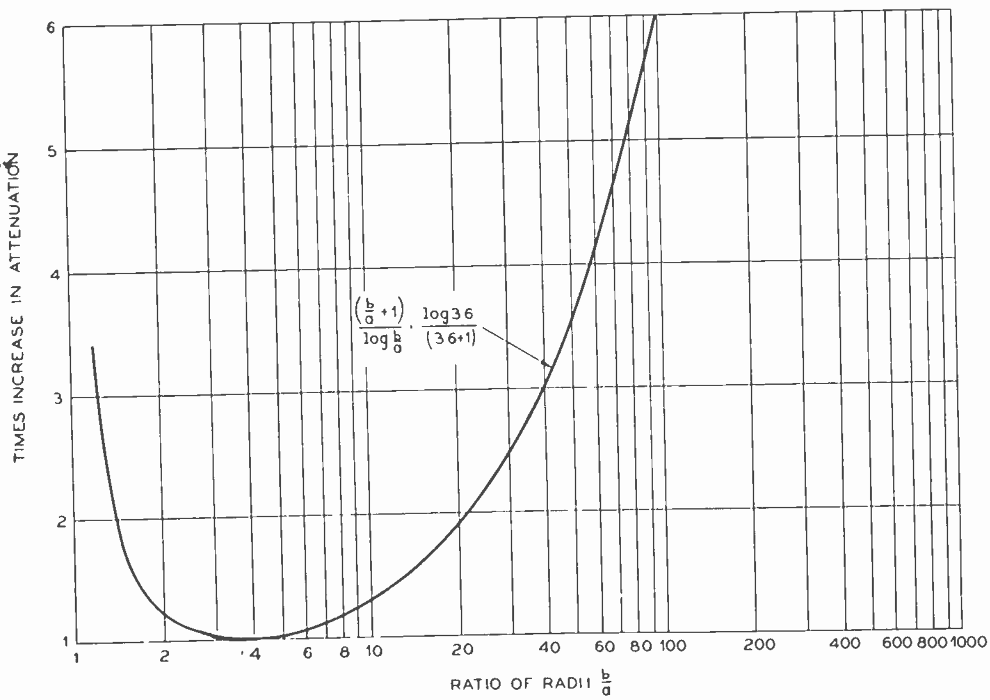

Fig. 4—The most desirable ratio (3.6) of outer to inner conductor and the penalty incurred in departing from this value.

This neglects leakage loss and assumes that both conductors are made of the same material. Upon minimizing with respect to a the optimum ratio:[6]

$$\frac{b}{a} = 3.6 \tag{4a}$$

[6] An experimental figure for the optimum ratio was given by C. S. Franklin in a British Patent, No. 284005. The above derivation for the optimum ratio was disclosed to the writers by E. I. Green and F. A. Leibe, American Telephone and Telegraph Company, New York City.

is readily obtained. This ratio corresponds to a characteristic impedance of 77 ohms. Fig. 4 gives the manner in which the attenuation varies as a function of b/a. Note that a moderate departure from the optimum ratio does not greatly increase the line losses.

So far it has been tacitly assumed that the conductors were exactly concentric. Eccentricity affects all of the line constants.[7] However, experience has shown that the departures from concentricity usually encountered in practice produce no appreciable increase in the attenuation constant of the line.

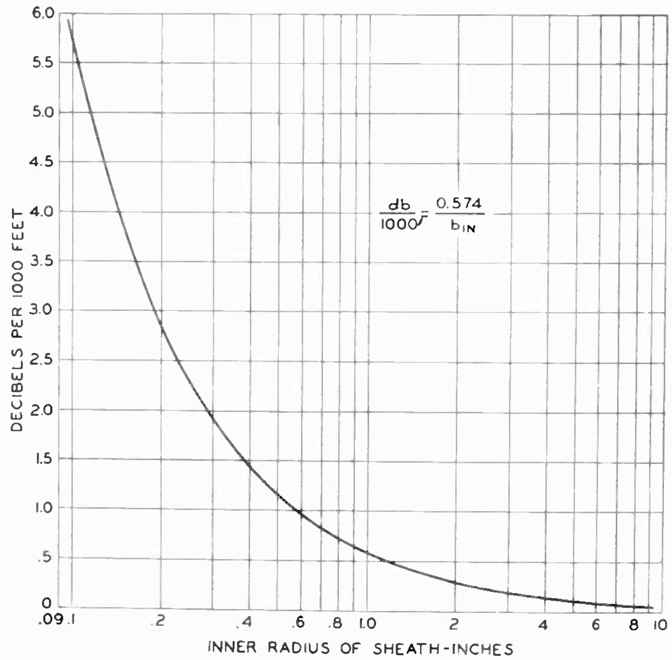

Fig. 5—Calculated losses at 20 megacycles expressed in decibels for concentric tube lines constructed from copper and employing the optimum ratio (3.6) of outer to inner conductor.

At commercial installations the actual power loss in the terminated line is measured directly in decibels and has invariably been found to agree with the predictions within the precision of such measurements which is about 0.5 db in field work. Certain of the lines referred to in this paper have been similarly tested in the laboratory under more favorable conditions and yielded agreements within 0.3 db where the total loss was of the order of 4 to 6 db.

Briefly summarizing it may be said that the attenuation in well constructed concentric lines is proportional to the square root of frequency, inversely proportional to the diameters (optimum ratio) and

[7] A. Russell, "Alternating Currents," vol. 1, p. 166, Cambridge Press.

proportional to the square root of resistivity. Numerically, the loss for copper lines of optimum ratio, neglecting leakage, is:

$$\frac{db}{1000 \text{ ft.}} = \frac{0.128 \sqrt{f_{mc}}}{b_{in.}}. \tag{5}$$

A plot of (5) for one particular frequency ($f=20$ mc) is shown in Fig. 5.

It is important to emphasize one precaution in the use of concentric lines. The high degree of isolation afforded by concentric tube

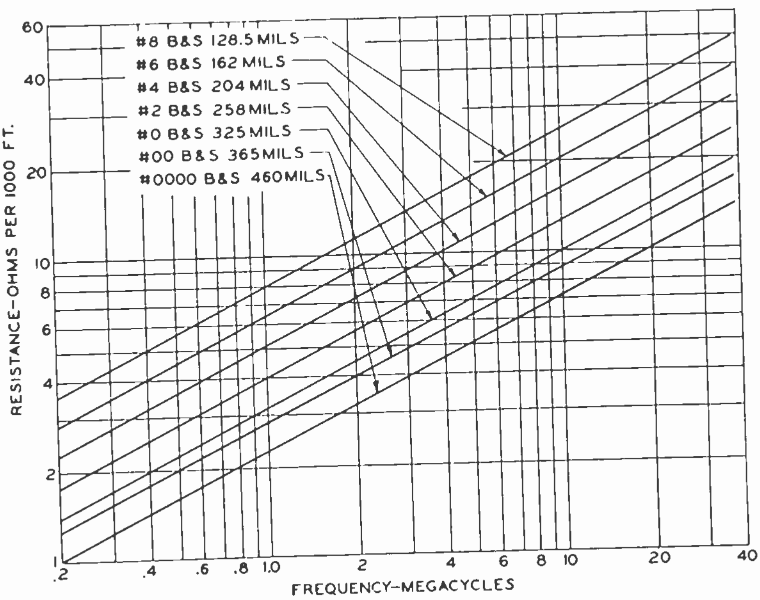

Fig. 6—Calculated radio-frequency resistance for several common sizes of solid copper conductors. Values are for one conductor only.

lines may be easily destroyed. Owing to pick-up from near-by antennas or from other sources, currents of appreciable magnitude may be flowing upon the exterior of the sheath. Spurious couplings between the antenna and the line or between the equipment and the line may introduce these currents into the shielded circuit. In this manner the discrimination of a receiving circuit against undesired signals may be destroyed. Also, the currents flowing upon the exterior of the sheath may destroy the directional characteristic of the antenna to which the line is connected. Grounds placed at frequent intervals are useful in reducing these currents. Sometimes it is both desirable and convenient to bury the line in the earth. Additional improvement is obtained by constructing the circuits which transform the antenna impedance to the line impedance so as to obtain rigorous symmetry to ground.

III. Open Wire Lines

The losses in open wire lines may not be determined in as simple a manner or with the degree of certainty that is possible with concentric tube lines owing to the complex nature of the electromagnetic field about open wire lines. The high-frequency resistance of one conductor may be obtained from the foregoing equation (3) by assuming that the radius of the outer pipe is infinite.[8] The characteristic impedance of balanced open wire lines is obtained with sufficient accuracy from:

$$Z_0 = 276 \log_{10} \frac{2D}{d} \text{ ohms} \tag{2e}$$

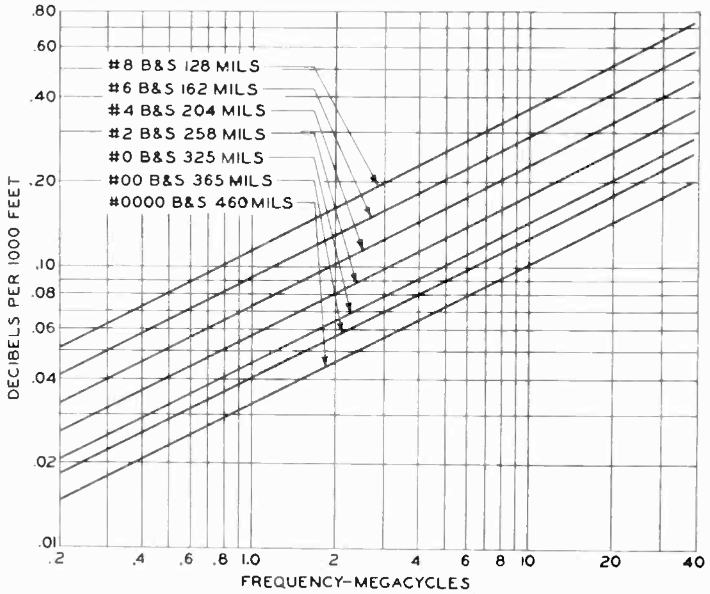

Fig. 7—Calculated attenuation expressed in decibels for copper losses in 600-ohm lines made up from common sizes of solid conductors.

in which D is the conductor spacing measured between wire centers and d is the wire diameter. Some typical results for the resistance of a single conductor appear in Fig. 6.

At first thought it would appear that, owing to the high resistance of a single conductor, the losses in open wire lines are higher than in concentric tube lines. In practical constructions, however, open wire characteristic impedances 5 to 10 times greater than those for concentric tube lines are easily obtained. For example, the loss of a 77-ohm

[8] Derivations for the resistance of single and multiple cylindrical conductors may be found in an article by S. Butterworth, "Eddy current losses in cylindrical conductors with special applications to the alternating-current resistance of short coils," *Phil. Trans. Royal Soc.*, vol. 222, p. 57, (1921).

concentric tube line is 6.38 times as great as that for a 770-ohm open wire line in which the wire diameter is equal to that of the inner conductor of the concentric tube line. Thus, the attenuation constant for a practical open wire line may be approximately the same as that for the larger practical sizes of concentric tube lines. Some typical computations appear in Fig. 7. A balanced two-wire line of 600-ohm characteristic impedance was chosen for the computations.

In Fig. 7 it was assumed that the proximity effect, that is, the redistribution of currents owing to the presence of the second conductor, is a correction of negligible magnitude. Only in the case of large conductors closely spaced does the proximity effect perceptibly increase the resistance. This may be seen from Fig. 8 which shows the increase

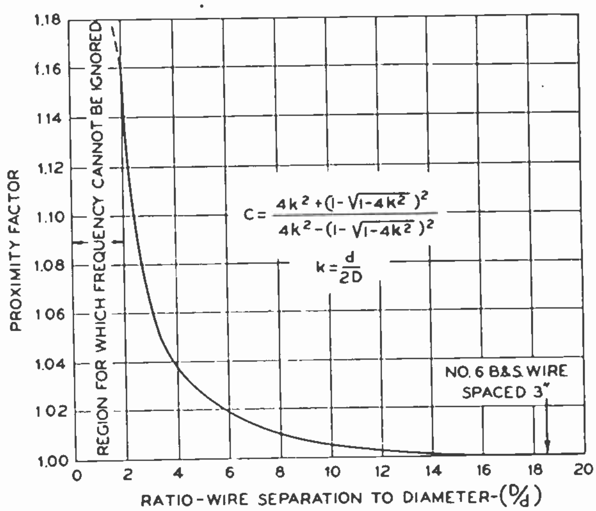

Fig. 8—Calculated values for the increase in copper losses due to the redistribution of current at close conductor spacings in balanced two-wire lines.

in resistance due to the proximity of the conductors. There are several excellent published articles upon this subject.[9,10,11,12]

The foregoing results give, of course, only the power dissipated in copper losses and tell nothing about radiation losses. If the line spacing is less than 1/10 of a wavelength and if the line length is more than 20 times the line spacing, the power radiated by a two-wire line terminated in its characteristic impedance is approximately:

$$\frac{P}{I^2} = 160\left[\frac{\pi D}{\lambda}\right]^2 \text{ watts/(amperes)}^2 \qquad (6)$$

[9] J. R. Carson, *Phil. Mag.*, ser. 6, vol. 41, p. 607, April, (1921).
[10] H. B. Dwight, *Jour. A.I.E.E.*, p. 203, March, (1922).
[11] H. B. Dwight, *Jour. A.I.E.E.*, p. 827, September, (1923).
[12] S. Pero Meade, *Bell Sys. Tech. Jour.*, vol. 4, no. 2, April, (1925). The equations given in this reference were employed in computing Fig. 8.

in which:

D/λ is the line spacing, and

I is the r-m-s value of the current in the line.

Under these assumptions the radiated power does not involve line length.

More accurate equations appear in an appendix to this paper. It may be concluded from (6) that the power radiated by a terminated line is,

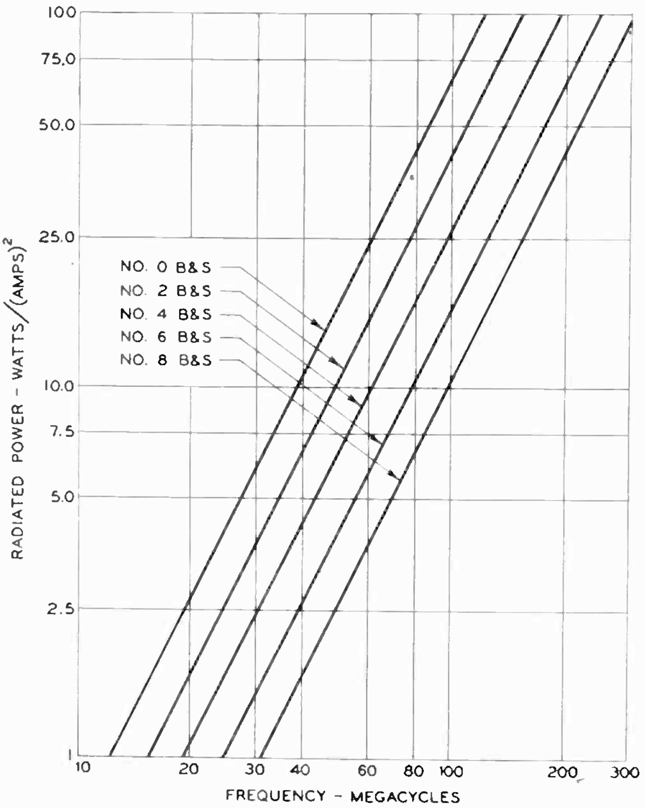

Fig. 9—Calculated power radiated by 600-ohm lines terminated in the characteristic impedance for several common conductor sizes. The line length is a factor of negligible importance for the above spacings and most practical lengths of line.

in magnitude, approximately twice that radiated by a doublet antenna of length equal to the line spacing. Thus, the most simple circuit with which it is possible to terminate an open wire line, a resistance of length equal to the line spacing, will radiate approximately one half as much power as the line. Therefore, considering both load and generator terminations, the total power dissipated in radiation may be approximately twice that given in (6). The equation is plotted on Fig. 9 for the cases of several 600-ohm lines constructed from practical conductor sizes. It may be seen from this figure that the power radiated by a practical terminated line is negligible as compared to the power transmitted by the line providing that operations are confined to wavelengths other than those in the ultra-short-wave region.

If the currents in the two wires are unequal or are not exactly 180 degrees out of phase there is an appreciable amount of power radiated by a two-wire line. Unbalances of this kind become evident when the driving voltages, measured to neutral, are incorrectly balanced and phased. Such unbalances also arise if the voltages induced by the antenna set up currents in the line which employ the two conductors in parallel.

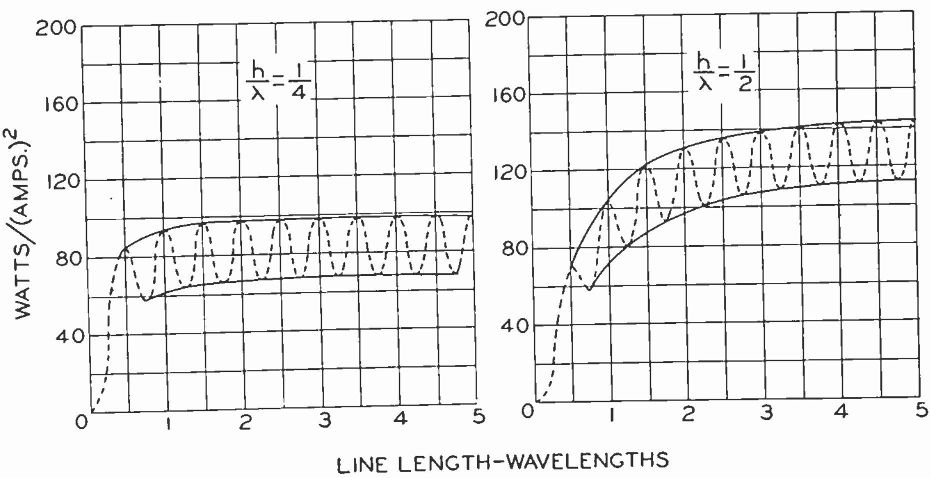

Fig. 10—Approximate power in watts radiated by an unbalanced current of 1.0 r-m-s amperes in a long two-wire line. Also, the power radiated by a single wire parallel to the earth for 1.0 r-m-s ampere line current. Two cases, $\frac{1}{2}$ and $\frac{1}{4}$ wavelength above ground are illustrated.

For the purpose of computation unbalanced currents may be considered as flowing in a single conductor parallel to a perfectly reflecting earth. The amplitude of the current in the single wire may be assumed to be the vector sum of the current values in the two conductors. This procedure ignores the mutual interactions of the balanced and unbalanced currents flowing in the two-wire line and hence, the results so obtained are not strictly correct. It is believed, however, that the error is small.

Based upon these assumptions the power radiated by unbalanced currents is approximately:[13]

$$\frac{P}{I^2} = 30\left[0.5772 + \log_\epsilon (2L) - \sin^2 (L)\left(1 - \frac{\sin H}{H}\right) - Ci(2L) \right.$$
$$\left. - 2Ci(H) + Ci(\sqrt{L^2 + H^2} - L) + Ci(\sqrt{L^2 + H^2} + L)\right] \tag{7}$$

[13] See Appendix.

in which:

P/I^2 is expressed in watts/(amps)^2,

I = r-m-s value of current at a position along the line of maximum current,

$H = \frac{4\pi h}{\lambda}$, $\frac{h}{\lambda}$ being the height of the wires above ground in wavelengths, and

$L = \frac{2\pi l}{\lambda}$, $\frac{l}{\lambda}$ being the length of the line in wavelengths.

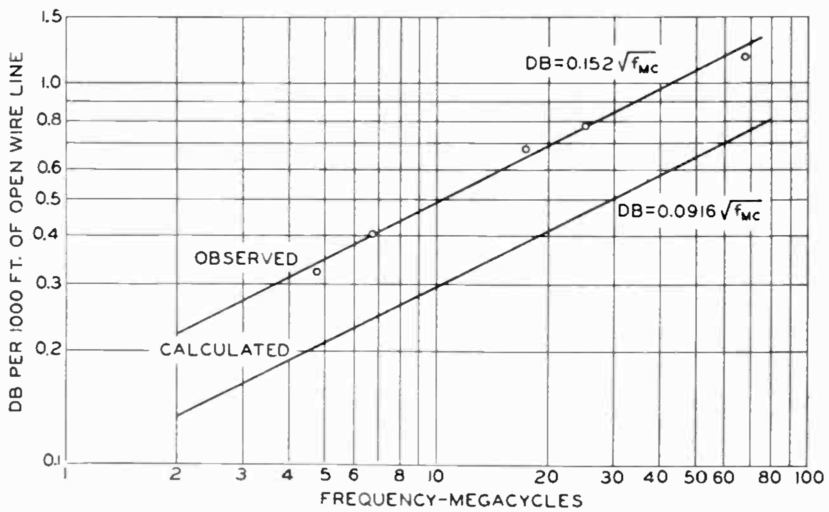

Fig. 11—Experimental observations of attenuation in a 600-ohm line comprising 0.162-inch copper conductors. The points are observed values. The lower curve is calculated only on the basis of copper losses.

The equation is plotted on Fig. 10 for two specific heights above ground and for various line lengths. Upon examining Figs. 9 and 10 it may be concluded that in practical constructions a thirty per cent unbalance in line currents radiates an amount of power roughly equal to that radiated by the balanced currents in the line.

It is our experience that losses due to current unbalances are appreciably greater and somewhat different in character from those indicated by (7). The discrepancy may reside in the assumptions employed in deriving the equation. In particular, the losses in the earth have been ignored. It may well be that the soil over which the line is erected introduces large losses in the line, particularly when the currents are unbalanced. Such losses would augment the attenuation constant of the line.[14] At least, the computations indicate the desirability of maintaining careful line current balances.

[14] John R. Carson, *Bell Sys. Tech. Jour.*, vol. 5, no. 4, October, (1926).

Some remarks upon the proper procedure for inserting the power losses due to radiation into the equations for the line may be of interest. Carson[15] has shown that the conventional solution of the transmission equation for guided waves on wires is incomplete and does not explain the phenomena of radiation. He shows that a "principal wave," and hence the currents in the conductor associated with this wave, travel along the conductors without sensible attenuation due to radiation. Radiation from the line results in the attenuation of an infinite number of "complementary waves." These are highly attenuated so that the radiation of energy is a phenomenon essentially associated with the terminals of the line or points of discontinuity which set up reflected waves.

It may be concluded from Carson's mathematical investigation that the radiation resistance is a term to be added to the impedance of the line at the terminals or points of discontinuity and that it does not appear in the propagation constant. On this basis, the power radiated by a practical balanced transmission line is negligibly small when compared to the power being transmitted by the line, except, perhaps for operation at the very short wavelengths.

Experimental data for the attenuation in open wire lines which are as complete as those already shown for concentric tube lines are not available for this paper. Some typical observations for 600-ohm lines constructed with No. 6 B & S semihard drawn copper wire appear in Fig. 11. The points are experimental observations. The lower curve was computed for the case 1830 e.m.u. copper resistivity. The observed values are about 66 per cent higher than the computed values.

The experimental procedure was as follows. A line 2000 feet long was carefully balanced and terminated by an iron wire line[16] for each of the experimental observations. By means of a portable calibrated indicating device the average currents for the one-half wavelength of line at the near end and at the far end were obtained. The attenuation in decibels was computed from average near-end and far-end current ratios.

It is difficult to explain the discrepancy between observed and computed values. If the resistivity employed in the computations were to be increased from 1830 to 5030 e.m.u. (a multiplication factor of 2.75) the computed curve so obtained would be in good agreement with the observed results. It is true that the wires were somewhat weathered. There is, however, little reason to believe that an appreciable amount of current flows in the oxide layer covering the wires. Effects of this

[15] John R. Carson, *Jour. A.I.E.E.*, p. 908, October, (1924).
[16] See Section VII.

kind would have been evident in the measurements upon concentric tube lines. It already has been mentioned that small current unbalances in the line may produce losses in the earth which increase the real part of the propagation constant. Possibly, losses of this kind may explain the discrepancy.

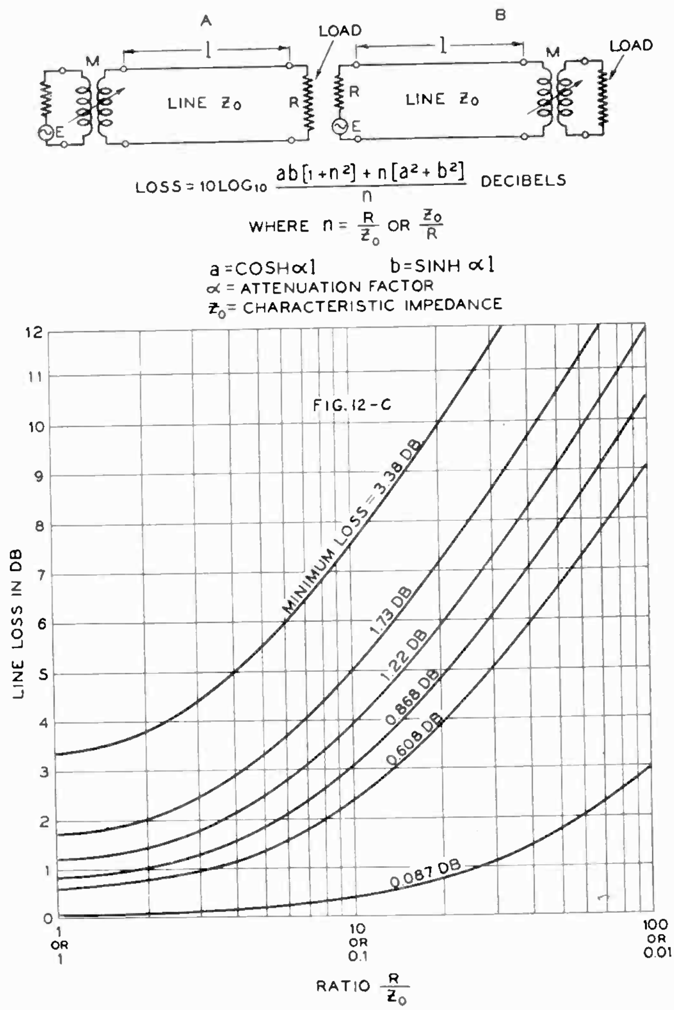

Fig. 12—Line loss as a function of the degree of matching. The several curves are designated in terms of the minimum line loss obtained for the case of a perfect match with the line characteristic impedance.

IV. Notes on Matching Impedances

It already has been mentioned that standing waves on a transmission line augment line losses. The penalty which is imposed by improper impedance matches may be seen from Fig. 12. This figure plots line loss as a function of the degree of matching for several attenuation factors. The line loss is computed from the ratio of the power dissipated in the load to the total power obtainable from the generator. The curves were obtained from conventional transmission line theory. For the purpose of simplifying calculations the line length is assumed to be an integral number of one-quarter wavelengths, thereby eliminating complex impedances. Otherwise the length of the line is immaterial, the product of length and attenuation per unit length being the criterion of loss.

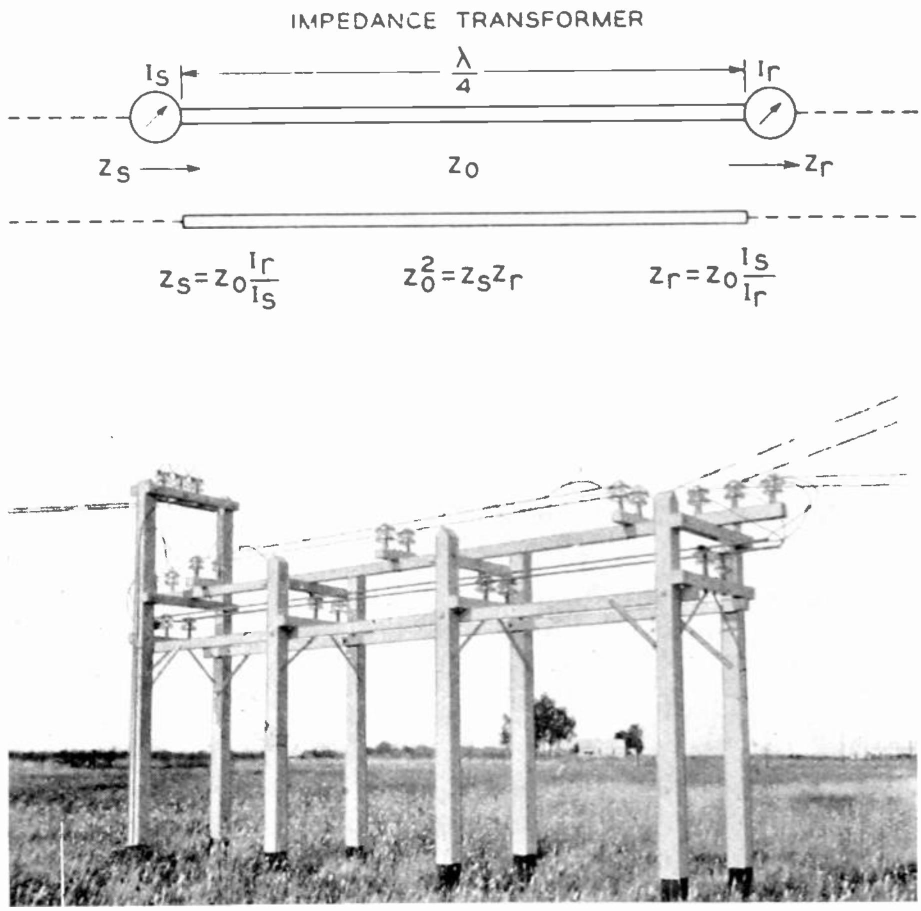

Fig. 13—The principles underlying the use of one-quarter wavelength bars as a transformer are shown in the upper diagram. The lower illustration depicts a commercial installation.

In the diagrams of Fig. 12A and Fig. 12B the circuit M is an adjustable ideal transformer. For every value of the resistance R the transformer M is assumed to be adjusted so as to maximize the load power. This process is equivalent to matching impedances at the terminals of the line adjacent to the transformers. It is of interest to observe that where R is not equal to the characteristic impedance Z_0, this adjustment does not yield an impedance match at the line terminals remote from the transformers. It is of further interest to observe that in the case where the load impedance is variable (Fig. 12B) the optimum adjustment is a compromise between a nonreflecting termination and an impedance match at the generator end.

Conventional tuned transformers may be employed to match the line impedance to the antenna and radio equipment impedances. In the transmitting case, tuned circuits are often found to be both bulky and costly. There are a number of schemes which employ a short section of line as a transformer element. These are feasible only at frequencies for which the wavelength is short. The circuits so provided are extremely simple and cheap.

In another paper[17] a scheme for employing a one-quarter wavelength section of line as a step-up or step-down transformer was described. Briefly the principle of operation is the fact that the sending end impedance Z_s and the receiving end impedance Z_r are related to the characteristic impedance Z_0 by the simple expression:

$$Z_sZ_r = Z_0^2. \tag{8}$$

Thus, by choosing the proper characteristic impedance any two real impedances may be matched providing these do not differ too greatly. Often the scheme is made workable by constricting the line spacing for a one-quarter wavelength section. Where a large difference in transformer line and transmission line spacing is undesirable the transformer line may comprise conductors of large diameter. A transformer set-up of this type is shown in Fig. 13.

There is another effective way for transforming line impedances by means of short line devices.[18] A complex impedance at the proper position along a partially terminated line is selected such that a shunt reactance at this position transforms the real part of the impedance to the surge impedance of the line at essentially unity power factor. The shunt could, of course, be a lumped reactance. It is found convenient to employ a short section of line for this reactance. A position along the line for the shunt reactance of either leading or lagging power factor may be chosen. In the former case the shunt reactance must be inductive and in the latter case capacitive. Computed shunt impedance positions for the two cases and the values of the shunt impedance in terms of line length for 600-ohm lines and various standing wave amplitudes on the unterminated section appear in Fig. 14. Actual settings correspond very well with the calculated settings.

V. Resistance and Attenuation Measurements on Transmission Lines

The following is a description of some of the measurement methods

[17] E. J. Sterba, Proc. I.R.E., p. 1184, July, (1931).

[18] Disclosed to the writers by P. H. Smith, Bell Telephone Laboratories, Inc., New York, N. Y.

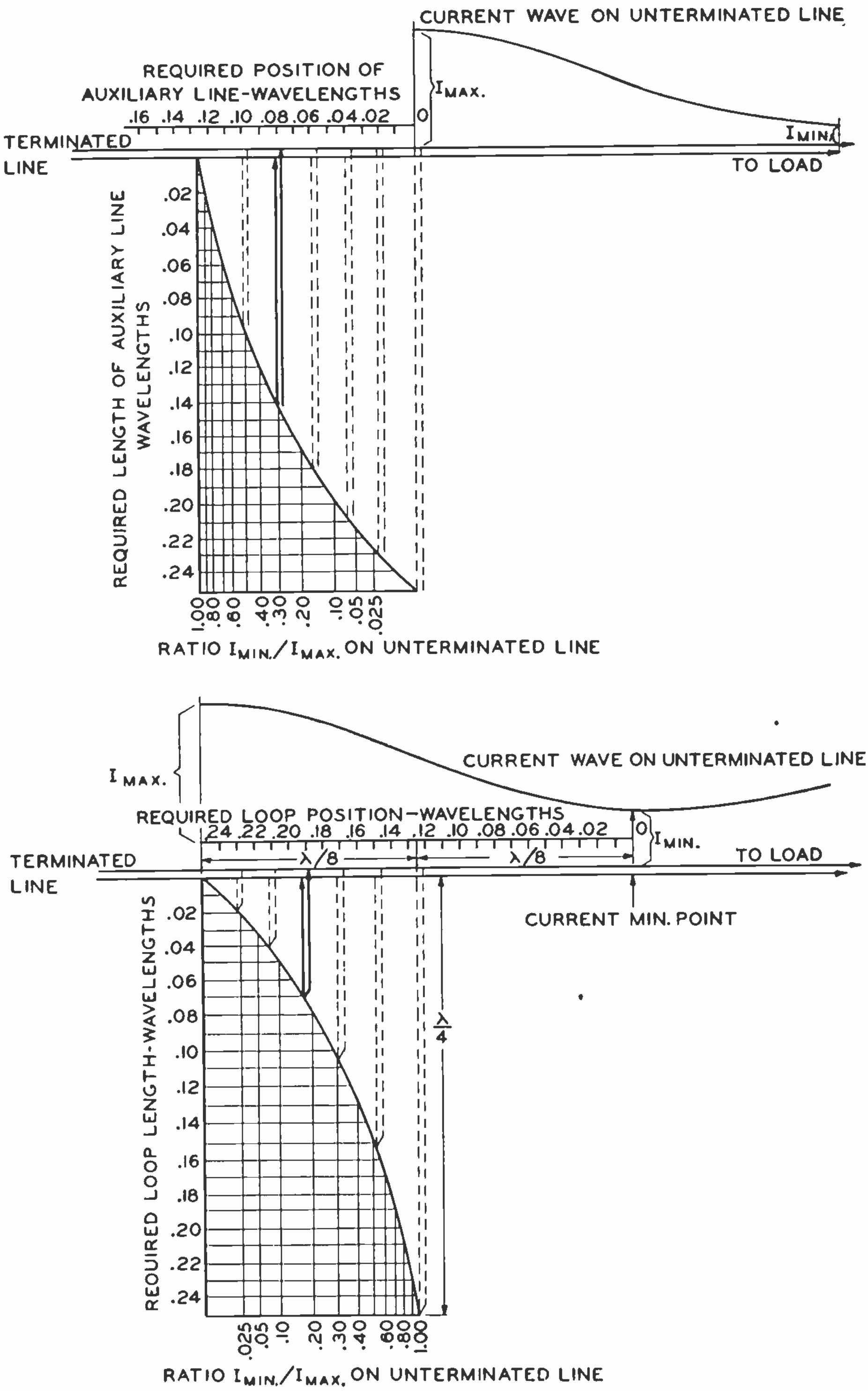

Fig. 14—The use and adjustment of an auxiliary line as a transformer element. The settings are computed for the case of 600-ohm lines. The position and length of the auxiliary line may be obtained from the curves for any given ratio of minimum to maximum currents on the unterminated portion of the line.

which have been found useful in the study of transmission lines. The schemes may not be applicable to every phase of the transmission line problem. However, it is hoped that they may suggest precautions to be observed in performing transmission line studies.

One scheme, very commonly employed, is to measure the attenuation along a transmission line by actual current measurements. This

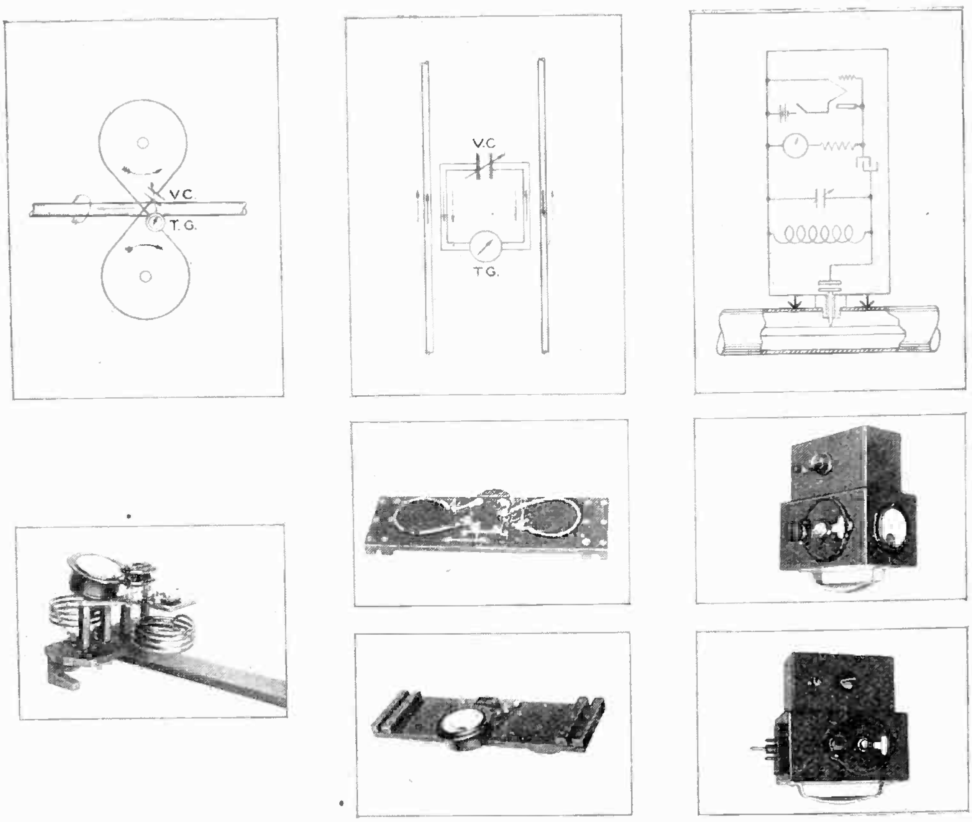

Fig. 15—Three types of portable indicating devices for observing the current distribution on transmission lines. The device to the left is sensitive chiefly to the current flowing in one conductor. The device in the center responds chiefly to balanced currents and is not sensitive to unbalanced currents. The device to the right is suitable for observing the voltage amplitudes on concentric tube lines.

method is particularly suited to measurements upon a long line terminated in its characteristic impedance. It has been found desirable to measure the current amplitudes at close intervals for at least a one-half wavelength section at the near end and the far end of the line. In this manner an average result which reduces observational errors and errors arising from standing waves of small amplitudes is obtained. From the ratio of the average sending end current I_s and the average receiving end current I_r and the average distance l between the two sections of line the attenuation per unit length is obtained from the definition:

$$\text{db} = 20 \log_{10} \frac{I_s}{I_r} \tag{9}$$

and since:

$$\frac{I_s}{I_r} = \epsilon^{Rl/2z_0} \tag{10}$$

therefore:

$$\frac{\text{db}}{l} = 4.343 \frac{R}{Z_0} \tag{11}$$

from which the resistance R per unit length may be obtained with a degree of accuracy depending chiefly on how accurately the characteristic impedance Z_0 is known.

The current distribution along the line is most conveniently ob-

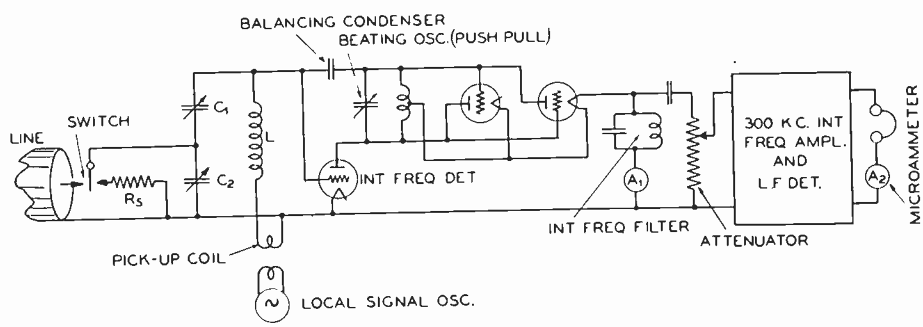

Fig. 16—Schematic diagram of apparatus for measuring line loss by means of a resistance substitution method.

tained by means of a portable indicating device. Three designs which have been found useful are shown in the following figures. The indicator to the left of Fig. 15 is used for measurements upon open wire lines. The manner in which it operates is evident from the figure. Note that except for closely spaced wires the device is sensitive only to the current in one side of the line. When this device is used the currents in both sides should be measured to assure there are no large current unbalances and for the purpose of averaging out any small unbalances. The device in the center of Fig. 15 is coupled to both sides of the line and is not sensitive to currents which employ the two-line conductor in parallel. It is useful where for other reasons the unbalanced currents cannot be reduced to a desirably low value. The device shown to the right of Fig. 15 is suitable for measurements on concentric lines. In order to employ this device openings at regular intervals are required in the outer sheath. An important precaution to be observed in employing this last device is that the shielding be sufficiently thorough to assure no pick-up from stray currents flowing upon the outside of the sheath.

It is of course essential that all portable devices of this kind extract a very small proportion of the power in the line otherwise the device becomes a source of reflection and spurious results are obtained.

Another method of measuring the attenuation of a line which is particularly useful in studying the effects of current unbalances is to employ a small portable horizontal antenna the impedance of which matches the characteristic impedance of the line. The antenna is connected in a short section and then in a long section of the line. It is essential that the height of the antenna above ground be equal for the two positions. Also, the location for the experiment should be such that the same ground losses are present for the two positions. The ratio of the antenna currents for the two positions and for the condition of equal power input is a measure of the total line losses.

One of the most satisfactory schemes for measuring line attenuation is the direct measurement of the line sending end impedance by means of the familiar resistance substitution method. It has been used extensively in measurements of concentric lines. For this purpose it is necessary to employ lines either open- or short-circuited at the far end and to restrict the measurements to lines which contain an integral number of quarter wavelengths.

Conventional transmission line theory indicates that under these conditions the impedance is either.

$$Z_1 = Z_0 \tanh\left(\alpha\frac{n\lambda}{4}\right) \tag{12}$$

$$Z_2 = Z_0 \coth\left(\alpha\frac{n\lambda}{4}\right) \tag{13}$$

where:

Z_0 = characteristic impedance,
α = attenuation factor; i.e., the real part of the propagation constant.
λ = wavelength, and
n = length in integral quarter wavelengths.

If n is even and the termination is a short circuit or if n is odd and the termination is an open circuit, (12) is employed. If n is odd and the termination is a short circuit or if n is even and the termination is an open circuit, (13) is employed. The attenuation factor is given by:

$$\alpha = \frac{1}{2}\frac{R}{Z_0} + \frac{1}{2}GZ_0 \text{ napiers per foot} \tag{14}$$

$$= 4.34\left(\frac{R}{Z_0} + GZ_0\right) \text{ decibels per foot.} \tag{14a}$$

where:

R = resistance in ohms per foot,
G = conductance in mhos per foot, and,
Z_0 = characteristic impedance in ohms.

For all the lines concerned with here tanh $[\alpha(n\lambda/4)]$ may be replaced

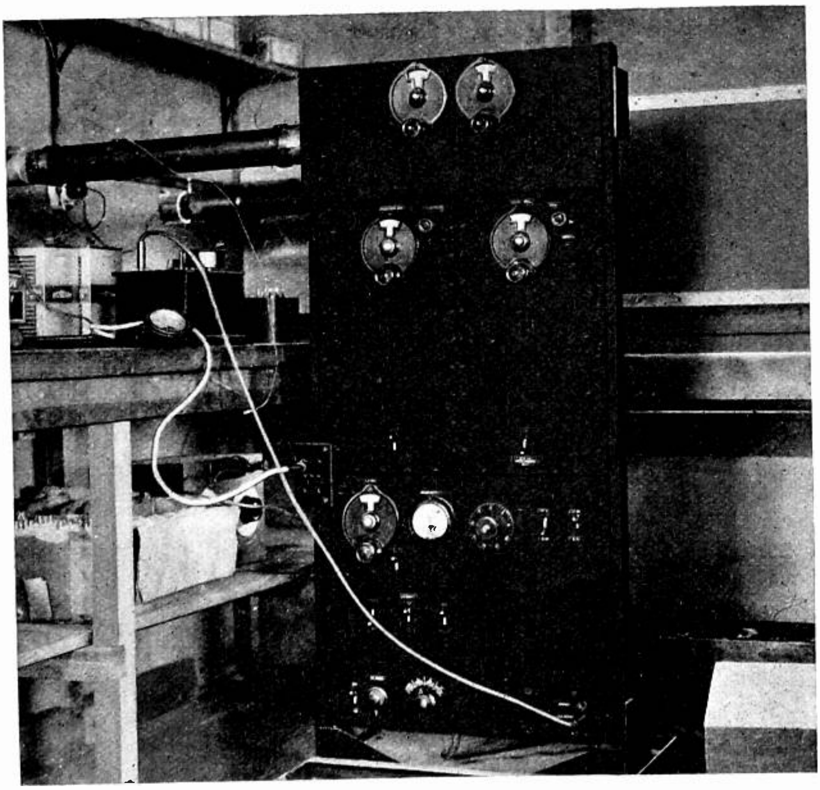

Fig. 17—Experimental set-up of apparatus for measuring line loss.

by $[\alpha(n\lambda/4)]$ without more than 1.5 per cent error. Thus, (12) and (13) reduce to:

$$Z_1 = \frac{R}{2}\frac{n\lambda}{4}\left(1 + \frac{GZ_0^2}{R}\right) \tag{12a}$$

[19] $$Z_2 = \frac{2Z_0^2}{\frac{n\lambda}{4}R}\left(1 - \frac{GZ_0^2}{R}\right). \tag{13a}$$

[19] It is assumed here that the conductance term in (14) is small compared with the resistance term.

Therefore, except for the small contribution of shunt conductance, Z_1 is independent of Z_0. In cases where G is negligible, the measurement of Z_1 gives directly the high-frequency resistance.

The method of measurement is shown schematically in Fig. 16 and an experimental set-up appears in Fig. 17. The modified high-frequency field intensity measuring unit[20] is a convenient indicating device and source of signal. The intermediate-frequency amplifier with its adjustable gain is very useful in maintaining a desirable level in the last detector which is the indicator. Returning to Fig. 16, the local signal oscillator is the source of voltage actuating the tuned circuit LC_1C_2 in which the substitutions are made. Loose coupling is desirable between the pick-up coil and oscillator. In measuring Z_1 which is a low impedance, the condenser C_2 is set at minimum capacity thus effectively making the comparison in a series circuit. In measuring Z_2 which is of the order of hundreds of ohms, C_2 is used to transform Z_2 into an appropriate series value consistent with selectivity and sensitivity.

The comparison resistances used for measuring (Z_1) may be fixed units and the line impedance obtained by interpolation. These comprise straight lengths of high resistance wire and range from a few tenths of an ohm to ten ohms. The wire size is chosen so that skin effect is negligible. Although these resistances possess appreciable inductive reactance at the higher frequencies the reactance usually may be safely tuned out.

It has been found practicable to employ continuously variable resistances made from hard drawing pencil leads equipped with spring clip contacts for measurements of high resistances such as (Z_2). The high resistivity of graphite makes it possible to obtain several thousand ohms in 5 or 6 inches, free of skin effect and with but little inductance.

The foregoing resistance substitution method has been found satisfactory for the purpose of measuring the characteristic impedance of lines. Two methods have been employed. One of these is the familiar procedure in which the sending end impedance is measured for the case of the line open- and short-circuited at the far end. The geometric mean of the two impedances so obtained is the characteristic impedance of the line.

Another scheme producing more precise results is also adapted to the foregoing resistance measuring method in that it requires that the line be some odd number of one-quarter wavelengths long. Such a line

[20] Readers not familiar with this measuring unit may refer to a paper by Friis and Bruce, Proc. I.R.E., August, (1926).

transforms to a different value a terminating impedance which is other than the characteristic impedance. By comparing a variable terminating resistance directly with the value to which it is transformed by the line a setting may be found for which the line functions as a one-to-one transformer. For this condition the value of the variable terminating resistance is the characteristic impedance of the line. Here again drawing pencil leads have been found to be satisfactory termination resistances when set by direct-current measurement methods.

In practice the foregoing resistance substitution method bring to light many slight irregularities. Variations of apparent characteristic impedance with frequency as much as 5 to 10 per cent have been found for concentric tube lines equipped with elbows, couplings, and similar fittings. It is believed that impedance variations of this order are to be expected from some such irregularities unless particular care is taken in the construction of the fittings. On the other hand, it has been found that a short straight length of carefully constructed concentric tube line is so smooth that its characteristic impedance may be employed as a calculable standard.

VI. Practical Construction Details

Open wire radio-frequency line construction is not very different from that employed in power practice. One outstanding difference is that line supports and insulators must be considered as individual irregularities spaced at intervals often greater than one wavelength. The effect of one such irregularity may be small. The total effect in a long line, however, is sometimes appreciable.

The body of the insulator, since it has a dielectric constant appreciably different from air and since its dimensions are comparable with the line spacing, is in itself a line irregularity. Tie wires or conductor clamps augment this effect. Cross arms and pins employed for mounting pin-type insulators also add to the effect, particularly during wet weather.

From the standpoint of line irregularities suspension-type insulators are more desirable than pin-type insulators. The latter construction, however, appears to be more practical because the lines are more rigid, sway less during wind storms, and because no intermediate spreaders are required to maintain the desired line spacing.

One other difficulty with open wire lines is the drift in velocity of propagation and surge impedance during rain and sleet storms.[16] Since a similar effect occurs in the elements of the antenna there is a decided drop in the efficiency of the combined antenna and line during rain and sleet storms. The effects of sleet may be reduced by heating the wires with sleet melting currents. The conductor size may be increased to reduce the effects of wet weather but this makes sleet melting more difficult.

There is an appreciable pick-up between balanced open wire lines on common supports. It appears desirable to separate lines to a common transmitter by at least 10 times the conductor spacing. Spacings greater than this may be required if two lines are to be operated simultaneously and in some cases it is more desirable to employ separate line supports in order to reduce the possibility of cross-talk difficulties. Of course any current unbalances in two parallel lines greatly increase the danger of cross-talk.

Concentric tube line construction is not as simple as open wire construction. Considering the transmitting case, there is a smaller safety factor for voltage overloads. Insulators are required to withstand high voltage gradients. Temperature changes with ensuing line expansions and contractions must be given consideration. It is these factors in addition to the added expenditure for copper which make concentric line construction more costly than open wire construction.

The first consideration in the design of a concentric line is the weight of the outer sheath. If the line is to be employed for high power transmitting purposes the voltage safety factor may be so low that accidental dents in the sheath may lead to breakdown. Obviously, there is a choice between a large diameter, lightweight sheath, and more rugged small diameter sheath without an appreciable difference in copper expenditure. Other factors which involve the remainder of the radio plant often determine the size of the outer sheath. We have found that for outer sheaths a diameter of 2.5 inches and a radial thickness of 0.0875 to 0.10 inch provides lines which are sufficiently rugged for transmitting 15 kw of modulated power at 16 meters wavelength.

Careful consideration needs to be given to the problem of protecting concentric lines from voltage overloads which may be brought about by accidental open or short circuits or by flashovers. Voltages of the order of 30,000 to 90,000 volts may easily be built up in this manner at the shorter wavelengths. Horn gaps are useful if located in the proper way. It is fortunate that conventional line input circuits are apt to be detuned in the event of an accidental open or short circuit on the line and that very little power may then be transmitted to the line.

Beads of high grade porcelain in diameters up to one inch are satisfactory insulators for low power and receiving lines. However, such simple insulators are not suitable for high power work. Owing to the volume of dielectric in large annular insulators sufficient heating may occur at the higher voltages to destroy the insulator. Insulators such as those described for line *B*, Fig. 1, have been found suitable at the higher voltages.

The air film between the insulator and the inner conductor lies in a region of steep voltage gradient. Even under what is considered normal operating voltage there may be enough corona in this region to produce heating of the insulator. It may be of interest to mention that a line approximately as described in *B*, Fig. 1, has been found satisfactory for normal operation at 16 meters for a carrier power of 15 kw. The line breaks down in the region of the insulator at 9000 r-m-s volts.

For transmitting purposes it has been found desirable to employ glazed insulators in concentric tube lines because dirt, soldering fluxes,

Fig. 18—A short section of three-quarter inch diameter line showing support for holding line in a sinuous form.

etc., acquired in assembly operations are more readily removed from glazed insulators.

There are a number of simple ways in which insulators may be held in place in concentric tube lines. For low power work and receiving purposes wire clips, rivets or even extruded metal ears upon the inner conductor, are satisfactory. As a rule these do not prove satisfactory at higher powers owing to high potential gradients at points and sharp edges. Small rings riveted or soldered upon either side of the insulator have proved satisfactory. Lines with soldered rings are more easily repaired. Care must be exercised, however, to prevent condensation of metal and fluxes in the pores of the insulator.

In open wire construction it is customary to accommodate line variations brought about by temperature changes by adjusting the sag of the conductors. Provisions for temperature variations in concentric tube lines are not so simple. The first obvious remedy is to employ lines buried at sufficient depth so that temperature changes are reduced to a slow seasonal variation. At the present time a buried 3/8-inch line has been in service for more than one year without developing faults. Without longer experience with concentric lines we would question the advisability of burying larger lines which are to be employed at high voltages due to the difficulty of finding faults should these occur.

A very simple scheme, suitable for small lines, is to reduce the

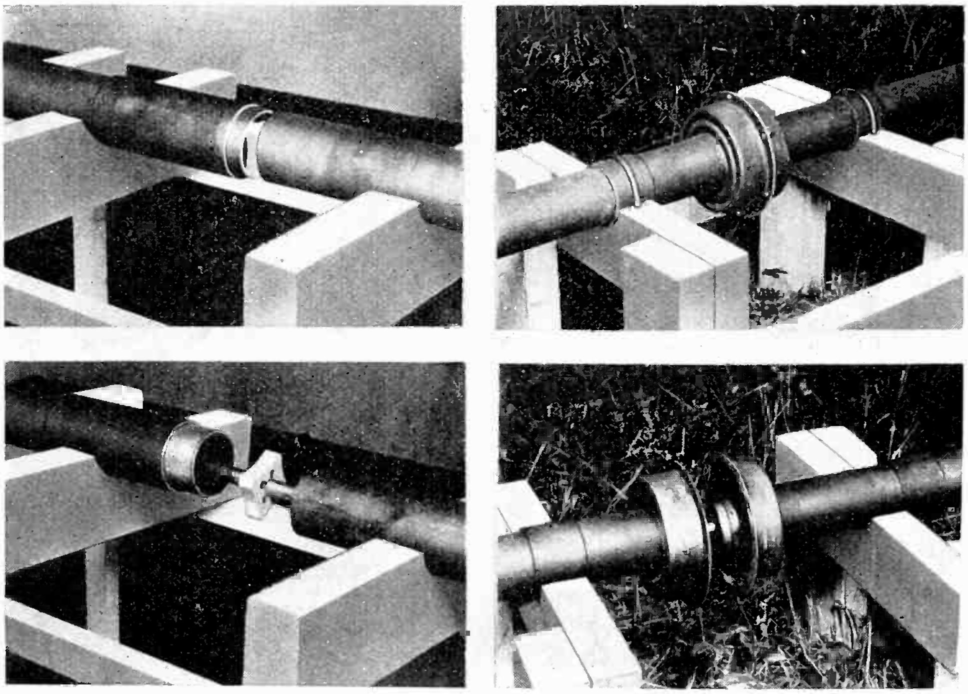

Fig. 19—Experimental expansion and lock joints for large sizes of concentric tube line.

effects of temperature variations by laying the line in a sinuous path as shown in Fig. 18. This construction permits the line to buckle slightly at the curves as the length varies and cumulative changes in length do not appear at the line terminals. The inner conductor is held loosely within the sheath so that it may buckle independently of the sheath. The outer conductor changes its length both at a different rate and at a different time from the inner conductor. With increasing temperature the sheath is at a higher temperature than the inner conductor. There is an appreciable time lag in heating of the inner conductor due to the heat insulation of the air space between the conductors. Small lines laid in a sinuous manner have been found remarkably free from mechanical breakdowns brought about by temperature variations of length.

Sliding joints may be employed to accommodate variations in line length brought about by temperature changes. It is very difficult to make such joints water-tight without recourse to expensive fittings. There is also the possibility of microphonic contacts which are particularly objectionable in receiving work.

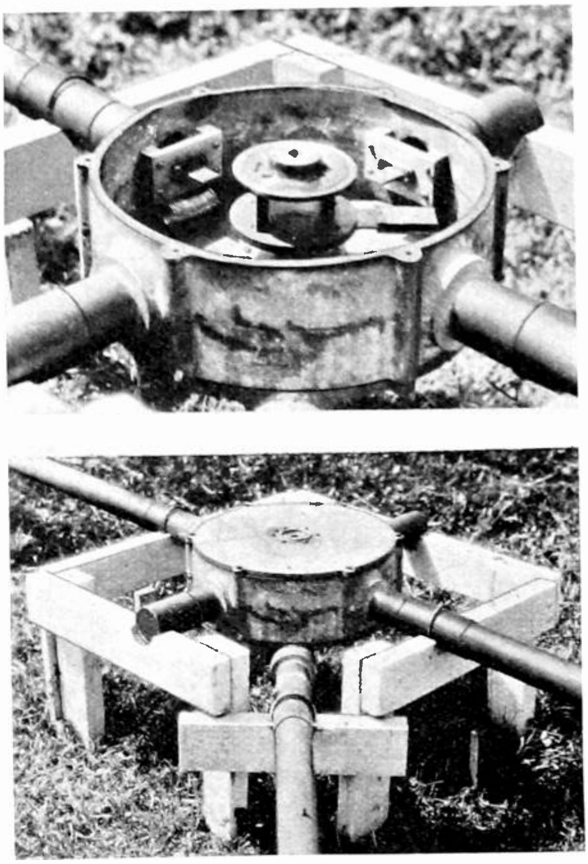

Fig. 20—An experimental selector switch for connecting several antenna lines to one transmitter line. The small coil antiresonates the capacity of the switch for the operating frequency associated with the particular contact to which it is connected.

The expansion joints shown to the right of Fig. 19 have been employed with some success. Dimensions and shapes should be chosen to minimize the irregularities in line impedance caused by expansion joints. It is a step in the right direction to maintain constant the ratio of conductor diameters at the joint. Even then, it was found that the irregularities caused by 10 such joints in a 600-foot line is observable, (approximately 10 per cent standing waves).

It is necessary that expansion joints be employed in conjunction with lock joints so arranged that no one joint is required to take more than a predetermined portion of the line expansion. One lock joint with an expansion joint 25 feet in either direction has been found to be a satisfactory length within which line variations are corrected. The lock joint proper (see Fig. 19) comprises an insulator of the same design as the intermediate insulators but made with an outer diameter equal to that of the outer sheath. It is held in place by a sleeve sweated to the sheath, the sleeve continuing the electric circuit. The insulator is also fixed to the inner conductor by means of rings. Since the lock joint is in a position symmetrical with respect to the two expansion joints it is required to withstand a shearing load brought about only by inequalities in the expansion of the conductors. In order to distribute the line expansion uniformly among the expansion joints it is necessary to clamp the outer sheath of the lock joint to a substantially braced support.

The joints described above are required to accommodate a total annual variation of 0.5 inch. After one year's period of experimental operation the few faults found in a line containing these joints were nearly all traced to faulty construction at the braced support which clamps the lock joint.

Copper pipe lines may be too costly to permit the installation of more than one line per transmitter. In such cases a selector switch is required if several antennas are to be associated with one transmitter. Some of the details of such a switch may be obtained from the experimental arrangement shown in Fig. 20.

The switch is an irregularity on the line and a source of undesired reflections. This difficulty may be corrected by making the design such that capacitive reactance of the switch predominates and then antiresonating this reactance with a suitable inductance. This scheme is effective providing the irregularity is not too great. In the latter event the corrective coil transforms the load impedance to a value different from the surge impedance so that the reflections arising from the mismatch are more serious than from the switch alone.

VII. Other Applications of Transmission Lines

In this section are described a number of transmission line applications to radio work some of which are feasible only at high frequencies because the wavelength is short.

Small concentric lines approximately 3/8-inch in diameter may be employed as radio-frequency wiring in radio stations. Such lines owing to the flexibility of the tubing may be snaked behind partitions in very much the same manner that armored or leaded conductors are installed. For this purpose refrigerator tubing has been found desirable because it is flexible and because it may be procured in long lengths. The inner conductor is insulated from the sheath by means of small porcelain beads spaced at intervals of approximately one inch. The beads are held in position by small metal ears extruded from the inner conductor. The beads fit loosely in the inner conductor so that the line may be bent into arcs as small as six inches radius. Construction details for small concentric lines may be obtained from Fig. 21.

Lines constructed from refrigerator tubing may be buried in the ground. Since only a few splices are necessary the possibility of faults

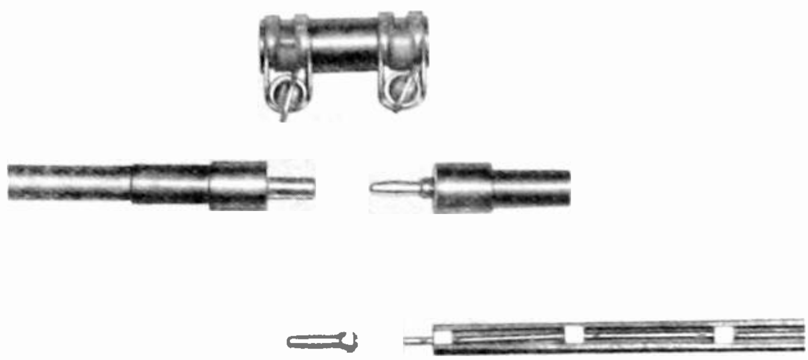

Fig. 21—Details of construction for small concentric lines suitable for station wiring and for patch cords. The plug-and-jack union is an effective scheme for temporarily connecting two small lines.

arising from water seeping into the line are correspondingly small. A buried line constructed in this manner has been in service for more than a year without developing faults.

A number of the above-described lines may be terminated upon a jack board and circuits set up with patch cords as in telephone practice. Of course, the beads in the patch cords may be more closely spaced to assure flexibility and freedom from short circuits. The scheme is particularly advantageous where, for operating reasons, it is useful to connect any station antenna to a particular receiving unit. A board set up for this purpose is shown in Fig. 22.

Concentric tube lines may be employed as standards of resistance when other standards become questionable. Since the agreement between theoretical and experimental values of radio-frequency resistance has been found very good at frequencies at high as 20 megacycles the theory may be considered adequate for much higher frequencies. One scheme for utilizing this situation so as to obtain an adjustable radio-frequency resistance will be described. The scheme utilizes the resonant properties of a section of concentric tube line of which the inner conductor is one-half wavelength long. The required resistance is obtained by a connection to the proper position on the inner conductor. The device is illustrated schematically on Fig. 23. It may be

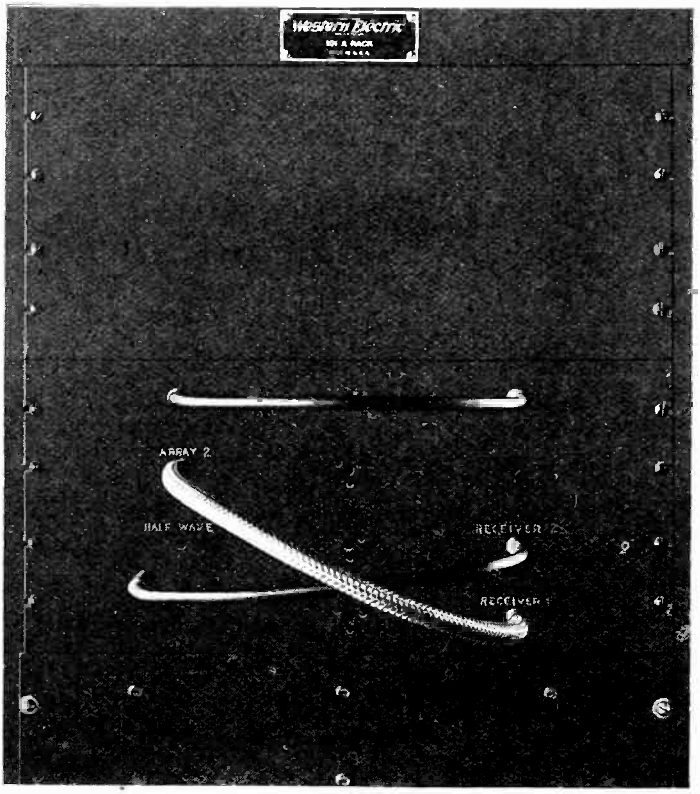

Fig. 22—An experimental radio-frequency jack board terminal for small concentric tube lines. Patch cords are constructed in the manner depicted in Fig. 21.

seen from the curves on this figure that the device is a means for obtaining a variable resistance which for most practical purposes is nonreactive. Additional advantages are that the device is rugged and that it may be designed to dissipate an appreciable amount of power.

For the purpose of testing transmitters and for other purposes in which the terminating network is required to dissipate several kilowatts, an iron wire line has been found to be of considerable utility. An iron wire line has the advantage that its impedance is almost independent of the frequency providing that the length of the line is sufficient. This impedance is very closely the characteristic impedance of the line. The far end of the line may be left either opened or closed. For operation at some one frequency the input impedance may be made more nearly equal to the characteristic impedance by means of a termination at the far end. For this purpose the scheme employing a short length of line as a parallel transforming impedance has been found very convenient.

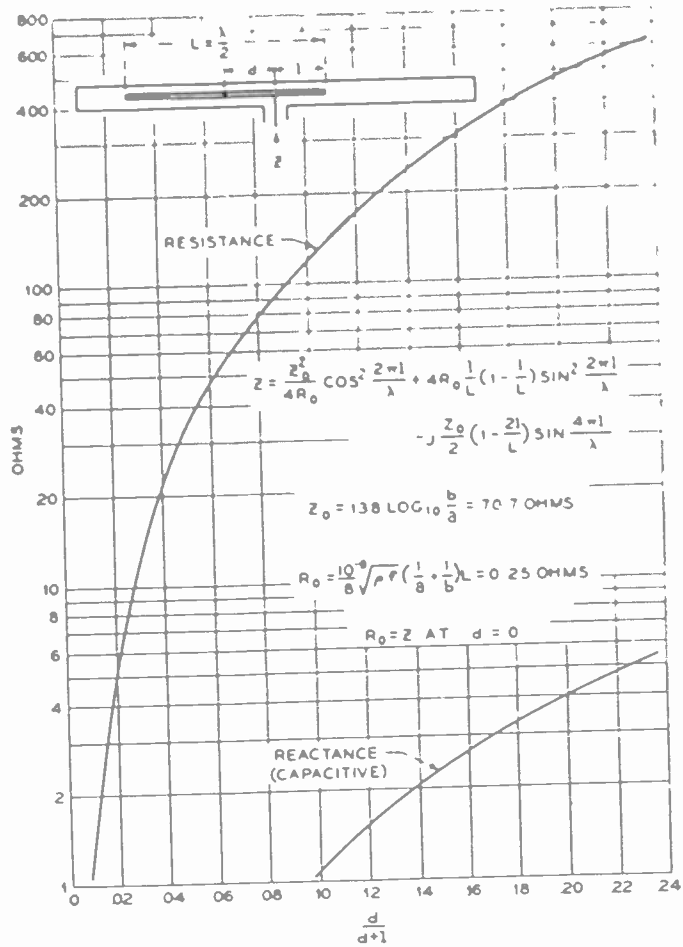

Fig. 23—A scheme for obtaining a calculable radio-frequency resistance standard which is essentially nonreactive and which is adjustable over wide limits of resistance.

An experimental attenuation curve for a 600-ohm line comprising 0.162-inch iron wire conductors is given in Fig. 24. The current entering the line was approximately one ampere. The measured resistivity for the iron is 12,300 e.m.u. The attenuation could be explained upon the basis that the permeability is 92.[21] This is not an unreasonable

[21] P. P. Cioffi, Bell Telephone Laboratories, New York City, found an initial permeability of 95 for a sample of the above wire.

value. In fact it may be very desirable to obtain the permeability of iron at radio frequencies by forming the material into a transmission line and observing the line attenuation and direct-current resistance.

An iron wire line 1600 feet long has been in use at the Deal Laboratories for several years for the purpose of testing transmitters. This line successfully dissipates 15 kw. at 20 megacycles.

VIII. Conclusion

In conclusion a few remarks on the relative utility of open wire and concentric tube lines may assist in selecting the most desirable con-

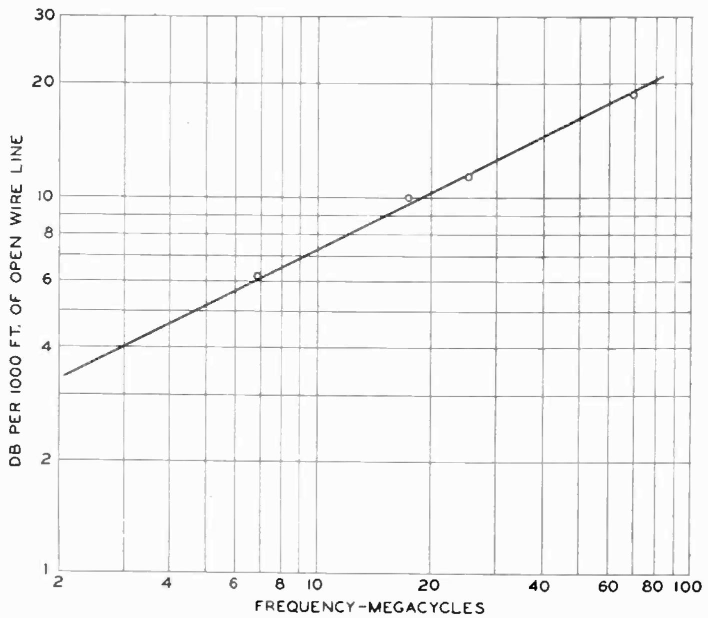

Fig. 24—The curve gives the attenuation in decibels for a balanced 600-ohm line constructed from No. 6 B & S iron wire.

struction for a particular service. A definite discrimination between the two is not readily made because the economics of the entire radio plant are involved.

Concentric lines are more costly than open wire lines. On the other hand, concentric lines permit the installation of a number of radio units within a single structure without incurring difficulties from cross-talk. The first cost and annual charges upon a compact installation may more than offset the cost of the lines when compared with an installation comprising several widely separated structures. Also, concentric lines may be constructed so as to be weatherproof.

There is little choice between the losses in open and concentric lines providing that a reasonable degree of current balance in the open wire lines is maintained. In order to obtain balances the open wire line terminal equipment both at the antenna and at the radio unit ends of the line must be carefully designed. The chief source of current unbalance difficulties resides in couplings between the antenna and an open wire line. These may be materially reduced but cannot be completely eliminated. Another source may be unbalances with respect to neutral at the radio unit.

Complete isolation of the antenna from the line can only be obtained with shielded lines. Similarly, complete isolation from static and other noise sources for which discrimination by the antenna is obtained can only be effected by shielded lines. This is particularly important in reception. In this case small concentric tube lines with losses as much as 2 db per 1000 feet may be used providing that the noise level is reduced by a corresponding amount.

We wish to acknowledge the helpful suggestions which have been received in the course of this work from H. T. Friis, J. C. Schelleng, and M. E. Strieby. Valuable advice on some of the mathematical questions encountered has been received from T. C. Fry.

Appendix

The following formulas for the power radiated by transmission lines were obtained by the conventional method of postulating the current distribution, calculating the electromagnetic fields and from the fields, the associated radiation by means of Poynting's theorem. As an independent check the same current distribution was postulated and the radiated power calculated following the methods of Pistolkors[22] and Bechmann[23]

Case I

An approximation for a line terminated in its characteristic impedance is a balanced two-wire line carrying a nonattenuated traveling wave. For this case the power radiated is:

$$P_1 = 120I^2\Bigg[\log_e(2L) - Ci(2L) + \frac{\sin(2L)}{(2L)} + 0.5772 - 1$$

$$- 2Ci(A) + \frac{\sin A}{A} - \frac{\sin(\sqrt{L^2+A^2} - L) + \sin(\sqrt{L^2+A^2} + L)}{2\sqrt{L^2+A^2}}$$

$$+ Ci(\sqrt{L^2+A^2} - L) + Ci(\sqrt{L^2+A^2} + L)\Bigg] \text{ watts} \tag{1}$$

[22] A. A. Pistolkors, Proc. I.R.E., p. 562, March, (1929).
[23] R. Bechmann, Proc. I.R.E., p. 461, March, (1931).

in which:

$$A = \frac{2\pi a}{\lambda}$$

$$L = \frac{2\pi l}{\lambda}$$

$$\frac{a}{\lambda} = \text{line spacing in wavelengths,}$$

$$\frac{l}{\lambda} = \text{line length in wavelengths,}$$

I = r-m-s value of current in each wire, and

$Ci(\ )$ = cosine integral.[24]

The equation simplifies considerably if it is assumed that a/λ is small so that:

$$\sin A \approx A \text{ and } L \gg A$$

under which condition:

$$P_1 = 160 I^2 \left(\frac{\pi a}{\lambda}\right)^2 \text{ watts.} \tag{1a}$$

The numerical constant in (1a) differs somewhat from a result published some time ago by Carson.[25]

Case II

An approximation for an unterminated line is a balanced two-wire line bearing standing waves of the form:

$$I_x = I \cos\left[\frac{2\pi x}{\lambda} + \frac{2\pi m}{\lambda}\right].$$

For this case the power radiated is:

$$P_2 = 60 I^2 \Bigg[\log_\epsilon (2L) - Ci(2L) + 2\cos M \cos (L - M)\frac{\sin L}{L} + 0.5772$$

$$- 2Ci(A) + \frac{\sin A}{A}[\cos^2 M + \cos^2 (L - M)]$$

$$- \cos^2 M - \cos^2 (L - M) - 2\cos M \cos (L - M)\frac{\sin\sqrt{L^2 + A^2}}{\sqrt{L^2 + A^2}}$$

[24] *See* Jahnke und Emde, "Funktionentafeln."
[25] John R. Carson, *Jour. A.I.E.E.*, p. 789, October, (1924).

$$+ Ci(\sqrt{L^2 + A^2} - L) + Ci(\sqrt{L^2 + A^2} + L) \Big] \text{ watts} \qquad (2)$$

in which,

$$M = 2\pi m/\lambda.$$

If as before it is assumed that the spacing is small and the line long the equation reduces to the following cases:

Case II-A

When the current is zero at both ends of the line, then:

$$\sin L = 0 \quad \text{and} \quad \sin M = \pm 1$$

and the radiated power is.

$$P_2 = 120 I^2 \left(\frac{\pi a}{\lambda}\right)^2 \text{ watts.} \qquad (2a)$$

This agrees with a result published by Manneback.[26]

Case II-B

When the current is zero at one end and maximum at the other end of the line, then:

$$\sin M = \pm 1 \text{ and } \cos L = 0$$

or,

$$\cos M = 1 \text{ and } \sin L = \pm 1$$

and the radiated power is:

$$P_2 = 80 I^2 \left(\frac{\pi a}{\lambda}\right)^2 \text{ watts.} \qquad (2b)$$

Case II-C

When the current is maximum at each end of the line, then:

$$\sin M = 0 \quad \text{and} \quad \cos L = \pm 1$$

and the radiated power is:

$$P_2 = 40 I^2 \left(\frac{\pi a}{\lambda}\right)^2 \text{ watts.} \qquad (2c)$$

The approximation for the power radiated by unbalanced currents is essentially the case of a long wire parallel to a perfect earth. The approximation may be obtained from (2) by assuming that power is radiated only in one hemisphere, which divides the numerical constant by a factor of two and by writing for a the quantity $2h$, h being the

[26] Charles Manneback, *Jour. A.I.E.E.*, p. 95, February, (1923).

height of the wire above ground. Equation (7) of the paper is written on the basis that ($\sin M = \pm 1$).

It is of interest to compare some of the above results with those for the case of a single conductor far removed from reflecting surfaces. If the wire is excited so as to bear standing waves of I r-m-s amperes maximum value the radiated power is:

$$P_3 = 30I^2\Big[0.5772 + \log_\epsilon (2L) - Ci(2L) - \cos^2 M - \cos^2 (L - M) + 2 \cos M \cos (L - M)\frac{\sin (L)}{(L)}\Big] \text{ watts.} \qquad (3)$$

If the wire is "terminated" so that there are no reflections from the ends a uniform current of I r-m-s amperes may be assumed to exist along the wire. In this case the power radiated is:

$$P_4 = 60I^2\Big[0.5772 - 1 + \log_\epsilon (2L) - Ci(2L) + \frac{\sin 2L}{2L}\Big] \text{ watts.} \qquad (4)$$

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A THEORETICAL COMPARISON OF COUPLED AMPLIFIERS WITH STAGGERED CIRCUITS*

By

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Summary—*Detuned or staggered single tuned circuits are compared theoretically with the so-called band-pass or coupled circuits. The networks in each case are compared with each other by expressing the ratio of the input to output voltage in terms of the amplifications, A_0 of a single tuned stage. It is shown that approximately the same results are obtained up to optimum coupling by either method. If very broad curves are desired coupled circuits give more amplification than staggered circuits. Resonance curves for each case are calculated. Some experimentally determined selectivity curves are given for staggered stages. These curve slopes bear out the theory given. Methods of obtaining the required detuning are discussed.*

Cascade amplification using a single tuned circuit as a coupling element between tubes has certain limitations as ordinarily used. The main limitation is the variation of the width of the resonance curve with the tuning frequency when the circuit is tuned over a frequency range two or three times that of the lowest frequency. The necessary width of the over-all resonance curve from a quality standpoint at the lower frequencies limits the number of tuned circuits which may be used. It is usually found that this number of tuned circuits is not sufficient for good high-frequency selectivity.

Various solutions of obtaining a more uniform resonance curve as the tuning frequency is varied have been proposed and adopted. Among the more important solutions are the use of the superheterodyne method and the utilization of the so-called band-pass or coupled circuits. It has been proposed also to stagger the stages but this possibility has not been investigated theoretically very extensively.

The advantages of the superheterodyne method are well known and will not be discussed here, except in so far as the results obtained by staggering the stages may be applied to a fixed frequency amplifier. The use of coupled circuits may or may not be a satisfactory solution to the problem of obtaining a uniform band width throughout a frequency range two or three times that of the lowest frequency to which the amplifier is tuned.

It is proposed here to compare single tuned circuit coupling elements slightly detuned from each other with coupled circuits using

* Decimal classification: R363. Original manuscript received by the Institute, March 13, 1931; revised manuscript received by the Institute, April 16, 1932.

either magnetic or capacity coupling. The equivalent circuit of two tuned circuits coupled by a common reactance is somewhat similar to that of two tuned circuits coupled by a tube as a study of Figs. 2A and 3 will show. The main difference is the presence of a source of energy in the case of two circuits coupled by a tube. Both circuits, however, have a common reactance so that it would be expected that somewhat similar results could be obtained by the use of either.

The staggered stages are compared with the coupled circuits by placing the amplification equations in similar forms. To do this, it is proposed to express the amplification equations of both types of cir-

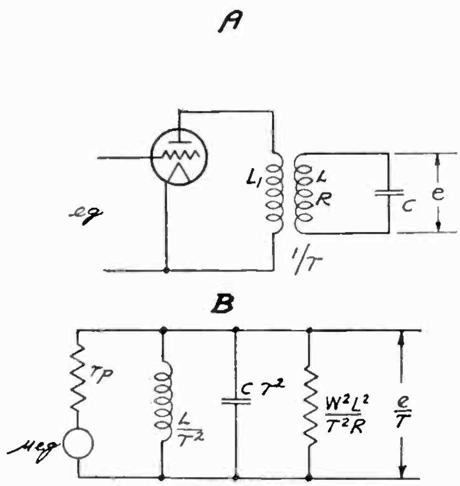

Fig. 1—Single tuned circuit.

cuits in terms of the amplification of a single stage having an untuned primary. The well-known amplification equations of the single stage are first placed in a form convenient for our purpose. Equations are next derived for coupled circuits and it is shown that the same equations may be used for either magnetic or capacity coupling. Equations are then derived for the case of circuits coupled through tubes so the effect of staggering may be found. This part of the work is a continuation of the study of a single tuned-input, tuned-output stage by Beatty[1] and that of the writer[2] of three tuned circuits coupled by two tubes. The two types of circuits are compared with each other using selectivity and amplification as a basis of comparison.

Single Stage Theory

Consider the circuit shown in Fig. 1A which represents a trans-

[1] R. T. Beatty, "The stability of the tuned-grid tuned-plate high-frequency amplifier," *Experimental Wireless and Wireless Engineer*, vol. 3, January, (1928).
[2] J. R. Nelson, "Circuit analysis applied to the screen-grid tube," Proc. I.R.E., vol. 17, no. 2; February, (1929).

former coupled stage having an untuned primary. The amplification or the ratio of e and e_g in the equivalent circuit Fig. 1B is

$$A_v = \frac{Tg_m}{T^2g + g_p + jT^2\left(\omega C - \frac{1}{\omega L}\right)} \tag{1}$$

where,

$$g_m = \mu/r_p$$
$$g_p = 1/r_p$$
$$g = \frac{R}{\omega^2 L^2}$$

which may be written as

$$A_v = \frac{Tg_m}{T^2g + g_p} \times \frac{1}{1 + jy} \quad \text{or} \quad A_0\frac{1}{1 + jy} \tag{2}$$

where,

$$y = \frac{T^2\left(\omega C = \frac{1}{\omega L}\right)}{T^2g + g_p}$$

$$A_0 = \frac{Tg_m}{T^2g + g_p}$$

A_0 is the amplification at resonance and may be written as

$$A_0 = g_m\frac{\omega L\frac{\omega L}{T}}{\frac{\omega^2L^2}{T^2}\cdot\frac{1}{r_p} + R} \tag{3}$$

considering $\omega L/T$ as the variable.
A_0 will be a maximum when

$$\frac{\omega^2L^2}{T^2} = r_pR\,. \tag{4}$$

y may also be written as

$$y = T^2\frac{1}{\omega L(g_p + gT^2)}\left[\left(\frac{\omega}{\omega_0}\right)^2 - 1\right] \text{ or } \frac{\omega L}{R + \frac{\omega^2L^2}{T^2r_p}}\left(\frac{\omega}{\omega_0}\right)^2 - 1 \tag{5}$$

where,

$$\omega_0^2 = \frac{1}{LC}.$$

Let,

$$\frac{\omega}{\omega_0} = 1 \pm \delta$$

$$y = \frac{\omega L}{R + \frac{\omega^2 L^2}{T^2 r_p}}(1 \pm 2\delta + \delta^2 - 1) = \frac{\omega L}{R + \frac{\omega^2 L^2}{T^2 r_p}}[\pm 2\delta + \delta^2]. \quad (6)$$

δ^2 may be neglected for small frequency differences. Other methods have developed an equation similar to (7) but with δ^2 neglected. This neglect of δ^2 has been the cause of calculated curves not agreeing with the measured curves at frequencies considerably removed from resonance. The selectivity defined as the ratio of the response at resonance to that at any other frequency is

$$S = 1 + jy. \quad (7)$$

Coupled Circuit Theory

Consider the circuits shown in Figs. 2A and B. The case for capacity coupling is derived in detail in Appendix I because the case for

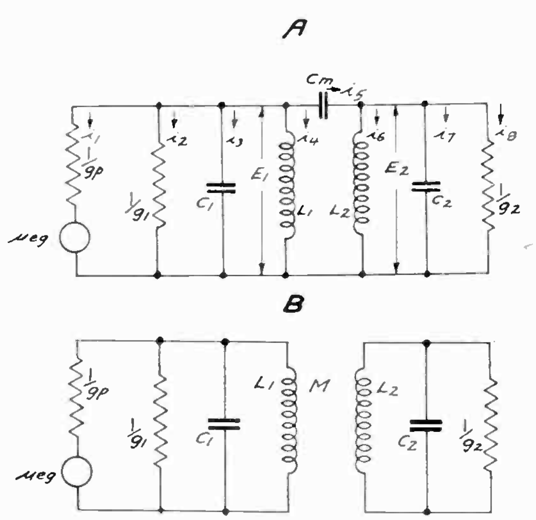

Fig. 2—Coupled circuits.
A—Capacity coupling.
B—Magnetic coupling.

magnetic coupling has been considered many times while capacity coupling has been somewhat neglected. It is shown in Appendix I that the amplification for either case may be written as

$$A_v = A_0 \frac{\omega M / R_2}{(1 + jy_1)(1 + jy_2) + \dfrac{\omega^2 M^2}{R_2 R_0}} \tag{8}$$

where ωM for the case of capacity coupling is

$$\omega M = \omega L \left(\frac{C_m}{C + C_m} \right) \tag{9}$$

and R_0 for both cases is

$$R_0 = R + \frac{\omega^2 L_1^2}{r_p} \cdot \tag{10}$$

The discussion of (8) will be deferred until the equations for tube coupled circuits are derived.

Tube Coupled Circuit Theory

Fig. 3 shows the equivalent circuit of the network which is going to be used in order to compare tube coupled staggered circuits with

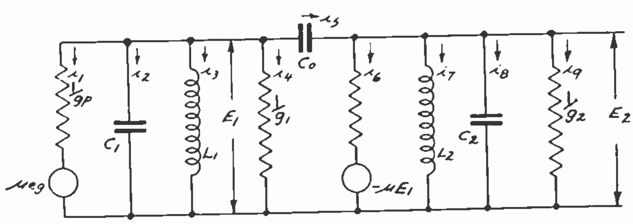

Fig. 3—Two-stage amplifier network, input not tuned.

reactance coupled circuits. The ratio of the voltage E_2 to the input voltage e_g is found in Appendix II and its value is given by (11).

$$A_v = A_0^2 \frac{1}{(1 + jy_1)(1 + jy_2) + jH} \tag{11}$$

where,

$$H = \frac{\omega C_0 g_m}{g^2} \cdot$$

If the first tube is not considered the relation between E_1, the voltage across the inductance L_1 which is $\omega L_1/B_1$ times the injected voltage e and E_2 may be placed in a similar form[1,2] except that A_0 appears to the first power and g_p will be absent in y_1.

$$A_v = A_0 \frac{1}{(1 + jy)(1 + jy) + jH} \cdot \tag{12}$$

Theoretical Comparison of Networks

So far we have derived equations for the ratio between input and output voltages for two tubes linked by two reactance coupled circuits, (8), and three tubes coupled by single tuned circuits, (11), or two tubes lined by single tuned circuits, (12). The above three equations are expressed in terms of the amplification obtainable with a single tube and tuned circuit. We are thus in a position to compare the results obtainable with coupled circuits to those obtainable with staggered single tuned circuits.

Equation (9) will be discussed first. Assume that $R_1 = R_2 = R$ and we have optimum coupling so $\omega^2 M^2 = R_2 R_0$ and that both circuits are tuned to resonance so $y_1 = y_2 = 0$. The value of (9) in terms of A_0 depends upon the value of $\omega^2 L_1^2/r_p$ in terms of R. Let $\omega^2 L_1^2/r_p = CR$ then R_0 becomes equal to $R(1+C)$. Equation (8) may then be written as

$$A_v = A_0 \frac{\sqrt{1+C}}{2}. \tag{13}$$

As was shown previously in the discussion for the single tuned circuit A_0 will be a maximum when $C=1$. Equation (13) then becomes

$$A_v = A_0/\sqrt{2}. \tag{14}$$

Next assume that C is very small, which condition could be realized by coupling the first tuned circuit weakly to the tube. The value of (13) then becomes

$$A_v = A_0/2. \tag{15}$$

In the latter case, however, A_0 will be less than it was in (14) so that the value of (15) will be less than that of (14). It is thus seen that the amplification obtained with two tuned circuits coupling two tubes varies between $1/2$ and $1/\sqrt{2}$ of that obtainable with a single tuned circuit.

There are two other cases, the first occurs when $\omega^2 M^2 < R_2 R_0$ and the second when $\omega^2 M^2 > R_2 R_0$. The value of y to make the value of A_v a maximum will be zero also when $\omega^2 M^2 < R_2 R_0$ but there will be two values of y to make A_v a maximum when $\omega^2 M^2 > R_2 R_0$. The performance of the circuits may be found in terms of A_0 for either case as was done in the above.

The selectivity of the coupled circuits compared to that of the single tuned circuits will be worked out next. Let $\omega^2 M^2/R_2 R_0 = K$. The amplification A_v at resonance as given by (8) becomes

$$A_r = A_0 \frac{\sqrt{(1 + C)K}}{1 + K} \tag{16}$$

At any other frequency A_r becomes

$$A_r = A_0 \frac{\sqrt{(1 + C)K}}{K + (1 + jy_1)(1 + jy_2)}. \tag{17}$$

Therefore,

$$S = \frac{K + (1 + jy_1)(1 + jy_2)}{1 + K}. \tag{18}$$

Thus, if K is unity the selectivity is approximately one half the square of that for a single tuned circuit, and if K is small the selectivity approaches the square of that of a single tuned circuit.

The effect of detuning or staggering two single tuned circuits will next be considered. Assume first that H is zero, (12), and that the circuits are detuned the same amount, and furthermore that the effective resistances are the same. Under these conditions $y_1 = -y_2 = y_0$ at the frequency, f_0, where the circuits are detuned the same amount. The conditions under which two peaks will be obtained in the amplification may be found quite readily. At the frequency f_0, A_r becomes

$$A_r = \frac{A_0}{(1 + jy_1)(1 + jy_2)} \quad \text{or} \quad \frac{A_0}{1 + y^2}. \tag{19}$$

When either value of y becomes zero the other becomes $\pm 2y_0$ so that the amplification becomes

$$A_r = \frac{A_0}{1 \pm 2jy}. \tag{20}$$

Thus when $1+y^2 >/ 1 \pm 2jy$ there will be two peaks and $1+y^2 = 1 \pm 2jy$ gives the critical value of y. It is thus seen that $y = \sqrt{2}$ gives the critical value and any value of $y > \sqrt{2}$ causes two peaks. The amplification for the critical value of y is

$$A_r = \frac{A_0}{3}. \tag{21}$$

The selectivity is

$$S = \frac{A_0}{3} \bigg/ \frac{A_0}{(1 + jy_1)(1 + jy_2)} \quad \text{or} \quad \frac{(1 + jy_1)(1 + jy_2)}{3}. \tag{22}$$

The above selectivity for frequencies considerably removed from resonance becomes approximately one third the square of the selectivity of a single tuned circuit. A comparison of (21) and (22) with (16) and (18) seems to indicate that better results are obtained with coupled circuits than with staggered circuits. The curve using staggered circuits with critical detuning will be somewhat wider than that for coupled circuits with critical coupling. Hence up to critical coupling detuning gives about the same results as coupled circuits. Beyond critical coupling, however, that is, for very broad curves, coupled circuits give considerably more gain. These points will be shown up better when typical resonance curves for the two cases are calculated.

The effect of regeneration when H is not zero in (12) will be considered next. The main difference mathematically between (9) and (12) is that the coupling adds a term to the real term in the denominator of (9) and the regeneration adds a term to the y term in (12). The term added by regeneration is always positive which will make the curve unsymmetrical if the circuits are detuned as the rest of the j term has a positive or negative value depending upon whether the frequency is higher or lower than the resonant frequency.

Let the circuits be detuned the same amount as before so that at resonance $y_1 = -y_2$ or in absolute value y_0. When the circuit is tuned to the high frequency A_v is

$$A_{v1} = \frac{A_0}{1 + j(2_y + H)}. \tag{24}$$

When the circuit is tuned to the low frequency A_v is

$$A_{v2} = \frac{A_0}{1 + j(2_y - H)} \tag{25}$$

and A_v at the frequency f_0 is

$$A_{vc} = \frac{A_0}{1 + (y_0^2 + jH)} \tag{26}$$

$$A_{v1} > A_{vc} \text{ when } |1 + y^2 + jH| > |1 + j(2y + H)|$$

or at the critical value

$$y^3 = 2y + 4H$$

hence for all values of H, y must be greater than the $\sqrt{2}$.

$$A_{v2} > A_{vc} \text{ when } |1 + 2 + jH| > |1 + j(2y - H)|$$

and at the critical value

$$y^3 = 2y - 4H.$$

Account has already been taken of the signs so that y must be positive and consequently any value of $H > 1/4$ makes $AV_2 > A_{vc}$ and if $H < 1/4$ y must be somewhere between 1 and $\sqrt{2}$ to make $A_{v2} > A_{vc}$.

The effect of regeneration then is to lower both the high-frequency amplification and the amplification at resonance and to raise the low-frequency amplification. If $H > 1/4$ the low-frequency amplification will always be greater than that at the resonant frequency regardless of the tuning. If the regeneration is not excessive the discussion given for H zero is not changed much so that no further discussion will be required. It is to be noted that the effect of regeneration becomes less as the detuning is increased.

Before going further the case for three tuned circuits and two tubes will be discussed briefly. There are so many possibilities that all that may be hoped for from a brief discussion is some general method of tuning to give a symmetrical resonance curve. It has been shown previously[2] that if the input of the first tube shown in Fig. 3 is tuned the relation between e_1, the voltage that would exist across the input circuit if there were no reaction, and e_3 is given by

$$A_v = \frac{1}{(1 + jy_1)(1 + jy)(1 + jy) + 2jH - H(y_2 + y_3)}. \tag{27}$$

If H is zero and any two are detuned the same amount one on one side and one on the other, the third should be tuned to the resonant frequency for a symmetrical curve and maximum amplification.

If H is not zero the low-frequency amplification will be greater than that at resonance and at the high-frequency if the third stage is tuned to resonance. The third stage should be tuned slightly higher than the resonance frequency and the less the detuning the more critical will be the tuning of the third stage.

TABLE I

Tube Parameters $r_p = 250{,}000$ ohms, $g_m = 1000$ micromhos
$f_0 = 600$ kilocycles
$L = 250$ microhenries
$R_1 = R_2 = 9.55$ ohms
$R_0 = R_1 + \frac{\omega^2 L_1^2}{r_p}$
R_2 in case $E = 11.8$ ohms

Curve	$\omega^2 M^2/R_0 R_2{}^1$	$\omega L/R_0$	$\omega L/R_2$	1	2	A_v Max.
A		80	100			67.7
B	1.1	80	100			34.4
C		80	100	0.005	0.005	39.0
D	2.43	80	100			36.1
E		80	80	0.0083	0.0083	27.6

Calculated Resonance Curves

Some calculated resonance curves are shown in Fig. 4. The values assumed are given in Table I. It was assumed that $\omega L/R$ remained constant over the frequency range covered. Curve A shows the curve for a single stage. Curve B shows the curve for a coupled stage with the coupling slightly greater than critical. In this curve and curve C, it is to be noted that the effective resistances are different due to the

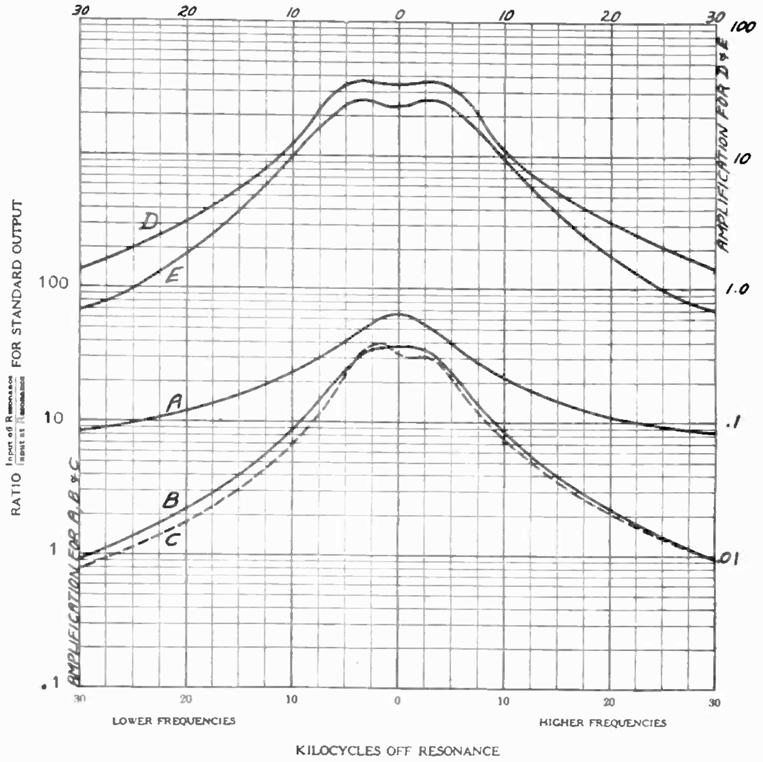

Fig. 4—Theoretical resonance curves.
A—Single stage
B—Coupled stage $\omega^2M^2/R_0R_2=1.1$
C—Two staggered stages to give same A_v as B.
D—Coupled stage $\omega^2M^2/R_0R_2=2.43$
E—Two staggered stages to give same width as D.

tube output resistance loading one circuit. Curve C shows the case of detuning the circuits to give approximately the same amplification. The curve is peaked on one side and the peak may be shifted from one side to the other merely by reversing the tuning of the stages. It is to be noted that there is very little difference in the width of the curves or in the selectivity.

Fig. $4E$ shows the case for the detuning carried past the critical value. The effective resistances were assumed to be equal in this case and as the factor H was assumed to be zero the curve is symmetrical.

Fig. 4*D* shows the case for the same coils used in a coupled circuit where the coupling is considerably greater than the critical value so as to give approximately the same width of curve. The main difference is in the application and the difference is in favor of the coupled circuits.

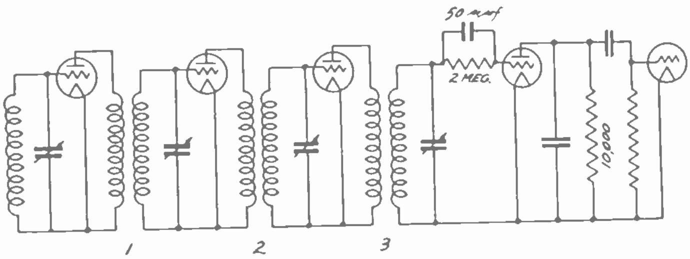

Fig. 5—Amplifier network used in obtaining experimental curves.

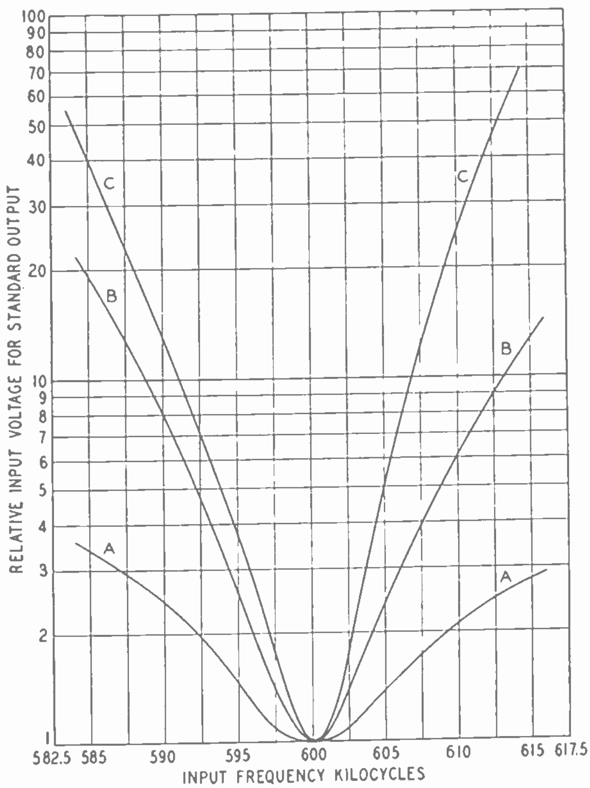

Fig. 6—Measured selectivity curves. Fig. 5 circuits in tune.
A—One stage.
B—Two stages.
C—Three stages.

Experimental Results

A commercial receiver was used in the following experimental work which consisted mainly in reasoning from the theory how to obtain certain results. The numbering of the tuned circuits together with the general layout is shown in Fig. 5*A*. Grid circuit power detector was used which caused considerable loading on the tuned circuit preceding the detector. Screen-grid tubes were used in the radio-frequency amplifier stages and a 27-type tube was used as a detector. The tube shields were removed so that considerable regeneration was present.

Selectivity curves were taken first with the stages in line. The results are shown in Fig. 6. The over-all curve for *C* is rather sharp for good quality and with one or more stages would give very poor results. An audio amplifier could be designed which would correct for the selectivity but there would be overcorrection at the higher carrier frequencies.

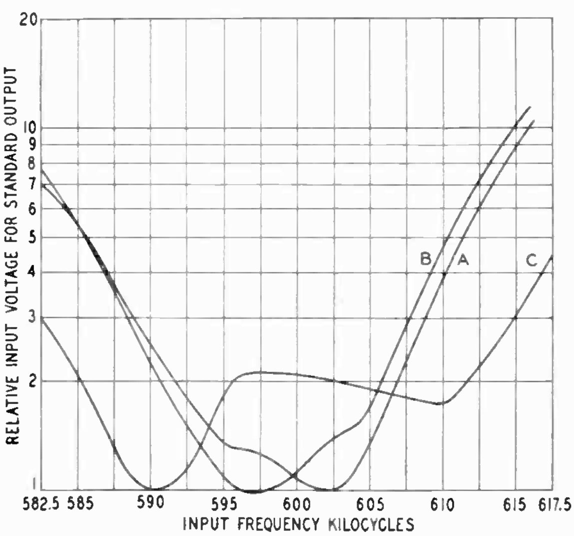

Fig. 7—Measured curves two stages. Fig. 5 circuits detuned.
A—No. 2 high 4 kc. No. 3 low 4 kc.
B—No. 2 low 4 kc. No. 3 high 4 kc.
C—No. 2 low 10 kc. No. 3 high 10 kc.

Two stages were used next. Each of these was detuned 4 kc from the frequency f_0, 600 kc. The results are shown by *A* and *B* in Fig. 7. The detuning was shifted in *A* and *B*, that is, one stage was tuned to a high frequency in one case and the other to a lower frequency, then the tuning was shifted. The shape of the curves were about the same in either case. It is to be noted that the curve shape is the same as that calculated in Fig. 4. It is thus seen that the curve *B* of Fig. 6 has been broadened considerably by detuning as shown in Fig. 7. The hump in the selectivity curve is caused by the circuit resistances not being the same. The results of detuning the stages 10 kc is shown by *C* of Fig. 7. This latter case is too broad but is given merely to show the possibilities.

Three stages were used next. The selectivity curve obtained by detuning two of the stages 4 kc on the high side and the other 4 kc on the low side is shown by *A* of Fig. 8. This curve is still fairly sharp but is an improvement over *C* of Fig. 6. The curve shape was improved by shifting the tuning as shown in *B* of Fig. 8. One circuit out of three should be tuned slightly on the high-frequency side of f_0, with the others tuned equally above and below resonance. This was done in *C* of Fig. 8. This was the best curve obtained so far and thus bears out the theory. If two of the circuits are detuned considerably the third

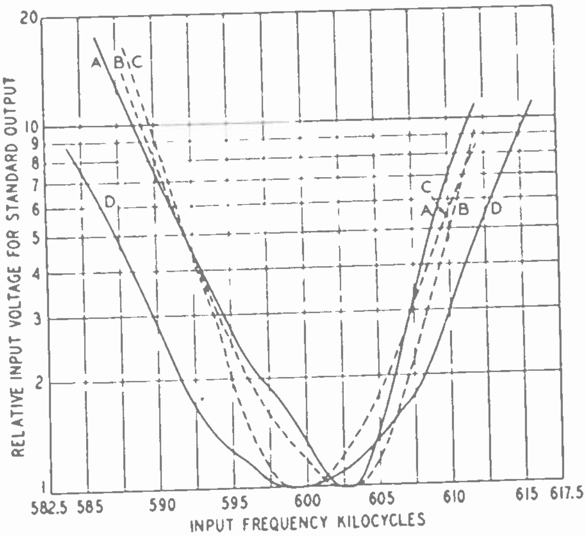

Fig. 8—Three stages of Fig. 5 staggered.
A—No. 1 high 4 kc. No. 2 high 4 kc. No. 3 low 4 kc.
B—No. 1 high 4 kc. No. 2 low 4 kc. No. 3 high 4 kc.
C—No. 1 high 1.5 kc. No. 2 low 4 kc. No. 3 high 4 kc.
D—No. 1 on resonance. No. 2 low 8 kc. No. 3 high 8 kc.

may be tuned to resonance without affecting the curve shape much as is shown by *D* of Fig. 8.

The curves given in Figs. 7 and 8 show that it is possible to detune circuits and obtain selectivity curves having the right shape. The correlative tunings for three stages are shown in Fig. 8. Briefly the correct method consists in tuning the middle stage slightly on the high-frequency side of f_0 and detuning the other two equal amounts. More circuits could be added thus making it possible to approach the square-bottom ideal selectivity curve.

Applications

The application to a fixed frequency amplifier is at once evident. It would not be desirable to detune the circuits at the low carrier frequency by adding capacity in an amplifier that covers several times the frequency range of the lowest frequency. The detuning could be realized in conventional amplifiers by designing the inductances to give the required results at the lowest frequency. The circuits could then either be aligned at the highest frequency or detuned the required amount and the differences in capacity would be negligible at the lower frequencies so that the design of the inductances would determine the detuning.

Detuning at the lowest carrier frequencies would make the ratio between the amplifications at the high and low frequencies greater than it is with the present conventional circuits which is bad enough. One or more circuits having an inverse frequency amplification characteristic to that of the conventional circuit could be used. Tuning the primary below the frequency band will give such a characteristic.

There are no real difficulties which would prevent this method of obtaining a more uniform frequency width of the resonance curve from being realized. The analysis given here shows merely the general results that may be obtained and it is not necessarily given for the purpose of advocating such an amplifier without further study. The possibilities of such an amplifier appear attractive, however. The curve may be widened at the lower frequencies, and may be made sharper at the higher frequencies than usually obtainable with conventional tuned circuits aligned because more tuned circuits may be used without affecting the quality too much at the lower carrier frequencies.

Appendix I

Derivation of Amplification Equations for Coupled Circuits

The circuits for magnetic and capacity coupled tuned circuits working between two tubes are shown in Fig. 2A and B. The ratio of the voltage E_2 applied to the grid of the second tube to the voltage e_g applied to the grid of the first tube is found for either case of coupling in terms of the amplification A_0 of a single tuned circuit working between two tubes. In the following analysis it is assumed that the effect of the resistance R of either tuned circuit may be represented by a resistance $1/g$ or ω^2L^2/R, shunted across the inductance L and capacity C. The following equations are derived from Fig. 2A.

$$i_1 + i_2 + i_3 + i_4 + i_5 = 0 \quad \text{1-A}$$

$$(E_1 - \mu e_g)g_p + \frac{E_1}{j\omega L_1} + E_1 j\omega C_1 + E_1 g_1 + (E_1 - E_2)j\omega C_m = 0 \quad \text{1-B}$$

$$E_1(g_p + g_1) + j\left(\omega C_m + \omega C_m - \frac{1}{\omega L}\right) - e_g g_m - Ej\omega C_m = 0 \quad \text{1-C}$$

or,

$$E_1 g_0(1 + jy_1) - e_g g_m - E_2 j\omega C_m = 0 \qquad \text{1-D}$$

where,

$$g = \frac{R}{\omega^2 L^2}$$

$$g_0 = g_p + g_1$$

$$y_1 = \frac{\omega(C_1 + C_m) - \dfrac{1}{\omega L_1}}{g_0}$$

$$-i_5 + i_6 + i_7 + i_8 = 0 \qquad \text{2-A}$$

$$-(E_1 - E_2)j\omega C_m + E_2(g + j)\left(\omega C_2 - \frac{1}{\omega L_2}\right) = 0 \qquad \text{2-B}$$

$$E_2(g_2 - j\left(\omega C_1 + \omega C_m - \frac{1}{\omega L_2}\right) - E_1 j\omega C_m = 0 \qquad \text{2-C}$$

or,

$$E_2 g_2(1 + jy) - E_1 j\omega C_m = 0 \qquad \text{2-D}$$

and,

$$E_1 = \frac{E_2 g_2(1 - jy_2)}{j\omega C_m} \qquad (3)$$

$$E_2\left[\frac{g_0 g_1}{j\omega C_m}(1 + jy_1)(1 + jy_2)\right] - E_2 j\omega C_2 = e_g g_m \qquad (4)$$

or,

$$\frac{E_2}{e_g} = \frac{g_m j\omega C_m}{g_0 g_2} \frac{2}{(1 + jy_1)(1 + jy_2) + \dfrac{\omega^2 C_m{}^2}{g_0 g_2}} \qquad (5)$$

or,

$$A_v = \frac{g_m}{g_0} \frac{j\omega C_m/g_2}{(1 + jy_1)(1 + jy_2) + \dfrac{\omega^2 C_m{}^2}{g_0 g_2}} \qquad (6)$$

but,

$$\frac{g_m}{g_0} = A_0$$

and,

$$j\omega C_m/g_2 = \frac{\omega L_2}{R_2}\frac{C_m}{C_m + C_2} = \frac{\omega M}{R_2}$$

where,

$$M = L_2\frac{C_m}{C_m + C_2}$$

$$A_v = A_0 \frac{j\omega M/R_2}{(1 + jy_1)(1 + jy_2) + \dfrac{\omega^2 M^2}{R_2\left(R_1 + \dfrac{\omega^2 L_1^2}{r_p}\right)}}. \tag{7}$$

We shall now show that the equations for the case of magnetic Fig. 2*B* may be expressed in the same form as those derived for Fig. 2*A*. This equivalence holds only over a narrow frequency band as will be shown later. It can be shown[3] that I_2 for Fig. 2*B* may be written as given below if $\omega C_1 r_p$ is considerably greater than unity, the usual condition.

$$I_2 = e_g\frac{g_m}{R_0}\frac{\dfrac{\omega M}{\omega C_0 R_2}}{\left(1 + \dfrac{jx_1}{R_0} + \dfrac{jx_2}{R_2} - \dfrac{x_1}{R_0}\dfrac{x_2}{R_2} + \dfrac{\omega^2 M^2}{R_0 R_2}\right)} \tag{8}$$

where,

$$x = \omega L - \frac{1}{\omega C}$$

$$R_0 = R_1 + \frac{\omega^2 L_1^2}{r_p}$$

let,

$$z = \frac{\omega L - \dfrac{1}{\omega C}}{R} = \frac{\omega^2 L^2}{R_2}\left(\frac{1}{\omega L} - \frac{1}{\omega^3 L^2 C}\right) = \frac{\omega L}{R}\left(1 - \frac{\omega_0^2}{\omega^2}\right).$$

$$I_2 = \frac{e_g g_m}{\omega C_1 R_0}\frac{1}{(1 + jz_1)(1 + jz_2) + \dfrac{\omega^2 M^2}{R_0 R_2}} \tag{9}$$

[3] J. R. Nelson, "Note on radio-frequency transformer coupled circuit theory," Proc. I.R.E., vol. 19, p. 1234; July, (1931).

if,

$$\omega C_2 = \omega C_1 = \omega C$$

$$A_v = \frac{E_2}{e_1} = -j\frac{g_{m}/\omega^2C^2}{R_1 + \dfrac{\omega^2L^2}{r_p}} \cdot \frac{\omega M/R_2}{(1+jz)(1+jz) + \dfrac{\omega^2M^2}{R_0R_2}} \qquad (10)$$

which is the same as for (7) except z occurs in (10) where y occurs in (7). The values of y and z are given below.

$$z = \frac{\omega L}{R}\left(1 - \left(\frac{\omega_0}{\omega}\right)^2\right)$$

and,

$$y = \frac{\omega L}{R}\left(\left(\frac{\omega}{\omega_0}\right)^2 - 1\right)$$

If the value of ω does not differ by more than ten per cent from that of ω_0 the expressions for y and z are approximately equivalent. Over this same range of frequency $1/\omega^2L^2$ is approximately equal to ω^2L^2 so that A_0 is approximately

$$A_0 \doteq \frac{g_{m}/\omega^2C^2}{R + \dfrac{\omega^2L^2}{r_p}}$$

Therefore, (10) may be written in the same form as (7).

$$A_v = \frac{j\omega M/R_2}{(1+jz_1)(1+jz_2) + \dfrac{\omega^2M^2}{R_0R_2}} \qquad 11)$$

We thus see that the same expression gives the amplification for both capacity and magnetic coupling provided the frequency does not differ by more than 10 per cent from the frequency to which the circuits are tuned.

Appendix II

Single Tuned Circuits Coupled by Tubes

Expressions have been previously derived[1,2] for a circuit similar to that shown in Fig. 3 except that voltage e was introduced in the coil L_1 instead of through the tube as shown. In the previous derivation the resistance $1/g_p$ and the generator μe_g were absent. It was desired in this work to have the ratio of the voltage E_2, applied to the succeeding tube, to that of voltage e_g applied to the input so the following equations of the network in Fig. 3 were derived.

$$i_1 + i_2 + i_3 + i_4 + i_5 = 0 \tag{1-A}$$

$$(E_1 - \mu e_g) + E_1 j\omega C_1 + \frac{E_1}{j\omega L_1} + E_1 g_1 + (E_1 - E_2) j\omega C_0 = 0 \tag{1-B}$$

$$E_1\left(g_p + g_1 + j\omega(C_1 + C_0) - \frac{1}{\omega L_1}\right) - E_2 j\omega C - e_g g_m = 0 \tag{1-C}$$

or,

$$E_1 g_a(1 + jy_1) - E_2 j\omega C_0 = g_m e_g \tag{1-D}$$

$$- i_5 + i_6 + i_7 + i_8 + i_9 = 0 \tag{2-A}$$

$$E_2 g_b(1 + jy_2) = - E_1 g_m + E_1 j\omega C_0. \tag{2-D}$$

Neglecting $j\omega C_0$ where added to gm

$$E_1 = \frac{E_2 g_b(1 + jy_2)}{g_m} \tag{3}$$

$$E_2 \frac{g_a g_b}{g_m}(1 + jy)(1 + jy_2) - E_2 j\omega C_0 = e_1 g_m \tag{4}$$

$$A_v = \frac{E_2}{e_g} = -\frac{g_m{}^2}{g_a g_b} \frac{1}{(1 + jy_1)(1 + jy_2) + \dfrac{j\omega C_0 g_m}{g_a g_b}} \tag{5}$$

$$A_v = A_0{}^2 \frac{1}{(1 + jy_1)(1 + jy_2) + jH} \tag{6}$$

where,

$$H = \frac{\omega C_0 g_m}{g_a g_b}.$$

BOOK REVIEWS

International Communications—The American Attitude, by Keith Clark. Published by the Columbia University Press, New York. 261 pages, 1931.

The growing interest in the development of international communication service is evidenced by the publication of a book bearing the above title, written by Miss Keith Clark, Assistant Professor of History and Government at Carleton College, Northfield, Minn.

The subject is treated in four chapters: Universal Postal Union, International Telegraph Union, Submarine Cables, and International Radio Union. Each chapter contains a résumé of the history of the international conferences dealing with that phase of the subject and reviews the agreements which have been drawn up in the form of Conventions or Treaties.

A section of each chapter is devoted to a discussion of the policy of the United States of America as indicated by the extent and nature of its participation in the several conferences and by its adherence or failure to adhere to the various Conventions.

A number of inaccuracies may be noted, such as the reference to an "International Radio Union," although there is formally no such organization. The author has made this book very useful, however, by including detailed references to official Government documents and to numerous other sources of information on the subjects treated.

*L. E. Whittemore

* American Telephone & Telegraph Company, New York.

Electricity. What It Is and How It Acts. Volume II, by Andrew W. Kramer. 290 pages, 112 figures. Published by the Technical Publishing Co., Chicago, Ill.

This volume is a continuation of Volume I of the same title which was devoted primarily to a discussion of electrical phenomena not involving radiation. In the present volume the discussion is extended to include radiation of various kinds. The method of presentation is simple and nonmathematical. Analogies are used freely and their use justified by the example of Faraday in visualizing the magnetic field. Each subject is treated from the most fundamental viewpoint so as to introduce the reader to some of the most recent theories of electrical phenomena. The existence of all forms of electromagnetic radiation from radio waves to cosmic rays is accounted for on the basis of electronic motion. The quantum theory and the wave theory of electromagnetic radiation are correlated by De Broglie's theory that light is both a particle and a wave at the same time. Classical experiments to check this and other theories presented in this book are outlined.

*S. S. Kirby

* Bureau of Standards, Washington, D. C.

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed on this page may be obtained gratis by addressing a request to the publisher or manufacturer.

A number of bulletins describing Esco products are available from the Electrical Specialty Company of Stamford, Conn. The line of Esco products includes high and low voltage motor-generators and dynamotors for radio transmitting service, plate power supply for automobile and yacht radio receivers, d-c and a-c motor rotary converters, and gasoline driven electric plants.

An illustrated loose-leaf folder issued by the General Industries Company of Elyria, Ohio, describes spring and electric motors for phonographs and combination phonographs and radio receivers. Numerous models, both for $33\frac{1}{3}$ and 78 r.p.m., are available. The electric motors are available for use on any commercial voltage and frequency.

Besides listing the various sizes of enameled copper wire manufactured by the Inca Manufacturing Division of National Electric Products Corporation, Fort Wayne, Ind., Bulletin No. 1 contains data on tests made on bare copper wire. Bulletin No. 2 deals with fabric insulated wires for radio purposes.

Audio and power transformers for use in radio receivers are illustrated in catalog 34R-3 of the Jefferson Electric Company, 1500 S. Laflin St., Chicago, Ill. Fuses, and accessories, also intended for use in radio receivers, are likewise described in this folder.

Wire products manufactured by the C. O. Jelliff Manufacturing Corporation, Southport, Conn., are illustrated in the 32-page catalog F. Most of the products illustrated are wire woven cloths or screenings. A separate folder describes, and includes a sample of, a wire cloth with protected edges manufactured for radio tube purposes.

Engineering Data is the title of a folder describing electro-dynamic speakers manufactured by the Jensen Radio Manufacturing Company, 6601 S. Laramie Ave., Chicago, Ill. The response characteristic of each speaker is included in the data given.

An instructive article on soldering is issued by the Kester Solder Company, 4201 Wrightwood Ave., Chicago, Ill., in the form of a pocket-size 32-page brochure entitled "Facts on Soldering." A separate folder describes various types of solder made for different purposes.

Two folders are available from the Meissner Manufacturing Company, 2815 West 19th St., Chicago, Ill., illustrating some of the various types of Meissner coil windings.

Wire wound resistors, commercially known as "Candohms" and manufactured by the Muter Company, 1255 South Michigan Ave., Chicago, Ill., are described in a folder giving the technical details of these resistors which are available in a variety of sizes.

RADIO ABSTRACTS AND REFERENCES

THIS is prepared monthly by the Bureau of Standards,* and is intended to cover the more important papers of interest to the professional radio engineer which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the "Classification of Radio Subjects: An Extension of the Dewey Decimal System," Bureau of Standards Circular No. 385, obtainable from the Superintendent of Documents, Government Printing Office, Washington, D.C., for 10 cents a copy. The classification also appeared in full on pp. 1433–56 of the August, 1930, issue of the PROCEEDINGS of the Institute of Radio Engineers.

The articles listed are not obtainable from the Government or the Institute of Radio Engineers, except when publications thereof. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R100. RADIO PRINCIPLES

R111.6 V. J. Andrew. The reception of frequency modulated radio signals. PROC. I.R.E., vol. 20, pp. 835–840; May, (1932).

The reception of frequency modulated signals by means of various adjustments of a tuned circuit is discussed nonmathematically.

R113 F. Ollendorff. Zur Frage der Wellenausbreitung in der Grossstadt. (On the question of wave propagation in the city.) *Elek. Nach. Tech.*, vol. 9, pp. 119–131; April, (1932).

The effect of the city on wave propagation is compared to that of a single isolated building with many receiving antennas. Maxwellian electromagnetic theory is used in the treatment of transmitted ground wave and reception in antenna systems.

R113.61 E. L. C. White. Automatic recording of Heaviside layer heights. *Nature* (London), vol. 129, p. 579; April 16, (1932).

Using the echo method of Breit and Tuve, a transmitter is caused to emit short pulse signals with a constant interval, which, together with their echoes, are arranged to give a stationary pattern on a cathode-ray oscillograph screen at the receiver.

R113.62 E. V. Appleton and G. Builder. Wireless echoes of short delay. *Proc. Phys. Soc.* (London), vol. 44, pp. 76–87; January 1, (1932).

An account of a simple method of producing short pulses of radio-frequency energy is given, together with notes on its application in the investigation of wireless echoes of short delay. Details of simultaneous visual and photographic methods of delineating such echoes are also described. The discussion of sample records and results serves as a basis for drawing conclusions concerning the relative advantages of the frequency-change and group-retardation methods of investigating the ionized regions of the upper atmosphere.

R113.62 Timing wireless echoes. *Wireless World and Radio Review*, vol. 30, pp. 459–460; May 4, (1932).

A brief survey of the nature of the problems to be studied during the Polar Year expedition.

R130 H. C. Steiner; A. C. Gable; H. T. Maser. Engineering features of gas-filled tubes. *Electrical Engineering*, vol. 51, pp. 312–317; May, (1932).

* This list compiled by Mr. A. H. Hodge and Miss E. M. Zandonini.

Fundamental principles, design features, and operating characteristics of gas-filled tubes. A bibliography is included.

R131 A. Gehrts. Raumladeströme von Oxydkathoden. (The space charge current of oxide cathodes.) *Zeit. für techn. Physik*, vol. 13, pp. 192–195; No. 4, (1932).

The relation between the calculated and measured space charge characteristic of the indirectly heated tube REN904 permitted the determination of the influence of the resistance of the oxide layer on the behavior of the characteristic.

R134 J. P. Woods. The calculation of detection performance for large signals. *Physics*, vol. 2, pp. 225–241; April, (1932).

It is a mathematically proved that a detector tube which has a broken straight line for its plate current grid voltage static characteristic will detect the standard radio signal without distortion, regardless of signal strength. The plate detection performance of a tube having a given curve for its static characteristic can be calculated from the extended power series which represents the characteristic. A Fourier analysis is used.

R143 W. Cauer. New theory of design of wave filters. *Physics*, vol. 2, pp. 242–268; April, (1932).

By a new mathematical theory of wave filters it is possible to realize as constant an image impedance as desired in the transmitting bands. In the new theory there are no mathematical differences in the consideration of proper attenuation or proper image impedance characteristics. The chief aim of the paper is to show the relation between the new theory and the older deisgn methods due to Zobel and others.

R148 E. Mallett. Apparent demodulation—Another viewpoint. *Wireless Engineer and Experimental Wireless* (London), vol. 9, pp. 248–252; May, (1932).

A simple mathematical treatment is given. It is shown that the "apparent demodulation" of the weak signal is in reality not a demodulation. The modulation of the weak wave is unaltered by the presence of the strong wave. The reduced acoustic signal is simply due to the fact that the modulated high frequency signal is not received at all during a part of the time, and during the rest of the time, it is not received at its normal strength.

R148 E. V. Appleton. Weak signal demodulation by a strong carrier. *Electronics*, vol. 4, p. 171; May, (1932).

A brief analysis of demodulation.

R200. Radio Measurements and Standardization

R200 A. E. Kennelly. Recent developments in magnetic units. *Electrical Engineering*, vol. 51, pp. 343–345; May, (1932).

To overcome confusion in the use of electromagnetic units, the International Electrotechnical Commission through its subcommittee on electric and magnetic units has been working for several years toward standardization in definition and nomenclature. Recent actions in that direction are discussed and a brief summary of earlier contingent actions is given.

R200 L. T. Robinson. Electrical units and their applications. *Electrical Engineering*, vol. 51, pp. 335–338; May, (1932).

A survey of the present status of electrical units. The fundamental units; mass, length, and time are said to be compared with a precision of 10^{-8}. The precision of other units is given.

R230 W. D. Oliphant. High-frequency coil measurements. *Jour. Sci. Inst.*,
×R240 vol. 9, pp. 121–122; April, (1932).

A simple, rapid, and accurate method of determining from a single series of experimental observations the inductance, self-capacity, and resistance of a coil at radio frequencies. A brief note is given on the measurement of self-capacity of such coils as high-frequency chokes.

R241 H. B. Brooks. The unit of electrical resistance. *Electrical Engineering*, vol. 51, pp. 338–341; May, (1932).

The mercury ohm, long recognized as a standard of resistance, with its elaborate technique and many disadvantages, has become obsolescent. The reasons for this together with a brief historical sketch of the ohm are presented.

R241 W. A. Fitch. Phase shift in radio transmitters. Proc. I.R.E., vol. 20, pp. 863–883; May, (1932).

A cathode-ray method of measuring phase shift is shown. Detuning of tank circuit causes phase shift. Effects of phase shift are treated mathematically, the most serious result being the creation of adjacent channel interference. The amplitude of the second order side band is roughly one per cent of the first order side band per degree of phase shift.

R243 M. Eppley. International standard of electromotive force. *Electrical Engineering*, vol. 51, pp. 341–343; May, (1932).

Although unremitting care must be exercised in their manufacture and handling, cadmium cells have proved to be convenient and reliable sources of standard e.m.f. Principal characteristics of these units and precautions to be observed in their use are pointed out in this article.

R243.1 J. L. McLaughlin. A linear electronic voltmeter. *QST*, vol. 16, pp. 18–19; May, (1932).

Construction details and characteristics of a vacuum-tube voltmeter having a 0–10- and 0–100-volt range.

R243.1 H. Kaden. Die Temperaturabhängigkeit von Messgeräten mit Trochengleichrichten und ihre Kompensation. (The variation with temperature of measuring apparatus containing dry rectifiers and its compensation.) *Hochfrequenz. und Elektroakustik*, vol. 39, pp. 115–122; April, (1932).

It is shown analytically that characteristics of a dry rectifier depend on three constants: the alternating resistance, the rectifying constant, and the temperature coefficient of resistance. Methods of compensation are given which include equations, circuit diagrams, and graphs.

R256 W. Lange. Über ein Eichmethod des Kondensator-mikrophons mit periodisch veränderlicher Ersatzkapazität. (On a calibration method for condenser microphones with a substitution arrangement.) *Hochfrequenz. und Elektroakustik*, vol. 39, pp. 133–136; April, (1932).

The effect of a known capacity change determines the membrane amplitude and thereby the noise pressure.

R261 H. A. Thomas. Developments in the testing of radio receivers. Paper read before the Wireless Section, Institution of Electrical Engineers on March 2, 1931. *Wireless Engineer and Experimental Wireless* (London), vol. 9, pp. 269–272; May, (1932).

The paper describes the improvements in the apparatus installed at the National Physical Laboratory for testing the performance of radio receivers. An attempt is made to draw up a schedule of tests by means of which the relative performance of different receivers may be assessed.

R262.3 S. I. Zilitinkewitsche. Der Widerstand des Röhrengenerators. (The resistance of vacuum-tube generators.) *Elek. Nach. Tech.*, vol. 9, pp. 132–136; April, (1932).

The physical meaning of the resistance of a vacuum-tube is discussed. Equations are devised for the effective resistance and the dynamic resistance of vacuum-tube generators.

R270 S. S. Kirby and K. A. Norton. Field intensity measurements at frequencies from 285 to 5400 kilocycles per second. RP429. *Bureau of Standards Journal of Research*, vol. 8, pp. 463–479; April, (1932). Proc. I.R.E., vol. 20, pp. 841–862; May, (1932).

Results of a large number of field intensity measurements are given. It is found that absorption increases with distance and frequency. Results are compared with Rolf's attenuation graphs and the Austin-Cohen transmission formula.

R300 Radio Apparatus and Equipment

R330 500 KW demountable valves. *Post Office Electrical Engineers Journal*, vol. 25, pp. 61–64; April, (1932).

Description of construction.

R330 G. Grammar. New tubes for class B audio. *QST*, vol. 16, pp. 14–15; May, (1932).

The '46 tube is especially adapted to class B amplification. With zero grid bias, and 400 volts plate potential only 6 milliamperes of current flow. The '82 mercury vapor rectifier is a full wave rectifier tube. It works well with the '46. Data are given.

R335 I. E. Mouromtseff. A new water-cooled power vacuum tube. Proc. I.R.E., vol. 20, pp. 783–807; May, (1932).

Description and operating characteristics of a new 200 KW vacuum tube.

R335 A new 6-volt output pentode. *QST*, vol. 16, pp. 20–21; May, (1932).

Characteristics of the Eveready Raytheon LA pentode. Its special feature is large output on small plate current and small input signal.

R337 W. Stockman. New field rectifier. *Wireless World & Radio Review*, vol. 30, p. 390; April 20, (1932).

A rectifying tube with a high voltage heater which allows one to dispense with a transformer is described.

R355.65 O. M. Hovgaard. Application of quartz plates to radio transmitters. Proc. I.R.E., vol. 20, pp. 767–782; May, (1932).

Disturbing elements encountered in application of quartz plates to broadcast and aircraft radio transmitters are discussed. A general procedure for minimizing such effects is considered from a circuit standpoint as well as in the light of practical experience. Apparatus which enables the operating staff to meet modern frequency requirements by monitoring the emitted carrier is described.

R357 N. C. Jamison. A cathode-ray frequency multiplier. *Physics*, vol. 2,
×R388 pp. 217–224; April, (1932).

A vacuum tube was constructed having a filament and slit arrangement for the production of a sheet of electrons, one pair of deflecting plates, and a receiving plate composed of a series of rectangular conductors. Adjacent conductors were insulated from one another and alternate conductors were in electrical connection. An alternating potential difference applied to the deflecting plates causes the electron beam to move across the surface of the receiving plate, current flowing first to one set of conductors and then to the other set. Tests with an alternating potential difference applied to the deflecting plates gave definite indications of receiving plate currents of two, four, and six times the frequency of the applied field.

R363 P. H. Osborn. A study of class B and C amplifier tank circuits. Proc. I.R.E., vol. 20, pp. 813–834; May, (1932).

An investigation was made of the relations between the constants L, C, and R of a power amplifier output tank circuit. Harmonic content was studied by means of a cathode-ray oscillograph. The results of the experimental data show that the power output and voltage and the efficiency depend only on the output circuit impedance and are independent of the L/C ratio.

R363 M. Wald. Elektrodynamischer Bandverstärker als Ersatz für Siebketten und Röhrenverstärker bei Tonfrequenztelegraphie. (An electrodynamic band-pass amplifier to replace band-pass filters and vacuum-tube amplifiers in audio-frequency telegraphy.) *Elek. Nach. Tech.*, vol. 9, pp. 91–111; March, (1932).

An electrodynamic band pass amplifier which allows selective reception and amplification of telegraph signals transmitted along a particular audio channel is described. The author shows that multistage devices of the same type are practicable and cites results of experimental tests.

R365.2 ×R385.5 A. L. Williams. Piezo-electric loudspeakers and microphones. *Electronics*, vol. 4, pp. 166–167; May, (1932).

A brief survey of progress since the 1931 meeting of the Institute of Radio Engineers, Chicago, Ill., in the commercial application of piezo-electric loud speakers and microphones.

R365.2 H. S. Knowles. Improved fidelity of two-speaker radio receivers. *Electronics*, vol. 4, pp. 154–156; May, (1932).

The advantages of using two loud speakers are given.

R365.2 P. Toulon. Extension du domaine des fréquences acoustiques reproduites avec fidélit par hautparleurs. (Extension of the range of faithful reproduction of acoustical frequencies by loudspeakers.) *L'Onde Electrique*, vol. 11, pp. 139–156; March, (1932).

The author takes up the idea of placing side by side several loud speakers in order to increase the fidelity of reproduction. Several precautions are necessary. The characteristics and arrangement of the loud speakers, and a description of the amplifiers are given.

R365.21 R. Raven-Hart. The importance of volume level. *Wireless World and Radio Review*, vol. 30, pp. 345–346; April 6, (1932).

The author discusses an important aspect of reproduction in relation to the original performance which is broadcast, and concludes that changes in volume level effected at the receiver must mar quality except where reproduction is at the same intensity as the original, and that the only alternative is to develop a new technique for broadcasting where the conductor himself makes the corrections necessary and not the control engineer or the listener at his receiver.

R385.5 A. L. Thuras; L. W. Giles; B. Leuvelink. A sensitive moving-coil microphone of high quality. *Bell Laboratories Record*, vol. 10, pp. 314–326; May, (1932).

A small microphone and its uses are described. It is very sensitive.

R386 L. R. Harris. Precision methods used in constructing electric wave filters for carrier systems. *Bell Sys. Tech. Jour.*, vol. 11, pp. 264–282; April, (1932).

This paper sets forth requirements which are met in the design of the filters for the type C carrier telephone system, and describes a new manufacturing adjustment made necessary by these requirements. This feature consists essentially of an inductance continually variable over a small range above and below its nominal value.

R388 G. Ulbricht. Eine neue Art Zeitproportionaler Kathodenstrahlablenkung. (A new kind of cathode-ray oscillograph time axis.) *Hochfrequenz. und Elektroakustik*, vol. 39, pp. 130–133; April, (1932).

A relaxation phenomena in screen-grid vacuum tubes is described. Utilizing this relaxation phenomenon by operating in the region where the characteristic curve has a negative slope, and by impressing a periodically changing voltage on the control grid, a voltage which provides a linear time axis for the cathode-ray tube, is obtained across a resistance in the plate circuit of the screen-grid tube. Experiments with this timing arrangement show that frequencies as high as 10 megacycles per second may be observed and photographed.

R388 G. F. Metcalf. A new cathode-ray oscillograph tube. *Electronics*, vol. 4, pp. 158–159; May, (1932).

A cathode-ray oscillograph tube with sensitive electrostatic control is described. Characteristics and some uses of the tube are given.

R800. Nonradio Subjects

536 E. J. C. Dixon. A resistance thermostat with light-sensitive cell

operation. *Post Office Electrical Engineers Journal*, vol. 25, pp. 65–67; April, (1932).

A method of using a resistance thermostat with a reflecting galvanometer and a light-sensitive cell to control the operation of the heater relay is described and practical circuits are given for working from power line. Sensitivity of device described is ±0.1 per cent but could be improved by using a more sensitive galvanometer.

537.65 R. C. Colwell. The vibrations of quartz plates. Proc. I.R.E., vol. 20, pp. 808–812; May, (1932).

It is pointed out that the mathematical theory of Chladni plates gives a general equation for the nodal lines of square plates. The equation is also applicable to quartz plates in so far as the forms of the nodal lines are concerned. It gives no information regarding the manner of vibration of the quartz plate. Some figures are shown.

621.313 Slow motion motor will run, indefinitely. *Electric Journal*, vol. 29, pp. 217–218; May, (1932). *Electrical Engineering*, vol. 51, pp. 319–320; May, (1932).

A motor is described which might have application as a frequency integrator, clock, or chronograph drive. Tbe clock consists of a series of concentric rings of iron which have teeth cut in both inside and outside of ring. The outside ring is mounted to rotate inside a toothed ring with revolving magnetic field. The number of teeth constituting a motor differ by small numbers, for example, 120–118. This difference in number of teeth causes a great speed reduction. The inner rim of the outer rotating ring acts as stator for a ring which rotates inside it. In this motor this wheel drives the minute hand. The inside of this ring acts as stator for still another rotor and the hour hand, etc. The process could be extended still further making a counting instrument of wide range for frequency.

621.313.7 J. A. Darbyshire. Rectifier circuits for measurement of small alternating currents. *Jour. Sci. Inst.*, vol. 9, pp. 123–127; April, (1932).

The properties are discussed of circuits designed for the study of the ripple that existed in a high tension circuit, used in connection with some electron diffraction experiments.

621.319.2 A. Rosen. The calculation of the propagation constants of uniform lines. *Post Office Electrical Engineers Journal*, vol. 25, pp. 67–71; April, (1932).

Equations and data for calculation purposes.

621.374.31 K. Howe. Automatic voltage regulation. *Electronics*, vol. 4, pp. 164–165; May, (1932).

A general description of the voltage regulating transformer.

621.374.7 ×537.65 V. v. Philippoff, Der piezoelektrische Oszillograph. (The piezoelectric oscillograph.) *Elektrotech. Zeits.*, vol. 53, pp. 405–408; April 28, (1932).

An oscillograph using the piezo-electric principle and especially useful at high frequency is described.

621.375.1 ×R339 L. R. Harness. Amplifier tubes for industrial applications. *Electric Journal*, vol. 29, pp. 233–35; May, (1932).

Purpose, operation, characteristics, and performance.

621.375.1 H. Strohmeyer. Zusammenarbeit von Elektronen und magnetischen Relais. (The operation of magnetic relays with electron-tube relays.) *Hochfrequenz. und Elektroakustik*, vol. 39, pp. 122–127; April, (1932).

The author points out that the operating time of a magnetic relay is determined by the time constant of the circuit which definitely limits quick relay action. If, however, the magnetic relay is operated in the plate circuit of a vacuum tube, the operating time of the combination may be increased or decreased to any desired value over a wide range. A vacuum tube having a high internal resistance and a large amplification factor is necessary where extremely fast relay action is desired.

621.375.1 C. B. Upp. A heavy duty industrial amplifier tube. *Electronics*, vol. 4, p. 162; May, (1932).

Characteristics and description.

621.385.96 P. Toulon. La reproduction des enregistrements inscrits sur pellicule cinématographique. (The reproduction of records recorded on moving picture film.) *L'Onde Electrique*, vol. 11, pp. 129–138; March, (1932).

The author examines the difficulties of accurate reproduction of frequencies by loud speakers. A description is given of primary amplifiers and power amplifiers, and the method used for reproducing accurately the sounds with loud speakers.

CONTRIBUTORS TO THIS ISSUE

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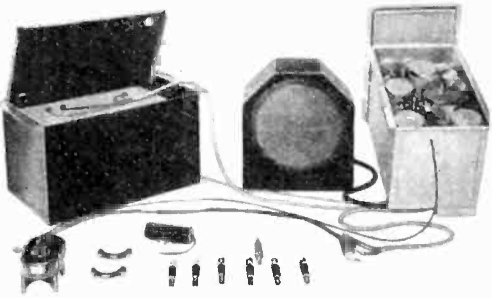
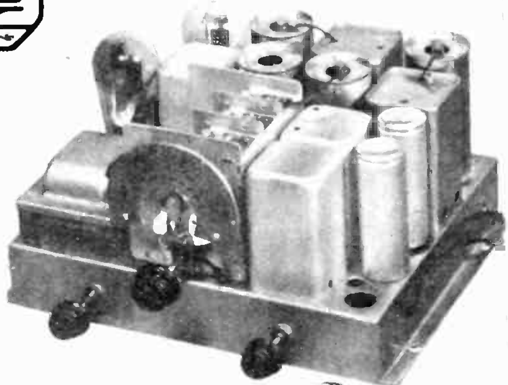
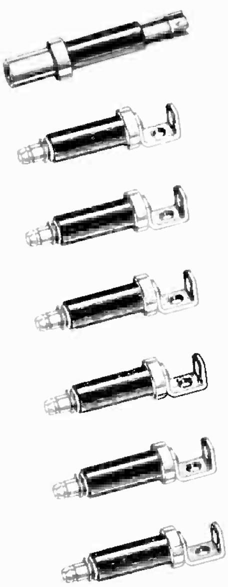
SOLID MOLDED RESISTORS

FOR MOTOR CAR RADIO — A·B — **FOR CONSOLE RADIO**

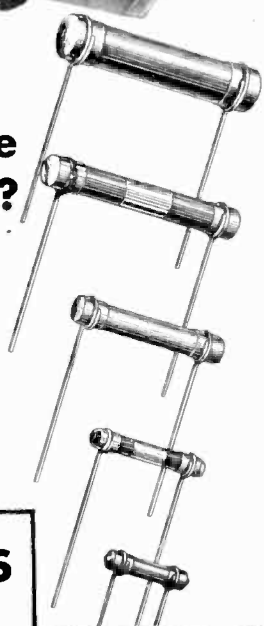






Do Your Sets go "Sour" in Service due to poor Resistors?

The satisfactory operation of your receivers depends to a great extent upon the accuracy of your fixed resistors. Bradleyunit Resistors are used by the world's largest radio manufacturers, because their resistance value is stable under varying conditions of load, temperature, and moisture. They are not subject to wide fluctuation due to long use. Don't risk the reputation of your receiver with poor resistors. Get an Allen-Bradley quotation on your next order.



BRADLEYOMETERS

Type A. Single Bradleyometer

The Bradleyometer is a potentiometer with approximately fifty solid resistance discs interleaved between metal discs.

The total number of discs can be arranged in accordance with any resistance-rotation curve.

One or more Bradleyometers can be arranged to operate with one knob. Mixer controls, T-pad and H-pad attenuators and other complex controls can be provided.

Type AA. Double Bradleyometer

Type AAA. Triple Bradleyometer

Interference from ignition systems in radio-equipped motor cars is suppressed with Bradley Suppressors. When used with suitable by-pass condensers in ignition circuit, shielded ignition cables are unnecessary.

Bradleyunit Resistors are made in five sizes, with or without leads, and are color coded to meet set manufacturers' specifications. These solid molded resistors are accurately calibrated and have great mechanical strength.

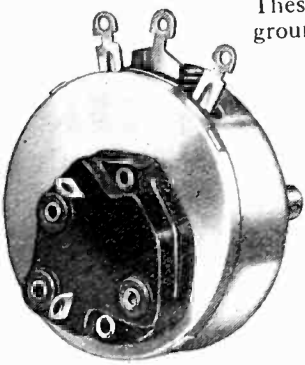
Allen-Bradley Co., 116 W. Greenfield Ave., Milwaukee, Wisconsin

ALLEN-BRADLEY RESISTORS
Produced by the makers of Allen-Bradley Control Apparatus

Introducing the new Nos. 90 and 95 Series Heavy Duty Units, available with or without "H" or "T" Type Snap Switches

The demand for heavy duty units capable of meeting the critical resistance gradients and heavy load requirements of today's receiving sets is fully met by these new volume and tone controls. An exterior view of the new No. 90 is shown at the left. The right-hand illustration shows the interior construction of our No. 95 Series.

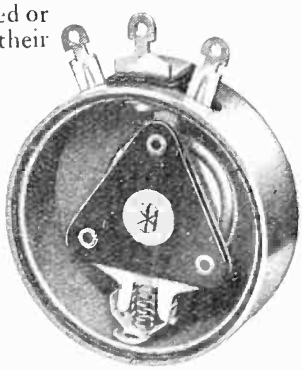
These units are made with insulated or grounded shafts, as desired. In their construction the same type of double wiping contact is used that has proved so completely successful in our No. 20 Series. This contact method insures quiet operation by permanently maintaining clean, bright contacting surfaces. These new units may be obtained without switch, or with choice of our Type T or Type H Switches incorporated. A tap on the resistance element at practically any position also may be obtained.



No. 90 Series with Type T S.P.S.T. Low Torque Switch

No. 90 Series Units

Equipped with a combination wire-wound and composition resistance element, the No. 90 Series is ideal for use as a tone control across the plates of a pair of 47's, for the hazard of resistance element burning out at high output levels is effectively eliminated. Shaft is insulated to withstand 2,000 volts. If grounded shaft is desired, specify No. 91 Control.



No. 95 Series, interior view, showing double wiping contact

No. 95 Series Units

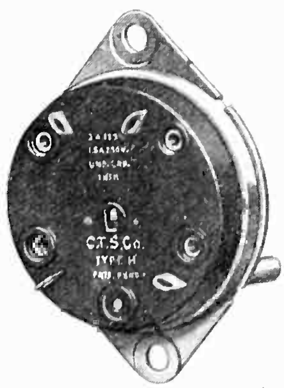
We can furnish the new No. 95 in materially higher resistance values than have heretofore been available in wire-wound radio control units. These resistance elements can be supplied in single or multi-section windings. Insulated shaft will withstand in excess of 2,000 volts. For grounded shaft unit, specify No. 96 Control.

CHICAGO TELEPHONE SUPPLY CO. OFFERS RADIO'S MOST COMPREHENSIVE LINE OF VOLUME AND TONE CONTROLS

No. 11 Snap Switch

New No. 11 Snap Switch

Has lowest operating torque of any commercial switch, smallest knob movement, and extremely low and uniform contact resistance. Positive kickoff. Double bearing "cold" cam cannot bind or wabble in operation. Electrostatic shield prevents hum pickup by volume control from live parts of switch. Supplied in S. P. S. T. type for panel mounting or in combination with our various types of volume and tone controls. Approved by Underwriters' Laboratories for 3 A., 125 v., 1.5 A., 250 v., AC or DC.



No. H Switch for panel mounting, D.P.S.T. or S.P.D.T.

No. H interior view

No. H Snap Switch

Like our No. 11, the No. H Snap Switch has low operating torque and low, uniform contact resistance. Supplied for panel mounting, as illustrated at left, or in combination with our various types of volume and tone controls. Made in D. P. S. T. or S. P. D. T. types. Bakelite housing; silver-plated contacts, with wide separation of all live parts from the cam. Approved by Underwriters' Laboratories for 3 A., 125 v., 1.5 A., 250 v., AC or DC.

CHICAGO TELEPHONE SUPPLY CO.

HERBERT H. FROST, Inc. • SALES DIVISION

ELKHART, INDIANA

CHANGE IN MAILING ADDRESS OR BUSINESS TITLE

Members of the Institute are asked to use this form for notifying the Institute office of a change in their mailing address or the listing of their company affiliation or title in the Year Book.

The Secretary,
THE INSTITUTE OF RADIO ENGINEERS,
33 West 39th Street,
New York, N.Y.

Dear Sir:

Effective .... please note change in my address
(date)

for mail as follows:

FROM

....
(Name)

....
(Street Address)

....
(City and State)

TO NEW ADDRESS

....
(Street Address)

....
(City and State)

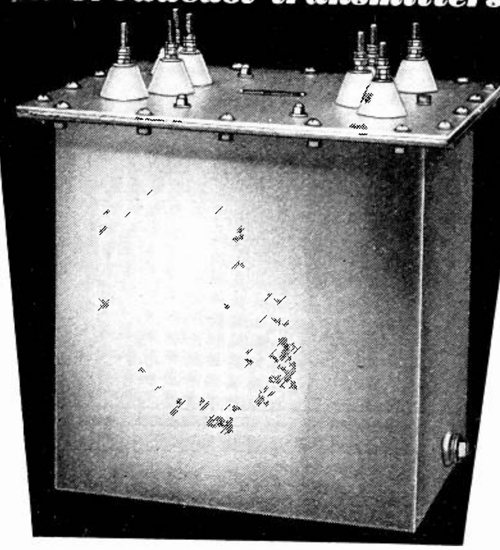
Please note following change in my Year Book listing.

....
(Title)

....
(Company Name)

....
(Company Address)

Transformers *for* Class B *amplification in broadcast transmitters*



Output transformer for use between push-pull, class "B" stage using 204-type tubes and a class "C" amplifier. Operating level 50d B; primary 1500/1500 ohms; secondary 4750 ohms; tested at 15,000 volts; oil insulated.

FOR six months AmerTran engineers have been studying all problems associated with Class B Amplifiers. With this experience as a background a complete line of audio-frequency transformers (input and output) has been especially designed for use with tubes suitable for Class B operation.

Large output transformers for use in Class B Amplifiers of broadcast transmitters are of the design illustrated above and have the following features:

1. Oil immersed with Isolantite bushings. This permits insulators testing at a voltage which is considerably in excess of any peak which might be experienced in actual practice.
2. Welded aluminum tank provides complete r.f. shielding.
3. Wire used in primary and secondary windings is of a size which insures low d.c. resistance and ample current capacity.
4. Primary sections are balanced within 0.5% and the same phase angle exists in each section.
5. Core laminations of the best quality high-permeability alloy are operated at a low density.
6. Coil structure insures low distributed capacity, low capacity coupling, and high inductance coupling.
7. High efficiency insured by excellent regulation, constant input impedance, and unusually satisfactory frequency characteristics throughout the band of 30 to 10,000 cycles.

Complete information on transformers for use with a specific type of tube will be mailed promptly on request.

AMERICAN TRANSFORMER COMPANY
178 Emmet Street **Newark, N. J.**

AMERTRAN
Audio Transformers

Cornell
TRADE MARK REG.

Type BC

This is a new and very popular model with two capacities on the same unit. The lug is common for both capacities. Both sections must be the same voltage. This is a most economical unit.

"CUB" CONDENSERS

The phenomenal growth of our Company since 1924 to its present enviable position as the leading manufacturer in the United States of fractional capacity paper dielectric capacitors has given us a sales volume with which we have been able to utilize automatic equipment for operations heretofore only accomplished by manual labor.

These savings we have passed along to our customers, and the consequence has been that our customers have been receiving the highest quality merchandise at prices consistent with economic conditions.

Many production economies may possibly be achieved by consultation with our engineers on your condensor problems.

WRITE FOR NEW CATALOG A

This new catalog describes our complete line of Paper Dielectric Capacitors for radio, television and ignition, and Resistors for radio and television. Write for a copy today.

CORNELL ELECTRIC MFG. CO., Inc.

Manufacturers of

CORNELL "CUB" CONDENSERS

Filter and By-Pass Condensers, Interference Filters and All Types of Paper Dielectric Capacitors and Resistors

LONG ISLAND CITY, NEW YORK

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

The Institute of Radio Engineers

Incorporated

33 West 39th Street, New York, N. Y.

APPLICATION FOR ASSOCIATE MEMBERSHIP

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

To the Board of Direction

Gentlemen:

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

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

....
(Sign with pen)

....
(Address for mail)

....
(Date) (City and State)

References:

(Signature of references not required here)

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

Address .... Address ....

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

Mr. ....

Address ....

City and State ....

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

ARTICLE II—MEMBERSHIP

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

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

ARTICLE III—ADMISSION AND EXPULSIONS

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

ENTRANCE FEE SHOULD ACCOMPANY APPLICATION

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

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

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

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

Business Address ....

Permanent Home Address ....

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

Education ....

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

TRAINING AND PROFESSIONAL EXPERIENCE

DATES	

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

Receipt Acknowledged............ Elected............ Deferred............
Grade .... Advised of Election........ . This Record Filed..........

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

EMPLOYMENT PAGE

Advertisements on this page are available to members of the Institute of Radio Engineers and to manufacturing concerns who wish to secure trained men for positions. All material for publication on this page is subject to editing from the Institute office and must be sent in by the 15th of the month previous to the month of publication. (July 15th for August PROCEEDINGS IRE, etc.) Employment blanks and rates will be supplied by the Institute office. Address requests for such forms to the Institute of Radio Engineers, 33 West 39th Street, New York City, N.Y.

MANUFACTURERS and others seeking radio engineers are invited to address replies to these advertisements at the Box Number indicated, care the Institute of Radio Engineers. All replies will be forwarded direct to the advertiser.

BROADCAST ENGINEER with six years experience on the design and construction of broadcast station apparatus and with some experience in development work on synchronization apparatus desires position with broadcast station or manufacturing concern. High school graduate. Age 26. Married. Will travel. Box 127.

COMMUNICATION ENGINEER with seven years experience including design and specifications for carrier telephone and telegraph equipment, cable testing and laying, and broadcast program circuits for Bell System and public utility corporation. Desires any position in communication industry, especially radio or electrical laboratory or studio control and operating work. B.S. in E.E. 1927. Age 26. Married. Will travel. Box 128.

PATENT ATTORNEY with ten years of patent law experience and six years of engineering background desires position with a corporation or firm of attorneys. Experienced in all branches of patent procedure and has organized a patent department. Is especially qualified for handling radio and sound system cases, as well as electrical cases generally, and also chemical and photographic matters. College graduate. Member I.R.E. Age 41. Married. Box 129.

BROADCAST OPERATOR with four and one half years experience as ship operator and two and one half years experience as announcer and broadcast station operator desires position operating broadcast, police, or airway station. Holds first class commercial license. High school graduate. Age 30. Will travel. Box 130.

EUROPEAN RADIO ENGINEER, one and one half years development of automatic receivers for high and low speed reception, and public address equipment and servicing, two years in charge of design and construction of amplifiers, five and one half years consulting engineer for European government. Familiar with European trade conditions and can speak several European languages. Desires position as European technical representative of American firm or as research worker. Age 31. Box 126.

Alphabetical Index to Advertisements

Piezo Electric Crystals

Does the frequency of your monitor comply with the new regulations of being within the plus or minus 50 cycle limits? If not, we are at your service to adjust your monitor to within those limits. *Ship your monitor to us* for either adjustment or grinding a new crystal if necessary. Our charge for this service is right, and will require but *seven* to *ten* days to perform this work. *Ask any broadcast radio engineer* what *he* thinks of our service.

CRYSTALS CRYSTALS CRYSTALS

Prices for grinding POWER CRYSTALS in the various frequency bands are as follows:

Frequency Range	
100 to 1500 Kc	$40.00
1501 to 3000 Kc	$45.00
3001 to 4000 Kc	$50.00
4001 to 6000 Kc	$60.00

The above prices include holder of our Standard design, and the crystals will be ground to within .03% of your specified frequency. If crystal is wanted unmounted deduct $5.00 from the above prices. Delivery two days after receipt of your order. In ordering please specify type tube, plate voltage and operating temperature.

Special Prices Will Be Quoted in Quantities of Ten or More

POWER CRYSTALS FOR AMATEUR USE

The prices below are for grinding a crystal to a frequency selected by us unmounted, (if wanted mounted in our Standard Holder add $5.00 to the prices below) said crystal to be ground for POWER use and we will state the frequency accurate to better than a tenth of one per-cent. IMMEDIATE SHIPMENT CAN BE MADE.

1715 to 2000 Kc band	$12.00 each
3500 to 4000 Kc band	$15.00 each

LOW FREQUENCY STANDARD CRYSTALS

We have stock available to grind crystals as low as 13 Kilo-cycles. Prices quoted upon receipt of your specifications.

PIONEERS: POWER CRYSTALS SINCE 1925

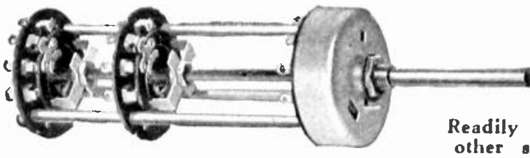
This record is proof to you that you will get the best there is in Piezo Electric Power Crystals. Get the best. It is more economical.

Scientific Radio Service

"THE CRYSTAL SPECIALISTS"

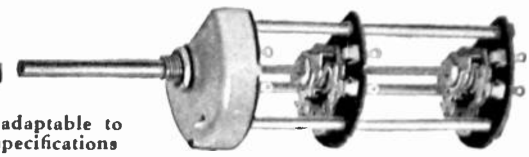
124 Jackson Ave., University Park Hyattsville, Maryland

SPECIAL ITEMS FOR RADIO MANUFACTURERS

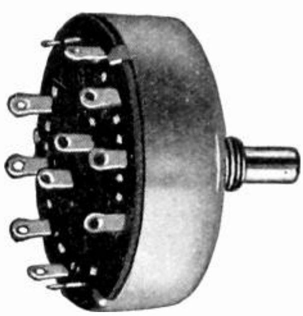


No. 7280A
3 gang, 4 positions
single circuit

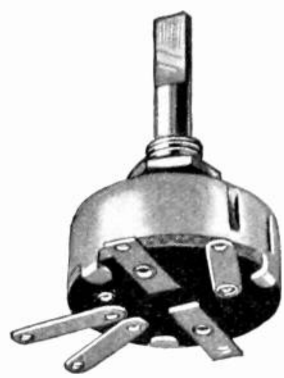
Readily adaptable to
other specifications



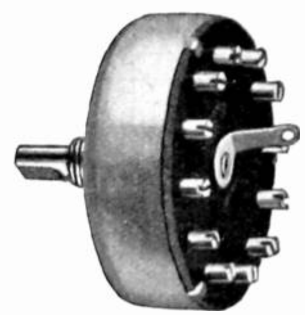
No. 7260A
3 gang, 2 positions
2 circuits



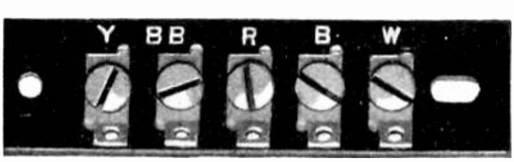
No. 7150A
5 positions, 2 circuits



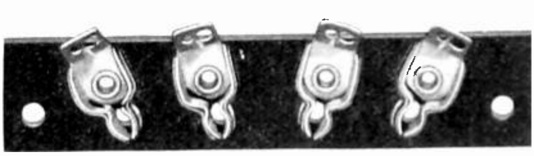
No. 7120A
Combined Acoustic and muting switch 2 or 3 positions



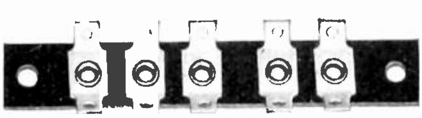
No. 7220A
12 positions, single circuit



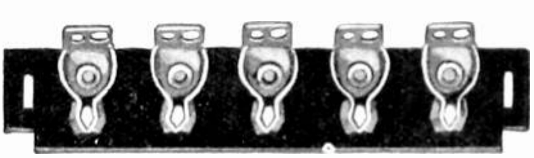
Terminal Strip
5 Lugs No. 6880A, 4 lugs No. 6990A



Terminal Strip
4 Clips, No. 6830A



Terminal Strip
5 Lugs, No. 6950A



Terminal Strip
5 Clips, No. A6310AA, 4 Clips No. A6320AA

SORENG - MANEGOLD CO.

1901-9 Clybourn Avenue, CHICAGO, ILL.

1646 W. Adams St.
Los Angeles, Cal.

701 Stephenson Bldg.
Detroit, Mich.

245-5th Ave.
New York City

Preprints Available

A LIMITED number of preprints of some of the papers presented at the technical sessions of the Twentieth Anniversary Convention of the Institute are available without cost to members of the Institute. They are as follows:

2. "Design of Resistors for Precise High-Frequency Measurements," by L. Behr and R. E. Tarpley.

3. "The Campbell-Shakelton Shielded Ratio Box." by L. Behr and A. J. Williams, Jr.

6. "Westinghouse Radio Station at Saxonburg, Pa.," by R. L. Davis and V. E. Trouant.

7. "Radio Dissemination of the National Standard of Frequency," (Abstract), by J. H. Dellinger and E. L. Hall.

8. "Radio Test Methods and Equipment," (Abstract), by W. F. Diehl.

9. "A New Circuit for the Production of Ultra-Short-Wave Oscillations," by H. N. Kozanowski.

10. "A New Water-Cooled Power Vacuum Tube," by I. E. Mouromtseff.

11. "The Precision Frequency Measuring System of RCA Communications, Inc.," by H. O. Peterson and A. M. Braaten.

12. "Kennelly-Heaviside Layer Studies Employing a Rapid Method of Virtual-Height Determination," by J. P. Schafer and W. M. Goodall.

13. "Transmission Lines for Short-Wave Radio Systems," by E. J. Sterba and C. B. Feldman.

15. "Note on the Measurement of Resistance at High Frequency," by P. B. Taylor.

16. "Dynamic Symmetry in Radio Design," by Arthur Van Dyck.

Requests for copies of any of the above should be made to the Institute office, 33 West 39th Street, New York City.

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