VOLUME 20 AUGUST, 1932 NUMBER 8

PROCEEDINGS
of
The Institute of Radio Engineers

Walter J Malone
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Institute of Radio Engineers
Forthcoming Meetings

CLEVELAND SECTION
September 16, 1932

DETROIT SECTION
September 16, 1932

LOS ANGELES SECTION
September 20, 1932

NEW YORK MEETING
September 7, 1932

SEATTLE SECTION
September 29, 1932

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 20 August, 1932 Number 8

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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.

PROCEEDINGS. The PROCEEDINGS is the official publication of the Institute and in it are published all of the papers, discussions, and communications received from the membership which are accepted for publication by the Board of Editors. Copies are sent without additional charge to all members of the Institute. The subscription price to nonmembers is $10.00 per year, with an additional charge for postage where such is necessary.

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MAILING. Entered as second-class matter at the post office at Menasha, Wisconsin. Acceptance for mailing at special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., and authorization was granted on October 26, 1927.

Published monthly by

THE INSTITUTE OF RADIO ENGINEERS, INC.

Publication office, 450–454 Ahnaip St., Menasha, Wis.

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33 West 39th Street, New York, N. Y.

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	Toronto 8, Ont., 425 Ashdale Ave.	Corless, C. J.
	Toronto 3, Ont., 86 Morningside Ave.	Nesbit, W. D.
	Toronto, Ont., 35 Nasmith Ave.	Sachs, M.
	Toronto, Ont., 497 Whitmore Ave.	Sparkes, P.
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	Chelmsford, Essex, Marconi College	Zayoona, J. H.
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	Columbus, 2163 Neil Ave.	Ross, A. J.
	Columbus, 231 16th Ave.	Smart, P. H.
	Columbus, 90 E. 12th Ave.	Welker, J. W.

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	Troy, 29 S. Walnut St.	Thomas, S. O.
Washington	Seattle, 4524 20th Ave., N.E.	Mendenhall, G. G.
	Seattle, 2621 41st Ave., S.W.	Scott, E. D.
	Seattle, Route 8, Box 428	Taylor, R. L.
Wisconsin	Milwaukee, 1929 W. Meinecke Ave.	Plautz, A.

APPLICATIONS FOR MEMBERSHIP

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 September 6, 1932. These applicants will be considered by the Board of Directors at its meeting on September 7, 1932.

For Transfer to the Member Grade

California	Los Angeles, 1365 Edgecliff Dr.	Lubcke, H. R.
Michigan	Detroit, 2000-2nd Ave.	Buchanan, A. B.
Missouri	St. Louis, Radio Station KMOX, 401 S. 12th St.	West, W. H.
England	Harrow Weald, Middlesex, "Fayrene," 2 Whitefriars Dr.	Henderson, F. A.

For Election to the Member Grade

Connecticut	South Manchester, 176 Wadsworth St.	Reinartz, J. L.
Illinois	Chicago, 100 W. Monroe St., 20th Fl.	Larsen, C. J.
New York	Brooklyn, c/o Cable Tube Radio Corp., 84–90 N. 9th St.	Lyle, A. E.
	New York City, 195 Broadway	Shackleton, S. P.
Pennsylvania	Emporium, Box 725	Miller, H. J.
England	London, N.W. 9, Standard Telephones and Cables, The Hyde, Hendon	Larnder, H.

For Election to the Associate Grade

California	Bolinas, R.C.A. Communications, Inc.	Forster, A. G.
	Coronado, Squadron VS-15M, Fleet Air Base	Walker, C. W.
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	San Pedro, U.S.S. Salt Lake City, c/o Postmaster	McGirr, W. P.
Dist. of Columbia	Bellevue, Radio Material School	Salisbury, L. C.
Georgia	LaGrange, 124 College Ave.	White, C. J., Jr.
Kentucky	Covington, c/o Station WCKY	Topmiller, C. H.
Massachusetts	Springfield, United American Bosch Corp.	Biernacki, F. L.
	Springfield, United American Bosch Corp.	Guertin, C. E.
Michigan	Royal Oak, 203 Allenhurst Ave.	Larime, L. H.
Missouri	St. Louis, c/o General Electric Supply Corp., 200 S. 7th St.	Jenkins, H. D.
	St. Louis, 200 S. 7th St.	King, H. T.
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	Tsinan, Dept. of Physics, Cheeloo University	Wu, C.
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	London, E. 15, 178 Plaistow Rd., West Ham	Clack, W. H.
	London, E. 8, 46 Shacklewell Lane, Dalston	Pritchard, W. H.
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Latvia	Riga, Rigas 2 Radiostacija	Akmentins, A.
New Zealand	Wellington, 36A Severn St., Island Bay	Wilkinson, K. T.
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S. Rhodesia	Salisbury, Box 1089, Beam Wireless Station	Tyrer, A. R.

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For Election to the Student Grade

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	Sac City, 225 S. 9th St.	Hoyt, C.
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	Madison	Jones, T. B.

Massachusetts	Allston, 1177 Commonwealth Ave	Wagner, H. M.
	Worcester, 964 Pleasant St	Morse, R. S.
Michigan	Adrian, R.F.D. No. 2	Clement, P. F.
	Grand Rapids, 1537 Broadway Ave	Rasikas, W.
New Jersey	Harrison, Development Lab., RCA Radiotron Co.	Schafer, E. W.
New York	Albany, 31 Hampton St	Di Lello, P. J.
	Albany, 31 Bogart Ter.	Wolberg, L.
	Buffalo, 46 Welmont Pl	Pries, L. F.
	New York City, 3812 Waldo Ave	Riesenkonig, H.
	Troy, 2216-15th St	King, P. B., Jr.
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	Puyallup, Route 2, Box 455	Herr, M. D.

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With deep regret the Institute records the death of

Louis Winslow Austin

on June 27th following an illness which confined him to a hospital after a surgical operation.

Dr. Austin's valuable work in the field of radio wave propagation is recognized throughout the world and for it he was awarded the Institute Medal of Honor in 1927. He was a charter member of the Institute and became its third president in 1914.

Many of his numerous contributions to scientific literature will be found in the pages of the Proceedings.

INSTITUTE NOTES

STANDARDIZATION

Technical Committee on Electro-Visual Devices

A meeting of the Technical Committee on Electro-Visual Devices of the Institute's Standardization Committee was held on June 16 in the office of the Institute. Those present were: J. V. L. Hogan, chairman; E. K. Cohan, Alfred N. Goldsmith, G. C. Gross (representing C. B. Jolliffe), E. F. Kingsbury, R. H. Marriott, A. F. Murray, Leslie Woods (representing W. E. Holland), and B. Dudley, secretary.

The committee reviewed the terms adopted at its previous meeting and also composed several new definitions.

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 although not with certainty in some places. The accuracy of the frequency is at all times better than one cycle (one in five million).

From the 5000 kilocycles any frequency may be checked by the method of harmonics. Information on how to receive and utilize the signals is given in pamphlets obtainable on request addressed to the Bureau of Standards, Washington, D. C.

The transmissions consist mainly of continuous, unkeyed carrier frequency, giving a continuous whistle in the phones when received with an oscillatory receiving set. For the first five minutes there are transmitted the general call (CQ de WWV) and announcement of the frequency. The frequency and the call letters of the station (WWV) are given every ten minutes thereafter.

Supplementary experimental transmissions are made at other times. Some of these are made with modulated waves, at various modulation frequencies. Information regarding proposed supplementary transmissions is given by radio during the regular transmissions, and also announced in the newspapers.

The Bureau desires to receive reports on the transmissions, especially because radio transmission phenomena change with the season of the year. The data desired are approximate field intensity, fading characteristics, and the suitability of the transmissions for frequency measurements. It is suggested that in reporting on intensities, the following designations be used where field intensity measurement apparatus is not used: (1) hardly perceptible, unreadable; (2) weak, readable now and then; (3) fairly good, readable with difficulty; (4) good, readable; (5) very good, perfectly readable. A statement as to whether fading is present or not is desired, and if so, its characteristics, such as time between peaks of signal intensity. Statements as to type of receiving set and type of antenna used are also desired. The Bureau would also appreciate reports on the use of the transmissions for purposes of frequency measurement or control.

All reports and letters regarding the transmissions should be addressed to the Bureau of Standards, Washington, D. C.

Institute Meetings

Buffalo-Niagara Section

The annual meeting of the Buffalo-Niagara Section was held at the University of Buffalo on June 22, L. G. Hector, chairman, presiding.

A report of the Nominating Committee recommending the reelection of the present officers was approved. These are chairman, L. Grant Hector; vice chairman, G. C. McNaab; secretary, E. C. Waud; and treasurer, G. C. Crom, Jr.

The technical portion of the program was devoted to a symposium on class B amplification. Dr. Hector in his paper outlined the theoretical differences between class A, B, and C amplifiers and demonstrated by means of an oscillograph the different forms of the plate current of the various types of amplifiers under normal and overload conditions.

He was followed by Mr. Crom of the Colonial Radio Corporation who discussed the design features necessary to obtain maximum power from the new type 46 class B tubes and demonstrated such an amplifier delivering approximately 30 watts power to a loud speaker.

J. S. Starrett of the RCA Victor Company then discussed the advantages of the class B amplifier as compared with pentodes, demonstrating both by listening to tests of the output of amplifiers of each type and seeing the distortion as indicated on a cathode ray oscillograph.

The fourth paper, by V. C. McNaab of the Wurlitzer Manufacturing Corporation, covered the design features necessary to secure good quality with limited power in the driver tube. A receiver of this type of small size and low cost was demonstrated.

Following the demonstrations, there was a lengthy discussion on these papers which was participated in by many of the eighty-three members and guests present.

Chicago Section

J. Barton Hoag presided at the April 29 meeting of the Chicago Section held at the Western Society of Engineers Headquarters, in Chicago.

Two papers were presented; the first on "Methods of Measurement of Frequency Deviations of Radio Transmitting Stations" was by John Sherman, member of the Chicago staff of the Supervisor of Radio, Department of Commerce. The paper was illustrated with slides and covered fully the interesting work this department of the government is carrying on. Many changes and improvements in broadcasting make this measurement highly important. A thorough discussion of the design and characteristics of the radio receivers and measurement equipment in operation in the engineering department was given.

The second paper of the evening on "Ignition Noise Suppression in Automobile Radio Receivers" was presented by F. W. Schor and covered a subject of considerable interest to those who come in contact with automobile radio equipment. The speaker pointed out some very interesting aspects of the subject and the very practical point of view from which he talked on the various problems brought forth considerable discussion from the two hundred members and guests in attendance.

The May meeting of the Chicago Section was held at the Hotel Stevens, J. Barton Hoag, presiding. This meeting was held during the R.M.A. Show and attracted a substantial number of visiting engineers.

A paper on "Radio Receiving Tube Developments" was presented by E. W. Ritter of the RCA Radiotron Company, and covered a number of the new tubes which have recently been announced. The substantial discussion which succeed the paper practically turned the meeting into a symposium on the subject and among those who spoke were Messrs. Arnold, Grimes, Hoag, Jarvis, Jones, Manson, Replogle, Ritter, Schmidt, Schor, and Stromeyer.

The meeting was attended by three hundred members and guests·

Cleveland Section

The April 22 meeting of the Cleveland Section was held at the Case School of Applied Science, Chairman E. L. Gove presiding.

S. E. Leonard, Engineer in charge of WTAM, presented a paper on "The Meaning of 'Microvolts per Meter'" in which he discussed the coverage of broadcast stations in general, and WTAM in particular. He exhibited a map of Ohio and western Pennsylvania showing the strength of the field from WTAM under various conditions of power and locations as plotted during the past several years. This map showed clearly the improvement in the uniformity of coverage obtained by the station since its removal to a more remote location and its increase in power above that originally permitted.

The speaker discussed in detail the empirical standards recently set up by the Federal Radio Commission as a basis for engineering testimony in hearings concerning broadcast stations. He pointed out that they represented more or less theoretical averages and are subject to many variations in special cases. Tables were presented to indicate the field intensities necessary for good service and also to show the average maximum good service range of broadcast stations of various powers.

The second paper of the evening, presented by J. R. Martin of the Case School of Applied Science, was "The Vacuum Tube in Oscillatory Circuits."

Conditions under which a vacuum tube circuit will oscillate were discussed and it was pointed out that a state of effective zero resistance must be reached before oscillations could be set up. By extending this effective resistance in a negative direction, the tendency toward sustained oscillation is increased. Of special interest was his development of the work to permit a very small tube to control a high power oscillating circuit. He stated, however, that theory and practice were somewhat at variance and that should the power circuit stop oscillating even for an instant, the control tube would be quickly destroyed by the power which would then be out of control.

The meeting was attended by forty-nine members and guests.

The May meeting held on the 20th was also at the Case School of Applied Science with E. L. Gove, chairman, presiding.

Two papers were presented, the first on "Considerations Involved in Broadcast Antenna Design" by Professor J. F. Byrne of Ohio State University and the second, "Theory and Practice of Thyratron and Cathode Ray Tubes" by Professor J. R. Martin of Case School of Applied Science.

Professor Bryne indicated that broadcast antenna design involved two considerations, engineering and economic. The former is based upon the necessity of establishing a strong signal at the surface of the earth and methods of determining antenna effectiveness by making a number of local field strength measurements were explained. These measurements show that for a given antenna power, maximum field intensity is a function of height and resistance at the base of the antenna. He stated that the greatest effectiveness of the vertical antenna was obtained when the height was five-eighths of the wavelength which increases effectiveness by modification of the polar radiation distribution. When the height of the antenna is small, the effectiveness can be increased by making the resistance at the base large in comparison with the ground resistance. This can be accomplished by adding a flat top which should not be more than three-sixteenths of a wavelength long.

Economic considerations indicate the feasibility of using antenna systems higher than most of those in service at present particularly in the case of high power stations operating at low frequencies. The added antenna cost can be justified by the increase in surface area coverage.

At the close of the paper the speaker demonstrated a field strength measuring set constructed at the Ohio State Uni ersity.

Professor Martin outlined the constructional details and methods of operation of thyratron and cathode ray tubes. A number of typical tubes were available for examination by the audience.

In discussing cathode ray tubes, he pointed out their value in the standardization of quartz crystals, in studying frequency and phase relationships, and in television experiments. The advantages of cathode ray tubes over mechanical oscillographs were pointed out. The particular tube displayed had a high voltage sensitivity and a low photographic sensitivity. Professor Martin also described a new tube of opposite characteristics in which the photographic film was introduced directly into the tube requiring the tube to be evacuated each time it was used.

The thyratron tube is filled with mercury vapor and offers a high resistance to current flow until ionization due to collision between the mercury molecules and electrons from the filament occurs. Then it becomes highly conductive behaving like a mercury arc which cannot be extinguished or controlled by the grid potential because the grid is sheathed by a film of positive ions. This tube has many industrial and experimental uses.

The meeting was attended by fifty members and guests.

Connecticut Valley Section

L. F. Curtis, chairman, presided at the March 30 meeting of the Connecticut Valley Section held at the Hotel Charles in Springfield, Mass.

David Grimes, License Engineer of the Radio Corporation of America, presented a paper on "Analyzing and Measuring Amplitude and Frequency Modulation."

In this paper the author covered the effects of frequency modulation in amplitude modulated signal generators. At some frequencies the frequency modulation in a generator which tunes across the broadcast band can be so great as to cause the carrier and one side band to disappear entirely, with consequent loss of accuracy in readings. Vector analyses of amplitude and frequency modulation were given, and after the discussion there followed an apparatus demonstration in which a highly selective receiver was employed to separate the side bands from the carrier of a typical signal generator. The effects of frequency modulation on the amplitude of carrier and side bands at different frequencies were demonstrated by vacuum tube voltmeter readings.

The paper was discussed by a number of the forty-two members and guests in attendance, of whom nine were present at the informal dinner which preceded the meeting.

The May meeting of the Connecticut Valley Section was held at the Hotel Charles in Springfield, with H. W. Holt, vice chairman, presiding.

"Recent Developments in Receiving Tubes" was the subject of a paper by P. T. Weeks, Chief Engineer of the Raytheon Production Company. Dr. Weeks' paper traced the development of the different types of receiving tubes grouped according to filament-voltage ratings, with comparisons of corresponding types in each group. The uses of the latest additions, such as the 46, 82, 58, etc., were explained. The talk was illustrated with lantern slides, showing curves for the various types, and a large number of tubes were available for inspection.

Several of the thirty-six members and guests in attendance participated in the discussion which followed presentation of the paper.

Detroit Section

A meeting of the Detroit Section was held at the Detroit News Conference Room with H. L. Byerlay, chairman, presiding.

The paper presented at this meeting on "Transmission of Voice and Radio Programs over Telephone Circuits" was by J. G. Braybrook of the Michigan Telephone Company, and B. E. Love of the American Telephone and Telegraph Company. Mr. Braybrook presented an interesting description of methods of transmitting voice and music over telephone lines. By means of an oscillograph he showed the complexity of the waves that must be transmitted. Some phonograph records were used to demonstrate the effect of cutting off frequencies in the range of 100 to 5000 cycles in the transmission of speech and music. The permissibility of substantial amounts of distortion in the transmission of intelligible speech was pointed out and contrasted with the much greater fidelity necessary for satisfactory transmission of music. The speaker closed with an explanation of some of the means used in minimizing distortion in telephone line transmission.

The second portion of the paper was presented by Mr. Love who described the cable loop system now in use in the transmission of radio programs over telephone lines for the three national networks. An interesting discussion followed and was participated in by several of the forty-four members and guests in attendance. The audience examined the equipment used for demonstration purposes after the close of the meeting.

Los Angeles Section

Mr. E. C. Schreiber, chairman, presided at a meeting held on June 21 of the Los Angeles Section held at the Mayfair Hotel.

A paper on "Standard Frequency Equipment, Their Types and Uses" was presented by W. L. Burnett of Rynerson and Burnett.

The author spoke chiefly of frequency measurements using a frequency standard consisting of a 50-kilocycle piezo electric oscillator and suitable multivibrators to obtain frequency-steps of 1,000 kilocycles, 100 kilocycles and 10 kilocycles. A final multivibrator permits the operation of a 100-cycle synchronous clock for checking against a standard of time.

The piezo electric oscillator is contained in a temperature control oven giving a high degree of temperature stability.

By using the double heterodyne method the author stated it is possible to check the frequencies of the transmitters in the broadcast spectrum to within a fraction of a cycle.

Forty-one members and guests attended the meeting and fourteen were present at the informal dinner which preceded it.

Pittsburgh Section

The Pittsburgh Section held a meeting on February 16 at the Fort Pitt Hotel with J. G. Allen presiding.

C. Williamson of the Physics Department of Carnegie Institute of Technology presented a paper on "Wave Forms of a Tube Oscillator and of Musical Notes" which covered the practical design of audio oscillators. The paper dealt with an explanation of what constituted a sine wave and why such was necessary in the design of oscillators for laboratory work. As a practical demonstration illustrating the points of the paper, an oscillator designed according to them was operated with an oscillograph to show how nearly the ideal had been reached.

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

The March meeting of the Pittsburgh Section was held on the 22d at the Fort Pitt Hotel and the presiding officer was J. G. Allen, chairman.

"The Use of Carrier Telephones on the Bell System" was the subject covered by R. T. Griffith, transmission engineer of the Bell Telephone Company of Pennsylvania. The paper proved to be an interesting and comprehensive discussion of the progress of carrier communication. At its conclusion, some general motion pictures of the radio activities of the Bell system were shown.

The attendance totaled thirty-six.

J. G. Allen presided at the May 17 meeting of the Pittsburgh Section held at the Fort Pitt Hotel. This was the annual dinner meeting at which the election of officers for the following year was held. In this election, R. T. Griffith of the Bell Telephone Company of Pennsylvania was elected chairman, J. G. McKinley of the West Penn Power Company became vice chairman, and C. K. Krause of the Duquesne Light Company was elected secretary-treasurer.

No formal paper was presented at this meeting and after a general discussion on current radio developments which was participated in by practically all of the twenty-six members and guests in attendance, the meeting was adjourned.

San Francisco Section

The annual meeting of the San Francisco Section was held on June 15 at the Bellevue Hotel with R. M. Heintz, chairman, presiding.

Prior to the presentation of the paper, the election of officers for the following year was held. The new chairman is C. V. Litton; the vice chairman, A. R. Rice of the Navy Department; and the secretary-treasurer, K. G. Clark of the office of the Supervisor of Radio.

The paper of the evening on "Present Status of Television" was presented by A. H. Brolly, chief engineer of Television Laboratories. At the conclusion of the talk, the meeting adjourned to the Television Laboratories where a demonstration of the Farnsworth system was put on for the benefit of those present.

The meeting was attended by one hundred and five members and guests of whom twenty-two were present at the informal dinner which preceded it.

Seattle Section

The Seattle Section held a meeting on May 26 at Guggenheim Hall with chairman L. C. Austin presiding.

A paper on "Acoustics" was presented by J. A. Johnson, an engineer with Electrical Research Products, Inc.

Prior to the delivery of the paper, a sound picture entitled "Principles of Acoustics" by H. Fletcher was shown.

The meeting was attended by fifty-five members and guests, a number of whom participated in the discussion which followed the presentation of the paper.

The June meeting of the Seattle Section was held on the ninth at Guggenheim Hall, chairman L. C. Austin presiding.

Two papers were presented at this meeting. The first on "Variable Frequency Modulation" was presented by Earl Scott and the second was on "Special Application of Pentodes as Amplifiers." This was presented by Ronald Boyles.

A number of the seventy-eight members and guests in attendance participated in the discussion of these papers.

Washington Section

A meeting of the Washington Section was held on June 9, 1932, in the Kennedy-Warren Building, H. G. Dorsey vice chairman, presiding.

A paper on "Design, Production and Inspection of Commercial Broadcast Superheterodyne Radio Receivers" was presented by C. E. Brigham, Chief Engineer, Brandes Laboratories, Inc.

The speaker ably discussed the manner in which the design, engineering development, production, and inspection of a modern commercial broadcast superheterodyne radio receiver was undertaken by the manufacturer.

A general discussion followed and was participated in by a number of the forty members and guests in attendance, of whom twenty-two were present at the informal dinner which preceded the meeting.

Personal Mention

E. H. I. Lee has been transferred from New York to Detroit as U. S. Supervisor of Radio.

Formerly with His Master's Voice Company in Montreal, W. G. Robinson has become a radio engineer for Erie Resistor of Canada at Toronto.

Major L. B. Bender of the Signal Corps, USA, has been transferred from Washington, D.C., to Fort Hayes, Columbus, Ohio.

Previously with the Technidyne Corporation Jacob Yolles has joined the License Division of the Radio Corporation of America in New York.

M. C. Batsel has been transferred from the RCA Photophone Company in New York to the RCA Victor plant in Camden, N.J.

J. E. Brown, Assistant U. S. Supervisor of Radio, has been transferred from Detroit, Michigan, to the New York City office.

Lieutenant W. P. Cogswell, USN, has been transferred from the Naval Air Station at Anacostia to the *U.S.S. Lexington* for sea duty.

Formerly vice president and European technical director of the International Standard Electric Corporation at London, G. Deakin has become technical director in charge of manufacturing and laboratory companies of the International Telephone and Telegraph Corporation at New York City.

Lieutenant J. B. Dow, USN, has been transferred to sea duty on the *U.S.S. Utah* from the Bureau of Engineering in Washington, D.C.

Captain N. H. Edes has been transferred from Canada to "A" Corps Signals at Aldershot, England.

Lieutenant Commander C. R. Holden, USN, has been transferred from the *U.S.S. Arizona* to the *U.S.S. Tennessee.*

Commander C. H. Maddox, USN, has been transferred from Washington, D.C. to the *U.S.S. Salt Lake City.*

Commander B. V. McCandlish, formerly Commander of the U. S. Naval War College at Newport, R.I. has been transferred to Naval Communications, Washington, D.C.

Formerly with the Columbia Pictures Corporation, D. F. Miller has become research engineer for Universal Pictures Corporation, Universal City, Calif.

Lieutenant G. B. Myers, USN, has been transferred from the U.S. Naval War College at Newport, R.I. to District Communications Office at Great Lakes, Ill.

Previously with RCA Photophone, M. O. Smith has joined the engineering staff of the National Broadcasting Company at San Francisco.

H. J. Tyzzer has become chief engineer of Pilot Radio and Tube Company of Lawrence, Mass.

Previously with Dubilier Condenser Corporation, G. J. Uzmann has joined the engineering staff of P. R. Mallory at New York City.

G. E. Webster has joined the engineering staff of the National Broadcasting Company of Chicago.

Formerly with Wired Radio, A. C. Wooldridge has become an engineer for American Radio News.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

TECHNICAL PAPERS

RECENT TRENDS IN RECEIVING TUBE DESIGN*

By

J. C. Warner, E. W. Ritter, and D. F. Schmit

(Research and Development Laboratory, RCA Radiotron Company, Harrison, N. J.)

Summary—*This paper gives a brief summary of the important steps in receiving tube design over the past ten years. The significance of new forms of grids and in particular the suppressor grid are discussed. Characteristics of new radio-frequency tubes containing suppressor grids are shown. Improvements in cathode and grid designs are illustrated by the characteristics of a new triode as well as two triple-grid tubes. A new tube for class B audio amplification is described together with a mercury vapor rectifier for supplying power to the class B amplifier.*

The functions of the vacuum tubes in a modern broadcast receiver are fundamentally much the same as in the tubes used in the receivers of ten and even fifteen years ago. Even at that time there were tubes which functioned as radio and audio amplifiers, as oscillators, as detectors, and as power output tubes. The same functions appear today with little variation. In fact the all-important reason for the use of tubes has not changed—the controlling of a power output which is greater than the power expended in the controlling action.

The advances in vacuum tube engineering and the multiplicity of modern types therefore are not due to the introduction of new functions but rather to the improvement of the functioning to meet the varied requirements of the numerous modern radio applications and to the development of new principles of tube design which have improved the functioning of tubes in these various applications. A brief summary of the outstanding steps of progress over the past ten years will serve to illustrate.

The first broadcast receivers depended upon a 6-volt storage battery for filament supply, and two types of tubes, or in many cases even one, the UV-201, served to fulfill all of the necessary receiving tube applications. In the light of present-day practice such tubes and their applications seem extremely crude. A transconductance of 300 micromhos was considered reasonably good and the output was so badly distorted that additional distortion due to the combination of uncer-

* Decimal classification: R330. Original manuscript received by the Institute, April 29, 1932. Presented before Twentieth Anniversary Convention, Pittsburgh, Pa., April 8, 1932.

tain transformer design with the absence of grid bias was hardly noticed.

The first marked improvement in tubes for broadcast receivers resulted from the change from the pure tungsten filament to a coated filament or a thoriated tungsten filament. Tubes which could operate on dry batteries were made possible by the use of these filaments (WD-11 and UV-199) and in the case of the UV-201 the change to the thoriated filament resulted in a 75 per cent decrease in filament power together with nearly a 100 per cent increase in transconductance.

Next came a realization that new tube designs would be needed to give the increasing undistorted output demanded for satisfactory operation of loud speakers. The first of these tubes was the UX-120, soon followed by the UX-171. Both of these tubes had amplification factors of about 3 and the unusually (for that day) high bias of 22 1/2 to 40 volts was required to give ample grid swing without drawing grid current.

The next important advance was in the design of tubes for alternating-current operation. The low voltage filament type of tube, UX-226, came into use for amplifier service, and the indirectly heated cathode type of tube, UY-227, for detector use. However, the advantages of the latter type soon led to its superseding the filament type.

Coincident with the development of the alternating-current type of tube was the application of the screen-grid principle, mainly for radio-frequency amplification resulting first in the 222 for battery sets followed by the 224 for alternating current. This brought about the elimination of troublesome regeneration and resulted in much higher gain per stage. The usual practice in receiver design then was to use screen-grid tubes in the radio-frequency stages, and triodes in the detector and first audio stages with a 171-A in the output stage, and these three types of tubes satisfied practically all of the needs of the set designer.

The perpetual demand for more power output soon left the 171-A far behind and the 245, with an output of 1.6 watts, came into use, first singly and then in push-pull circuits. This satisfied for a time but was later overshadowed by the advent of the output pentode, 247, with its 2.5 watts output. Here the increase in output was gained not by an increase in the size of the tube or an increase in its operating voltages but by a change in the fundamental design.

At this time the home receiver was receiving the major part of the attention of both set and tube engineers. The possibilities of expanding the battery receiver and automobile receiver fields focused attention on new tubes for these applications and led to two complete new lines of tubes, the 2-volt dry battery tubes and the 6-volt heater-cathode automobile tubes.

One of the disadvantages of the ordinary screen-grid tube was the limited grid swing which could be accommodated. This resulted in excessive distortion and cross-modulation when high signal voltages were applied to radio-frequency stages, and made a satisfactory volume control extremely difficult. In an outstanding contribution to tube engineering, Ballantine and Snow[1] eliminated these troubles by the development of the "variable-mu" type of tube. In its screen-grid form (235 and 551) this tube has come into almost universal use in alternating-current home receivers. More recently the variable-mu automobile and 2-volt (239 and 234) tubes have been developed.

Altogether the developments summarized have resulted in a total of about 25 distinct types of tubes all of which are to some extent in use today although only about 15 types are in active use in modern set designs. In addition there are another dozen or more of minor variations which have had a limited use.

This great multiplicity of types is often thought of as an undesirable state of affairs which should be corrected by a gradual reduction in types much as has been done in other fields, notably with incandescent lamps. It must be remembered however that radio tube standardization faces a particularly difficult problem in the diversity of power sources in the three major fields of application combined with a certain degree of difference in the standards of performance as well. Furthermore, radio receiver engineering is still in a state of change and so long as there continue to be new types of circuits and new circuit requirements, together with new kinds of tube characteristics, it must be expected that new tubes will continue to emerge from laboratory and factory to contribute their part in the advance of the radio art as a whole.

The Suppressor Grid

Many advances in tubes have been closely related to new grids of one sort or another or to new ways of using grids. Examples are the screen-grid and the nonuniform grid employed in the variable-mu tube, and recently the suppressor grid is playing a major part in tube design. The purpose of the suppressor grid in the output pentode is well known. More recently the suppressor grid has been used in radio-frequency tubes.[2] The advantages of this use are several. The screen

[1] Ballantine and Snow, "Reduction of distortion and cross-talk in radio receivers by means of variable-mu tetrodes," Proc. I.R.E., vol. 18, p. 2102; December, (1930).

[2] E. W. Ritter, "The r-f pentode," *Electronics*, January, p. 10, (1932).

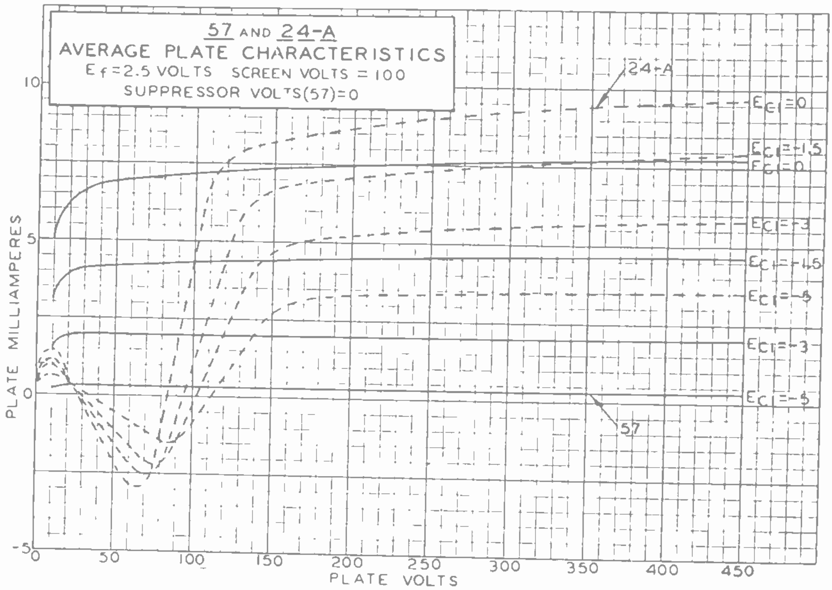
57 AND 24-A
AVERAGE PLATE CHARACTERISTICS
Ef=2.5 VOLTS SCREEN VOLTS = 100
SUPPRESSOR VOLTS(57)=0
24-A
57
PLATE MILLIAMPERES
PLATE VOLTS

Fig. 1

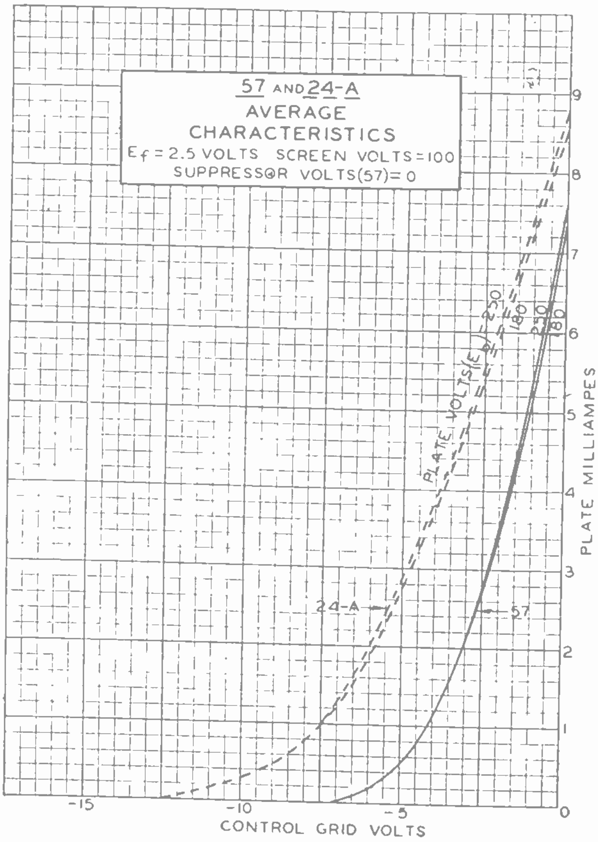
57 AND 24-A
AVERAGE
CHARACTERISTICS
Ef = 2.5 VOLTS SCREEN VOLTS=100
SUPPRESSOR VOLTS(57)= 0
PLATE VOLTS
24-A
57
PLATE MILLIAMPES
CONTROL GRID VOLTS

Fig. 2

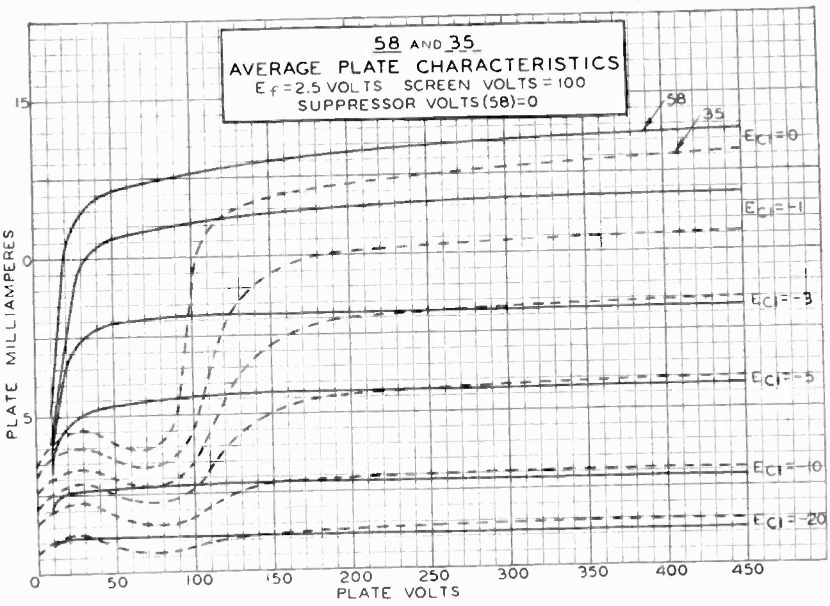
58 AND 35
AVERAGE PLATE CHARACTERISTICS
E_f=2.5 VOLTS SCREEN VOLTS=100
SUPPRESSOR VOLTS(58)=0
58
35
E_{C1}=0
E_{C1}=-1
E_{C1}=-3
E_{C1}=-5
E_{C1}=-10
E_{C1}=-20
PLATE MILLIAMPERES
PLATE VOLTS

Fig. 3

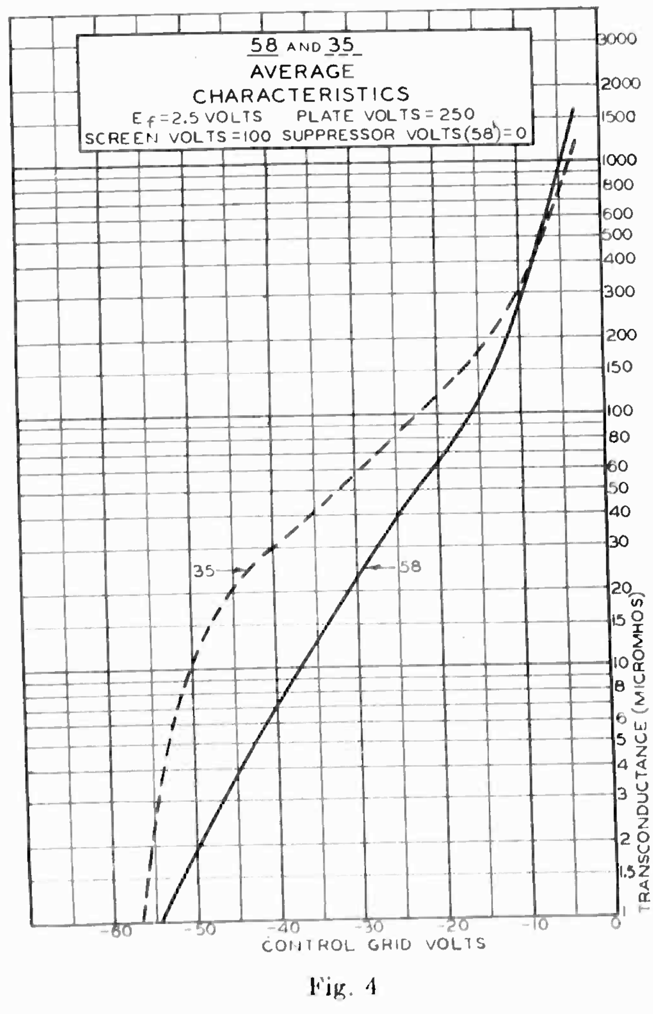
58 AND 35
AVERAGE CHARACTERISTICS
E_f=2.5 VOLTS PLATE VOLTS=250
SCREEN VOLTS=100 SUPPRESSOR VOLTS(58)=0
35
58
TRANSCONDUCTANCE (MICROMHOS)
CONTROL GRID VOLTS

Fig. 4

voltage may be set at the best value without reference to the plate voltage. The plate voltage swing is independent of the relation between plate and screen voltages and so may be made much larger than in the ordinary screen-grid tetrode. Secondary emission currents from screen to plate or plate to screen are substantially eliminated, thus straightening out the characteristics and also eliminating that part of the total tube hiss which in the ordinary screen-grid tube originates in

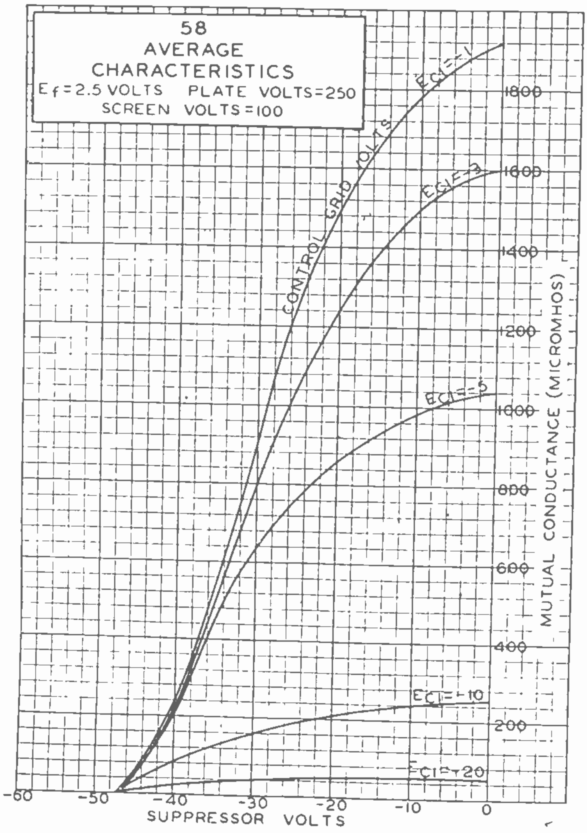

Fig. 5

the secondary emission current from screen to plate. Last, the plate resistance is greatly increased, which is particularly important in a variable-mu type of tube where the nonuniform control grid tends to lower the plate resistance.

The advantages of the suppressor grid, together with other design improvements, have been utilized in two new tube designs, one having a sharp cut-off like the 224 and the other a variable-mu like the 235 or 551. Figs. 1 and 2 show a comparison of the usual characteristic curves of the 224 and the new triple-grid tube, RCA-57, and Figs. 3 and 4 give comparisons of static characteristics and transconductance of the 235 and the RCA-58. The effect of the suppressor grid on the shape of the curves is self-evident. The higher transconductance of the newer types is due to improvements in the design of the cathode and the control grid.

Control of Characteristics by Means of the Suppressor Grid

In the output pentode and also in the first radio-frequency pentode with suppressor (239) the suppressor is connected internally to the

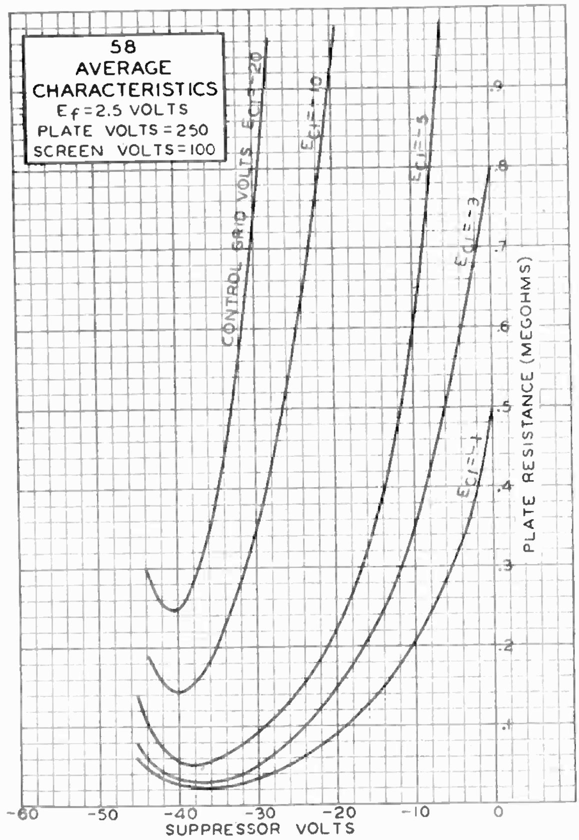

Fig. 6

cathode. In the 57 and 58 the lead to the suppressor is brought out to a sixth pin on the base. It therefore becomes possible to utilize this grid as an additional control element. In the older use of the pentode with the suppressor at the same potential as the cathode there is negligible space charge between screen and suppressor. However, if the suppressor voltage is made sufficiently negative, space charge will build up between these two grids and the characteristics of the tube approach those of a triode with the suppressor as a control grid. The regular control grid then loses much of its transconductance with respect to the plate current; i.e., the voltage gain falls off, and at the same time the plate resistance falls to a low value since with space charge between screen and suppressor the plate can draw electrons through the suppressor alone more easily than through all three grids. Figs. 5 and 6 show this type of control on the transconductance and plate resistance of the 58, and the possibilities of utilizing this feature in the control of gain and fidelity are considerable. Other circuit possibilities making use of the sixth element are evident but have not yet been fully developed.

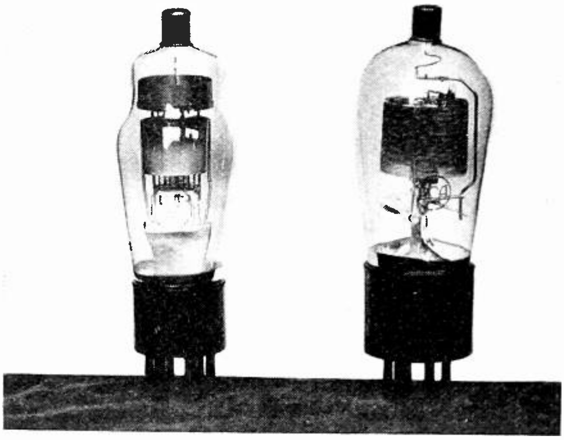

Fig. 7

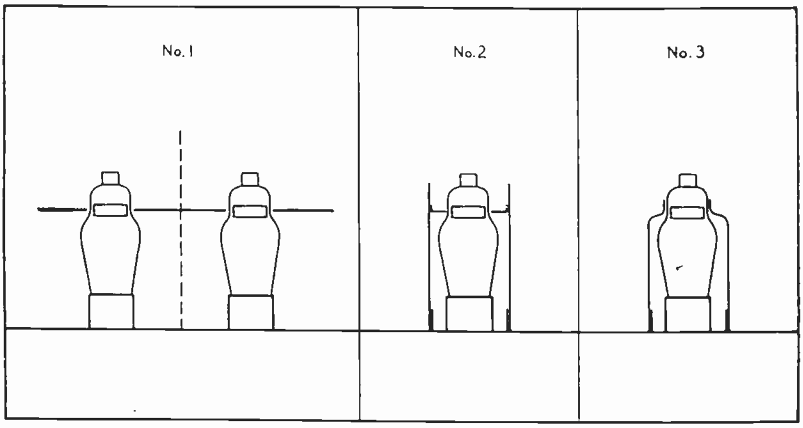

Fig. 8

Shielding Designs

The internal shielding design utilized in the 57 and 58 is considerably different from the design employed in the 224 and the total plate-to-ground capacitance has been noticeably reduced. Also the plate-to-screen capacitance has been reduced so as to improve the operation on short waves. Fig. 7 shows the 57 mechanical design in comparison with the 224. The external shielding must of course conform to the requirements of the internal design and the comparative capacitances are shown by the following tabulation. Fig. 8 shows the three shielding methods used in obtaining the measurements on the 57.

TABLE I

	224	57		
		No. 1 shield	No. 2 shield	No. 3 shield
Grid-plate capacitance	0.008	0.008	0.008	0.0045
Input capacitance	5.5	4.6	4.6	4.6
Output capacitance	11.5	7.5	8.8	8.8

Cathode Efficiency

For a number of years the cathodes of alternating-current tubes have operated at 2.5 volts and 1.75 amperes. Improvements in design have made possible a reduction of 40 per cent in this current without any harmful effect on tube characteristics. The two radio-frequency pentodes already mentioned contain 2.5-volt, 1-ampere cathodes, and an additional example is the RCA-56 triode.

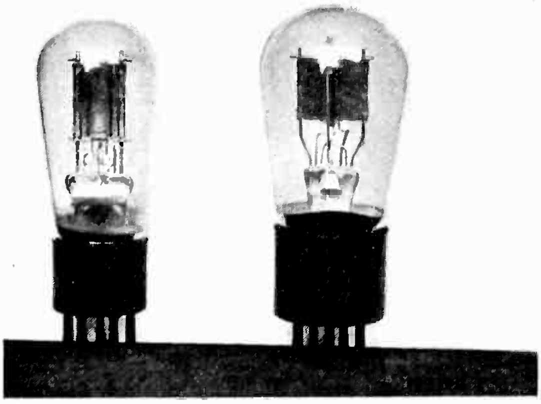

Fig. 9

This tube is somewhat similar to the 227 but due to the new cathode and grid design the transconductance has been considerably increased. The amplification factor is 14 compared with 9 for the 227. Plate resistances are about the same for both tubes.

The smaller structure of the 56 is shown by comparison with the 227 in Fig. 9 and the characteristics are compared in Fig. 10.

Improvements in Output Tubes

Throughout the entire development of broadcast reception there has been an almost continuous demand for more and more power output. The earlier output tubes were of course simple triodes and these were made in increasing size until with the 250 design it became apparent that the point had been reached where the high cost of the tube and its power supply seriously limited the use of the tube. The output pentode resulted in an increased efficiency with lower grid swing but brought in other difficulties in the way of high production of odd harmonics and the requirement of a high load impedance.

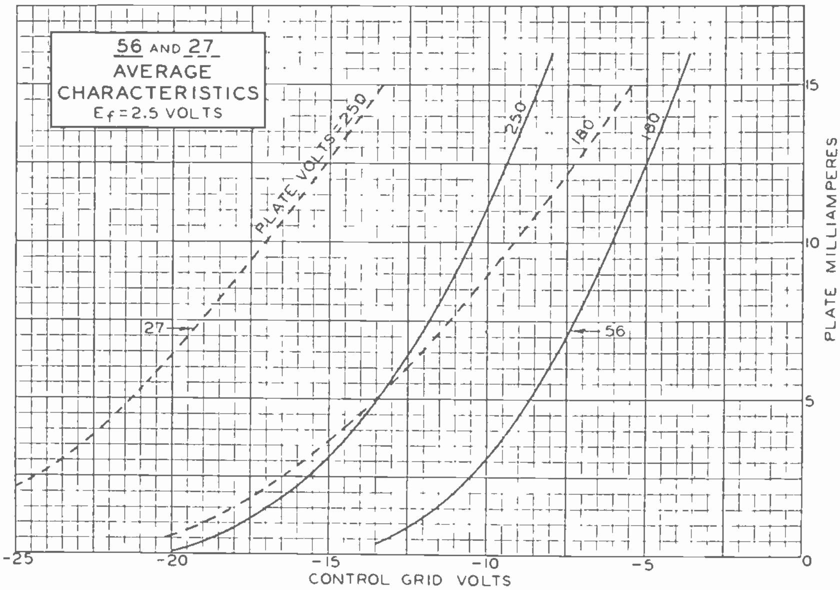

Fig. 10

Recently the so-called class B or "positive grid swing" system[3] of amplification has been receiving more and more attention[4] and is coming into use in broadcast receivers.

This system is analogous to the use of class B radio-frequency stages in broadcast transmitters. For example, a 5-kw transmitter may be used as the exciting stage for a 50-kw, class B output stage. This analogy brings out one important point—the control circuit of the class B stage requires an appreciable amount of driving power. This is an uncommon requirement in receiving circuits and necessitates an entirely new transformer design.

[3] I.R.E. Standards, No. 7202, p. 71, (1931).

[4] L. E. Barton, "High audio power from relatively small tubes," Proc. I.R.E., vol. 18, p. 1131; July, (1931).

A class B amplifier tube operates with its control-grid bias adjusted to give approximate cut-off of plate current. Plate current flows essentially only during the positive half-cycle of the exciting voltage. In a radio-frequency stage the harmonics are eliminated by the usual tank circuit but in an audio amplifier it is necessary to use two

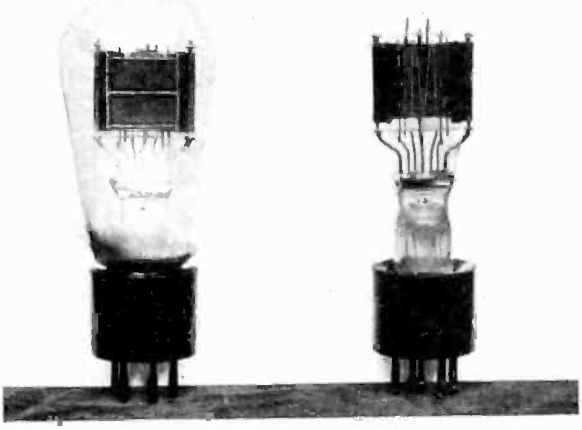

Fig. 11

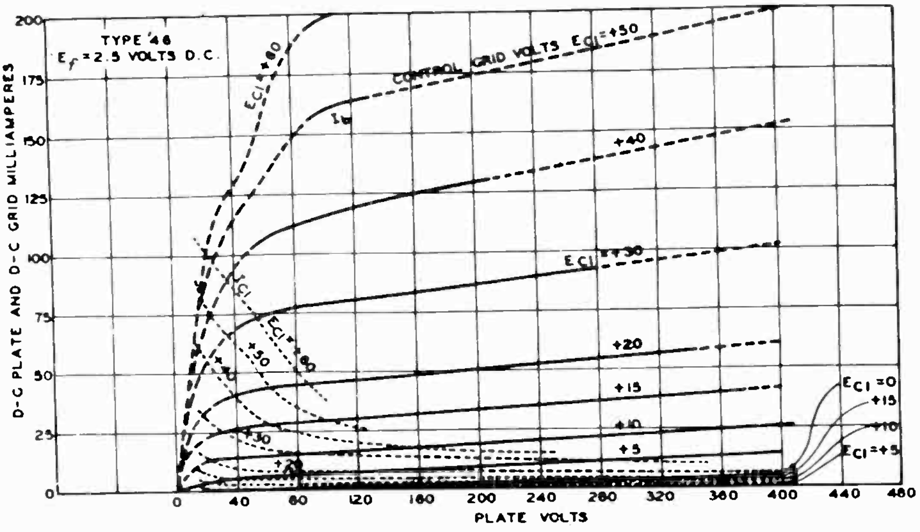

Fig. 12—Average plate characteristics. Class B operation.

tubes, each one supplying an approximate half sine wave to the output circuit.

One of the essentials of a good class B amplifier tube is that the grid-voltage, plate-current characteristic should be as steep as possible and should have a sharp cut-off. It is further desirable to have a uniform load on the driving circuit and, naturally, as light a load as possible. All of these requirements lead to a design of tube which will operate at zero bias and of course have a high amplification factor. It has been suggested by L. E. Barton that the high-mu and sharp cut-off can best be obtained by the use of two grids, one inside the other, and connected together to form the control electrode. In a flat structure this gives an improved cut-off, and for a given amplification factor requires less grid current than a closely-wound single grid.

In reducing the grid current to a low value the possibilities of utilizing secondary emission from the grid naturally come to mind and

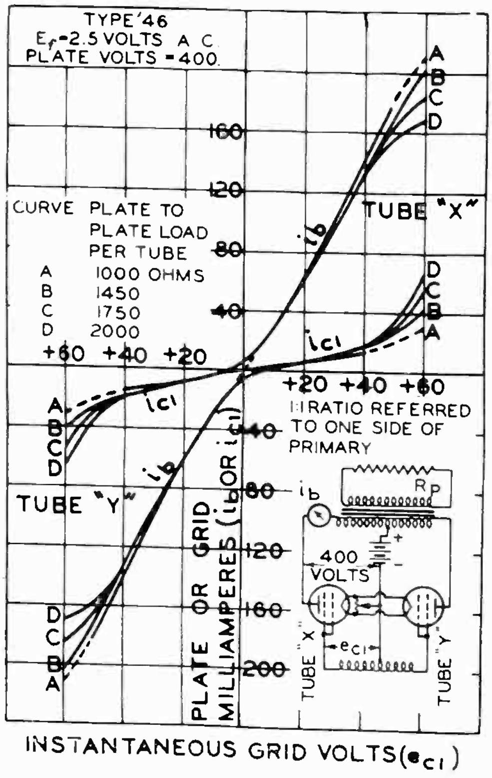

Fig. 13—Dynamic transfer characteristics. Class B operation.

it is possible to gain considerable reduction in grid power by this means. However, the rapid change in secondary emission during the life of the tube and the difficulty of producing uniform secondary emission make this method uncertain and unreliable at the present time.

Fig. 11 shows a tube which has been designed to give an output of over 20 watts per pair when driven to full output. Static plate characteristics are shown in Fig. 12, and dynamic transfer characteristics in Fig. 13.

Perhaps the greatest advantage of the class B system is due to its high plate efficiency. This is important not so much in the saving of power consumption in the receiver but in the design of the tube itself. The following table gives comparative figures on a typical class A triode, a pentode, and the class B tubes.

TABLE II

	245	247	46 (pair)
E_b volts	250	250	400
I_b milliamperes	34	32	116
Input watts	8.5	8.0	46.0
Output watts	1.6	2.5	20.0
Per cent plate efficiency (full output)	18.8	28.8	43.5
(1/2 output)	9.4	15.6	30.5
(1/10 output)	1.9	3.1	13.1

It may be said that a comparison of the 47 at 250 volts with the 46 at 400 volts is unfair to the former, but it must be remembered that they are similar in plate area and bulb size. In class B operation as the signal voltage increases the plate dissipation also increases while in class A operation the plate dissipation decreases as the signal increases. Therefore, class A amplifier tubes must be designed to dissipate the

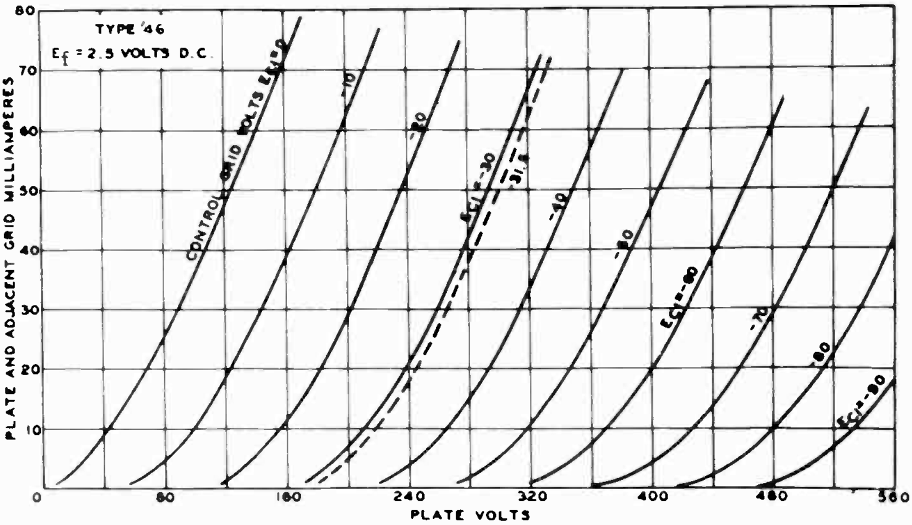

Fig. 14—Average plate characteristics. Class A operation.

maximum power continuously because of comparatively long periods of zero or low signal input. Since maximum output is required only when the modulation nears 100 per cent, the average plate dissipation of class B tubes is very low when compared to that required by class A tubes designed to deliver the same peak output. An example may be used to illustrate this. Assuming a pentode design, which will deliver a peak output of 20 watts, the plate structure and bulb must be designed to be capable of dissipating about 60 watts continuously or 30 watts per tube when push-pull is employed. This necessarily requires tubes to operate at high plate voltages in order to keep within reasonable limits in filament design. High plate voltages require an expensive rectifier system as well as an expensive tube. When class B tubes are employed relatively lower plate voltages may be used, reducing the cost of the tube and power pack. The average power delivered by an amplifier when used on broadcast reception is very low compared to the peak power when the modulation reaches 100 per cent. Measurements on average broadcast reception indicate that the average audio input voltage may be only 20 per cent of that required for peak output. When using class B tubes capable of delivering 20 watts peak power, the

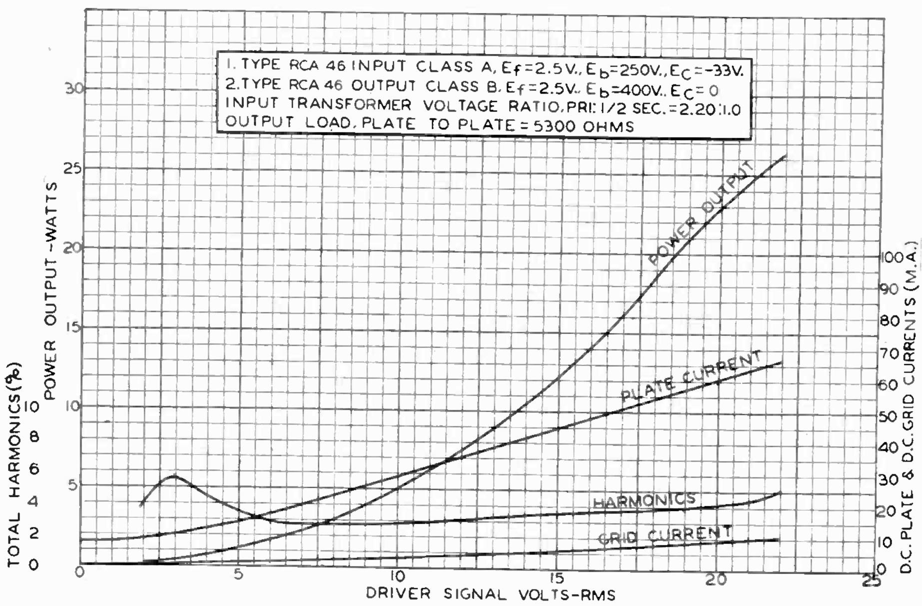

Fig. 15

average power output will then be 0.8 watt. The average plate dissipation will be 5.2 watts per tube as compared to 30 watts per tube for a similar class A amplifier.

As mentioned before the 46 tubes have two grids which have been brought out to separate grid terminals on the base. This was done to allow the application of this tube as a driver for the output stage. By connecting the second grid to the plate, the tube can be operated as a class A amplifier at a plate voltage of 250 volts with a control bias of 33 volts. The characteristics of the tube used in this way are illustrated in Fig. 14.

By selecting the proper input transformer various tubes may be used as drivers. For moderate outputs of 5 to 6 watts a single 56 may be employed: for outputs of 12 watts two 56 tubes may be used in push-pull with 300 volts on the output stage. By increasing the plate voltage on the output tubes, the output can be increased to 18 watts. To obtain 20 watts or more, a 46 tube may be used to drive the two 46 tubes operated at 400 volts as indicated in Fig. 15. The circuit conditions are shown on the illustration.

Power Supply for Class B Stage

The fact that the direct-current load on the rectifier is not constant but varies with the signal on the class B stage introduces new problems in rectifier design. Too high regulation tends to produce bad regeneration effects in the receiver, and also seriously limits the output of the output stage. This leads naturally to a consideration of the possibilities

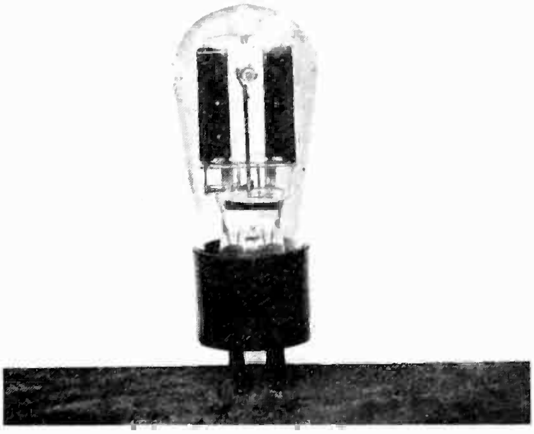

Fig. 16

of using a mercury vapor rectifier instead of the high vacuum tubes such as the 80. The characteristics of mercury vapor rectifiers are well known and they have come into almost universal use in modern transmitters.[5] Their use in receivers has been held back by their well-known tendency to produce highly damped high-frequency oscillatons which feed back into the radio-frequency stages of the receiver and produce serious noise disturbances. However, the use of simple high-frequency filters combined with modern methods of shielding have made the use of the tube perfectly feasible, and the mercury tube has become an essential part of the alternating-current class B system.

Fig. 16 shows a mercury vapor tube designed for receiver service, and Fig. 17 a static curve of the 80 in comparison with the mercury vapor tube.

[5] Steiner & Maser, "Hot-cathode mercury-vapor rectifier tubes," Proc. I.R.E., vol. 18, p. 67; January, (1930).

In addition to the advantages of low regulation the inherently high efficiency of the mercury vapor tube is of some advantage and permits the use of a smaller bulb than is necessary for a high vacuum tube of the same current rating. In fact the high efficiency of the mercury tube and class B amplifier, together with the reduced power consumption and small size of the other tubes described, contribute considerably to the

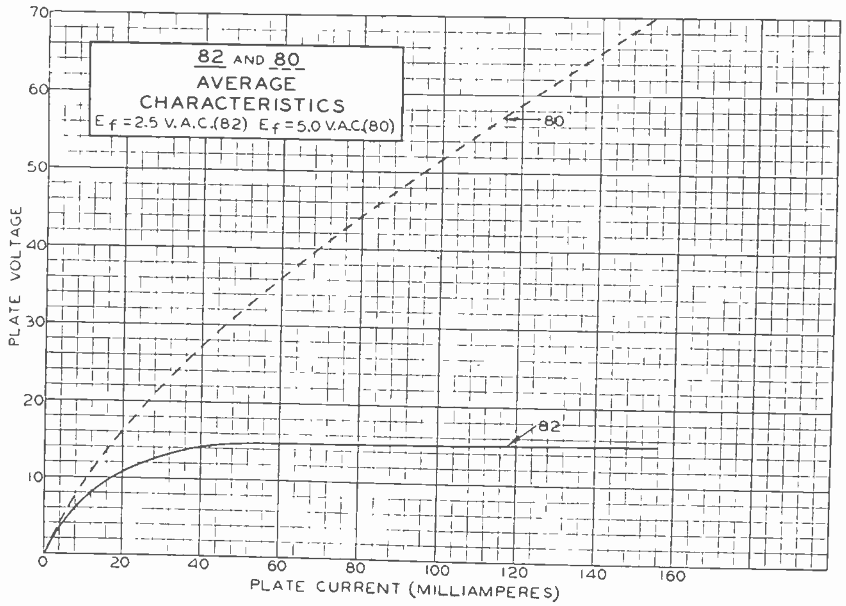

Fig. 17

aim of the modern receiver designer in producing a design which will occupy small space, require moderate power consumption, and still have the best possible performance characteristics.

Conclusion

In conclusion the writers wish to acknowledge the valuable assisance of Messrs. T. M. Shrader and J. M. Stinchfield in the preparation of this paper. Mr. Shrader is responsible for many of the design features of the tubes described, and Mr. Stinchfield has supplied much of the information on the performance of the tubes under operating conditions.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

ANALYSIS AND REDUCTION OF OUTPUT DISTURBANCES RESULTING FROM THE ALTERNATING-CURRENT OPERATION OF THE HEATERS OF INDIRECTLY HEATED CATHODE TRIODES*

By

J. O. McNally

(Bell Telephone Laboratories, Inc., New York City)

Summary—*This paper discusses the disturbance currents in the output circuits of indirectly heated cathode triodes, introduced by the use of alternating current in the heaters. It indicates that the disturbance currents are introduced into the output circuit by (1) the electric field of the heater, (2) the magnetic field of the heater current, and (3) the resistance between heater and grid and between heater and plate, and the capacitance between heater and grid and heater and plate.*

The outputs due to the electric field between cathode and plate are produced by the "grid" action of the heater and heater leads. The frequency of the output is chiefly that of the heater supply. The outputs are shown to be effectively reduced by electrostatically shielding the heater.

Disturbance currents of the frequency of the heater supply, and of double this frequency are shown to be produced by the magnetic field. The double-frequency component is shown experimentally to be proportional to the square of the heater current. The following means of reducing the magnetic field are discussed: (1) the adoption of a heater geometry which produces a smaller field in the space between the cathode and the plate, (2) the use of a magnetic shield around the heater system, and (3) the use of a lower current, higher voltage heater.

The ways in which disturbance currents are introduced by leakage resistances and capacitances between heater and grid and heater and plate are indicated, and experimental verification is given for the case of resistance between the grid and heater.

Use has been made of this disturbance current analysis in the development of an extremely low disturbance output tube, which is described.

The advantages to be obtained by operating the cathodes of the vacuum tubes in radio receiving equipment, public address systems, and talking motion picture reproduction systems directly from alternating-current lighting circuits have long been recognized. The use of alternating current for heating the cathodes produces objectionable disturbance currents of the frequencies of the power supply and its harmonics in the output of the amplifying system. These currents enter the system by induction between circuit elements and directly through the vacuum tubes. It has been found practicable to reduce the disturbance currents produced through cir-

* Decimal classification: R161.5 Original manuscript received by the Institute, April 29, 1932. Presented before Twentieth Anniversary Convention, Pittsburgh, Pa., April 8, 1932.

cuit coupling to any desired level by suitable arrangement of the circuit elements and adequate shielding. The disturbance currents originating in the tube have not been reduced to as low levels as required in some audio-frequency amplifying systems.

Prior to the time when the indirectly heated cathode type of tube became available, operation of the cathode on alternating current was necessarily restricted to the filamentary type of tube. By proper selection of filament voltage and current and by the adoption of certain operating conditions for these tubes it was possible to obtain balancing actions which could be made to reduce the disturbance currents in the outputs by very appreciable amounts.[1] The conditions necessary for the best degree of balance vary from tube to tube and are fairly critical with circuit variations, with the result that the optimum conditions of operation are infrequently obtained in the usual circuit. Because of the amount of disturbance introduced, even the best practical design in the filamentary types of tubes does not permit the satisfactory use of alternating current where amplification from relatively low levels is required.

The introduction of the indirectly heated cathode into the vacuum triode made immediately possible a further extension of the direct use of alternating current as the source of the cathode energy. A heater and cathode unit consisting of a hairpin of tungsten wire mounted in a cylindrical insulator of magnesia or its equivalent, with a tightly fitting nickel sleeve surrounding the insulator upon which is deposited the active cathode material, has been generally standardized in triodes for broadcast radio receiver use. The heater element of such a tube can be operated on alternating current in radio-frequency stages without the introduction of fundamental and higher frequency disturbance currents from the heater supply. However, its use in audio-frequency circuits of flat frequency characteristic down to that of the heater supply is, in general, limited to circuits with gain of the order of 50 db following the first tube operated on alternating current. If such a tube employing alternating current for the heater supply is used in amplifiers with flat frequency characteristics of appreciably greater gain, disturbances from the heater supply are too great to be tolerated.

The amplifying units of sound recording and reproducing systems have over-all gains of the order of 100 db. With a system having this amount of gain, it is possible to use alternating-current supply in heating the cathodes of all tubes only by a sacrifice in the frequency characteristic of the amplifier below approximately 150 cycles or by the toleration in the output of a high level of extraneous noise arising

[1] See bibliography, item 8.

from the heater disturbance currents in the tubes of the earlier stages.

The advantages of using alternating current for cathode supply in high quality sound reproduction amplifiers as well as in public address systems, radio broadcast speech input equipment, and other high gain audio-frequency amplifiers, made desirable the study of heater disturbance current levels and an investigation of means of sufficiently reducing them to permit the use of alternating-current heating on all tubes in such systems. In order to make alternating-current cathode heating generally applicable, its disturbances in the plate circuit of the first tube should be of the same order of magnitude as that of resistance and thermionic emission noises. Alternating-current heating could then be applied to any amplifier whose gain was not limited by these fundamental noise sources.

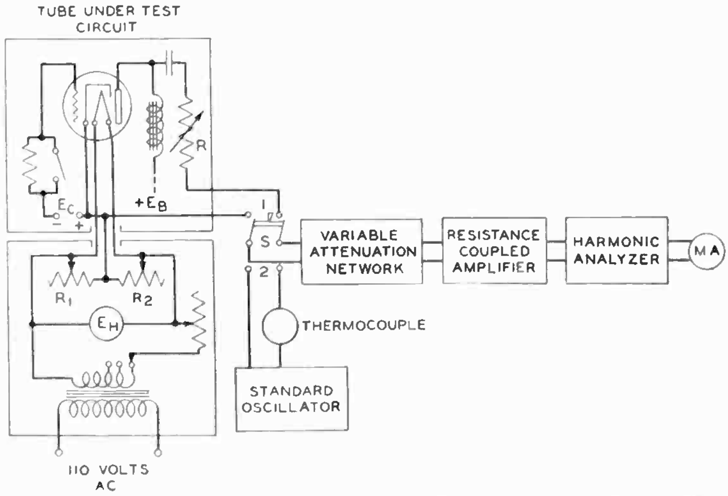

Fig. 1—Schematic diagram of disturbance current measuring circuit.

Measuring Equipment

As a first step in these studies, aparatus was assembled in which the outputs of the frequency of the heater current and its harmonics could be measured separately. The tube, whose disturbance current is to be measured, is placed in a circuit indicated in Fig. 1 as the "tube under test" circuit. The output from the tube is fed through switch *S* into the variable attenuator and hence to the resistance coupled amplifier. The output of this amplifier is applied to the input of the harmonic analyzer which permits the separation and measurement of the fundamental and harmonics of the frequency supplying the cathode power. The analyzer was designed for frequencies between 50 and 350 cycles per second. The frequency of the alternating current used for the cathode supply was approximately 60 cycles per second. The sensitivity of the apparatus was such that a readily readable output was obtained from the analyzer with a current 120 db below 1.0 milliampere (10^{-9} ampere) flowing through a resistance of 600 ohms in the input of the resistance coupled amplifier. In order to evaluate the analyzer readings, a standard oscillator is provided, the known output of which may be fed through the attenuator and amplifier and into the analyzer.

It will be noted that the tube under test obtains its heater power from a transformer, suitable voltage control being obtained by taps and rheostats in the secondary side. The two resistances, R_1 and R_2, shown connected in series across the heater are two 100-ohm decade resistance boxes. Their common point is connected to the common point of the grid and plate circuits. By keeping the sum of the two resistances a constant, a potentiometer is provided by which the cathode can be connected to any potential point along the heater circuit. The position of the common point of the two resistances is indicated by the ratio a/b where a is the resistance between the common point and one side of the heater arbitrarily selected and b is the total resistance of the potentiometer. With a/b either zero or unity, one side of the heater is connected directly to the cathode and the other side varies in potential with respect to the cathode by an amount equal to the heater voltage. With a/b equal to 0.5 the heater is effectively connected to the cathode at its center point and the two heater leads vary in potential with respect to the cathode by equal amounts in opposite phase. In the shunt feed plate circuit the reactance of the choke coil is high and that of the blocking condenser is low compared to the load resistances used. A variable load resistance, R, has been provided and a resistance mounting with a switch, for short-circuiting when desired, has been placed in the grid circuit.

In the assembly of this apparatus in order to prevent pick-up, the heater circuit was separated from the rest of the tube under test circuit by placing each circuit in a copper lined box. To prevent inductive pick-up it was found necessary to place the heater transformer at a distance of several feet from the choke coil in the plate circuit of the tube under test.

The amplifier preceding the harmonic analyzer contained three resistance coupled stages, resistance coupling being used to eliminate the possibility of interference entering through the iron-core coils otherwise necessary. In the harmonic analyzer three tuned circuits discriminate against the unwanted frequencies and the necessary gain is obtained by two shield-grid tubes. A direct-current meter in the plate circuit of a triode operating near the point of plate current cut-off gives a measure of the output. It was found to be necessary to enclose the coil in the first tuned circuit within a magnetic shield in order to reduce sufficiently the interference experienced with other apparatus in operation in and near the laboratory.

Checks were made to note that the modulation in the amplifier was sufficiently small so as not to produce errors in the amounts of harmonics measured. Careful checks were also made to show conclusively that no disturbance currents were being introduced through parts of the circuit other than in the tube itself. In one test the heater of the tube under test was operated on direct current and the outputs of 60 and 120 cycles per second were found to be practically zero. With the heater still operating on direct current, the heater transformer supplied current to a dummy heater in the form of a short length of resistance wire placed as near as possible to the heater terminals of the tube socket; the output of the analyzer continued to remain at zero.

Disturbance Currents in Output of Triodes with Standard Hairpin Heater

The curves of Fig. 2 exhibit typical results of measurements of the fundamental and second harmonic disturbance current outputs as a function of the position of the cathode-heater common point (the a/b ratio) for tubes having an indirectly heated cathode of the usual type of hairpin heater. The measurements were taken with the tubes operating under normal conditions. The plate potential is 135 volts, the grid potential is -6.0 volts and the average plate current is 5.0 milliamperes. The heater voltage is 2.0 volts and the heater current is 1.6 amperes. The measurements were made with a load resistance equal to the plate resistance of the tube and with one or two hundred ohms resistance in the grid circuit. The outputs of fundamental and second harmonic disturbance currents are expressed in decibels below one milliampere. Here 80 db below 1.0 milliampere indicates an output ripple of approximately 1.0 millivolt; 100 db below 1.0 milliampere corresponds to approximately 0.1 millivolt. It will be noted that in most cases the output of fundamental frequency varies with changes in the ratio a/b. Although the minimum level does not occur as regularly at a/b equal to 0.5 as has been observed in tubes having the filamentary type of cathode, there is a general similarity in the relation. The level of second harmonic output is unchanged by variations in a/b. Measurements were also made of the third and fourth harmonic output currents. The level of the third harmonic generated in the tube was so low that it was masked by the presence of third harmonic in the power supply.

120 db below 1.0 milliampere represented an upper limit for the fourth harmonic output.

Measurements were made of disturbance current outputs as a function of the plate potential, grid potential, and of the impedance into which the tube operates. It was found over a considerable range of these parameters that the magnitude of the disturbance currents was proportional to the amplification obtained in the stage and was

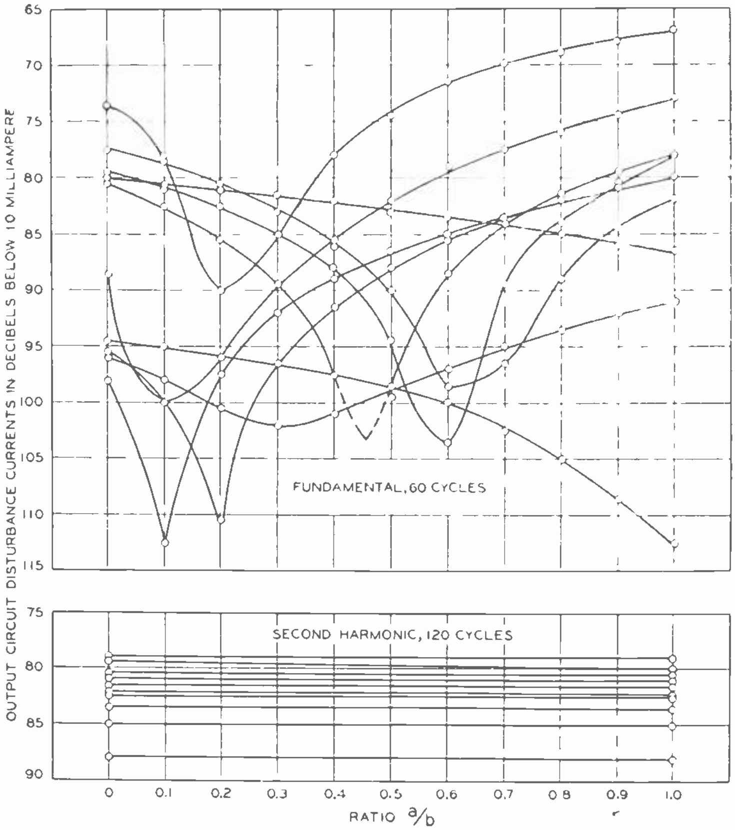

Fig. 2—Disturbance currents in typical indirectly heated cathode triodes used for radio reception.

independent of these operating parameters except as they affect the amount of the stage amplification. The level of the disturbance currents is affected materially by the value of resistance in the grid circuit as will be shown later.

Source of Disturbance Current Pick-up

It may be shown that the disturbance currents are introduced into the output circuit by:

1. Electric field of the heater
2. Magnetic field of the heater current

3. Resistance between heater and grid and between heater and plate, and capacitance between heater and grid, and between heater and plate.

Disturbance Currents Produced by the Electric Field of the Heater

The electric field of the heater element in the space between the cathode and anode will affect the electron current to the plate in precisely the same manner as does the electric field of the control grid. It is to be expected, therefore, that a disturbance current will be found in the plate circuit which is due to the grid action of the heater element.

With one point of the heater circuit connected to the cathode, the electric field of the heater at each point in the cathode-anode space will

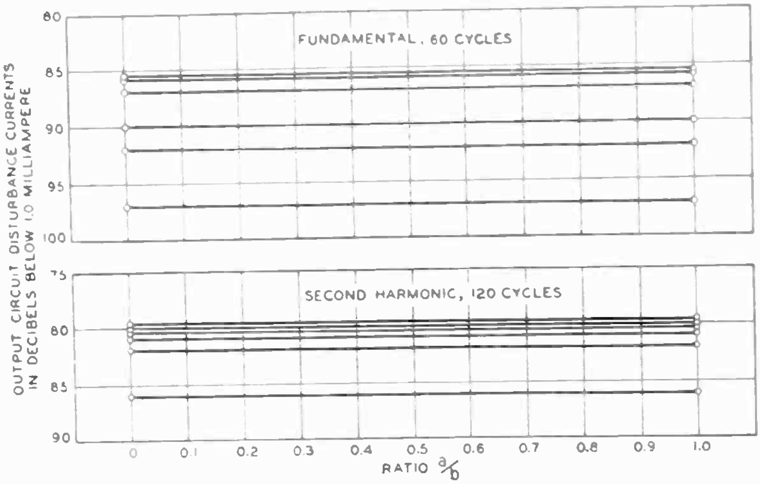

Fig. 3—Disturbance currents in indirectly heated cathode triodes having shields to reduce the electric field of the heater in the cathode-plate space.

be the sum of the fields due to each segment of the heater element. As the common point of the heater and cathode is shifted along the resistance across the heater, the value of the field will change. It would be expected when the common point was located at the symmetry mid-point of the heater circuit that the electric field would have its minimum value.

An examination of the disturbance currents in the output circuit as the common point of the heater and cathode is shifted along the resistance across the heater bears out, in a general way, this expectation. Results on typical standard tubes given in Fig. 2 show that the magnitude of the 60-cycle current in the output circuit does vary with the position of the common point. A definite minimum is shown in most cases. Since in the tubes under test the 60-cycle disturbance current is due to factors in addition to the electric field, the expected characteristic variation of the disturbance current due to shift of the common point is masked to varying degrees in the different tubes.

It should be possible to eliminate substantially the electric field of the heater in the cathode-anode space by adequate shielding of the heater circuit. The nickel sleeve upon which the active material of the cathode is placed acts as such a shield for a portion of the heater circuit. Experimental tubes were constructed in which further shielding was provided. This shielding was made so complete that the electric field of the heater in the cathode-anode space should be substantially zero. The results of disturbance current measurements of six such tubes are shown in Fig. 3. In these tubes there is no variation in the

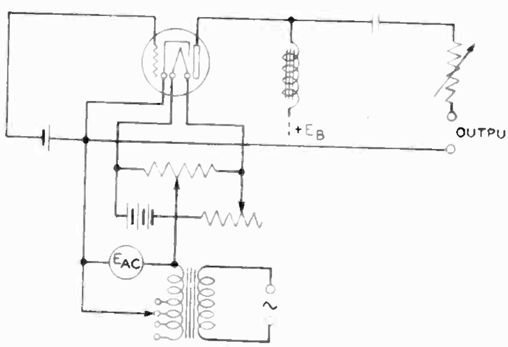

Fig. 4—Circuit for the study of the electrostatic effect in the presence of large disturbance outputs due to the magnetic field.

60-cycle current as the position of the common point is shifted. It, therefore, seems reasonable to conclude that the disturbance currents due to the electric field are substantially eliminated.

The disturbance currents in these specially shielded tubes were due mainly to the magnetic effect of the heater current. In order to get a somewhat better indication of the adequacy of the shielding, an experiment was performed in which the magnetic effect of the heater current was eliminated. The cathode was heated by direct current and a variable alternating voltage was applied between the heater and cathode. The circuit arrangement is shown in Fig. 4. In this experiment the electric field of the heater was different from that when the heater was operated on alternating current. In this case the entire heater circuit was, of course, at one alternating potential with respect to the cathode. However, the adequacy of shielding can well be tested in this manner. In Fig. 5 are given such measurements on the shielded tubes of Fig. 3 and on unshielded tubes that are similar to them in all resects except for the shielding. It will be seen that the shielding has reduced the level of the 60-cycle disturbance current by approximately 40 db.

In our discussion of the electric field disturbance current we have so far limited ourselves to the 60-cycle current. It will be noted that in the

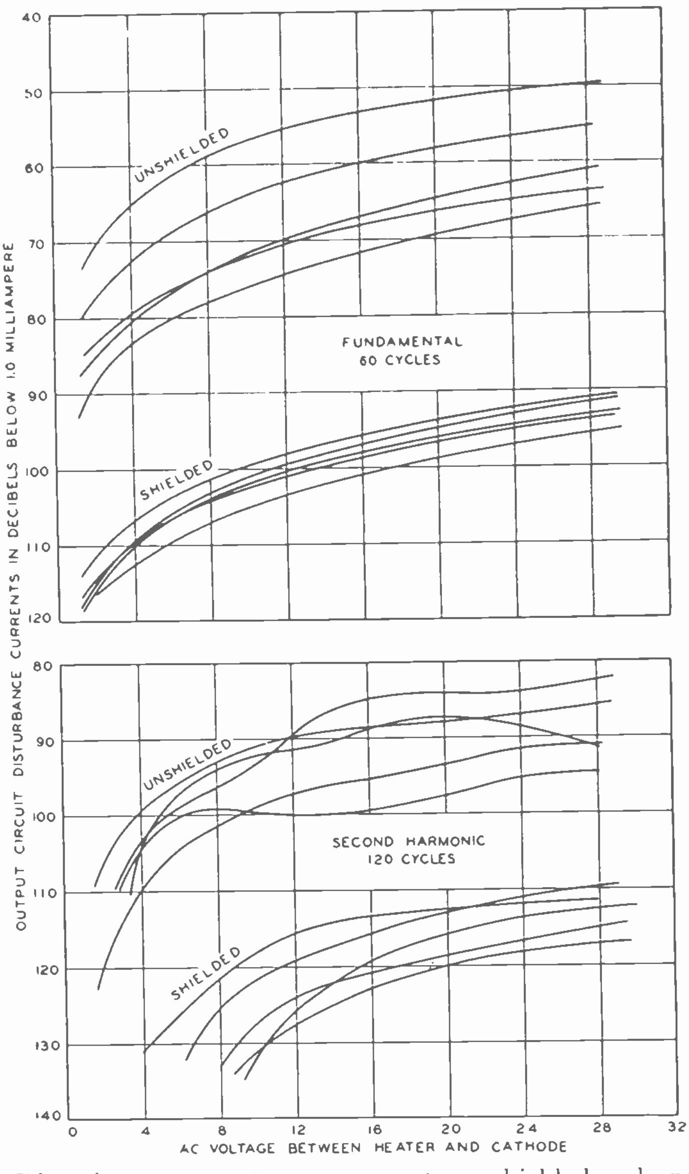

Fig. 5—Disturbance current measurements on shielded and unshielded tubes with the circuit of Fig. 4.

data of Fig. 5, second harmonic disturbance currents are presented and that their absolute value is shifted approximately 30 db by the shielding. While the data given for typical tubes in Fig. 2 do not indicate any variation in the second harmonic current with the position of the common point, we have observed a number of cases of standard heater type tubes where such a variation was present. The evidence from such standard tubes, as well as the data given in Fig. 5, indicates clearly that second harmonic disturbance currents may be due to the electric field effect. The presence of such second harmonic currents is to be expected from fundamental considerations. The grid action of the heater circuit varies in a nonlinear way as the effective voltage of the heater system is changed with respect to the cathode. This nonlinearity of the grid action would be expected to produce second harmonic components in precisely the same manner as they are produced in the familiar case of μ modulation with the standard control grid.

These experiments have indicated the presence of disturbance currents due to the electric field of the heater and have roughly established their magnitude in the case of tubes of standard construction. The substantial elimination of disturbance currents due to the electric field of the heater by adequate shielding of the heater circuit in the experimental tubes, has indicated a possible means of controlling the level of disturbance currents due to the electric field of the heater in the design of standard tubes.

Disturbance Currents Produced by the Magnetic Field of the Heater Current

The magnetic field of the heater current in the space between the cathode and anode will affect the electron current to the plate. The electrons will be deflected by the magnetic field according to the magnetic field force relations.[2] The deflection of electrons by this field causes a double-frequency change in the electron space charge condition which results in a second harmonic component of disturbance current in the anode circuit. Due to asymmetries in the space charge system, the two changes in space charge per cycle of the heater current are not equal. The inequality in the two changes will produce a disturbance current in the plate circuit of the same frequency as that of the heater current.

In order to observe the effect of the magnetic field on the disturbance current output, experimental tubes were constructed with the cathode arrangement shown in Fig. 6. In this arrangement a hairpin heater of the usual type is mounted in a two-hole insulator. A platinum cylinder is placed over the outside of the insulator and is welded to the upper end of the hairpin heater. An electrical connection from the bottom end of this cylinder is made through a lead wire in the press. A second cylindrical insulator fits tightly over the platinum cylinder and the nickel sleeve coated with the thermionically active material

[2] See bibliography, item 10.

is placed on the outside of this insulator. The cathode unit is mounted in a structure having electrical characteristics and electrostatic shields identical with those of the group from which the data of Fig. 3 were taken.

The cathode can be heated to an operating temperature by two different heater systems. In the first system, voltage is applied across the terminals of the hairpin and the platinum cylinder plays no part as a heater unit. In the second system, the two terminals of the hairpin are connected together and the heater voltage is applied across the

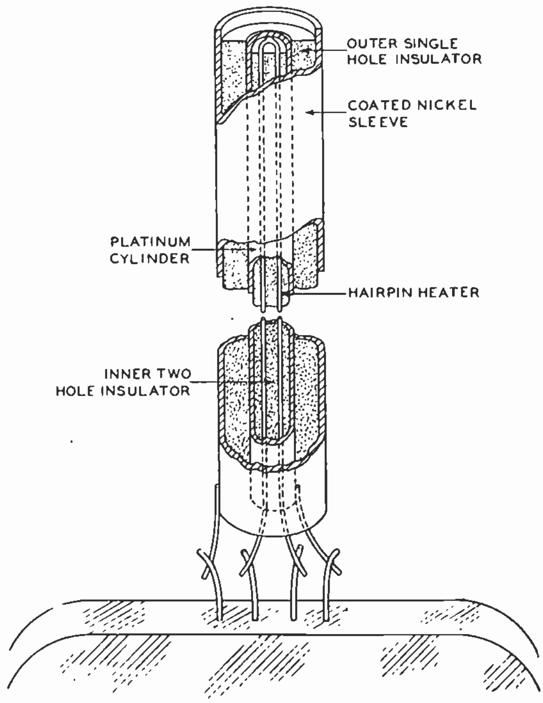

Fig. 6—Experimental heater arrangement for the study of the magnetic effect.

cylinder in series with the two heater legs connected in parallel. Platinum was used in the current cylinder, since it was the most convenient metal that would withstand the necessary temperatures during pumping. The computed magnetic field intensity in the space between the cathode and anode for the second system is approximately one-fourth of that of the first. Due to the fact the platinum cylinder was formed from sheets of insufficient width, the cylinder was not complete, and the field, therefore, was not reduced by the computed amount.

The disturbance current outputs for a number of such tubes were measured with both heater systems and the current due to the electric field was found to be substantially eliminated by the electrostatic shield. The fundamental output is produced by assymmetrics that are random and no exact relation between the magnetic field and the fundamental is predictable. In the tubes measured, the fundamental decreased a maximum of 22 db and increased in one tube as much as 9.5 db in changing from the first to the second heater system. Similar measurements made on the second harmonic output indicated a decrease in level in all tubes ranging from 11 to 15.5 db. If the current cylinder had been complete, the reduction would have been approximately 22 db.

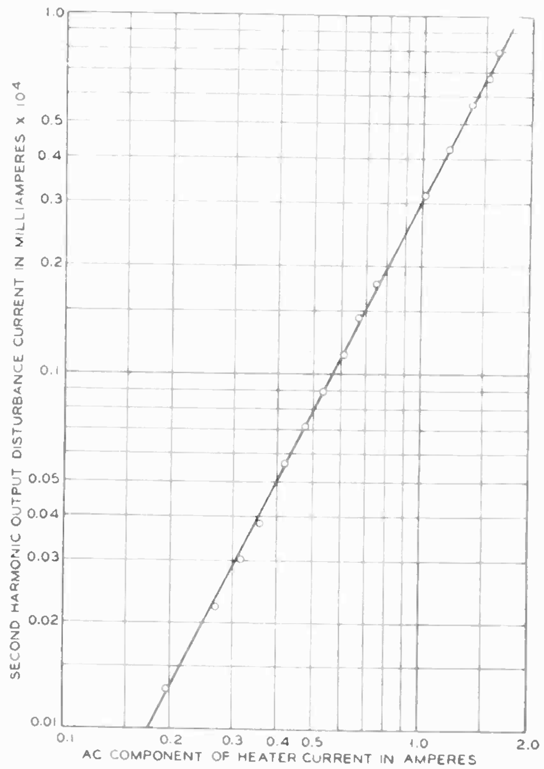

Fig. 7—Second harmonic output disturbance current produced by the magnetic field as a function of the alternating-current component in the heater.

From fundamental considerations it is expected that the second harmonic current would be proportional to the square of the magnetic field. An experimental verification of this square law has been obtained. The heater of one of the tubes, from which the data of Fig. 3 were obtained, having a cathode sheath and an electrostatic shield, was operated on a combination of alternating and direct current.

The alternating-current component was varied from the normal heater current of 1.6 amperes to approximately 0.2 ampere. The direct-current component was adjusted so that at each point the total effective heater current was 1.6 amperes. In Fig. 7 the second harmonic disturbance current in the output circuit is plotted as a function of the alternating component of the heater current. It will be observed that on the log-log coördinate system the points fall along a straight line whose slope is approximately two.

It is evident that in order to reduce the disturbance currents due to the magnetic field of the heater current it is necessary to reduce the magnetic field. Three general methods for this reduction suggest themselves.

1. The selection of a heater element of such geometry that the magnetic field in the space between cathode and anode is decreased.
2. The placing of a magnetic shield around the heater system.
3. Obtaining the necessary heating power for the cathode with a higher voltage and lower current heater unit.

The simplest form of heater is a single wire coaxial with the cathode cylinder. The magnetic field at the surface of the cathode is inversely proportional to the radius of the cathode cylinder. To decrease the magnetic field of such a heater the radius of the cathode is increased. For a given cathode area an increase in its radius necessitates a corresponding decrease in its length. Heat losses from conduction at the two ends and uniformity of cathode temperature determine the maximum practicable diameter. Some flexibility in design is permitted by spiraling the heater. Employing a spiral diameter sufficiently small compared to that of the cathode, the field at the cathode surface is substantially the same as that due to a straight conductor in the axis of the cylinder. The optimum practicable heater of this type offers less possibility in the reduction of the magnetic field than any of the other heater arrangements considered.

The hairpin type heater is probably the simplest form in which partial neutralization of magnetic field is obtained by the form given to the heater. In this unit the magnetic field produced by one leg is neutralized in part by that of the other leg. In this type of heater the maximum field at the surface of the cathode is proportional to $c/(r^2-c^2)$, where c is one-half the distance between the legs of the heater wires, and r is the cathode radius. As c is decreased and r increased, the magnetic field is reduced. The mechanical difficulties in making the twin-bore insulator and increased tendency for its fusion as c is decreased place definite limits on this dimension. The same considerations hold in determining the optimum value of r that hold in the single-wire heater. For the low voltage heaters commonly used in radio receiving tubes the optimum arrangement results in a reduction in second harmonic output due to the magnetic effect of approximately 10 db over that obtained with the optimum single coaxial heater. The tubes on which data were given in Fig. 2 have a hairpin heater corresponding approximately to this optimum design.

A heater composed of two concentric conductors carrying the same current in opposite directions will give substantially no magnetic field at the surface of the cathode. This ideal heater can be approached in a closely wound spiral with a single wire return in its axis. Two spirals of the same diameter with their windings interlaced also give a very small magnetic field at the surface of the cathode. Experimental tubes having heaters of optimum practicable design of these two types have been made. The results obtained indicated some improvement in the second harmonic output produced by the magnetic field over levels noted in comparable hairpin heater structure. The improvement was of the order of 15 db. There are serious manufacturing and quality control difficulties with such heater units.

The magnetic field in the space between cathode and anode for the hairpin type of heater can be reduced by surrounding the heater unit with a magnetic shield. Considerable portions of the shield must operate at a temperature of approximately 750 degrees C. It is necessary to find a material which will maintain a high permeability at this temperature. Nickel loses its magnetic properties at approximately 340 degrees C and iron loses its magnetic properties at approximately 750 degrees C. The magnetic properties of cobalt are satisfactory at 750 degrees C, but it is difficult to work cobalt down to the required dimensions. An alloy of iron and cobalt that would maintain a high permeability at a temperature well in excess of the operating temperature of the cathode was prepared.

Experimental tubes with heaters of the hairpin type and having a magnetic shield of this material placed between the insulator and the cathode nickel sleeve were constructed. The second harmonic disturbance current due to the magnetic field was approximately 20 db less in tubes with the magnetic shield than in identical tubes with no shield.

The magnetic field is proportional to the heater current. A reduction in magnetic field can, therefore, be obtained by the use of a heater in which the current component of the power is small and the voltage component correspondingly large. Halving the heater current and retaining the same heater and cathode geometry reduces the second harmonic magnetic component to one-fourth or by approximately 12 db.

An increase in heater voltage raises the level of the electric field components of disturbance current. This must be taken into account in the consideration of heater unit design.

Disturbance Currents Produced by Resistance and Capacitance Between Heater and Grid and Heater and Plate

Heater circuit voltages are introduced into the grid circuit and into the plate circuit through resistance and capacitance between the heater and each of these elements. A circuit diagram is shown in Fig. 8a which indicates the circuit paths. For simplicity, one side of the heater is shown connected to the cathode and the resistances and capacitances from the heater to the other elements are connected to the opposite side of the heater. The grid-heater circuit is shown schematically in Fig. 8b. The heater voltage is indicated as a generator from

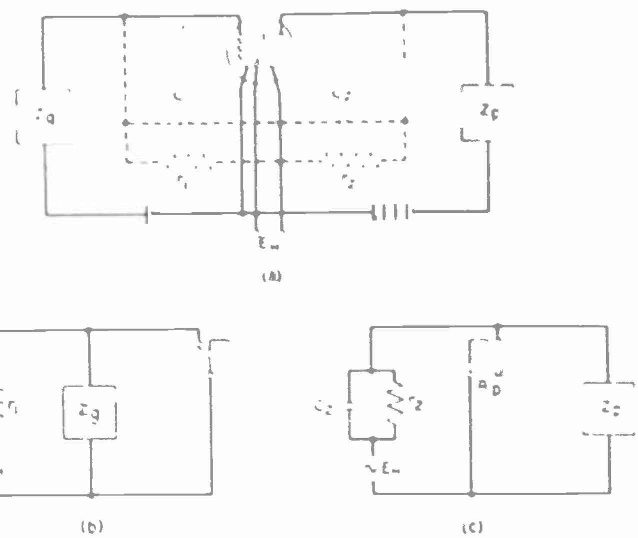

Fig. 8—Diagrams of disturbance current circuits between heater and grid and heater and plate.

which current flows through the grid circuit impedance, Z_g, and through the resistance in parallel with the capacitance between the grid and the heater. The voltage drop across Z_g is applied directly across the grid and cathode so that this voltage amplified will appear in the plate circuit.

The plate-heater circuit is shown schematically in Fig. 8c. The heater voltage is again shown as a generator. It will be noted that in this circuit Z_p is in parallel with the output resistance of the tube, R_p, and that the disturbance voltage is not amplified.

In the usual case the heater will be connected to the cathode at some point other than the end, which makes it necessary to consider two driving voltages and two sets of resistances and capacitances between heater and grid and plate.

An experimental tube was constructed with extremely high insulation and the capacitance between grid and heater was made negligible by bringing the grid lead out of the top of the tube and by having no grid support in the stem press. A resistance of 1000 ohms was placed in the grid circuit and resistances varying in value from 30,000 ohms to 90 megohms were connected from one side of the heater to the grid. The cathode was connected to the opposite side of the heater. Fundamental output disturbance currents were measured for the different values of grid-heater resistance. The results of these measurements are given in Fig. 9. This current was also computed from the relation

$$\frac{E_H Z_g}{r_1 + Z_g} \cdot \frac{\mu}{R_p + Z_p}$$

It will be seen that the experimental points lie substantially on the computed line which give concrete experimental verification of the circuit relations indicated in Fig. 8.

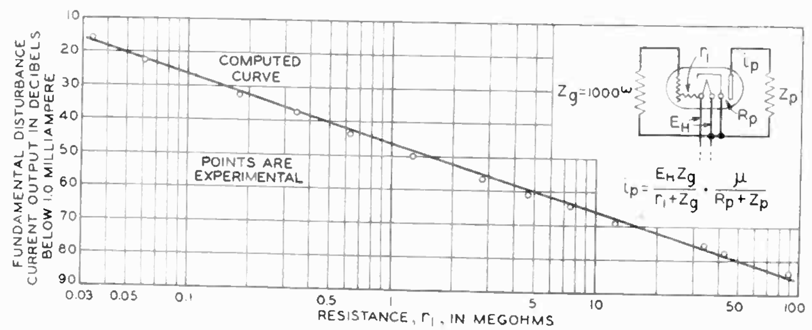

Fig. 9—Fundamental disturbance output current as a function of the grid-to-heater resistance.

A Tube with Extremely Low Level of Disturbance Currents in the Output Circuit

These studies of disturbance currents produced in the output circuit by alternating current in the heater element have been used as a background for the design of a tube with a sufficiently low level of disturbance currents in the output circuit to be satisfactory for use in all voice frequency amplifying systems.

In order to get the required low level of the disturbance currents due to the magnetic field, a low current, high voltage heater was adopted. This heater is a fine wire wound as a close spiral and shaped in the form of a hairpin. It is mounted in a twin bore insulator. The voltage across the heater is 10 volts and current is 0.3 ampere.

To obtain sufficiently low disturbance currents due to the electric field, substantially complete shielding of the heater circuit from the glass press to the top of the hairpin is employed. This shielding is obtained by suitable extensions of each end of the cathode cylinder. The shielding arrangement can be seen in the photograph, Fig. 10.

The capacitance between the heater and grid is materially decreased by the complete shielding of the heater system above the press, and by making the connection for the grid through the top of the bulb. No grid support wires enter the stem press, the grid being supported

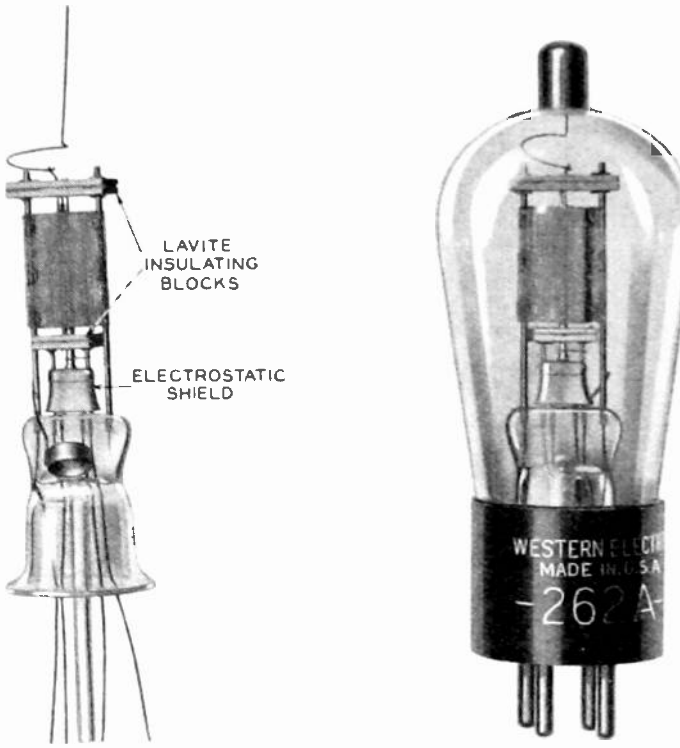

Fig. 10—No. 262A mount and completed tube.

at top and bottom by the lavite insulators shown in Fig. 10. The lead through the top of the bulb is terminated in a cap as is common practice with the control grid in shield-grid tubes. The direct grid-to-heater capacitance has been reduced to approximately 1/1000 of the value common in indirectly heated cathode tubes used in radio reception. No special precautions have been necessary in reducing the heater-to-plate capacity, although some reduction is effected by the heater shielding. This capacity is approximately 1.0 $\mu\mu$f.

The electrostatic shield covering the heater leads effectively prevents the deposit of metal vaporized during the pumping process, and of material vaporized from the cathode during life, from forming a leakage path on the stem press between plate and heater leads. With the tube elements at operating temperature, the resistance between heater and plate is maintained at a value greater than 100 megohms. With the grid removed entirely from the stem press, the insulation between grid and heater is considerably greater than 100,000 megohms.

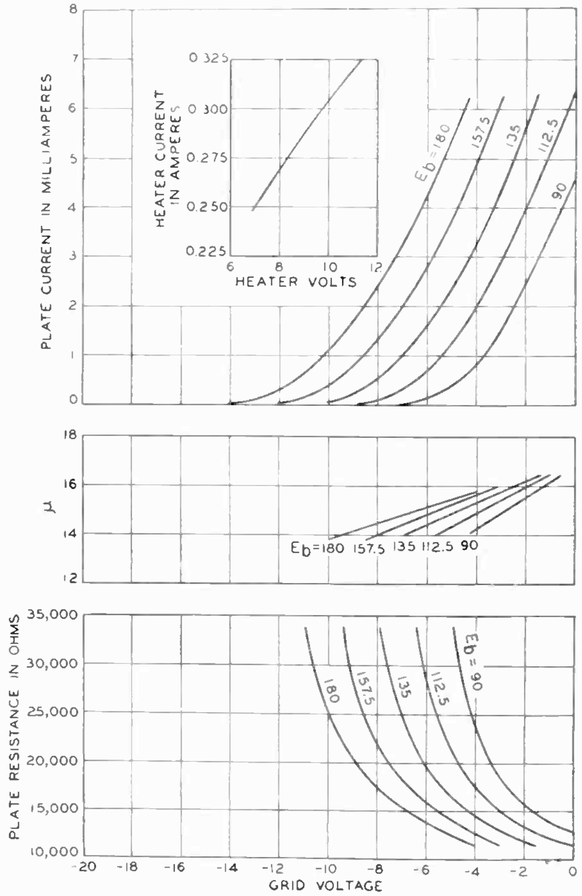

Fig. 11—Electrical characteristics of the No. 262A tube.

When 2.0 megohms are inserted and taken out of the grid circuit, with the cathode connected to the mid-point of the heater circuit, the average change in fundamental output for a representative number of tubes is less than 1.0 db. For the same cathode-to-heater connection the corresponding increase in fundamental output was 25 db for similar tubes in which the grid lead was taken through the stem to the usual prong in a standard five prong base.

The electrical characteristics of the tube are given in Fig. 11. The tube is normally used with a plate potential of 135 volts and a grid bias of -4.5 volts. Under these conditions the plate current is 3.0

milliamperes, the output impedance 15,000 ohms, and the amplification factor 15. The tube is satisfactory for use with a plate potential of 180 volts and a plate current of 10.0 milliamperes. The tube has been given the Western Electric code number 262A. The mount and the completed tube are shown in Fig. 10. Distribution curves of disturbance currents in the output for typical tubes under normal conditions of operation are shown in Fig. 12. The input resistance used here is less than 100 ohms. These data indicate that for $a/b=0.5$, the level of fundamental disturbance current for all tubes is lower than 95 db below 1.0 milliampere (0.27 millivolt), and the level of second harmonic disturbance current for all tubes is lower than 105 db below 1.0 milliampere (0.084 millivolt).

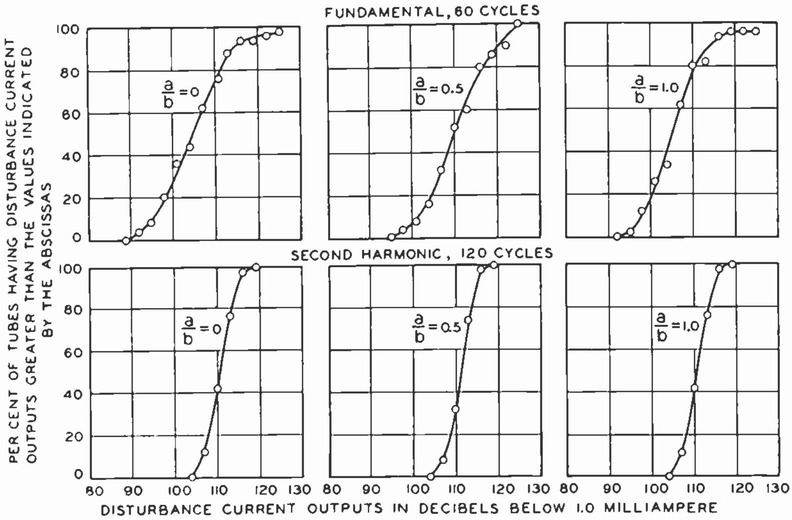

Fig. 12—Distribution of disturbance output currents in a representative number of No. 262A tubes picked at random.

The level of output noise from sources other than the alternating-current heater is of interest. The noise level was measured for representative tubes in a voice frequency amplifier of flat frequency characteristic. The heater was operated on direct current. With an input resistance of less than 100 ohms, the noise level in the output circuit varies between 118 and 127 db below 1.0 milliampere. This noise is principally due to the shot effect from the cathode. With 2 megohms in the grid circuit, the noise level in the output circuit is approximately 105 db below 1.0 milliampere. This noise is substantially all due to the resistance noise of the grid circuit.

In order to obtain a sufficiently high resistance between grid and cathode and grid and plate, and to maintain this value throughout life, special lavite blocks at the two ends are used for mounting the tube elements. These blocks are shown in Fig. 10. They are so designed that there is not a continuous path between any two of the tube elements on the side of the blocks facing the tube elements. This makes it impossible to obtain leakage paths from vaporized metal or material from the cathode.

It is essential that a tube for use in the first stages of a high gain voice frequency amplifier have the lowest microphonic response characteristic possible. In applications which subject the tube to mechanical shock, with even the best possible cushioning and shielding, microphonic disturbance currents may be introduced. Every precaution has been taken in the detailed mechanical design of this tube to make microphonic pick-up a minimum. The top and bottom insulators essential to the maintenance of high resistance have been a material structural aid in reducing microphonic noise levels.

Equipment has been developed in which tubes can be given reproducible mechanical shocks and the microphonic noise currents in the output circuit measured.[3] In this measuring apparatus the microphonic output levels from the No. 262A tube are approximately 20 db lower than from the standard 2.0-volt heater type of tube for radio reception and approximately 10 db better than specially designed filamentary tubes for low microphonic noise level, such as the Western Electric No. 264A tube.

Acknowledgment

The author wishes to express appreciation to Mr. M. J. Kelly and Mr. H. A. Pidgeon for consultation and suggestions during these studies and especially to Mr. Kelly for consultation in preparation of the manuscript.

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[3] A description of this equipment and information concerning microphonic noise from various types of tubes will be published in the near future by Mr. H. A. Pidgeon and Mr. D. B. Penick of the Bell Telephone Laboratories.

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Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

MODERN RADIO EQUIPMENT FOR AIR MAIL AND TRANSPORT USE*

By

A. P. Berejkoff and C. G. Fick
(Radio Engineering Department, General Electric Company, Schenectady, New York)

Summary—*The general requirements for aircraft radio equipment for air mail and transport use are discussed. This is followed by a discussion of the factors which were considered in the design of an aircraft radio telephone and telegraph equipment. A description is then given of the mechanical and electrical features of a specific equipment.*

In the course of the last four years, during which aviation radio has developed along with our aviation systems, the requirements for aircraft radio transmitting apparatus have been pretty well established.

Some operating companies have built their communication systems around radiotelephone operation from plane-to-ground and ground-to-plane while others have selected cw telegraph communication for their systems. The conditions governing the choice of the most suitable system are many and varied. Of these, the most important are the length of the flight or the distance between route radio stations, the terrain, weather, and transmission and reception conditions.

For radiotelephone operation, two general classes of equipment are being used. One of these classes provides a 50-watt carrier, and the other provides a 10- to 15-watt carrier. Both have provisions for complete telephone modulation.

Where cw telegraphy is chosen for the communications system, the circumstances mainly governing its choice are that communication distances required are so great that to provide sufficient radiotelephone power would result in an impractical weight, size, and power drain for the equipment, and also that the average atmospherics and transmission conditions are such as to make telephone communication well-nigh impossible.

The general power output requirements for cw telegraph systems have been found to be from 10 to 20 watts. This amount of power in telegraph operation has provided regular plane-to-ground communication of several hundred miles even under the most adverse conditions.

* Decimal classification: R526. Original manuscript received by the Institute, March 24, 1932. Presented before Twentieth Anniversary Convention, Pittsburgh, Pa., April 8, 1932.

The chief advantages of the cw telegraph system are the small size, light weight, and low power drain of the equipment, while the chief disadvantage of the system is the relative inconvenience of communication by telegraphic code. In some systems, the radio communication by cw telegraph is carried on as one of the duties of the copilot, while in others, a radio operator is carried. In the latter instance, it has been argued that the total equipment weight including an operator and a small cw equipment provides such excellence and reliability of service that it is fully justified.

Having stated the very general requirements for the three classes of aircraft radio equipments, it is now planned to discuss the design of one type equipment, namely, a radiotelephone transmitter equipment for air mail and transport use. The important factors which must be considered in the design of such an equipment will be discussed, and finally there will be given a description of a specific equipment, the General Electric type RT-76-A, which has been brought out for the airmail and transport service.

In the design of aircraft radio equipment, the fundamental requirements to be considered are:

(a) Low weight with sufficient ruggedness to stand hard service.
(b) Small dimensions.
(c) Low power drain.
(d) Ability to operate into various types of antennas.
(e) Capability of emergency operation.
(f) Simplicity of adjustment and operation.
(g) Accessibility of construction to allow rapid servicing.
(h) Conservative operation of the circuit elements to insure low maintenance costs.
(i) Low initial investment cost for the equipment and installation.

Of first importance in the design of the transmitter is the selection of vacuum tubes. Since the power output desired was at least 50 watts, the UV-211 tube was chosen. This tube is nominally rated at 100 watts output so that by its use, a considerable amount of overload capacity was available. Also the UV-211 tube has proved its reliability and ruggedness in many other branches of radio service, and accordingly its use in aircraft seemed entirely justified.

The selection of the UV-211 type tube predicates the use of 1000 volts direct current for plate supply. This plate supply can be obtained from a number of sources namely:

(a) Wind-driven direct-current generator with constant speed propeller.
(b) Engine-driven direct-current generator.

(c) Dynamotor operating from the airplane's 12-volt storage battery.

(d) Alternating-current generator, driven by a self-regulating propellor or by the airplane engine. The alternating current is generated at low voltage and is stepped up and rectified by the conventional circuits.

Of the possibilities for obtaining the vacuum tube plate supply, the dynamotor has found most general use. Its advantages are mainly that it is independent of the airplane engines, it is not required to be mounted in the air stream where it would increase the parasitic drag, and since it is not dependent on the motion of the airplane or its engines, it provides emergency operation together with the possibility of easily checking the radio equipment's operation before the take-off. The dynamotor's chief disadvantage is of course, the fact that it causes a heavy drain on the aircraft storage battery, necessitating that the battery be continuously charging from a low voltage generator attached to the airplane engine.

Of the other two direct-current machine possibilities, the engine-driven generator is perhaps more generally used. With either the engine-driven or wind-driven generators there is of course, no possibility of emergency operation when the airplane is down, and checking of the radio equipment operation before the take-off is relatively difficult.

The use of alternating current with rectification and filtering has been discussed in a previous paper presented before the Institute.[1] This type of operation, although possessing certain advantages, is not finding very wide use in the commercial field.

The General Electric RT-76-A transmitter has been designed primarily for dynamotor operation although it can by slight modification be adapted to operation from an engine-driven direct-current generator.

In selecting a radio-frequency power amplifier circuit for the transmitter, there were three possible methods of obtaining the required output and modulation capabilities. They were:

(a) Class B power amplifier requiring modulated radio-frequency input.

(b) Grid bias modulated amplifier.

(c) Class C power amplifier requiring plate modulation.

The class B power amplifier circuit has been described elsewhere.[2] The main advantage of this type circuit is that by its use, modulation can be accomplished at a power level lower than the output. This re-

[1] Miner, Proc. I.R.E., vol. 19, pp. 59–77; January, (1931).
[2] Fay, Proc. I.R.E., vol. 20, pp. 548–569; March, (1932).

sults in the use of smaller tubes and lower filament and plate drain for the modulators, a very important consideration in the design. The disadvantages in the use of class B power amplification, however, outnumber the advantages. In Fig. 1 is shown the typical operation of a UV-211 as a class B amplifier. Although a peak power of 200 watts is possible, thus allowing the use of a 50-watt carrier output with 100 per cent modulation, it may be seen that some distortion would be introduced by reason of the curvature of the characteristic at high modulation. As a matter of fact, the UV-211 is rated at only 40 watts carrier

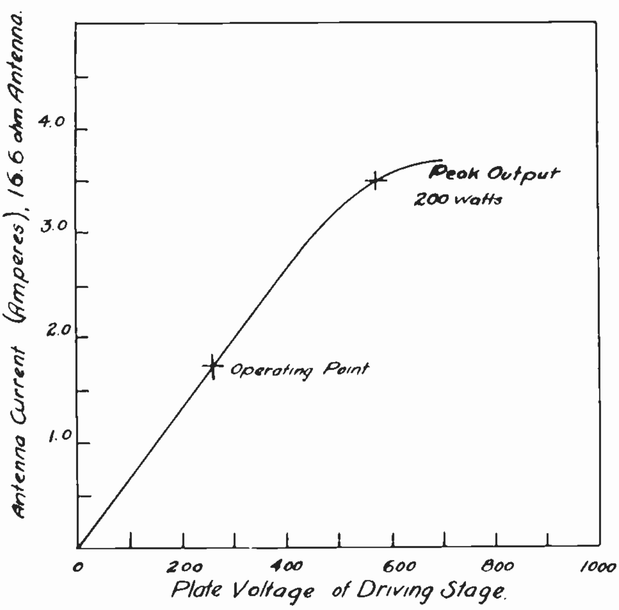

Fig. 1—Typical operation of a UV-211 as a class B amplifier. Drawing stage, UX-210.

output for class B service. Thus, to provide at least a 50-watt carrier plus provision for losses in the antenna tuning system, and provision for overload capacity, two UV-211 tubes would have to be used in the amplifier stage. A further disadvantage of class B operation is that the driving stage must be completely modulated. This means that an intermediate amplifier must be inserted in the tube line-up for the purpose, since it would not be possible to generate the radio-frequency oscillations and produce complete modulation in a single stage. The introduction of an intermediate amplifier stage would require an additional tuning control and space for its tuned circuit. Also, with the plate supply of 1000 volts available, the smaller tubes which could be used in the generation, intermediate, and modulator stages would all have to be operated at lower voltage than the power amplifier. This would necessitate the use of power wasting, voltage dropping resistors.

The bias modulated power amplifier falls in practically the same category as the class B amplifier in so far as its utility in the present equipment is concerned. The bias modulated system does not require the input radio-frequency excitation to be modulated, since modulation is obtained by varying at an audio-frequency rate, the axis of a constant input radio-frequency voltage instead of, as in the class B system, maintaining a constant axis and supplying an amplitude modulated radio-frequency excitation. The elimination of the modulated radio-frequency input requirement for bias modulation would allow

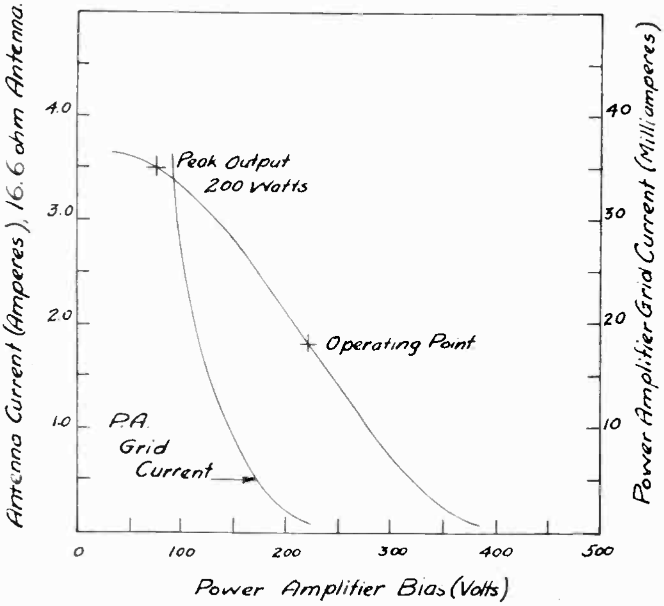

Fig. 2—Typical operation of a UV-211 as a bias-modulated amplifier.

the intermediate stage to be dispensed with. The system is limited in output capability in the same manner as the class B amplifier, and in addition requires special precautions in the design of the modulator circuit, since the load on the modulator varies considerably over the audio-frequency cycle. Fig. 2 shows typical operation of the UV-211 tube in a bias modulated power amplifier circuit. Note particularly the rapid change of grid current with bias voltage.

The class C type of power amplifier circuit was chosen for use in the RT-76-A transmitter because of its obvious advantages over the other systems mentioned. The UV-211 is rated at 100 watts output for class C operation. Thus there is ample provision for a 50-watt carrier plus antenna tuning system losses, plus considerable overload capacity without operating the tube beyond its rating. Also the class C amplifier is simplest in operation since it is not critical to adjustments of bias voltage, radio-frequency excitation, or power output loading.

The consideration of a driver circuit brought forth the question of whether a crystal oscillator or master oscillator type of circuit should be used. With the crystal oscillator, there was the possibility of attaining a frequency stability of approximately 0.025 per cent. On the other hand, there was the disadvantage of having to use the crystal oscillator tube at a relatively low voltage, thus necessitating a power wasting potentiometer for its low voltage plate supply. There was also the necessity of introducing between the crystal oscillator tube and power amplifier, an intermediate amplifier circuit capable of doubling

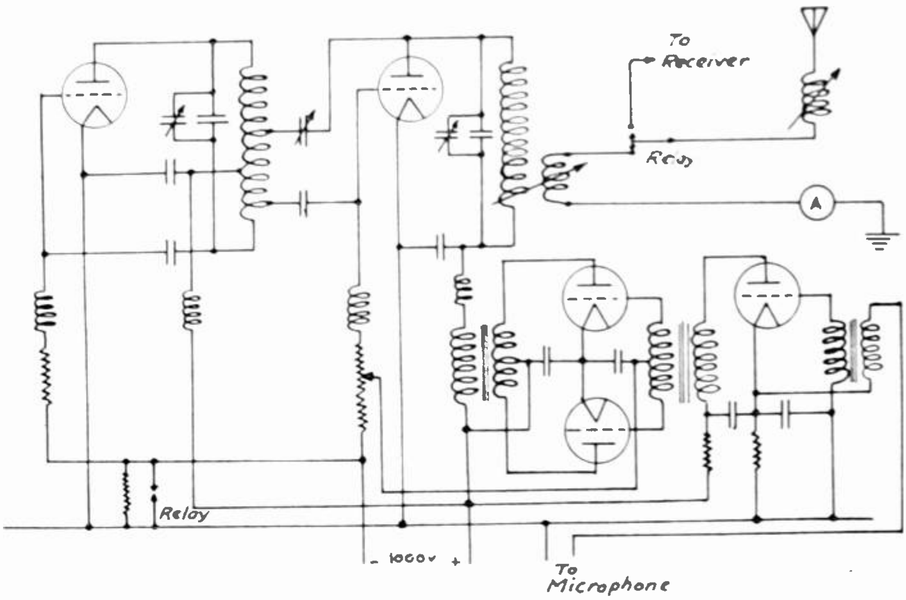

Fig. 3—Schematic diagram, RT-76-A transmitter.

the crystal frequency, since some of the output frequencies desired were too high for practical fundamental frequency operation of the crystals.

In order to provide a linear relationship between power amplifier plate voltage and output load current up to a plate voltage of 2000 (the condition for complete modulation), the UV-211 tube requires a driving power of from 5 to 10 watts. This amount of power would be difficult to obtain from the frequency doubling stage without exceeding the limits of conservative operation of the crystal and doubler stages.

The master oscillator type of circuit was selected for use, because of its inherent simplicity and reliability. A UV-211 tube was used for this purpose also, since it could be operated at the same plate voltage as the power amplifier. Furthermore, the use of a lightly loaded, high power master oscillator protects against frequency variation due to antenna detuning, and against frequency modulation caused by variation of the power amplifier input impedance with modulation.

In the design of the master oscillator circuit, a frequency stability of at least 0.05 per cent was sought. This value, well within present-day requirements, represents a stability which can be attained by careful design. Master oscillators with better stabilities than the value given above are of course possible, but the attendant complexities are such as to eliminate their consideration in aircraft radio.

After the selection of the radio circuits, the modulator was next considered. The class B type[3] of modulator system is admirably suited

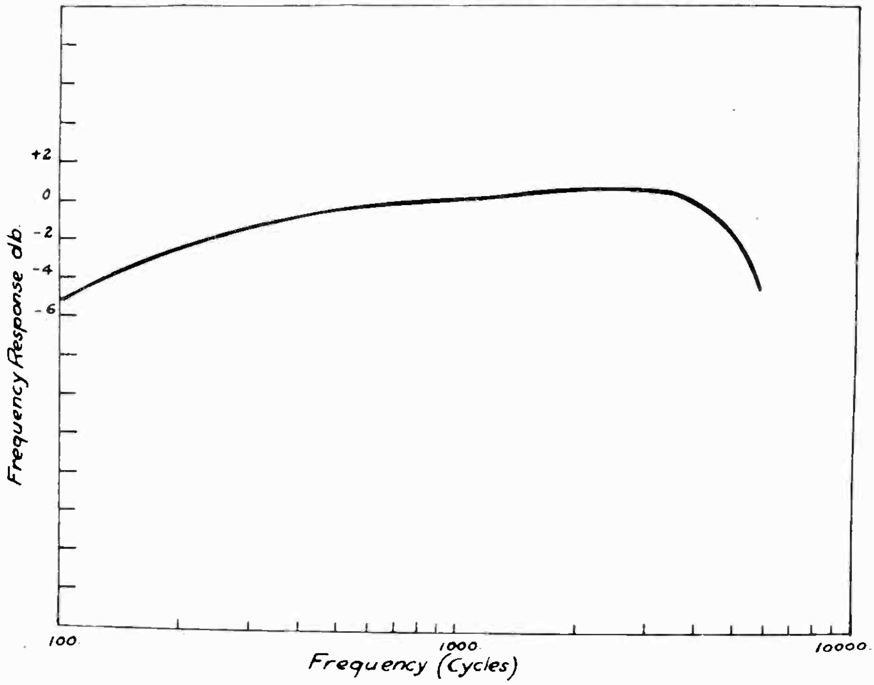

Fig. 4—Frequency response, RT-76-A transmitter.

to use in aircraft equipments because of the fact that it provides the maximum power output for a given tube line-up, and also it has the great advantage that no plate power is required unless modulation is supplied.

Two UV-211 tubes were selected as modulators, and for a speech amplifier to supply the necessary excitation, a single UX-841 was used. The single stage of speech amplification was found to supply ample grid voltage swing and driving power for the UV-211 modulators since they were called upon to supply only approximately 50 watts of audio power. The chief reasons for selection of the UX-841 were the ease of obtaining its bias, since it is a high amplification factor tube, and the fact that its operating plate current, obtained from the 1000-volt supply through a dropping resistor, is low.

[3] Barton, Proc. I.R.E., vol. 19, pp. 1131–1150; July, (1931).

The antenna tuning system was designed and built as a unit separate from the transmitter, mainly for reasons of flexibility of operation, and reduction of over-all transmitter dimensions. It was believed that in certain aircraft installations where space is at a premium, it would be considerably easier to locate and mount two smaller units than to find space for a single large unit containing both the transmitter and antenna tuning equipment.

Fig. 3 shows a schematic diagram of the circuits used in the RT-76-A transmitting equipment. The tuned circuits of the master oscil-

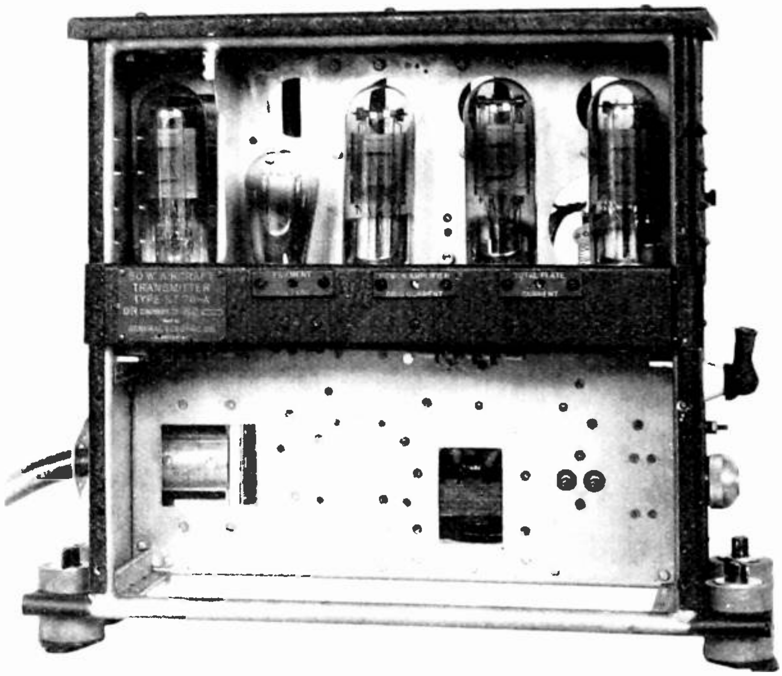

Fig. 5—Front view of RT-76-A transmitter with tube shield off and tuning unit out.

lator and power amplifier, together with the necessary radio-frequency chokes, and antenna coupling inductance are mounted in a replaceable tuning unit shown in Fig. 6. Two tuning units are included with each equipment, one for the 3000-kilocycle band of aircraft frequencies, and one for the 5000-kilocycle band. In addition to this, the design has been carried out to the point that tuning units are available for any frequencies in the band, 1500 to 6000 kilocycles.

Sufficient mutual inductance is provided for feeding antennas of resistances as high as 16 ohms. The coupling to the power amplifier tuned circuit is variable, and is controllable from the front panel of the tuning unit.

A "Send-Receive" relay is included in the transmitter unit. For telephone operation, the relay is controlled by a push-switch located on the microphone, while for cw operation, the "Send-Receive" switch may be mounted on or conveniently near the transmitting key. In addition to turning the transmitter carrier on or off, the "Send-Receive" relay transfers the antenna from transmitter to receiver and vice versa. A set of contacts is also provided, which may be used for making the receiver less sensitive when the transmitter carrier is on.

The transmitter is designed to be operated with the usual antinoise carbon microphone. The output of the microphone required for com-

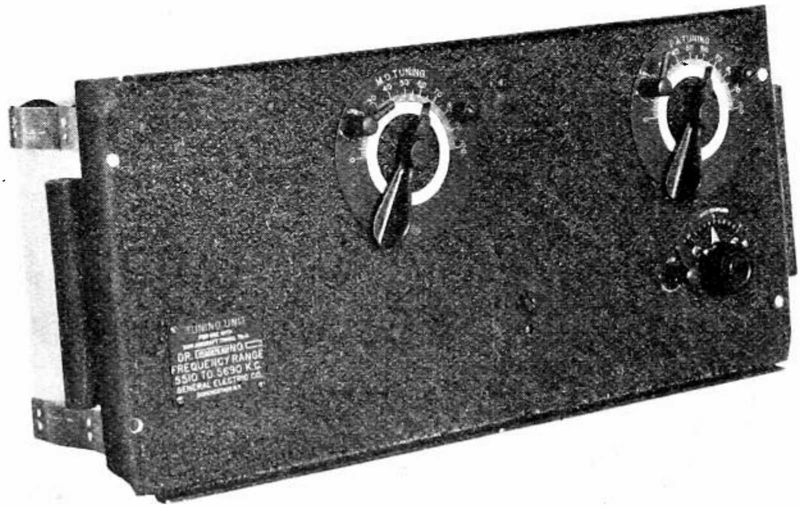

Fig. 6—Tuning unit for RT-76-A transmitter.

plete modulation of the transmitter is 0.25 volt. Fig. 4 shows the over-all frequency response of the RT-76-A transmitter.

In practice, the transmitter may be located in any convenient position in the airplane. When set on frequency, it does not have to be touched, except for inspection or resetting to a different frequency. Control of the transmitter is provided by the control box shown in the complete equipment photograph, Fig. 7. This control box contains the dynamotor "Start-Stop" switch with a telltale indicating lamp, and a "cw-phone" switch. This latter switch, when in the "cw" position, cuts out the modulators, and allows the transmitter to be operated for cw telegraphy. For cw telegraphy, the carrier output power may be raised to 75 watts without exceeding the rated input to the transmitter.

The power drain of the equipment when operating on telephone with a 50-watt carrier output is:

275 to 300 milliamperes at 1000 volts direct current
15 amperes at 12 volts direct current

Since the class B system of modulation is used, the total plate-current drain varies with modulation. The above values are given for average conditions. Assuming a dynamotor efficiency of 60 per cent, the total drain for the complete equipment is approximately 54 amperes from the airplane's 12-volt storage battery.

Fig. 8 shows the general appearance of the transmitter unit ready for use, and Fig. 5, the same transmitter with the top shield of the front panel removed to provide access to the tubes, and with the tuning unit withdrawn from its compartment in the bottom section. It can be

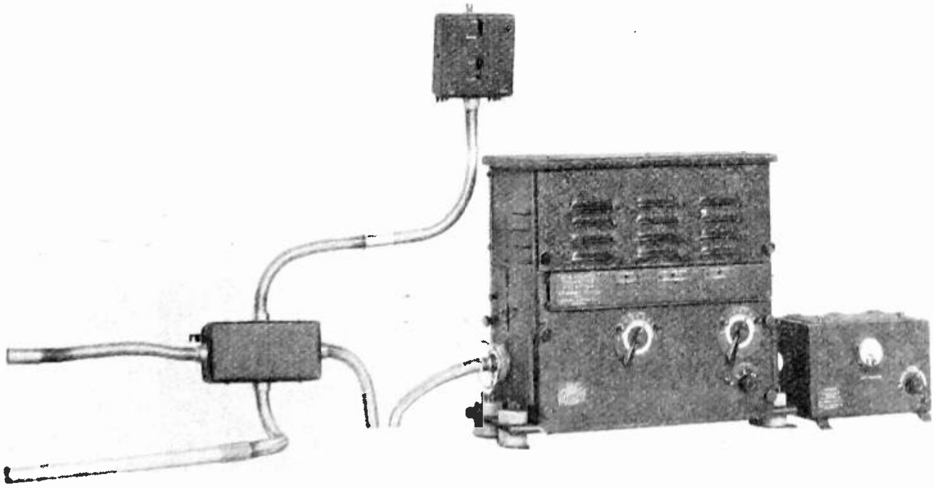

Fig. 7—Complete aircraft transmitter equipment type RT-76-A.

seen from this picture that in general construction, this transmitter consists of a rectangular skeleton frame made of chrome-molybdenum steel tubing of the kind generally used for construction of the fuselage structure in the majority of commercial airplanes. The front and rear bottom tubes are extended beyond the confines of the transmitter to afford convenient means for mounting the transmitter on conventional shock absorbers of sponge rubber. The completed frame is cadmium plated to protect it from corrosion. The center duralumin strip fastened to the front of the frame affords support to some of the fixed circuit elements, and to the tube shelf. The remainder of the circuit elements are mounted on the vertical duralumin shield extending from top to bottom of the transmitter unit, and separating it into front and rear compartments.

The plug-in tuning unit is completely encased by a shielding box.

This permits storage of the unused tuning systems without danger of damage to them, and also serves as a protection against dirt and dust which might cause electrical breakdowns. The shielding box is readily removable, exposing the circuit elements for inspection. As can be seen from Fig. 6, the front of the tuning unit serves as a part of the transmitter panel when in place.

Since all of the connections to the tuning unit are carried through plugs which engage with the jack elements mounted in the transmitter, the change in the frequency band is accomplished without changing a

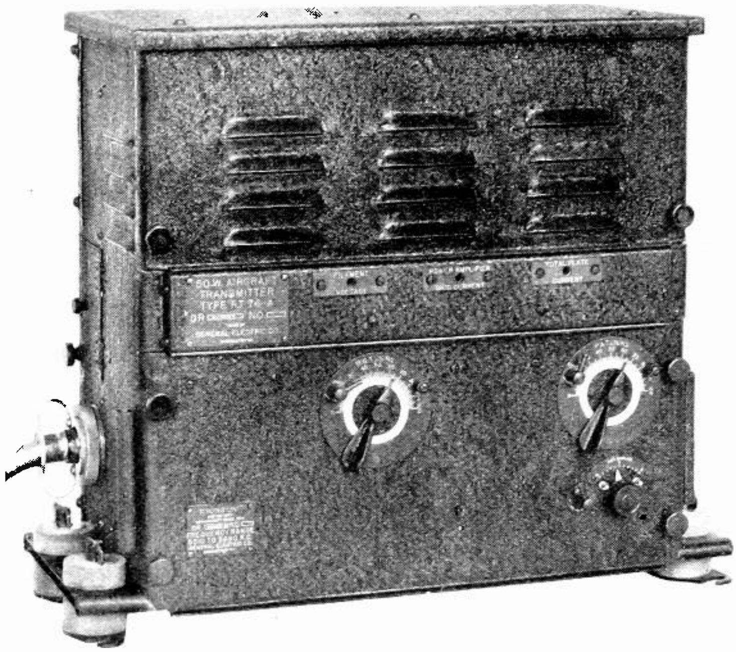

Fig. 8—Transmitter, type RT-76-A.

single wire connection. Guides provided in the bottom of the compartment in which the tuning unit is inserted secure proper alignment between the plug and jack elements, and add to the facility with which a quick interchange of tuning units can be accomplished.

The circuit fuses used in the transmitter are easily replaceable through the door in the left side of the transmitter, and the relay contacts may be inspected through the door on the right side of the unit.

To mount the transmitter in an airplane, four especially designed studs are provided. They are fastened permanently to the supporting framework in the airplane, and need not be removed when the transmitter is taken out for inspection or servicing. To do so, it is only necessary to disconnect the antenna and ground leads, remove the two connection plugs, and free the rubber mounts by depressing the rubber and throwing the retaining crossbars hinged in the ends of the studs into vertical position. After this, the unit can be freely lifted out of its place in the airplane.

The skeleton frame of the transmitter is enclosed on all sides by quickly detachable aluminum shields, and as a result of this type of construction, a complete accessibility of all parts is secured, making for unusually convenient servicing of the unit. As can be seen from the illustrations, the side shields have louvres and the top has overhanging edges, concealing the space between the frame and the top. This arrangement provides an adequate ventilation for internal parts of the transmitter, and yet affords protection against splash or rain so that the transmitter can be used in airplanes of the open cockpit type.

The cable outlets from all of the component units which this equipment comprises are furnished with a length of flexible braided tubing, electrically connected to the cases and grounded. This flexible braided tubing makes it possible to arrange the interconnecting wires in straight-line runs between the units and to enclose them in varying lengths of aluminum tubing. The ends of the flexible tubes are slipped over the aluminum tubing to provide an uninterrupted electrical shielding harness over all of the wires.

The total weight of the equipment described including the interconnecting shielded wires, plugs, dynamotor, control and junction boxes, and the transmitter with antenna tuning unit is under 90 pounds. The weight of the transmitter unit itself is 37 pounds and its over-all dimensions are $20\frac{1}{4}$ inches $\times 17\frac{3}{4}$ inches by $8\frac{3}{4}$ inches.

PLANNING THE NBC STUDIOS FOR RADIO CITY*

By

O. B. Hanson

(National Broadcasting Company, Inc., New York City)

During the past year much publicity has been given to Radio City so that the public is already informed in a general way about this project and the aspirations of its founders. It is so planned that when completed it will occupy three blocks, from Forty-eighth to Fifty-first Streets, between Fifth and Sixth Avenues, New York City, and will house an entertainment, musical, and radio broadcast center of unsurpassed size and grandeur, widely proclaimed as Radio City because of the predominance of radio activities in the center.

From time to time there have appeared in various periodicals, plot plans and architectural renderings of the project, some of which are included here, to refresh the memory of the reader. Fig. 1 is an architectural rendering of the project as a whole, as viewed from above Fifty-first Street and Fifth Avenue, and clearly shows the relative positions of the buildings.

This paper is chiefly concerned with the Central Tower Building, seventy stories and 836 feet in height, designed to house the broadcast studios and offices of the National Broadcasting Company. The architectural rendering of this building alone, as viewed from Forty-ninth Street and Fifth Avenue, is shown in Fig. 2. It contains 1,978,000 square feet of rentable area. For convenience of reference it is referred to on the architects plans as Building No. 1, or the Broadcast Building. Due to the large spans necessary in a building designed to house broadcast studios it will be obvious that it is impractical to support such a tower seventy stories high directly over the broadcast studios. Therefore a section of the building devoted to studios has been designed in the lower floors just back of the tower, a maximum height of twelve stories. This is clearly evident in Fig. 3, an aërial view of the Broadcast Building from Sixth Avenue and Forty-eighth Street. By reference to Fig. 4 it will be seen how the studios have been placed to avoid interference with this mass of steel which supports the weight of the tower, shown by the heavy columns. The tower section is primarily the office section of the building and the NBC offices are to be

* Decimal classification: R612.1. Original manuscript received by the Institute, April 16, 1932. Presented before New York meeting, March 2, 1932.

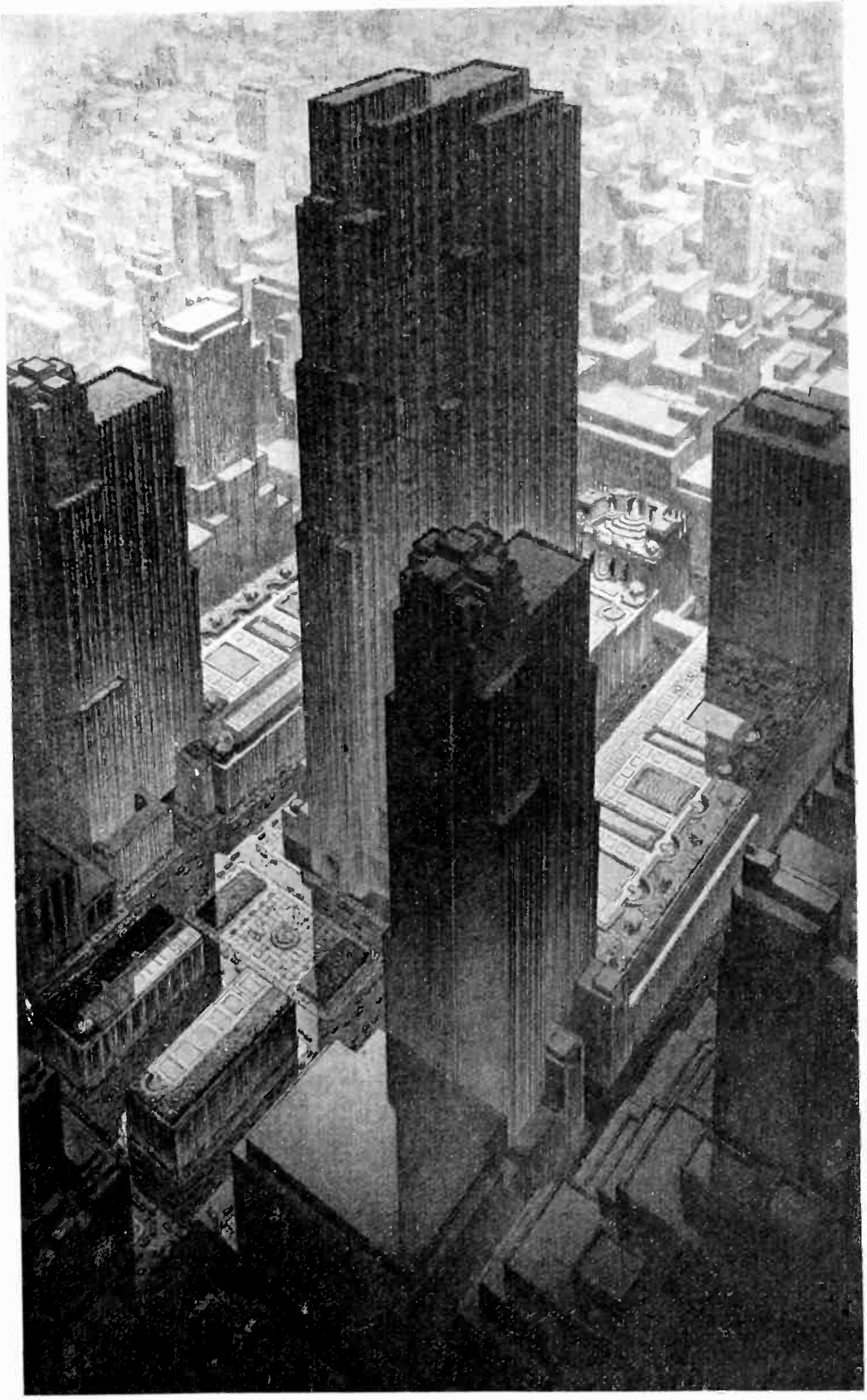

Fig. 1

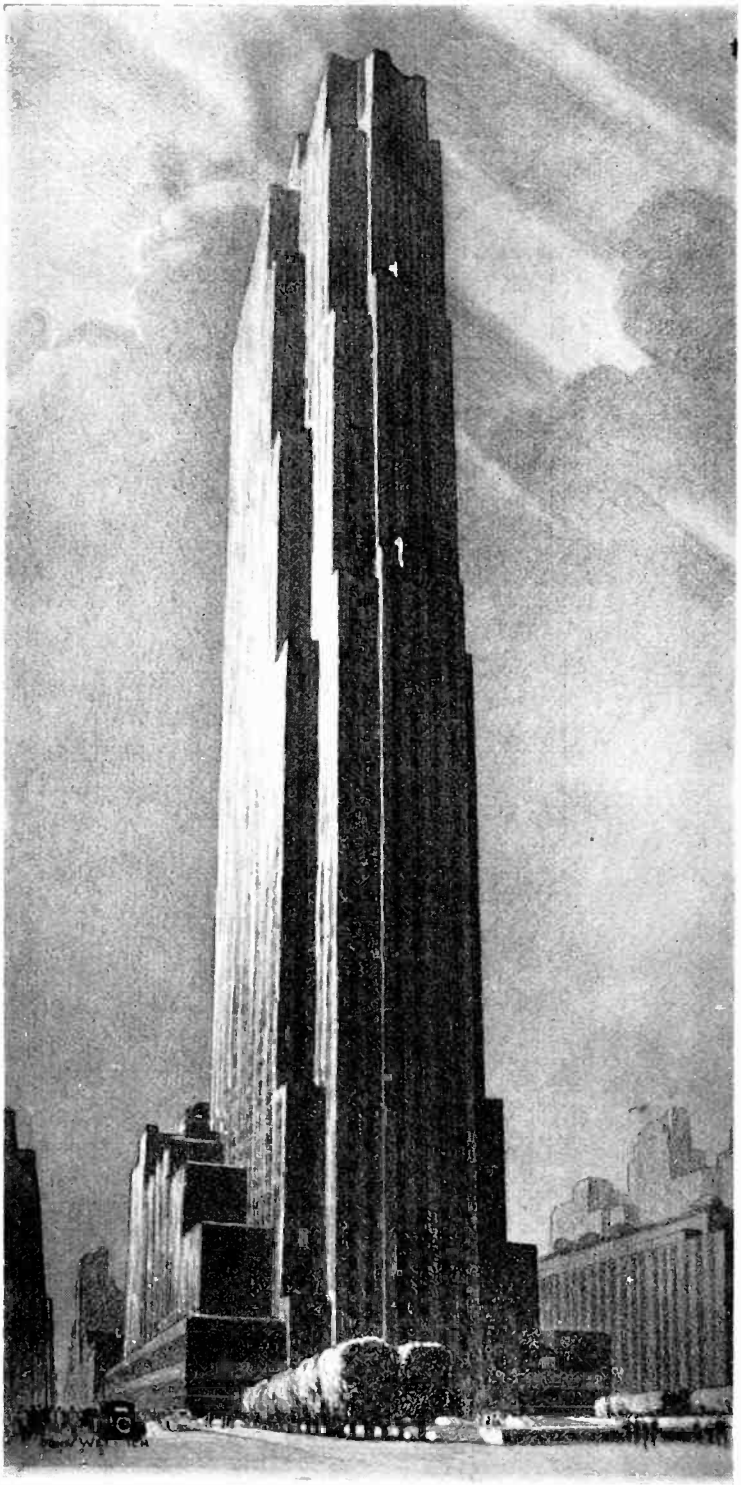

Fig. 2

located on the lower floors (fourth to eighth), adjacent to the studio section and served by the regular office elevators.

In planning this project, it has been estimated that 27 studios, and 5 audition rooms and other appurtenant rooms will be required, and that to house these studios and offices, approximately 500,000 square

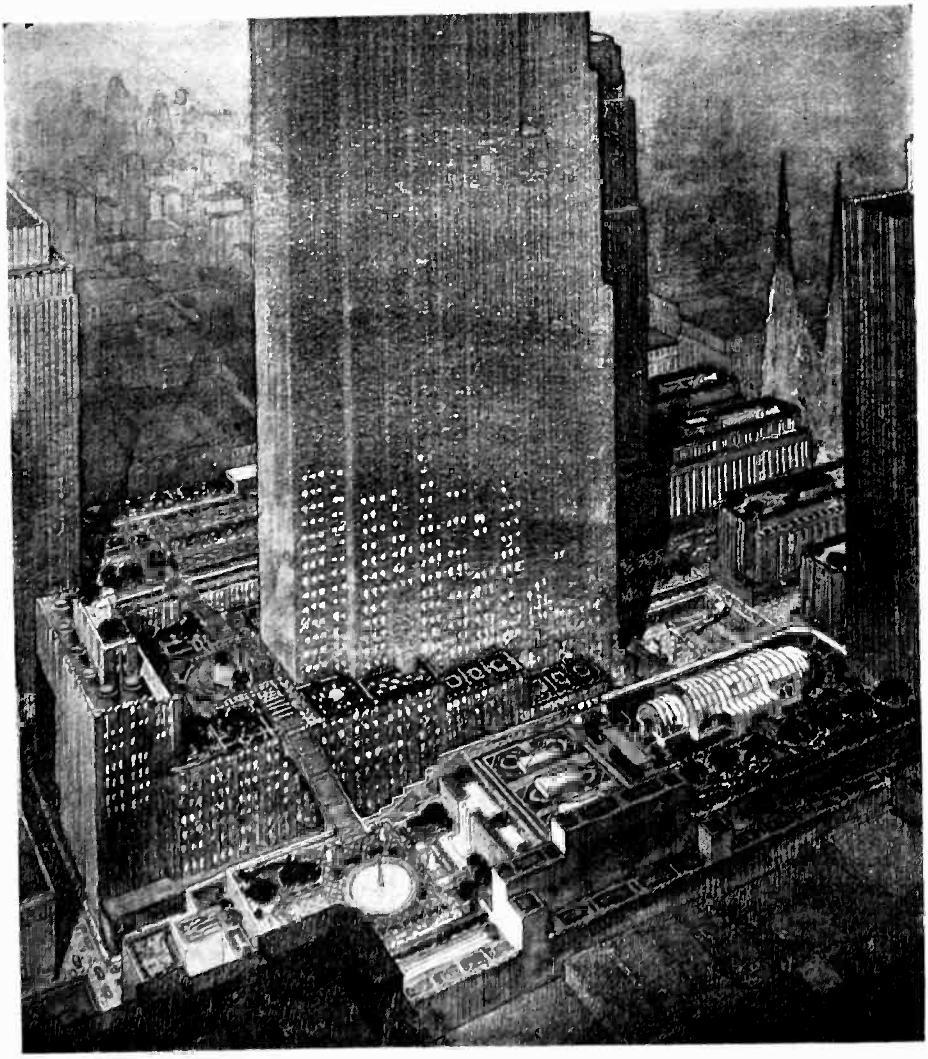

Fig. 3

feet would be needed and approximately 370,000 required for the studio section alone. In the designing of this layout, certain fundamental principles underlying the engineering and traffic problems of broadcast studios have been adhered to. They will not be enlarged upon here, as they are covered in a previous paper by the writer.[1] The

[1] O. B. Hanson and R. M. Morris, "The design and construction of broadcast studios," PROC. I.R.E., vol. 19, pp. 17–34; January, (1931).

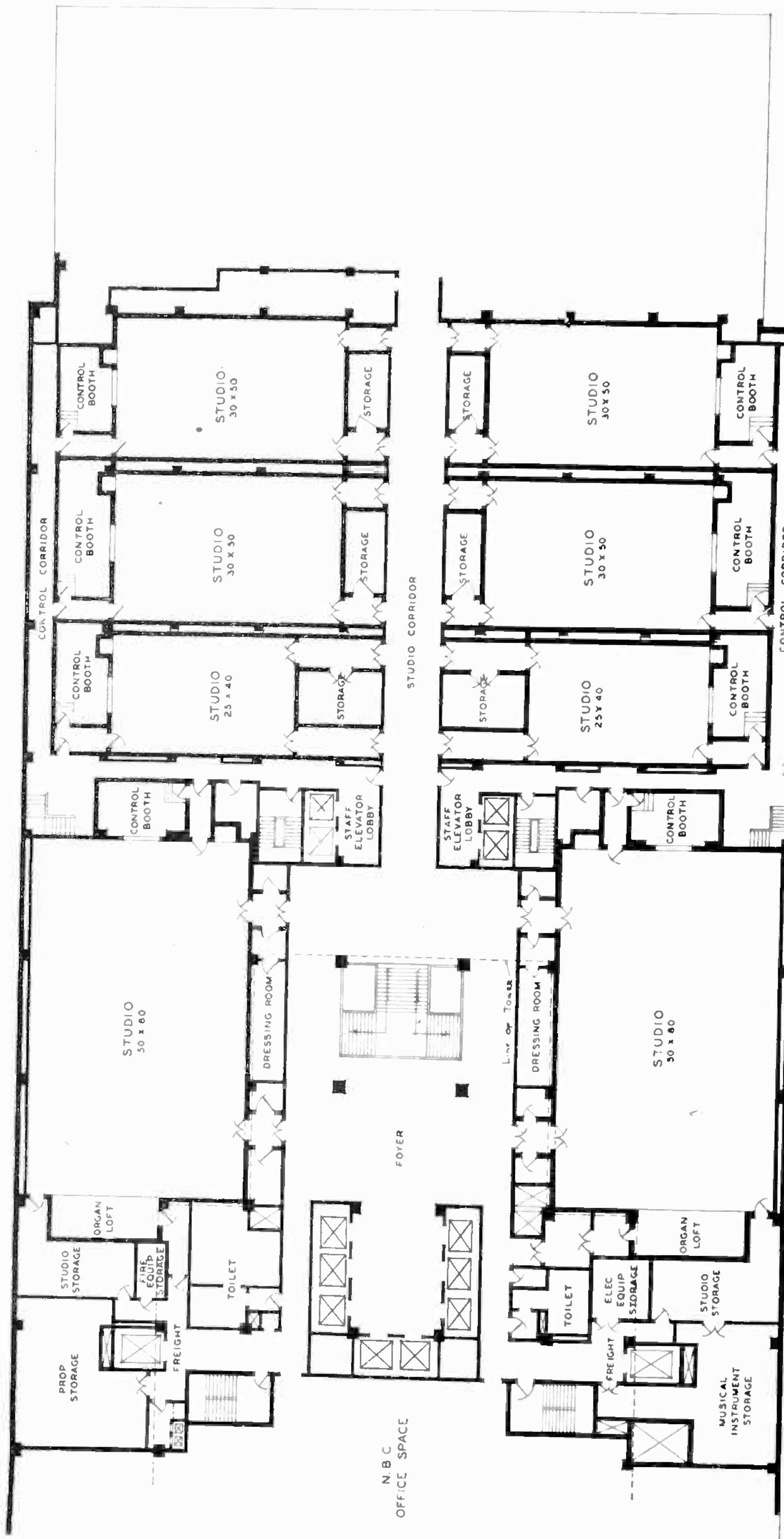

Fig. 4—Typical studio floor plan, National Broadcasting Company—Radio City. Fourth and seventh floors.

method of centralized control referred to in that paper has been applied to this project. In all previous studio layouts the studios were located on a single floor so that it has been necessary to modify this principle in its application here, for in this instance the studios are located on three different levels; namely, the 4th, 7th, and 9th floors, the whole project occupying from the 3rd to the 11th floors in the studio section of Building No. 1. Fig. 5 is a longitudinal cross section through the building, showing the locations of the studio floors with respect to each other and with respect to the main control room, the latter being located on the 6th floor, sandwiched between the two main studio floors.

It will be observed that all the studios on the 4th and 7th floors are two stories in height and that on the 9th floor there is one studio planned to be the largest in the world, which will be three stories in height. By reference to Fig. 4, showing the fourth floor or lower studio bank which is again duplicated on the seventh floor, it will be observed that the studios are symmetrically placed and that the entrances thereto open into a large foyer which provides the communications for the artists, performers, and musicians. The main lobby here and the main studio elevators occupy that part of the building which falls directly beneath the mass of the tower, as within this area the steel is necessarily heavy and no large spans can be provided. The building directly over the studios themselves is but twelve stories in height.

Heretofore in designing studio layouts wherein the studios were all on the same level, the engineers and production staff, together with the control rooms and main control room have been centrally located with the studios around them, and the artists' approach to the studios on their extreme outer side. In this project it was impossible to carry out that scheme as the main approach to the studios was centrally located itself. However, the plan has been carried out vertically rather than horizontally. By reference again to Fig. 5, it will be noticed that the central main control room, the main apparatus and equipment room, power supply, and offices of the operating staff, production staff, and traffic department have been centrally located on the sixth floor. Ready access can be obtained from this floor to the three studio floors by means of private elevators operated for the sole use of the operating and production staffs. On each of the typical studio floors, shown in Fig. 4, these elevators open into private corridors which communicate directly with the individual control and monitoring booths of each studio. They are so laid out that at no time is it necessary for any of the production staff to enter upon the communicating corridors

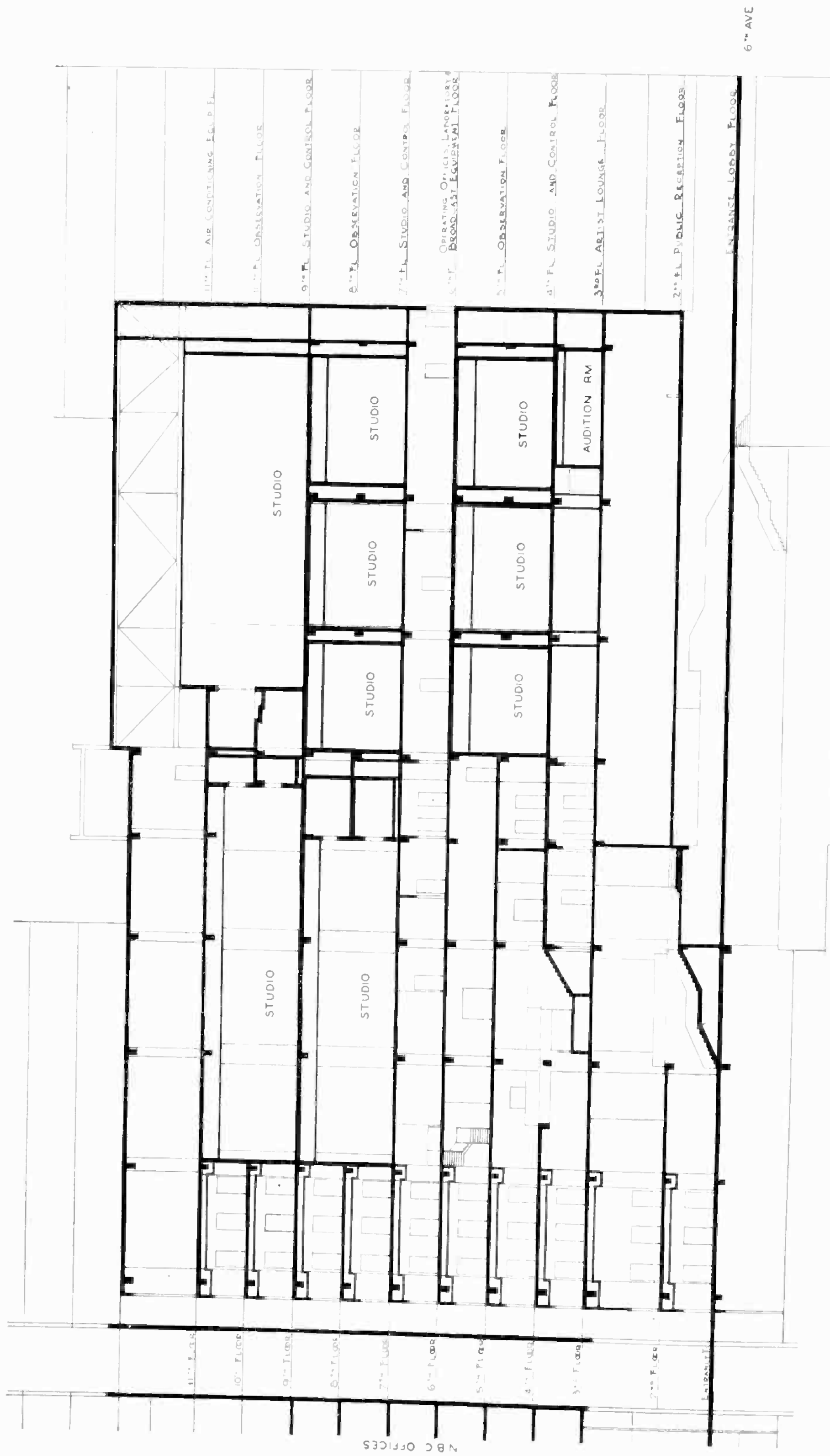

Fig. 5—Longitudinal section of NBC studio building—Radio City.

and the lobbies provided exclusively for the artists, performers, and musicians. It will be further noted that artists and performers coming from the main studio elevators can, at a glance, see the entrances to all studios on the floor and that a point of control for each floor is placed directly in front of the elevator lobby. The musicians and performers thus have ready ingress and egress from one studio to another through the main foyer, without crossing the communication corridors of the engineering and production staffs.

It is anticipated from past experience in the handling of musicians and guests that several thousand people will be present in this studio block at the same time, and for this reason a serious traffic problem must necessarily be solved in the planning of this layout.

A logical question no doubt arises as to why so many studios are required for the operation of a dual network broadcast system such as is now operated by the NBC when it is necessary to keep only two programs on the air simultaneously. Our statistics show that on the average, four hours of rehearsal are required for every hour of actual broadcasting. This means that when one broadcast studio is on the air, four others are being occupied by rehearsals, and a sixth is in preparation for the following program. However, as productions vary in size and type, a variety of different sized studios must be provided to meet adequately the needs of the various productions. Further, when dual channel operation is encountered, a minimum of twelve studios must be provided, a few extra being desirable in order to permit a choice of size to suit simultaneous broadcasts. It is necessary to provide three or four audition studios which are in continuous operation by the program staff in their search for and selection of new talent. In our present quarters at 711 Fifth Avenue there are nine studios and three audition rooms, together with two small rooms used for the purpose of monitoring outside pick-ups, out-of-town and foreign broadcasts, which gives the equivalent in apparatus of a fourteen-studio layout for the operation of two channels, and even under these circumstances, operations are hampered because of a lack of choice of proper sized studios to fit all cases simultaneously. In network broadcasting it is frequently necessary to split networks and transmit four programs simultaneously, which further complicates the studio problem. With this explanation, and allowing for the natural growth of the industry, it can be readily appreciated that the 27 studios planned for Radio City are not too numerous, as might appear at first.

In addition to the needs of musicians and performers, who will be passing in and out of the building, it is planned to care for the clients and the NBC's guests adequately. During the past decade, a considerable proportion of the radio audience have visited the studios of their local broadcast stations, and large numbers have visited the studios of the NBC in the several cities in which we operate. This is a natural interest and curiosity on the part of our listeners who through their radio sets act as hosts in their homes to our artists and performers, and it is only reasonable that they expect us to reciprocate, and we therefore gladly welcome their visits to our studios. In Radio City, considerable space is being provided for adequately receiving and entertaining these visitors.

By reference to Fig. 5 it will be noticed that on floors 2, 5, 6, 8, and 10, public reception rooms and observation galleries are provided in connection with practically every studio. In the larger studios, the observation galleries are planned with theater seats so that the guests may be comfortable while watching the progress of a broadcast production. Fig. 6 shows a typical guest floor and the locations of these observation galleries with respect to the studios. In nearly all cases these observation galleries are separated from the studios by sound-insulated transparent partitions but in some instances the galleries are opened into the studios. It will be noticed also on this floor, which is typical of the fifth and eighth, that a private observation room is provided for the sponsor of a program and his staff of assistants, critics, and guests. It will be noted also that every available inch of space has been put to use and that the plan is dovetailed together in a symmetrical manner, which indicates that the problem has been thoroughly studied so that no space is wasted. As many of our guests are naturally interested in the technical end of broadcasting, an observation room has been provided on the sixth floor, directly in front of the main elevator lobby and separated from the main control room by plate glass windows so that these visitors may see the technical apparatus and the staff in operation.

In addition to the large studios shown on the fourth, seventh, and ninth floors, there are a number of small studios on the ninth floor, (Fig. 7), especially designed to handle speakers and small productions. In view of the increasing number of child artists a special reception and lounge room has been provided for them on the ninth floor, together with a suitably arranged studio for children's productions. Studio 9-H, the largest studio of the group, has floor dimensions of 80×130 feet, with an average ceiling height of 35 feet, and is equipped with a balcony on the tenth floor which will seat some three hundred visitors. The balconies in this case open into the studio. Studio 9-G, on the ninth floor, is the second largest studio and is being constructed for the presentation of dramatic and other productions, which require

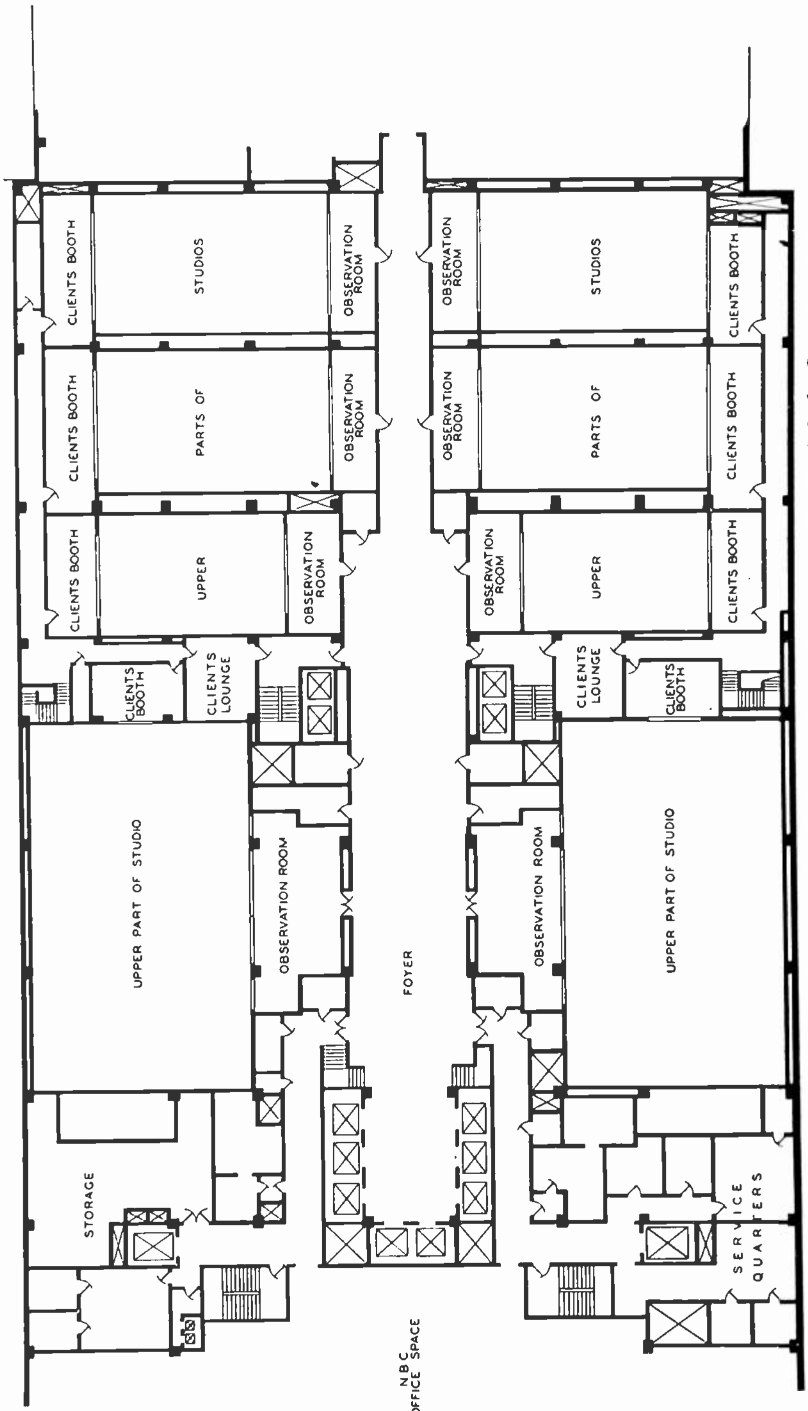

Fig. 6—Typical observation floor plan. Fifth and eighth floors.

a stage and a local audience. If, and when, television reaches a point where its operation is comparable to the making of motion pictures, these two studios would prove invaluable because of the large floor area available.

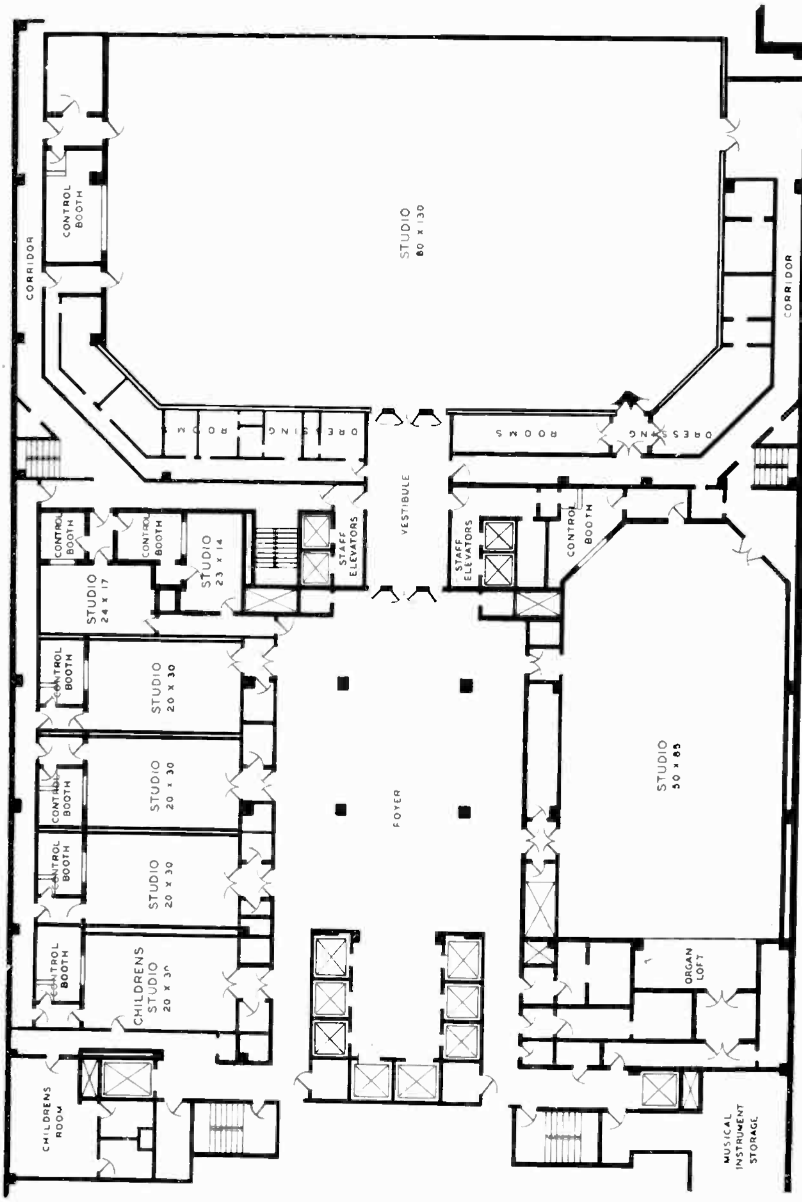

Fig. 7—Studio layout. Ninth floor.

On the tenth floor (Fig. 8) there are provided four studios grouped around a central control room for the purpose of arranging dramatic productions for simple television broadcasting. The various scenes can be set up in these studios, arranged in proper sequence so that the television camera or scanner located in the control room can be rotated and focused on any one of the four studios, in order. This group of studios can also be used for the production of drama without vision, wherein it is necessary to place the orchestra in one studio, the principal actors in another, crowd scenes in another, and sound effects in the

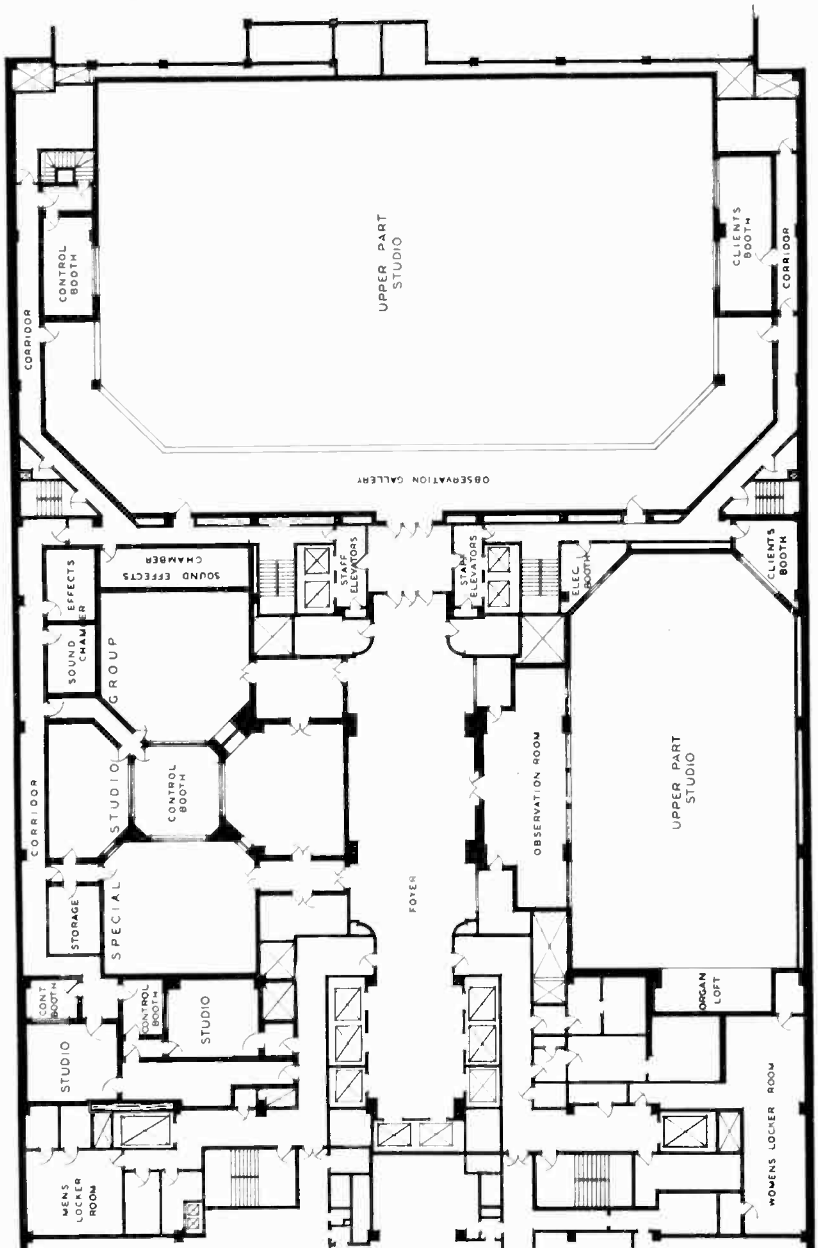

Fig. 8—Studio layout. Tenth floor.

fourth, mixing these four pick-ups electrically in the common control booth to obtain the desired effect. This set-up is not necessary in all productions but it was felt that one such group should be provided to handle certain complicated productions.

Paramount in the design of all studios is the sound insulation, and a system of soundproofing similar to that which has been used in the New York and Chicago studios is planned, that is to say the walls, floors, and ceiling of each studio will be floated free from the building structure. This will also apply in the case of control rooms and other appurtenant rooms. Without such an insulating system it would be impractical to build studios in steel buildings as the steel framework provides an excellent transmission system for sound and mechanical vibrations.

In the construction of broadcast studios in Europe the engineers have barred the use of steel framework in the sections of their buildings which are to house studios, because of the difficulty of sound control where steel is involved as a framework. The British and Germans have both been careful in this respect. In this country, however, the problem is forced upon us and we cannot avoid it. It has been successfully solved by complete isolation of each studio. With such construction, sound attenuation of the order of 60 db, is readily obtainable, and as each studio is similarly isolated, an attenuation of 100 db or more between studio units can be obtained. No suitable apparatus is yet available that will measure accurately attenuation of this order with sound levels normally attained in the studios.

Second in importance to sound insulation is air conditioning, without which windowless studios would be inoperative. The air conditioning plant designed for the NBC studios when complete will probably be the largest and most complicated plant in the world and will cost something like a million dollars. Obviously, special precautions will have to be taken with this air conditioning plant to prevent the transmission of sound from one studio to another through ventilating ducts and the control of noise generated in the ventilating machinery itself. Special streamline air supply outlets are being designed to reduce the air rush in the studios, the total noise from the air conditioning to be less than a sound level of plus 8 db. In some instances the air will be completely changed in the studios eight times an hour. This calls for the handling of large volumes of conditioned air, and complicates the problem of noise from air rush. Those familiar with air conditioning plants involving refrigeration, dehumidification, cooling, individual temperature, and air flow control, etc., will have an idea of the complexity of this problem.

In addition to the foregoing, the problems of decoration to obtain proper psychological effect on the performers are being studied at the present time together with the coördinating of this decoration with the acoustical treatment of the studios. Inasmuch as this study is still in process, little can be said regarding it in this paper. It is planned, however, to provide acoustical treatment in such form that the time period of a studio may be varied at the will of the studio staff by the control of the mechanical operation of acoustical units from a switchboard in the monitoring booth. In anticipation of television, all of these studios will be electrically shielded and provided with suitable lighting facilities, to illuminate scenery for the proper operation of television cameras. As we have not yet been confronted with the actual operation of television on a large scale, the exact form which some of these facilities will take is a matter of conjecture at this time.

In this paper nothing has been said about the technical apparatus and the technical facilities provided from the microphone to the listener. Although the elements of this system are in the planning now, nothing of real value can be offered concerning them at this time. This matter will form the basis for a future paper.

In conclusion, the writer wishes to say that the work already done on this project is the outcome of more than a year's efforts on the part of the engineering staff of the NBC, in cooperation with the architects and engineers of the Rockefeller Center. We have reason to believe that our combined efforts are successfully handling the problems as they arise, and we look forward to the completion of the project with great enthusiasm.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

INVESTIGATIONS ON GAS-FILLED CATHODE RAY TUBES*

By

MANFRED VON ARDENNE

(Berlin-Lichterfelde-Ost, Germany)

Summary—*In considering the function of initial concentration and ray focusing it is found, for tubes using slow electrons (up to 4000 volts), that both a Wehnelt cylinder and a molecular concentrator are necessary. The action of molecular concentration has been studied thoroughly, and the static relation between pressure, beam current, and anode potentials are explained. The use of plate potentials higher than about 5000 volts was found to be unsuitable. Special phenomena accompanying such concentration are taken up, including the effect of ionic oscillations and the problem of lack of sharpness with high-frequency deflections. A remedy for the former was found in a grounded metallic outside coating, and for the latter an improvement was found by filling with light gas (hydrogen, helium). The processes in the gas space and the deflection of electrons were then investigated. The various methods for reducing secondary illumination are described. In conclusion, the reaction of the Braun tube upon the circuit was taken up and the avoidance of distortion and origin displacements by means of circuit precautions are described.*

THE conditions necessary for the formation of a very narrow beam of cathode rays in oscillograph tubes with incandescent cathodes have been studied. It is necessary to have initial concentration of the diverging rays at the cathode, and also concentration of the stream of electrons along the entire path that they follow between cathode and screen.

The initial concentration can be done in various ways: from the practical viewpoint it can be done by using a supplementary diaphragm, or by an electrostatic method. The use of diaphragms has the disadvantage of causing a large current loss, as only a small part of the emitted electrons pass through, most of them being caught by the diaphragm and thus must be considered as lost. In gas-filled tubes this current loss has a minimum value that depends on the geometric arrangement of the electrodes and other factors. If potentials are selected so that this minimum loss is not exceeded, the arrangement is ineffective in producing concentration of the ray of electrons passing through the diaphragm. In order to cause preliminary concentration in this manner it is therefore necessary to use high-powered anodes and cathodes.

The other method of initial concentration, by electrostatic focusing, as is done in the classical form of the concave reflecting cathode

* Decimal classification: R388. Original manuscript received by the Institute, October 14, 1931. Translation received by the Institute, April 8, 1932.

and in the Wehnelt cylinder, has the decided advantage that the emission from the entire effective cathode surface is utilized, and there is no additional loss in power because of captured electrons. This is important because it makes simple apparatus possible and involves no danger in connecting to the supply line. In order to show this effect the paths of the streams of electrons have been made visible by gas excitation in the accompanying illustrations. Fig. 1 shows, as an example, an unconcentrated group of cathode rays such as is obtained with a cathode consisting almost of a point, if there is no electrostatic concentration. Any part of the cathode surface that emits somewhat more strongly, produces a component group of cathode rays that

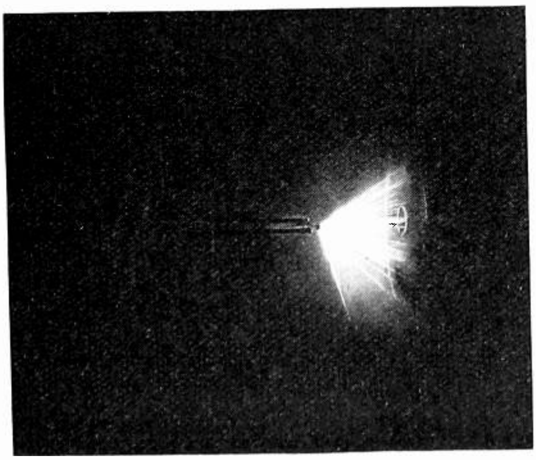

Fig. 1—Beam formation without a Wehnelt cylinder.

leaves the cathode in the normal direction. The result is a cone of radiation from which the anode screens out and passes only the central component rays. By using a concentric cylinder with a negative bias, all components are forced to the center, and the grouping can be improved as desired. Fig. 2 shows a partial concentration that is still insufficient and is produced by a slightly negative bias. Fig. 3 shows ideal focusing with a stronger bias, in which the entire primary stream of rays passes through the orifice to the anode without a loss by screening out. In gas-filled tubes with incandescent cathodes, this method of preliminary concentration has the advantage that the negative cylinder serves to attract the ions, and thus it is possible to use higher anode potentials without greatly reducing the life of the oxide surface on the cathode.

Preliminary concentration between cathode and anode alone is not sufficient. If no concentrating forces act later on the individual beams in the group, there is a dispersion because of the mutual repulsive action of the negative charges. Beam concentration theoretically can take place as a result of external agents, possibly an intense electric or magnetic field, or by spontaneous concentration of the beam electrons, as can be observed with very high velocities or in the presence of gas residues. External, coaxial, magnetic concentrating fields are generally used with high tension oscillographs, especially since the work of Du-

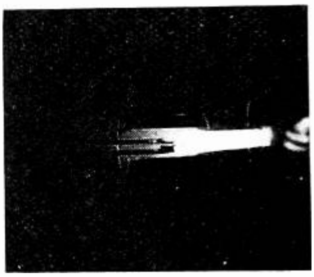

Fig. 2—Electrostatic concentration with a weak bias.

four and Rogowski. As is known, the concentrating coils in the cathode ray of H. Busch[1] throws an optical picture of the cathode on the screen. In low-tension oscillographs, the concentrating coil offers practically no advantages; its operation cannot be explained without a certain spontaneous concentration of an electrodynamic nature, and this is vanishingly small at low potentials. Electrostatic fields can hardly be

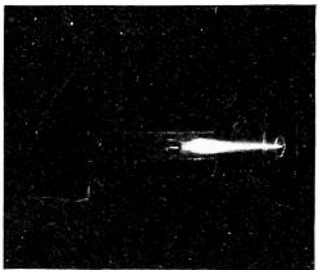

Fig. 3—Electrostatic concentration with strong negative cylinder bias.

used to concentrate the beam. In long negatively charged cylinders the cathode beams undergo a not inconsiderable breaking action.

Spontaneous concentration can take place with very high velocities by electrodynamic attraction of the convection currents. In high-tension oscillographs the beam is stable even at pressures[2] below 10^{-5} mm. With slow speed cathode rays (up to about 4000 volts)[3] this effect

[1] H. Busch, *Annalen der Physik*, vol. 81, p. 974, (1926).
[2] *See* E. Sommerfeld, *Arbeiten Inst. Aachen*, vol. 3, p. 107, (1928).
[3] Editorial note: The velocity of electrons due to potentials of these values applied to the anode or other accelerating electrodes.

is no longer sufficient since with these velocities there is mutual repulsion of the beam electrons and it is necessary to have a means of overcoming the space charge in the beam, such as that produced by the presence of gas residues[4] at a pressure of about 10^{-3} mm.

Gas Concentration and Its Limits

The phenomenon of gas concentration is shown in Fig. 4. This shows very slow electrons (200 volts)[3] in argon. We see plainly that the beam, as before, is made up of individual component beams. Because of the somewhat more divergent emission angle at the cathode, these component beams are not exactly parallel in their subsequent paths, and nodes and loops form. This phenomenon is rarely observed at

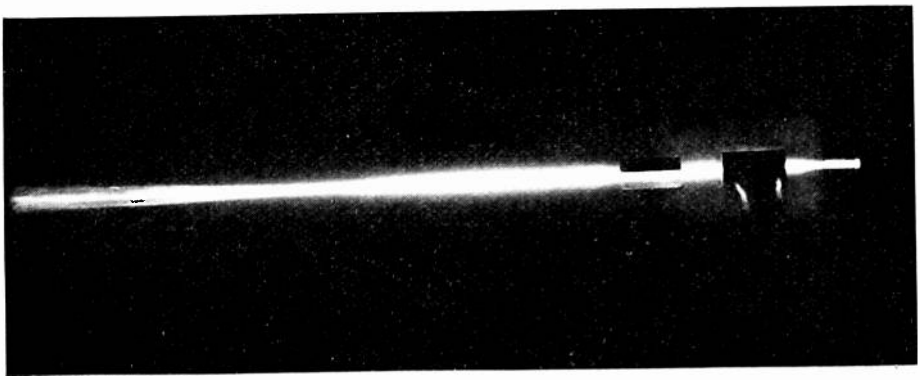

Fig. 4—200-volt beam in inert gas under 10^{-3}-mm pressure.

greater velocities. However, there is a certain natural lack of clearness (a halo) at very low velocities, and if the pressure is decreased the beam stops abruptly in the middle of the gas space.

The explanation of the gas concentration will be mentioned briefly. Johnson[4] assumes that the inert positive ions, the final product of the gas atoms ionized by electronic impact, remain in the beam path, while the secondary electrons that are formed diffuse from it.

The following observations can be understood from this point of view. At low pressures stronger primary beam currents are necessary, and conversely. The beam velocity influences the concentration in a complicated manner. With very low velocities (up to 300 volts)[3] (and especially at the end of the beam) the primary energy is no longer sufficient for ionization and the ionization stops. At very high anode potentials (above about 1000 volts) the concentration of ions diminishes greatly, so that the ions are forced out of the beam path. The effect

[4] *See* J. B. Johnson, *Phys. Rev.* vol. 17, p. 420, (1920), and *Jour. Opt. Soc. Amer.*, vol. 6, p. 701, (1922).

of the anode potential is approximately constant between these two limits.

The action of the positive and negative space charge densities can be found by measuring the plate characteristic (see below) and also in other ways. It should be about 10. In other respects the ion density fluctuates within wide limits depending on the magnitude of the beam current and pressure. In many cases it is possible to overcome lack of point sharpness resulting from low ionic density, by increasing the beam current by greater heating of the cathode. (For example, see the section below on high-frequency deflection.)

The strong influence of the beam current strength on the concentration is of practical importance in all methods for the modulation of

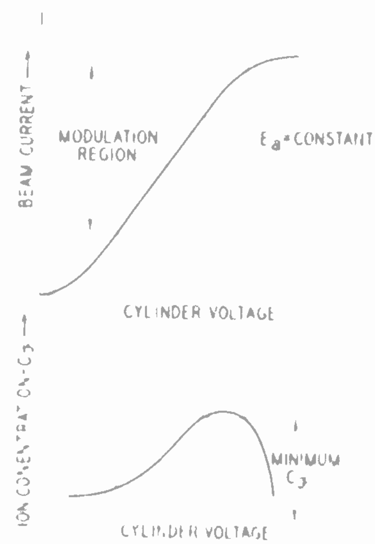

Fig. 5—Wehnelt cylinder luminosity control.

brightness, using a variation in the cathode supply. Conditions for the control of brightness by means of Wehnelt cylinder potential (shown in Fig. 5) are typical of this. The spot brightness, that depends uniformly on the beam current strength, except for extreme values, increases with an increased negative bias potential on the cylinder as shown in Fig. 5a. As the effective cathode area increases at the same time, the ionization also increases constantly. On the other hand, the ion concentration, that is, the number of ions per unit of area, has a maximum because the beam is already diffused if the biases are too small (Fig. 5b). The greatest spot sharpness is only found above a minimum ion density and thus, as shown, the control of the brightness range is greatly restricted. This is particularly true with very high anode po-

[5] M. v. Ardenne, *Fernsehen*, No. 5, (1930).

tentials in which, because of the projection of the ions from the beam path, the concentration is less. The available modulation range therefore becomes smaller and smaller.

According to the above relationships, the use of pure gas concentration without external fields is reserved for low-tension tubes. The maximum limit for the beam velocity depends to a very great extent on the optional pressure of the gas. With hydrogen it is not far above 5000 volts and drops to about 4000 volts with argon.

Abnormalities in Gas Concentration

On working with heated cathode ray tubes using the gas concentration effect and anode potentials of some 1000 volts (constructed in the author's laboratory[6]) two phenomena which are noticed upon deflecting the cathode ray deserve special mention. The effect is shown in Fig. 6. These are ion oscillations. If the cathode ray, such as is used

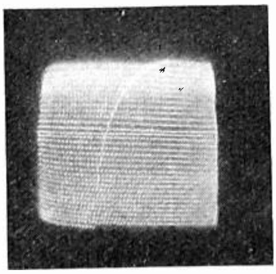

Fig. 6—Line screen disturbed by ion oscillations.

for example in making a line screen in television, is allowed to pass over the surface of the screen so that the velocity in the direction of the lines is a minimum of 100 m/sec., undesirable additional oscillations will appear in the deflections as shown in Fig. 6. These interfering oscillations can be reproduced as soon as the anode potential exceeds about 2000 volts. The frequency of the interfering oscillations is well-defined and changes only slightly with the pressure or current strength. It is of the order of magnitude of 50,000 cycles. The direction of the oscillation amplitudes is always radial from the center of the screen. The beam acts as if the beam velocity were modulated a few per cent at this frequency.

As it has been definitely established that this phenomenon was related to processes in the tube, the following explanation seems logical. The ions with rapid collision electrons ($v > 2000$ volts) make an oscillating movement in planes through the beam, whose frequency must be determined essentially by the kinetic temperature velocity of the

[6] Introduced in U. S. A. by General Radio Company, Cambridge, Mass.

agreement with present ideas as to its method of operation. On the basis of these results and the adaptibility to higher anode potentials, special hydrogen-filled tubes have been made.

Processes in the Gas Space

It is not easy to give information on the current variation in the gas space surrounding the beam. It must be assumed basically that there is a continuous formation of ions and just as many secondary electrons along the beam, while at the same time a number of primary electrons are deflected from the direction of the beam after impact. The secondary electrons are considerably slower and have a tendency to attach themselves to the positive ions. This may result in luminescence of the

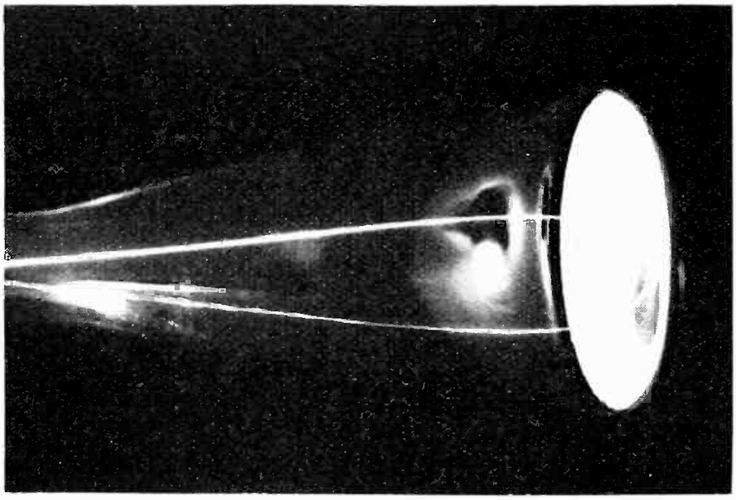

Fig. 10—Simultaneous deflection of stray electrons and beam by a magnetic field.

gas along the path of the beam. An essential and technically important effect of the electron dispersion is the secondary illumination of the screen. Regardless of the location of the beam, when there is a gas charge the entire surface of the screen is illuminated with about 1/100 of the brightness. This secondary illumination can be due only to considerably lower velocities of diffused electrons. An interesting test may be cited in connection with this assumption. Fig. 10 shows a beam path that appeared under the action of a small magnetic pole held in front of the screen. This is a very long time exposure, so that the secondary illumination has sufficient exposure. We see that not only has the beam been pushed away from the surface of the screen, but the electrons causing secondary illumination also have been subjected to the same action because the parts of the screen in the vicinity of the transformation point are obviously dark. From this alone it can be concluded that the secondary electrons do not have a direction much different from that of the beam. They are absent at a sharp angle to the beam. Therefore, the secondary illumination seems to be an effect of the refracted beam electrons. Further considerations agree with experimental results, and show that the ratio of the brightness of secondary illumination to focal illumination is a minimum in heavy gases.

Fluorescent Screen

During the course of the investigations it developed that it is not easy to find a fluorescent screen that combines the following properties; high illumination power, suitable spectrum (photographically or optically active or both), very short secondary illumination time, suitable graduation curve, and absolute resistance to combustion with strong beam currents and also, obviously, the screen must be homogeneous and uniformly dense.

First it was found that the lower part of the graduation curve of illumination is proportionally straighter as the substance forming the fluorescent screen is finer, and as more surface is used. The sensitivity to very slow electrons is then especially high, and the ratio of spot light and secondary light is less good. Particularly for television purposes, the corresponding reduction in the brightness interval is highly undesirable and therefore fluorescent screens have been developed with absorbing surfaces that retard slow electrons to the point where they become ineffective. Secondary illumination can be suppressed almost completely by using such screens and by filling with a heavy gas. Great difficulties were encountered in securing satisfactory heat resistance of the screen. Strong cathode rays cause such a heat development in a minimum space that the fluorescent screen substances can be caused to glow thermally. In the older screens with water glass as the binder, there was combustion, brown coloration, and fatigue and exhaustion of the illuminating power. Sodium salts were blamed for the discoloration, as they were reduced in the cathode ray with the precipitation of sodium. By using a different method of preparation it is now possible to obtain a fluorescent screen that will resist a stationary 3000-volt beam without discoloration and, because of better conduction of heat from the crystals, also without essential fatigue phenomena. About the luminous spot there is a kind of light halo caused by repulsion of the beam electrons resulting from the strong screen charge. With very low velocities of about 300 volts the retarding action of these charges may become so great that the beam no longer has sufficient energy when it reaches the screen. This case is shown in Fig. 11. However, removing the screen charges in low voltage tubes does not offer sufficient advantages to justify the increased difficulties and expense in manufacture. The slight increase in the point brightness that can be obtained by conduction will be more than counterbalanced by the reduced optical transparency that is unavoidable when placing a conducting layer on the screen.

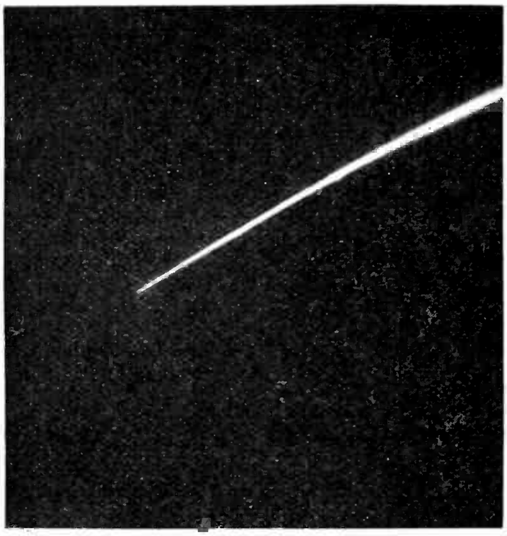

FIG. 11—Retardation of a slow (100-volt) cathode ray by screen charges.

THE BRAUN TUBE IN THE CIRCUIT

One of the great advantages of the Braun tube is that an extremely small control force is sufficient to deflect the beam. But it is already known that although the power consumption is small, it does not re-

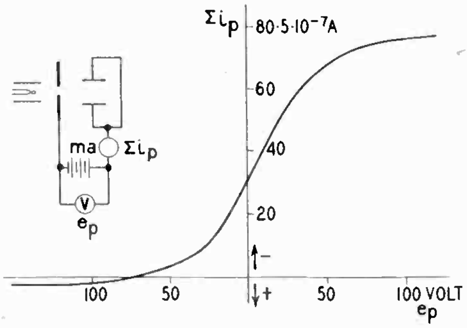

FIG. 12—Total plate current and plate bias.

main constant with the different positions of the beam. If the Braun tube is to be used as a measuring instrument in connection with generators with high internal resistance, this property of the variable resistance certainly must be taken into consideration. In the following we shall give some measurements of the characteristics in the plate circuit of the author's tubes.

In Fig. 12 the plate circuit is short-circuited and we measured the total current i_p, which is received by the plate if there is a bias e_p with respect to the beam. At about -100 volts on the plates the ion currents stop. At $+100$ volts the diffusion current of the beam electrons, that are drawn to the plates, is saturated. The electron current for control purposes, as we see, is about 100 times as large as the positive current (see above).

The natural characteristic of the plate circuit is shown in Fig. 13 in the connection that is most commonly used, in which one plate is connected to the grounded anode and the other plate is brought to the measuring potential as compared with the former. The three curves

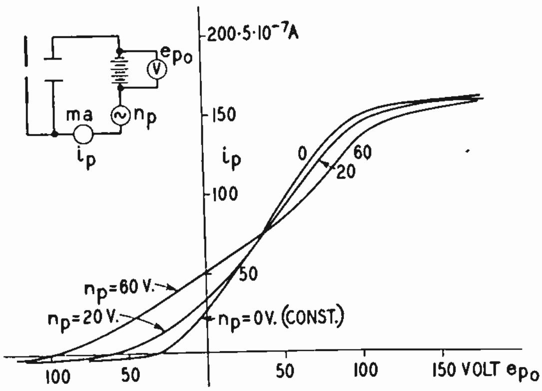

Fig. 13—Directional characteristic of the current across the deflection plates.

represent directional characteristics i_p (e_p) obtained with an additional alternating potential (e_p) at 0, 20, and 60 volts. The static characteristic, as is seen, shows a pronounced bend with a 0-volt plate potential. This corresponds to the stationary position of the beam. Normally the plate voltage fluctuates in this region and therefore rectification should be expected in the plate circuit which, with high external generator resistances, can cause a displacement of the origin in the direction of deviation. This source of error must be taken into consideration when making amplitude measurements or plotting oscillation center points. The external resistance of the generator being tested should not be more than 10^5 ohms, or else the 0 point will be displaced.

Fig. 14 also shows a plate circuit characteristic obtained with an anode potential of 1500 volts. The resistance curve can be constructed from the characteristic in the usual manner, and is plotted in Fig. 13. We see that the tube can be replaced by a pronounced e.m.f. of about 50 volts (natural oscillation of the stray electrons) and a strong variable internal resistance that, with great deviation, can increase to 10^7 ohms but drops to a tenth of this amount in the rest position that is most frequently used. If we use a potential of 100 volts (see Fig. 15) for maximum deflection, we find that a normal cathode ray tube requires a power of about 0.01 watt for deflection. The relation between beam deflection and plate voltage is given by the measurement shown in Fig 15 with the deflection as ordinate and the plate voltage as abscissa Contrary to expectations, the curve does not consist of one straight line but has distinct slopes. Apparently the tube is less sensitive near the origin. It has made a deflection of about 1 cm (at ± 25 volts), and

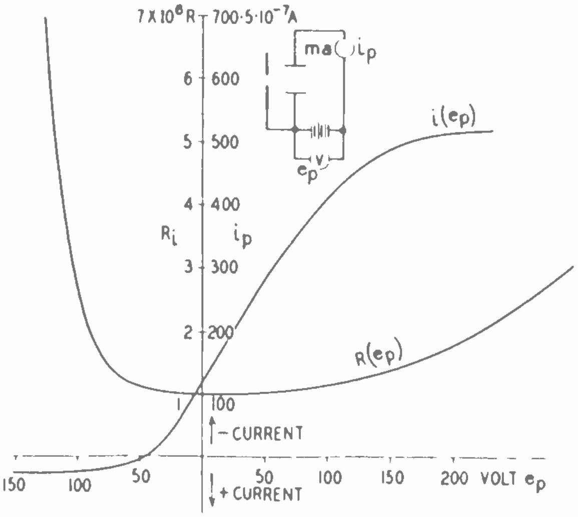

FIG. 14 Resistance of an argon-filled Braun tube

thus it has greater sensitivity following a linear law. It seems logical to explain such action in connection with the circuit. If we assume that the generator has an internal resistance, the direction of the deflection error can be explained using the R_i curve in Fig. 14.

But it is found that even with the extremely strong sources of potential the cathode ray oscillograph retains the characteristic control error shown in Fig. 15, although to a reduced extent. There are always breaking effects on the beam in the vicinity of the origin, that is, there are curve errors. This phenomenon appears particularly clearly if the line screens are to be recorded at constant speed as shown in Fig. 8. The control error then acts as a coordinate intersection that appears brighter than the vicinity in the center of the screen.

This control error seems to be a space-charge effect. The mechanism of the deflection does not act as if a negative beam charge moves in the vacuum between the plates. Rather, the space charge (surrounding the beam), which is very irregular because of great ion density in the vicinity of the anode where the plate is located for opticogeometric reasons, prevents the field strength acting on the beam from being identical with that given by theory. The effective field strength is smaller than that calculated, because of the field screening by the space charge in the center of the plate, and is a minimum there. It seems probable that this explanation is correct because it is possible to reduce greatly the control error if the deflecting plates are given a negative bias as compared with the beam (anode). Diffusion currents to the plates become unnoticeable according to the characteristic in Fig. 12, and one approximates the theoretical case of a concentrated space charge.

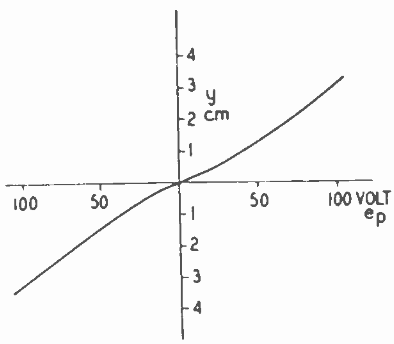

FIG. 15—Typical curve for cathode ray deflection as a function of the deflecting potentials.

As the result of this circuit investigation it can be stated that the gas-filled Braun tube on a maximum generator resistance of 100,000 ohms represents a load that can be disregarded, and that it does not cause noticeable curve distortion with negative-bias plates. If this precaution is not observed, there will be strong regenerative effects and rectification as well as apparent distortion, so that the oscillograph will not agree exactly with the voltage supplied to the plate circuit.

This paper can only give a few of the many detailed experiences that have been obtained with commercial types of low voltage tubes during many years of manufacturing and testing. But as many details and difficulties have been discovered and worked out, it seems possible to design these gas-filled cathode ray tubes as measuring instruments for quantitative work, and to apply them as television receivers.

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JOAK	Tokyo (Japan)	870		10000	WENR
1YA	Auckland (N.Z.)	900		4100	
2GB	Sydney (Aust.)	950		4000	
3YA	Christchurch (N.Z.)	980		3500	KDKA
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A number of authors[1,2,3,4] have discussed the effect of the earth's magnetic field upon the relative absorption of various frequencies in the Kennelly-Heaviside layer. The general conclusion reached is that a maximum in absorption of the sky wave should take place about 1400 kc. Pedersen[4] shows graphically that this effect is quite broad (as may be seen from equations of Nichols and Schelleng)[2] and that for waves sent out from the transmitter at low angles from the horizontal, this effect will be quite small at night. This is in agreement with observed results and it can be concluded that over a wholly dark path such selective absorption effects are probably not sharp.

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in the southern hemisphere, morning transmissions of U. S. stations could regularly be observed after 11 P.M., N.Z.T., until about the time that the transmitting station came into daylight.

Ordinarily U. S. stations could be heard about three or four nights out of each week (based on such transmissions as were made during periods in which they could be observed), the "good" and "poor" nights apparently appearing at random. In general, as the daylight zone below the antarctic circle was approached, received signals diminished in intensity, disappearing entirely after the daylight zone had been entered more than about 100 km.

Reception over long distances was characterized by slow fading, the intensity ranging from minimum to maximum in periods usually of the order of three to five minutes. The rapid fluttering usually characterizing high-frequency transmissions over long distances was not in evidence. On one occasion during a special broadcast, the transmissions of KDKA (980 kc, Pittsburgh) and W8XK (11,880 kc, Pittsburgh) on the same program were observed simultaneously. KDKA was subject to slow fading from zero to fairly high values while the high-frequency waves of W8XK were subject to rapid fluttering but never fell to zero.

No field intensity measurements were made, but aural observations in Dunedin indicated that peak intensities occasionally reached values comparable to the intensities of lower power New Zealand stations located within a few hundred kilometers. For example, WENR (870 kc, 50 kw, Chicago) was observed on a few occasions during the winter at Dunedin to be of nearly the same strength as 3YA (980 kc, 0.5 kw, Christchurch) 300 kilometers distant.

Heterodyne beat interference was frequently noted between foreign stations and occasionally between a foreign station and a local station. For example, a heterodyne beat was frequently noted between 3YA (Christchurch, N. Z.) and KDKA (Pittsburgh) both on 980 kc; and between 2YA (Wellington) and WGN (Chicago), both operating on 720 kc, although in the latter case it was never very strong. Heterodyne beats strong enough to be troublesome were noticed on a large number of frequencies while the modulation level of the stations involved was below the heterodyne beat level. VVC, Calcutta, and 3LO, Melbourne, for example, gave a sufficiently strong heterodyne beat to make reception on this channel difficult except from a local station. Fortunately, where large time differences between stations exist, such interference is usually limited to special or late transmissions from the station to the eastward.

Data in Tables I and II show that the stations regularly observed have frequencies distributed quite uniformly throughout the broadcast

frequency band. Considering the power, best transmission conditions might possibly favor the higher frequencies, though a consideration of other factors such as time differences and distances render such a conclusion doubtful. At least it can be concluded that the sky wave over long distances with a wholly dark path is not subject to any marked increase in absorption in any portion of the frequency spectrum between 500 and 1500 kc.

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and central and western U. S. stations regularly. During the summer in the southern hemisphere, observations were more difficult, sunset occurring after 7:30 P.M. (N.Z.T.) in latitude 45° south (3 A.M., E.S.T.; 12:00 midnight, P.S.T.) as most U. S. stations had concluded their transmissions before this hour. En route to the antarctic about 1-1/2 hours were gained due to change in longitude. This made possible observation of western and central U. S. stations. During the summertime

frequency band. Considering the power, best transmission conditions might possibly favor the higher frequencies, though a consideration of other factors such as time differences and distances render such a conclusion doubtful. At least it can be concluded that the sky wave over long distances with a wholly dark path is not subject to any marked increase in absorption in any portion of the frequency spectrum between 500 and 1500 kc.

A number of authors[1,2,3,4] have discussed the effect of the earth's magnetic field upon the relative absorption of various frequencies in the Kennelly-Heaviside layer. The general conclusion reached is that a maximum in absorption of the sky wave should take place about 1400 kc. Pedersen[4] shows graphically that this effect is quite broad (as may be seen from equations of Nichols and Schelleng)[2] and that for waves sent out from the transmitter at low angles from the horizontal, this effect will be quite small at night. This is in agreement with observed results and it can be concluded that over a wholly dark path such selective absorption effects are probably not sharp.

[1] E. V. Appleton, *Proc. Phys. Soc.* (London), February, (1925).
[2] Nichols and Schelleng, *Bell Sys. Tech. Jour.*, April, (1925).
[3] A. H. Taylor and E. O. Hulbert, *Phys. Rev.* vol. 27, pp. 189–215, (1926).
[4] P. O. Pedersen. "The Propagation of Radio Waves," Chap. VIII and Chap. XI pp. 235–237.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

A NEW TYPE OF ULTRA-SHORT-WAVE OSCILLATOR*

By

I. E. Mouromtseff and H. V. Noble
(Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pa.)

Summary—*A new oscillator for generating ultra-short waves, in which the conventional tank circuit is replaced by a portion of concentric transmission line, is described. Electrically, the tube structure forms an integral part of the transmission line. For this the quantity VC/L for the tube must be the same as for the rest of the line, C and L being capacity and inductance per unit length of the line. A comparison of the closed oscillating circuit with the "standing wave oscillator" shows distinct advantages of the latter in the region of ultra-short waves. The mode of connecting the load to the oscillator is discussed. Some of the oscillator characteristics are given for wavelengths of 5 and 3 meters with 15- and 12-kw output, respectively. Mention is made of a marked physiological effect of ultra-short-wave fields.*

NUMEROUS therapeutic and industrial applications of ultra-short, or so-called quasi-optical waves have been under development during the last few years. This has advanced a new problem in the high-frequency engineering art: development of tubes and circuits for ultra-high frequencies, corresponding to waves below 8 meters. The fact is that the limitations inherent in the conventional power tubes and circuits prevent their use for generating ultra-short waves, as they refuse to oscillate at ultra-high frequencies, and have too low power output and efficiency. The cause of this is clear. The natural way to decrease the frequency of a conventional oscillator is to decrease the lumped capacity and inductance of the oscillating circuit, according to the equation

$$f = \frac{1}{2\pi\sqrt{LC}}. \tag{1}$$

Physical limits in this case are defined by C_a equal to the interelectrode capacity of the tube, and by L_c equal to the minimum inductance of the conductors connecting plate and grid outside as well as inside the tube. Therefore, tubes specially designed for ultra-short waves must have electrodes of comparatively small area, and a structure short in the axial direction. This results in limiting the power dissipation allowed for a tube; hence, the limitation in power output. The difficulty of securing large outputs from ultra-short-wave tubes is further augmented by the comparatively low efficiency which rapidly decreases

* Decimal classification: R339×R133. Original manuscript received by the Institute, March 24, 1932. Presented before Twentieth Anniversary Convention, Pittsburgh, Pa., April 9, 1932.

as frequency increases, because of various losses in the tube circuit, including a very large loss by radiation. Thus, large tubes with a large interelectrode capacity and a long structure cannot be used for generating waves below 7 or 8 meters; while tubes with small capacity and inductance, because of their size, are limited in power output.

However, one can find an interesting solution of the same problem in such a way that the large interelectrode capacity and the long structure of a tube are no longer objectionable and have no direct bearing on the frequency of oscillations generated by the tube.

Fig. 1

Let us imagine that a tube, instead of being attached to a conventional closed oscillating circuit, is connected by its plate and grid to a section of transmission line $AA'BB'$ (Fig. 1). Generally, if a transmission line is terminated by an impedance Z_t, its input impedance Z_i is given by

$$Z_i = Z_0 \frac{Z_t + jZ_0 \tan \dfrac{2\pi l}{\lambda}}{Z_0 + jZ_t \tan \dfrac{2\pi l}{\lambda}} \tag{2}$$

where λ is the wavelength, l is the length of the line, and Z_0 is the *characteristic impedance* of the line. For a line of negligible resistance and leakage conductance

$$Z_0 = \sqrt{L/C} \tag{3}$$

L and C here being the inductance and capacity of the line *per unit length.*

For a line open at the end $A'B'$, in which case $Z_t = \infty$, we have

$$Z_i = \frac{Z_0}{j \tan \dfrac{2\pi l}{\lambda}} \cdot \tag{4}$$

Finally, if the line has a length equal to $\lambda/2$

$$Z_i = \frac{Z_0}{0} = \infty . \tag{5}$$

It shows that for $\lambda = 2l$ an open transmission line, viewed from the tube is analogous to a parallel resonant circuit. An a-c e.m.f. of frequency corresponding to $\lambda = 2l$ applied across one end of such a system produces standing waves in both conductors, with voltage nodes in the middle and voltage antinodes at the ends of these conductors (Fig. 2). The current waves have their nodes at the ends and their anti-

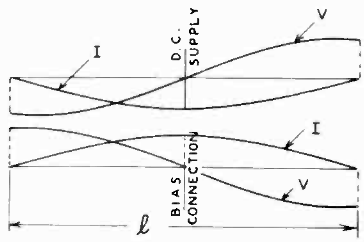

Fig. 2.

nodes in the middle of the system. Once started, the oscillations will continue even in the absence of an e.m.f. until their energy is exhausted by dissipation in the resistance of the conductors, in radiation, or in the load circuit, if any. The first two losses are usually negligible compared to the load. The phase relation at the ends of a parallel wire system set in oscillation is such that a tube connected, as shown in Fig. 1, can sustain the oscillation if d-c voltage is supplied to the plate, and the grid is properly biased. Standing waves are a result of the direct and reflected waves propagated in such a manner that every impulse originated by the tube returns to it after traveling back and forth along each conductor over the distance $2l$. This requires an interval of time

$$T = \frac{2l}{v}. \tag{6}$$

Here, v is the velocity of propagation of electric waves along the conductors and is very nearly equal to the velocity of light. On the other hand, an impulse can be produced by the tube only on a negative half cycle of plate voltage and positive swing of the grid. Hence, T is also the time of one oscillating cycle, and

$$2l = vT = \lambda. \tag{7}$$

Thus, the length of waves generated in a parallel wire system depends only on *the length of the wires*. Therefore, such a system can be used for production of ultra-short waves.

However, if waves of the order of 5 meters, or below, are to be generated, the leads connecting the tube to the transmission line begin to cause trouble by introduction of freak inductances and capacities, which break the uniformity of the system and give birth to unwarranted reflections of oscillatory energy. Fortunately, these difficulties can be removed by making the tube (and its connecting leads) electrically an integral part of the transmission line.[1] Considering that, normally, power tubes have a cylindrical structure, the most logical way to achieve the above requirement is by the use of *concentric cylinder transmission lines* (Fig. 3). The plate and the grid must then be of such diameters that the characteristic impedance $\sqrt{L/C}$ along the tube will be the same as that for the main part of the oscillator. This can be readily achieved by a proper choice of the ratio of anode and grid diameters. Indeed, the characteristic impedance for a concentric cylinder is given by the equation

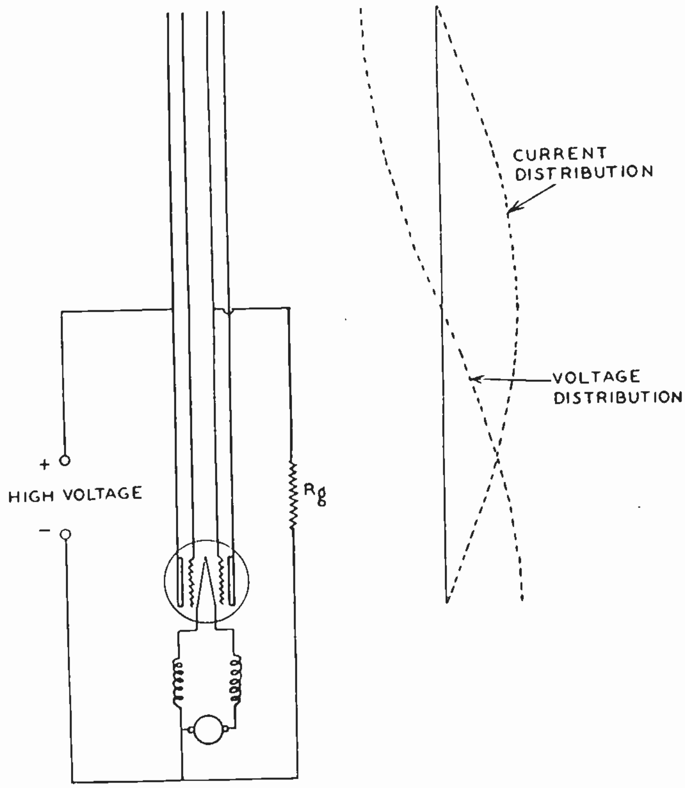

Fig. 3

$$Z_0 = 60 \log \frac{D}{d} \tag{8}$$

where D is the diameter of the outer, and d of the inner cylinder. There-

[1] First suggested by C. A. Boddie, then with Westinghouse Research Laboratories.

fore, for the equality of impedances the ratio of the anode to the grid diameters must be the same as that of the outer and inner pipes of the oscillator itself.

Fig. 4 shows a standing wave oscillator actually built for 2.8-meter waves. A special tube, the AW-200 designed for it, is seen projecting

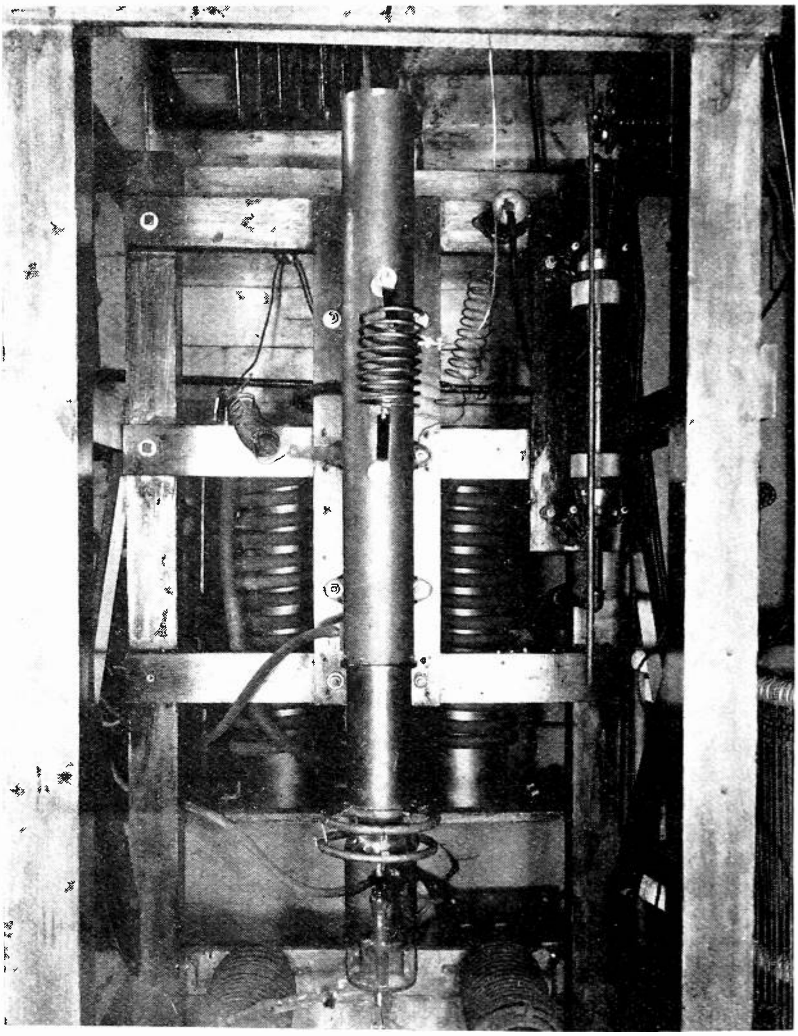

Fig. 4

from the lower end of the oscillator, and is also shown separately in Fig. 5. Notwithstanding the large size of the tube, which is 1 meter long and has an anode 3 inches in diameter, and notwithstanding a large plate-to-grid capacity amounting to 23 $\mu\mu$f, extremely short waves can be generated by this tube. The outstanding features of the tube are: the introduction of the grid and filament from opposite ends of the tube; completely symmetrical grid structure supported from a heavy reëntrant copper thimble; a tubular outer grid lead of about the same diameter as the grid itself; and a grid capable of withstanding large currents and voltages. The anode is made of a section of copper pipe. It is water-cooled and can easily dissipate 30 kw. The water

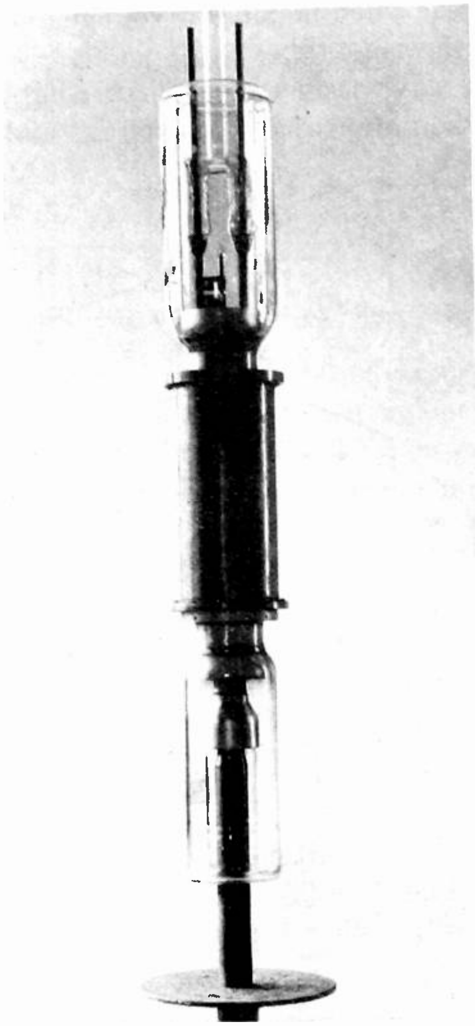

Fig. 5

jacket surrounding the anode is 5 inches in diameter and is directly connected to the outer cylinders. The ratio of diameters of the concentric pipes is approximately 2:1, and so is the ratio of the anode and grid diameters.

From inspection of the curve of voltage distribution along the conductors (Fig. 2) it follows that d-c plate voltage as well as grid bias can conveniently be supplied at the midpoint of the system, where the r-f potential is zero (Fig. 3).

In order to have some insight into the difference in performance of a standing wave oscillator as compared to a conventional Hartley circuit, in which the lumped capacity has been reduced to the inter-electrode capacity of the tube, C_a, let us calculate the amount of oscillating energy for the same maximum voltage, V_0, across the tube in both cases, using the same tube.

For a standing wave oscillator the circulating volt-amperes will be found under an assumption that both voltage and current are distrib-

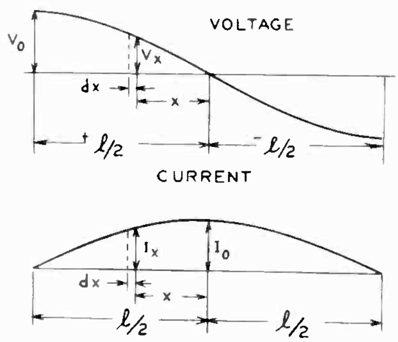

Fig. 6

uted along the conductors according to the sine law. Then, the maximum of potential energy, W_p, corresponding to the maximum voltage, V_0, can be calculated (Fig. 6) as

$$W_p = \frac{CV_0^2}{2}\int_{-(l/2)}^{+(l/2)} \sin^2 \frac{\pi}{l} x dx = \frac{ClV_0^2}{4}. \tag{9}$$

In the last expression, C is capacity per unit length and l is the length of the transmission line. A similar expression can be obtained for maximum kinetic energy W_k, when the current is a maximum and voltage is zero at every point

$$W_k = \frac{LlI_0^2}{4}. \tag{10}$$

L in the formula is inductance per unit length. Assuming, further, that resistance and leakage conductance of the line are zero, we can write:

$$\frac{ClV_0^2}{4} = \frac{LlI_0^2}{4} \tag{11}$$

from which expression we obtain

$$I_{eff} = \frac{I_0}{\sqrt{2}} = \sqrt{C/L} \times \frac{V_0}{\sqrt{2}} = \frac{V_{eff}}{Z_0} \qquad (12)$$

Therefore, the volt-amperes in the sytem are

$$(VA)_s = I_{eff}V_{eff} = \frac{V_0}{Z_0\sqrt{2}} \times \frac{V_0}{\sqrt{2}} = \frac{V_0^2}{2Z_0} \qquad (13)$$

A corresponding expression for a closed circuit having a lumped capacity; namely, that of the tube, C_g, is

$$(VA)_c = \pi f C_g V_0^2. \qquad (14)$$

By division of (13) by (14) we find

$$\frac{(VA)_s}{(VA)_c} = \frac{1}{2\pi f C_g Z_0} = \frac{X}{Z_0}. \qquad (15)$$

From (13) we can see that the flywheel effect of volt-amperes, in the case of a standing wave oscillator, is independent of the frequency of oscillation, but depends on the characteristic impedance of the system, Z_0, which, for a concentric cylinder line, is determined by the ratio of diameters of the outer and inner cylinders (equation 8). By making this ratio closer to unity we can increase Z_0, and hence, the flywheel effect for any frequency. This, of course, will involve the tube structure, in which the grid diameter must be increased. In a closed circuit, the flywheel effect (14) increases with frequency and tube capacity, C_g, but the highest possible frequency is limited by the conditions discussed at the beginning of this paper (1). Using practical figures in (15) for an AW-200 tube generating 5-meter waves, we can compare the two circuits as to the flywheel effect as follows:

$$\frac{(VA)_s}{(VA)_c} = \frac{10^{12}}{2\pi \times 6 \times 10^7 \times 23 \times 42} = \text{approximately } 2.8. \qquad (16)$$

A higher value of volt-amperes in a standing wave oscillator, while contributing to the stability of oscillation and elimination of harmonics does not create any concern as to the circuit losses, because heavy concentric pipe leads and tube parts can easily take care of a considerable amount of oscillating energy without appreciable loss.

Another interesting comparison between a standing wave oscillator and a closed circuit can be drawn from determining the highest practical frequency limit in either case, if the same tube is used in both cases. With a standing wave oscillator the lowest limit for generated waves depends on the external physical dimensions of the tube. Indeed, the whole grid end of the tube with the glass blank attached to it is located inside the outer pipe and must not project beyond the neutral line of the oscillator, where d-c voltage supply wires are to be connected (Fig. 7). Thus, for an AW-200 tube in its present shape the shortest possible wavelength is somewhat below 3 meters (2.8 meters).

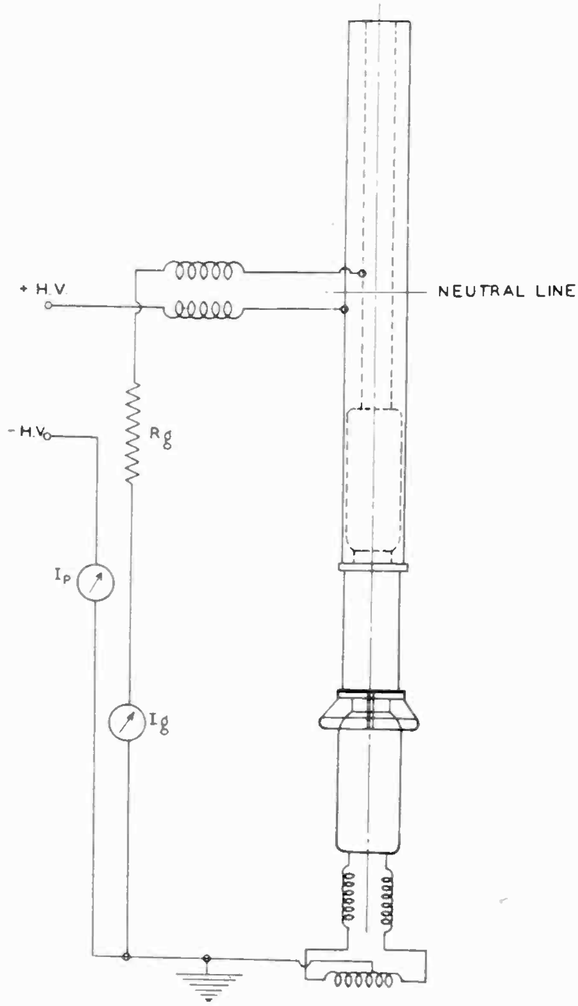

Fig. 7—Oscillating circuit using AW-200 tube.

If the same tube were used in a closed circuit of minimum dimensions, shown in Fig. 8, the shortest possible wave has been calculated to be about 6 meters, at which, however, only a very low output can be secured.

A Loaded Standing Wave Oscillator

If a standing wave oscillator is terminated by an ohmic resistance, R, equal to the characteristic impedance of the line, Z_0, no self-oscillations are possible in such a system. This follows from (2), which in a case of $Z_t = Z_0 = R$ gives the value of the input resistance

$$Z_i = R. \tag{17}$$

The whole system behaves as pure ohmic resistance with no reflected waves to sustain the oscillation. The reflection factor is in this case

$$f_r = \frac{V_r}{V_i} = \frac{R - Z_0}{R + Z_0} = 0 \tag{18}$$

where V_r and V_i are the amplitudes of the reflected and incident waves, respectively. With $R \neq Z_0$, the same factor is different from

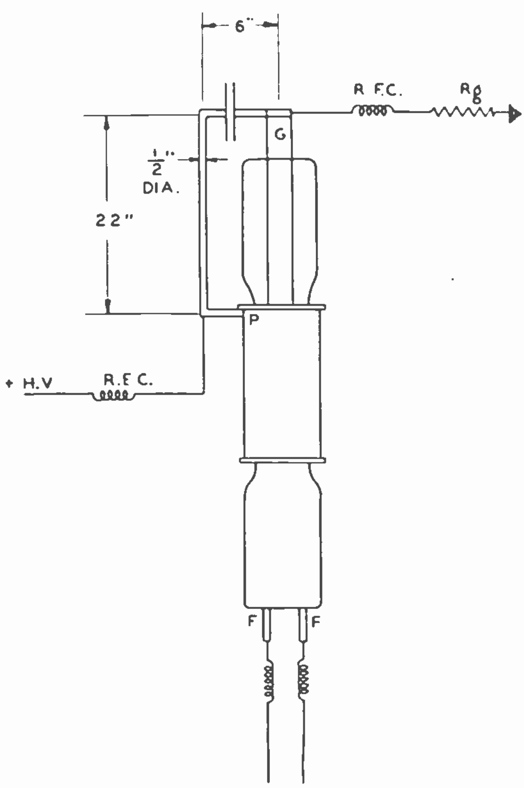

Fig. 8—Shortest connections for Hartley oscillator. $\lambda = 6$ meters.

zero. The energy reaching the load end is partly consumed and partly reflected, creating standing waves on the conductors, resulting in sustained oscillations. From (18) it follows that the greater the load resistance, R, as compared to the characteristic impedance, Z_0, the closer f_r is to unity; the greater is the amplitude of the reflected wave, V_r; the more pronounced is the oscillation; and the smaller percentage of energy is consumed in the load during each cycle. The ratio of the energy consumed in the resistance during one half cycle, P_L, to the amount of energy P_t transferred in the circuit during the same time is the logarithmic decrement, δ, of the system. From this definition, and keeping in mind (9), we can write:

$$\delta = \frac{P_L}{P_t} = \frac{V_0^2 \times T}{2R \times 2} \div \frac{CIV_0^2}{4} \qquad (19)$$

or,

$$\delta = \frac{2Z_0}{R}. \qquad (20)$$

Assuming that for stable oscillation, $\delta = 0.25$ we can find from (19) the minimum ratio of volt-amperes/watts and R/Z_0:

$$\frac{VA}{\text{Watts}} = \frac{R}{Z_0} = 8. \qquad (21)$$

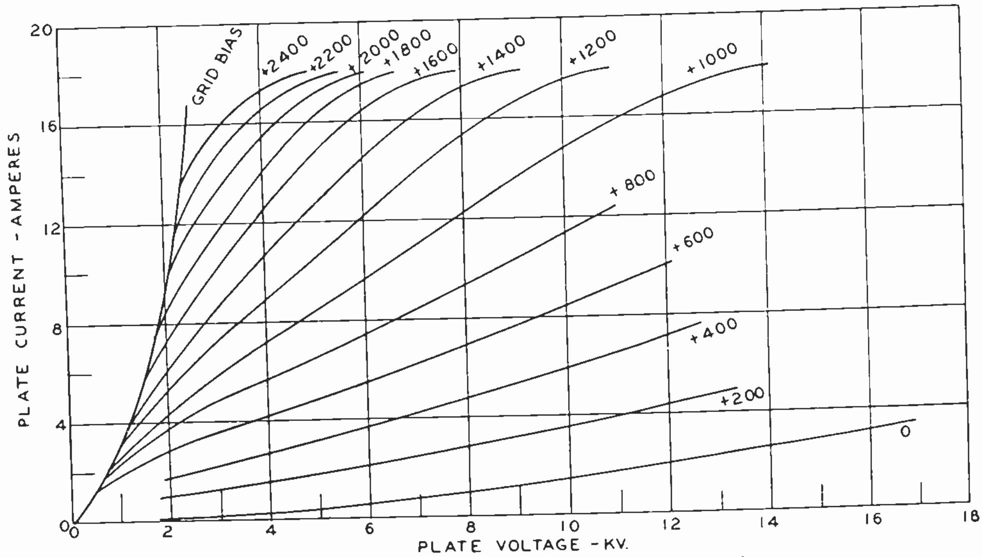

Fig. 9—AW-200 plate characteristics. $\mu = 30$.

This, or higher ratio must be maintained for smooth oscillation and for elimination of high-frequency harmonics from the voltage/time curve. For a practical case of a line with the ratio of cylinder diameters $D/d = 2$, the characteristic impedance $Z_0 = 42$ ohms (8). Hence, the minimum load resistance which can be allowed is 336 ohms. Further, one must consult the tube characteristics to find whether this figure corresponds to a good operating condition for the tube. Due to the symmetry of the system, the effect of a load connected across the receiving end of the line is equivalent to that of an equal load connected directly across the plate and grid of the tube. With filament at zero radio-frequency potential, the plate-filament voltage is half of the plate-grid voltage. Hence, an equivalent resistance, R', between the plate and filament, which will produce the same effect as a load across the ends of the line, must be $1/4(R)$. In our practical case $R'_{min} = 84$ ohms. From the tube chart (Fig. 9) one can easily conclude that with such a load the tube operation will be extremely inefficient. With 10 kv on the plate, the optimum load resistance for the tube, R_p, is about 800 ohms. Therefore, the load resistance across the end of the oscillator must be 3200 ohms, which makes the ratio of volt-amperes/watts proportionally greater.

Actually, it is more practical to tap the load at some points a and a' near the neutral line (Fig. 10), where the voltage swing is smaller.

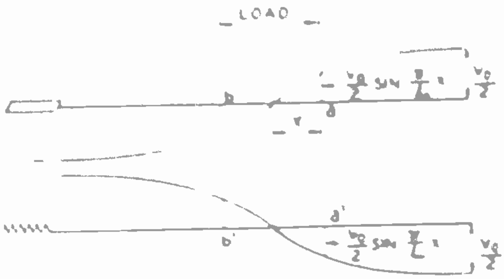

Fig. 10

Correspondingly, the new load resistance, R, must be chosen so that it will match the same optimum resistance for the tube, R_p, which can be calculated from the relation

$$\frac{V^2}{2R} = \frac{V^2 \sin^2 \frac{\pi}{l} x}{2R} \tag{22}$$

or,

$$R = R \sin^2 \frac{\pi}{l} x. \tag{23}$$

No matter what an actual load circuit may be, it is usually more convenient to locate it at some distance from the oscillator and couple the two by a single wire, or parallel wire transmission line. The duty of the latter is the transfer of power from the oscillator to the load circuit with as little loss as possible. In this case, standing waves, causing radiation and more pronounced resistance loss, are an undesirable phenomenon. Therefore, this transmission line must be "perfectly terminated," and the load circuit must be "matched" to the impedance of the line by means of an adequate coupling arrangement. If it is achieved successfully, the input resistance of the transmission line, which at the same time is the real load on the oscillator, will be equal to the characteristic impedance of the line Z_0' and will be purely ohmic in character.

A parallel wire transmission line of practical dimensions has a characteristic impedance of the order of several hundred ohms. It has to be connected at two symmetrical points such as *ab* on the oscillator either on the outer, or inner cylinder (Fig. 10). From voltage relation, one can readily calculate the equivalent load on the tube effected by the transmission line connected at the chosen points. Generally, it

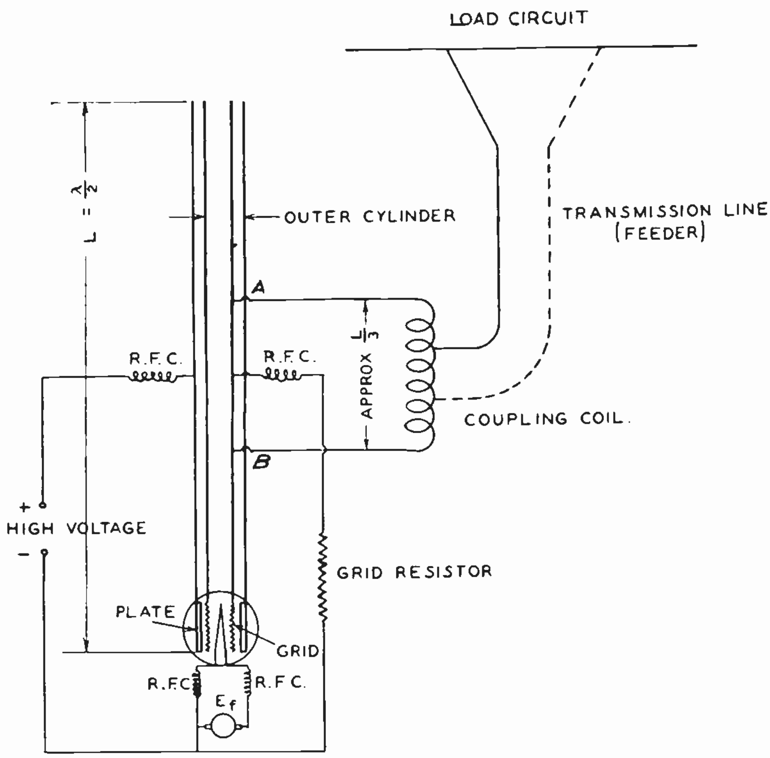

Fig. 11

will not correspond to the optimum operating condition for the tube. However, if instead of directly connecting the load transmission line to the oscillator at points *A* and *B*, we bridge over the same points by a coil, or by a small tank circuit (Fig. 11), and tap the coil at two symmetrical points for attaching the load, we can match the desired tube operating condition perfectly. The coil impedance for frequencies considered is very high; that of the tank circuit is infinity. Therefore, these additional members will not greatly affect the whole arrangement, though the tank circuit will introduce an additional flywheel effect into the transmission system.

One can add that the use of two tubes, one at each end of an oscillator, allows an increase in the output and contributes to the stability of performance as is the case with every push-pull circuit.

Experimental Results

An arrangement similar to that of Fig. 11 has been used for measuring the power output and efficiency with various adjustments of the oscillator. First, the output circuit usually one or two dipoles was

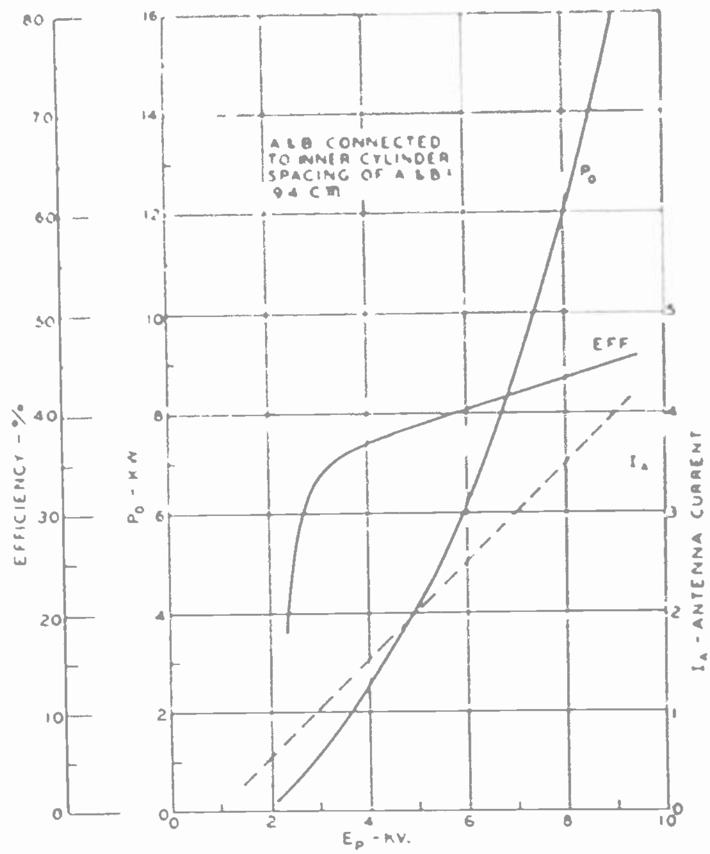

Fig. 12—Standing wave oscillator. $\lambda = 5.5$ meters.

tuned to the frequency generated by the oscillator. Then, the coupling of the load to a parallel wire, eventually to a single wire transmission line was adjusted, until no standing waves were detected on it. Finally, an optimum coupling of the transmission line to the oscillator was found. The power output for various conditions of plate voltage and grid bias was calculated by subtracting the power dissipated in the tube from the total plate input, as measured on a plate d-c voltmeter and ammeter. The power dissipation was reckoned from the temperature rise of the anode cooling water and the rate of water flow. Preliminary experiments showed that the power loss in the oscillator with no load was negligible. Figs. 12 and 13 give the power output and efficiency as a function of variable plate voltage for 5.5- and 2.8-meter waves, respectively.

A dotted straight line in Fig. 12 indicates that the antenna current varies proportionally with the plate voltage, a condition necessary for

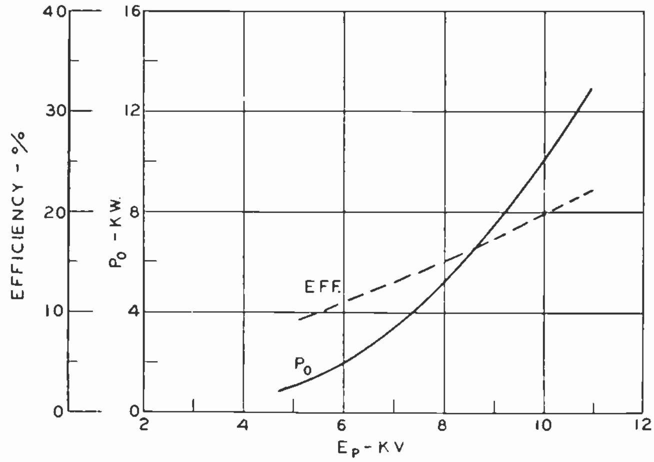

Fig. 13—Standing wave oscillator. $\lambda = 2.8$ meters.

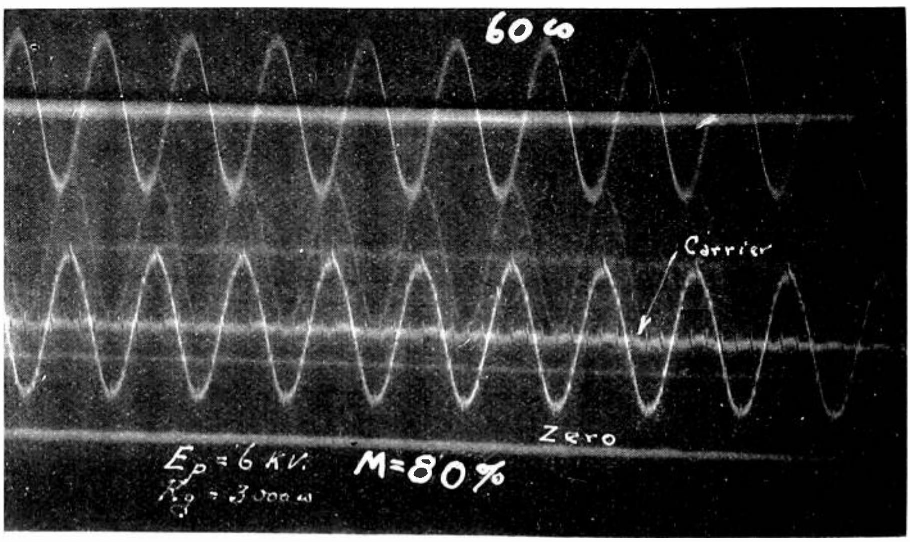

Fig. 14

distortionless plate modulation of the oscillator. As to the realizable percentage of modulation, which can be effected, one first can notice with dissatisfaction that with a slow variation of the d-c plate voltage the oscillation goes out before the plate voltage reaches zero. The actual value of the critical voltage depends on various elements of the circuit adjustment. Fortunately, the oscillator behaves differently, when the plate voltage variation is effected rapidly, which is the case when an actual modulating voltage is superimposed upon the d-c voltage. Figs. 14 and 15 represent oscillograms taken on a standing wave oscillator modulated with 60 cycles. The first picture corresponds to an 80 per

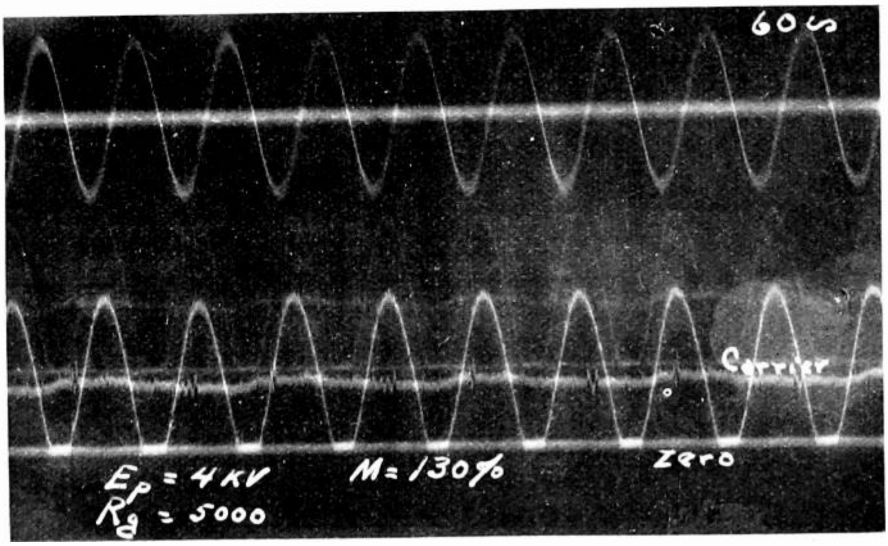

Fig. 15

cent modulation; the second was taken at 30 per cent overmodulation. From the latter, one can see that the oscillation goes off and starts up again at practically zero plate voltage.

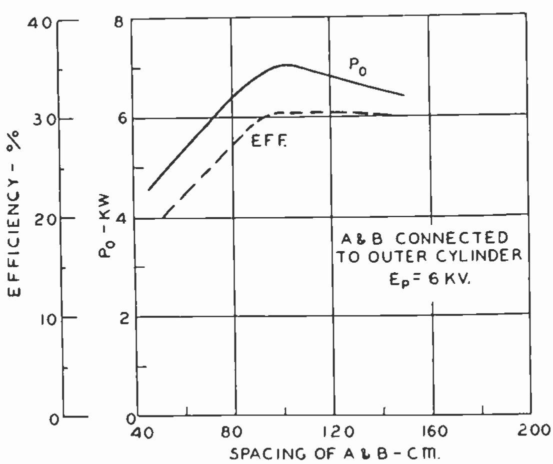

Fig. 16

Figs. 16 and 17 show how power output and efficiency change when the distance of points to which the load is coupled from the neutral line of the oscillator be varied. Attaching the load to the inner cylinder gives slightly greater output and considerably better efficiency.

Physiological effects of such powerful oscillations of the frequencies considered—60 and 100 megacycles—are possibly worth while mentioning. In the vicinity of the load circuit and as far as 20 or 30 feet away, a person feels a very pronounced diathermic effect in the limbs, particularly while standing on a concrete floor, or on a metal surface; also when holding a hand on a metal conduit, or stretching hands in the direction of the field. With a proper location in the field,

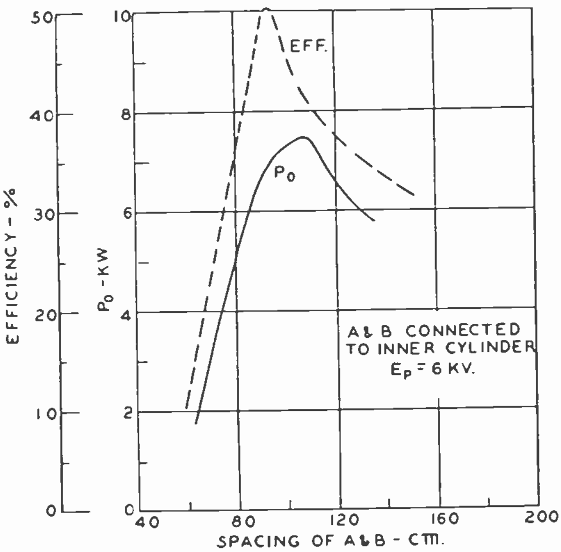

Fig. 17

the blood temperature of a person can be raised by 1° C within 1 minute. It is no wonder that an operator, if he is not careful enough, feels drowsy and depressed after a few hours of work in such a field. Bugs and other insects, if located between the plates of the condenser in a tank circuit die almost instantaneously. Various food, as rolls, hot dogs, etc., are roasted or cooked within a fraction of a minute. Rolls and bread are toasted from within. One can light a 40- to 100-watt lamp merely by holding it in the field of the oscillator.

Conclusion

The standing wave oscillator gives a means of generating powerful oscillations of very high frequencies. It can find application in every field where high frequency and high power combined are of importance. A particular interest can be attached to the possibility of using it for therapeutic, physiological, and biological problems—premises so far unexplored, perhaps just because of a lack of sufficiently powerful generators in the important frequency band.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

THE ACTION OF SHORT-WAVE FRAME AËRIALS*

By

L. S. Palmer

(Professor of Physics, The University College, Hull, England)

and

L. L. K. Honeyball

(The University College, Hull, England)

PART I

THE THEORY OF THE ACTION OF A TUNED RECTANGULAR FRAME AËRIAL

By

L. S. Palmer

Summary—*Part I (1) An electromagnetic wave incident on a tuned rectangular frame aërial comparable in dimensions with the length of the wave is considered to cause a circulating current which may be resolved into a "direct" current component due to the primary action of the wave, and an "indirect" current component due to the current in adjacent parts of the frame.*

This treatment leads to the conclusion that the resultant current is dependent on the dimensions of the frame and on the angle of incidence of the wave.

(2) For each particular width of the frame less than one wavelength there are at least two critical heights for which the frame current will be a maximum. That is, for each particular width less than one wavelength there are at least two critical frame areas for best reception, and any variation of these areas will result in decreased signal strength.

(3) The resonant current which circulates in a tuned frame aërial under the influence of an electromagnetic wave is not the maximum current that can be produced in the frame by the incident wave. By formatising the frame, that is, by properly adjusting the height and width of the frame, the current may be increased to a very much larger value.

Part II (1) The transmitter, which is capable of working on wavelengths from 7.54 to 8.80 meters, is described and the chief difficulties in its design are discussed.

The construction and mode of operation of the receiving frame are next described. The frame is capable of expanding or contracting in either or both dimensions. The maximum size is about 8.5 meters square. Tuning is accomplished by shunting the frame with a copper strip in parallel with a condenser. The current is measured by a vacuum thermojunction and microammeter.

(2) Measurements are described which enable the direction of wave propagation to be made parallel to the plane of the frame. Preliminary measurements were made to study the nature of the polarization and inclination of the electric vector of the wave.

(3) The result of a series of measurements of the critical frame dimensions are recorded, and compared with data predicted from the theory discussed in Part I.

* Decimal classification: R125.3×R325.3. Original manuscript received by the Institute, March 14, 1932.

(4) The conclusions resulting from the foregoing investigation are enumerated and may be summed up in the statement that the maximum frame current only results when the frame is both tuned and formatised.

A mechanical graph for solving the transcendental equations which determine the formatising conditions is explained in an appendix.

1. Introduction

IN RECENT years many papers have been published on the action of frame aërials and, although it may not have been explicitly stated, the formulas deduced have been based on the implicit assumption that the frame dimensions were negligible compared with the length of the wave. With the ever-increasing applications of short waves the use of such formulas may lead to serious errors and, in fact, to quite inaccurate conceptions of the action of frame aërials. For example, it is usually assumed that the current produced in a frame aërial by an incident wave or the radiation from a frame aërial is proportional to the square of the area of the frame,[1] but this does not hold good when the frame dimensions and the wavelength are of the same order of magnitude. With short waves comparable with the frame dimensions, an *increase* of frame area may cause a *decrease* of received current.

Preliminary experiments carried out at Manchester in 1927 indicated that, in order to get maximum received current in a rectangular frame, the area of the frame had to be very carefully adjusted to any one of several critical values which were found to be dependent on the length of the wave in use. Furthermore, for any given wavelength the critical area varied with the angle of incidence of the wave.

That the angle of the wave to the plane of the frame affects the signal strength is clearly indicated in the ordinary directional properties of a rotating frame used as a wireless beacon; but that the signal strength should also vary as the frame (whether square or rectangular) rotates in its own plane is an effect which has not hitherto been anticipated. It is, in fact, not true that the received current depends only on the change of magnetic flux threading the frame. The e.m.f. so produced may be comparable in magnitude with that induced in one limb of the frame by the current in an opposite limb. Thus the resultant current will be due to the superposition of these two e.m.f's. When they are comparable in magnitude, their relative phases may be such that the resultant current is either abnormally large or abnormally small. It is shown in the next section that these phase differences depend not

[1] E. B. Moullin, "The fields close to a radiating aërial." *Proc. Camb. Phil. Soc.*, vol. 25, pt. IV, p. 491; October, (1929).

only on the frame impedance but also on the frame dimensions. Hence a variation of dimensions or shape, even with proper tuning, will result in changes of frame current which may be of the same order of magnitude as the well-known current changes produced by variation of tuning.

In view of this fact, the word "formatising" (from the Latin "*formare*," to shape or to form) is suggested for the process of changing the dimensions of a frame aerial in order to obtain maximum current. This is then distinguished from the quite different process of "tuning," which also affects the current value. Thus only when a frame is both tuned and formatised is the current a maximum. The formatising conditions for a frame are as critical as the tuning conditions and in order to avoid confusion between effects due to tuning and effects due to formatising, the following theory will deal only with tuned frames; that is, with frames the reactances of which are zero. This condition is also easy to fulfill in practice.

The preliminary experiments with tuned frames carried out in 1927 showed that the received current could be increased or decreased many thousandfold by changing the width or height of the frame by only a few centimeters. In some instances doubling the frame area caused the current to decrease to an almost negligible amount, even though the frame was kept properly tuned.

The conditions for formatising a tuned rectangular receiving frame will now be determined. The conditions will be different if the frame is to be used for transmission. This question is now being investigated.

2. Theoretical Conditions for Formatising a Receiving Frame

In the first instance the problem will be simplified by considering the frame $P\ Q\ R\ S$ (Fig. 1) to be oriented so that its plane is parallel to the direction of wave propagation $A\ B$. It will be assumed that the wave is plane polarized so that its electric vector is also in the plane $P\ Q\ R\ S$. The current variations will then be distinct from those given by the ordinary "figure-of-eight" polar diagram.

Let the direction of propogation AB make an angle γ with the side PQ of the frame. Then the passing wave will induce in the vertical wires PS and QR and in the horizontal wires PQ and RS e.m.f's which will differ in phase by an amount depending only on the respective "effective" distances apart of the parallel wires. This assumes the transmitter is distant at least several wavelengths from the frame. From the figure the "effective" distance is seen to be PT for the vertical wires, and PT' for the horizontal wires. $PT = PQ \cos\gamma$ or $W \cos\gamma$ where W is the frame width; and $PT' = PS \sin\gamma$ or $H \sin\gamma$ where H

is the frame height. Since the frame is tuned, the phase differences between the e.m.f's will be the same as the phase differences between the currents. Hence the currents resulting directly from the incidence of the waves on the vertical wires will differ in phase by $2\pi W \cos \gamma/\lambda$ and similarly the currents in the horizontal wires will differ in phase by $2\pi H \sin \gamma/\lambda$.

If the transmitter be situated a distance from the frame very large compared with the frame dimensions then the current amplitudes will depend only on the transmitter constants, the distance between the transmitter and the frame, the angle γ and the ohmic resistance of the frame. We will call these currents the "direct" current components.

Now any current in the wire PS (say) will produce a current in any neighboring wire as long as the two wires are not perpendicular to each other. Thus there will be in the wire QR an induced current due to the

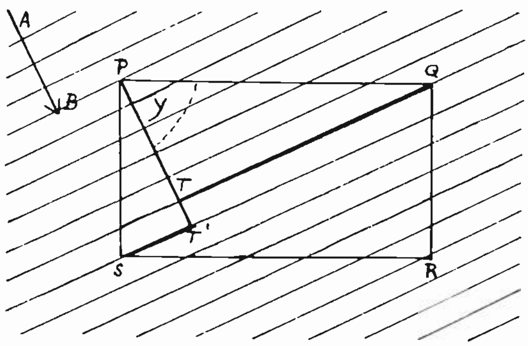

Fig. 1

current in PS and vice versa. Similarly the currents in the horizontal wires will induce additional currents in each other. These currents may be called the "indirect" current components. The resultant current in the frame will be the vector sum of the direct current due to the wave and the indirect current due to the current in the neighboring parallel limb of the frame.

It is therefore necessary to determine the amplitudes of and phase differences between the indirect currents in the opposite limbs of the frame. To do this let the current in the wire PS be $I = I_0 \sin \omega t$. Then, if we look upon the wire PS as a tuned Hertzian oscillator, it will have an electric moment $-\int I_0 \sin \omega t \, dt$, resulting in a field a distance W away such that the electric vector E of the field is given by:

$$E = AI_0 \left\{ \frac{1}{W^3} \cos(\omega t - a) + \frac{1}{vW^2} \frac{d}{dt}[\cos(\omega t - a)] + \frac{1}{v^2 W} \frac{d^2}{dt^2}[\cos(\omega t - a)] \right\}$$

where A is a constant, v is the velocity of the electromagnetic waves in the medium, and $a = 2\pi W/\lambda$, λ being the wavelength.

Hence,

$$E = \frac{AI_0}{W^3}\left[a^2 + (1 - a^2)^2\right]^{1/2} \sin\left[\omega t - (a - \phi + \pi/2)\right]$$

$$= E_0 \sin\left[\omega t - (a - \phi + \pi/2)\right] \qquad (1)$$

where

$\phi = \tan^{-1} a/(1 - a^2)$ and E_0 is a monotonic function of a.

This field produces a current I' in any conductor not lying perpendicular to E, and the induced current will be in phase with E if the conductor be tuned to the frequency $\omega/2\pi$. Hence the phase difference between the current I in PS and the indirect current I' produced in the parallel tuned conductor QR is:

$$-(a - \phi + \pi/2).$$

Thus the phase differences between the direct-current component in the vertical wire QR and the initial current in the wire PS is $2\pi W \cos\gamma/\lambda$ or $a\cos\gamma$ and between the indirect-current component in QR and the current in PS is $-(a - \phi + \pi/2)$. Similarly the corresponding phase differences between the currents in the horizontal wires are $a' \sin\gamma$ and $-(a' - \phi' + \pi/2)$, respectively. Now if it be assumed that the amplitudes of these component currents are approximately independent of a or a', then the resultant current will be a maximum when the phase differences between its components are equal. That is, when,

$$a(1 \pm \cos\gamma) - \phi + \pi/2 = 0 \qquad (2a)$$

and when,

$$a'(1 \pm \sin\gamma) - \phi' + \pi/2 = 0. \qquad (2b)$$

The alternative signs depend on whether the limb nearer to or remote from the transmitter is being considered.

It has already been shown that the amplitude of the direct component is independent of a (or a') as long as the transmitter is a great distance from the frame, and the experimental results show that the function E_0 in (1) may also be taken to be approximately independent of a (or a') over the range of values of W (or H) used in the experiments described in Part II.* Hence the solutions of (2) will give approximately the critical values of W and H for which the frame current will be a maximum. With the minus sign (2a) leads to a critical value of W which was too small to be tested satisfactorily with the frame aërial described in Part II. Consequently this paper will deal only with

* This assumption has received further support from the recent paper on "The Spreading of Electromagnetic Waves from a Hertzian Dipole" by J. A. Ratcliffe, L. G. Vedy and A. F. Wilkins in the Wireless Proceedings of the I. E. E., (London) vol. 7, No. 20, p. 71; June, (1932).

positive values of $\cos\gamma$ and $\sin\gamma$, that is, with values of γ less than 90 degrees. Equations (2) reduce to the transcendental equations:

$$\tan a(1+\cos\gamma) = (a^2-1)/a \tag{3a}$$

and

$$\tan a'(1+\sin\gamma) = (a'^2-1)/a'. \tag{3b}$$

In the case when $\gamma=0$ degrees and the electric vector of the wave is parallel to PS the solutions are:

$$a = 2.05,\ 3.75,\ 5.35,\ \text{etc.},$$

and

$$a' = 4.45,\ 7.7,\ \text{etc.},$$

leading to values of PQ (or W) $=0.33\lambda$, 0.60λ, 0.85λ, etc., and of PS (or H) $=0.71\lambda$, 1.22λ, etc. These solutions may be determined by the graphical method outlined in the appendix.

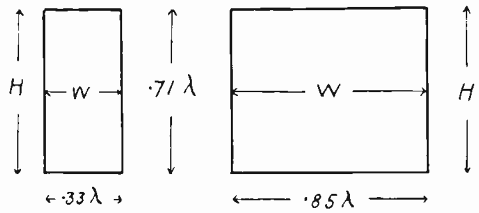

Fig. 2—Two critical dimensions of a frame aërial corresponding to optimum received current when γ equals zero.

There is an ambiguity arising from the fact that the solutions include values of a and a' for both maximum and minimum currents, but experiment shows that odd solutions give maximum current conditions and even solutions give minimum current conditions.

Thus we conclude that for a tuned rectangular frame oriented in the plane of wave propagation and with its sides parallel to the electric vector, the received current will be a maximum when the frame has either of the critical shapes shown in Fig. 2, which is roughly to scale.

But these are not the only critical values for the case when $\gamma=0$ degrees, because the preliminary experiments with this value of γ indicated that current maxima occurred for *any* width of the frame provided that the height was suitably adjusted. This fact can be explained by extending the foregoing theory as follows:

Suppose the frame width be *increased* so that the phase angles between the two component currents in the vertical wires be not 0 but $+\psi$ (say) then,

$$a(1+\cos\gamma) - \phi + \pi/2 = +\psi \tag{4a}$$

and the new width may be considered to act as an added inductance producing the undesirable phase lag ψ.

If now, the height be *decreased* just sufficiently to produce a phase lead of $-\psi$ between the component currents in the horizontal wires of the frame, then the effect is similar to introducing a capacity of sufficient magnitude to compensate for the inductive effect of increasing the frame width. Thus,

$$a'(1 + \sin\gamma) - \phi' + \pi/2 = -\psi \tag{4b}$$

and the *total* "reactance" effect is the same as before but the frame dimensions are quite different. The dimensions for different values of ψ can now be calculated from the reduced equations:

$$\tan\left[a(1 + \cos\gamma) - \psi\right] = (a^2 - 1)/a \tag{5a}$$

and,

$$\tan\left[a'(1 + \sin\gamma) + \psi\right] = (a'^2 - 1)/a'. \tag{5b}$$

This has been done for $\gamma = 0$ degrees, 15 degrees, 30 degrees, and 45 degrees, and the results are shown in Fig. 3. The graphs for $\gamma = 60$ degrees, 75 degrees, and 90 degrees are the same as those for 30 degrees, 15 degrees, and 0 degrees respectively but the values of W/λ and H/λ are interchanged.

The coördinates of any point on the graphs give the frame dimensions (W/λ and H/λ) for which the lag (or lead) between the currents in the sides is just compensated by the lead (or lag) between the currents in the top and bottom. Hence for these dimensions the resultant frame current will be a maximum.

The two points A refer to the frames depicted in Fig. 2 for which $\gamma = 0$ degrees.

Only two graphs are shown for each value of γ because calculations were not made for values of W exceeding one wavelength. It follows from these graphs that for any given frame width less than one wavelength there are at least two heights for which the frame aërial current will be a maximum, whilst for a width of about half a wavelength there may be three critical heights indicated by the points B in Fig. 3. Similarly, there may be more than two critical widths for certain values of the height; one case, for a height of 0.8λ (approximately), is shown by the points C in Fig. 3.

Since both the height and the width have critical values dependent on the wavelength, it follows that the area of a frame for maximum current is also critically dependent on the wavelength. By calculating the product HW/λ^2 for various points on the graphs, it may be shown that the frame of maximum area is not necessarily that for which $H = W$. In the case of $\gamma = 0$ degrees, for example, the maximum values of HW/λ^2 are about 0.24, when $H/\lambda = 0.6$ and $W/\lambda = 0.4$, and 0.73 when $H/\lambda = 1.22$ and $W/\lambda = 0.6$. For $\gamma = 45$ degrees the maximum areas are $0.15\lambda^2$ when $H = W = 0.39\lambda$, and $0.50\lambda^2$ when $H = W = 0.72\lambda$ and the frames of maximum area are square for this value of γ only. For $\gamma = 90$ degrees the maximum areas are the same as those for $\gamma = 0$ degrees but the values of H and W are interchanged. Thus with the

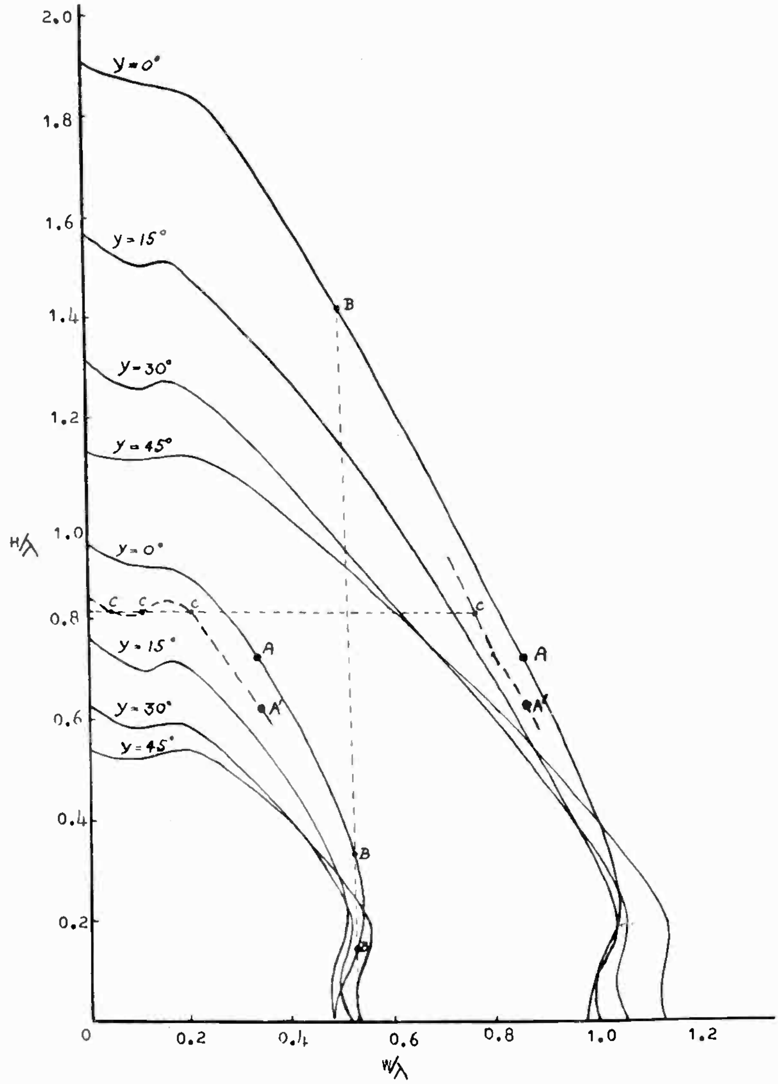

Fig. 3—Graphs showing frame dimension for which lag or lead angles between currents in sides are compensated by corresponding lead or lag angles in top and bottom.

smaller frames the maximum critical areas have values of the ratio H/W which vary from 0.6/0.4 (or 1.5) when $\gamma = 0$ degrees to 0.4/0.6 (or 0.66) when $\gamma = 90$ degrees, while in the case of the larger frames, the maximum critical areas have values of the ratio H/W which vary from 1.22/0.6 (or 2.0) to 0.6/1.22 (or 0.5) for the same range of γ.

Thus, if the values of W and H are such that the frame current tends to be a maximum, (i.e., they are given by points on the graphs of Fig. 3) then the resulting frames only have large areas when one dimension is not greater than twice the other.

This indicates that, although the frame dimensions may be given by points on the graphs, the resulting maximum currents will not be of equal value, but will be largest for those frames which have the largest areas. This was confirmed by experiment (see Fig. 10) which shows that very flat and wide or very high and narrow frames, even when of correct critical dimensions, received much less current than those for which the values of H and W were more nearly equal.

Since the values of the maximum critical areas decrease from $0.24\lambda^2$ and $0.73\lambda^2$ when $\gamma=0$ degrees or 90 degrees to $0.15\lambda^2$ and $0.50\lambda^2$ when $\gamma=45$ degrees, it follows that, for a given wavelength, the maximum critical area of a frame oriented with one diagonal parallel to the electric vector is less than the maximum critical area of a frame oriented with two sides parallel to the electric vector. Thus we arrive at the unexpected conclusion that, if a frame were critically adjusted for a wave arriving from a given direction, the signal strength should decrease if the frame were rotated *in its own plane* although its dimensions (and area) remained constant.

The most important of the foregoing deductions concerning a frame with its plane parallel to the direction of propagation will now be stated before describing the experimental work in Part II.

The first conclusion is, that for efficient reception with a tuned rectangular frame aërial comparable in dimensions with the length of the wave the frame must be formatised; that is, the width W and the height H of the frame must have certain critical values which depend on each other, on the wavelength and on the angle γ between the side of the frame and the electric vector of the wave. The formatising conditions are given by the odd solutions of the equations:

$$\tan |a(1 + \cos\gamma) - \psi| = (a^2 - 1)/a$$

and

$$\tan |a'(1 + \sin\gamma) + \psi| = (a'^2 - 1)/a'.$$

The second conclusion is, that there may be at least two formatised frames with widths less than one wavelength.

For an angle of incidence γ equal to 0 degrees or 90 degrees the areas of the two best formatised frames (i.e., frames with the largest critical areas) are $0.24\lambda^2$ and $0.73\lambda^2$, whilst when $\gamma=45$ degrees the critical areas are $0.15\lambda^2$ and $0.50\lambda^2$.

The third conclusion is, that the current in a formatised frame, whether square or not, will vary as the frame rotates about an axis perpendicular to its own plane. The current will pass through maximum and minimum values twice in every rotation.

We may now summarize the main principle which has been developed in the foregoing discussion. In the same sense that the phase lead due to the capacity of an oscillatory circuit may be annulled by an equal phase lag due to the inductance of the circuit with a resulting maximum circuit current; so a phase lead (or lag) due to the width of a frame aërial may be annulled by an equal phase lag (or lead) due to the height of the aërial, with a resulting maximum frame current.

To describe this process of adjusting the dimensions of a frame aërial in order to obtain maximum current, it is convenient to use the word "formatising." We may then speak of a "tuned" and "formatised" frame aërial or of one which has been "tuned," but is "deformatised," and so on. "Formatisation" may be defined as the process of obtaining maximum frame aërial current by the adjustment of the frame dimensions.

PART II

EXPERIMENTAL INVESTIGATION

BY

L. S. PALMER

L. L. K. HONEYBALL

Before it is possible to test the foregoing theory, three practical problems present themselves for solution. First, it is necessary to work with a short-wave transmitter in order that a frame comparable in dimensions with the wavelength shall not be too unwieldy.

Second, it is desirable to be able to alter the angle γ at which the waves are incident on the frame. This might be done by moving the oscillator round the frame keeping it in the plane of the frame, or by rotating the frame about an axis perpendicular to its plane.

Third, the frame must be so designed that it is capable of expansion and contraction in both dimensions.

1. THE TRANSMITTER

In order to design a transmitter to work with adequate power output on wavelengths less than ten meters, it is necessary to overcome the abnormally high losses which occur at these frequencies. This difficulty was not serious in the present work because measurements could be carried out at relatively short distances. The essential factors for the present purpose are that the received current shall be large enough to actuate a vacuum thermojunction and that the distance between the transmitter and receiving frame shall be at least several wavelengths. The distance actually used in the majority of the measurements was about a hundred meters.

A more serious difficulty arose from the fact that in order to get a large oscillatory current in the transmitter inductances it is desirable to have a flat oscillatory circuit. The inductance must therefore be small and consequently will be a poor radiator. This difficulty was largely overcome by using copper tubes for the oscillatory circuit in-

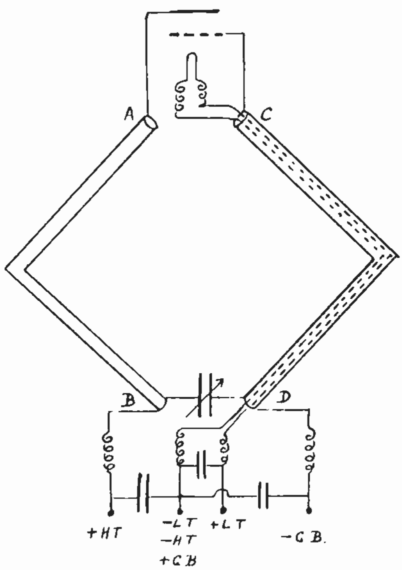

Fig. 4—Schematic of oscillator used in experiments.

ductances. These tubes, even when of considerable size, had sufficiently small inductance values to keep the oscillatory circuit reasonably flat.

A modified Hartley circuit was employed in which the copper tubes were used for the plate and grid inductances. (*AB* and *CD* in Fig. 4.) These tubes were connected at their upper ends by two valves used in parallel, whilst their lower ends were connected by a tuning condenser. The tubular inductances formed a frame about a meter square which constituted the radiating portion of the circuit. Suitable condensers and chokes confined the high-frequency currents to this radiating circuit. The filament leads were passed down through the grid tube which effectively screened them. The radiating unit, consisting of the components shown in Fig. 4, was mounted so that it could be used on the ground or it could be elevated and rotated on the end of a fifteen-foot scaffold pole (Fig. 5). The power to filaments, grids, and plates was supplied by four land lines which approached the radiating circuit on the side remote from the receiving frame. No reradiation could be detected from these lines. It was found that the elevation of the oscillator was of no use for altering the angle γ because the receiving frame was close to the ground where the wave was found to be elliptically polarized and practically independent of the elevation of the oscillator. The effect of this peculiarity of the field is considered below when discussing the experimental results.

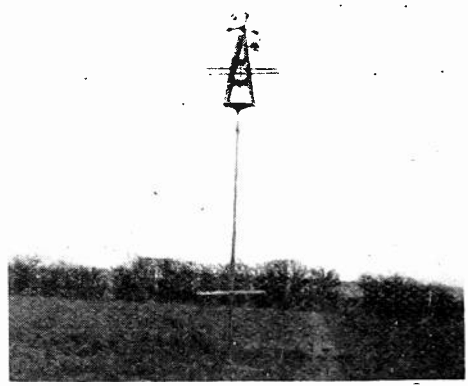

Fig. 5—View of short-wave transmitter used in experiments.

The wavelength range of the oscillator was from 7.54 to 8.80 meters. These wavelengths were checked by measuring the position of potential nodes and antinodes on Lecher wires.

2. The Receiving Frame

The receiving frame was arranged to hang between two sixty-foot scaffold poles which were erected in an open field at a distance apart of thirty feet. The top and bottom wires of the frame were horizontal and the sides vertical. In order to change the dimensions readily the following arrangement was devised. The frame was made of two lengths of bare flexible stranded copper wire. The longer section ran from *D* to *C* via *A* and *B* (Figs. 6 and 7) and the shorter wire stretched directly from *D* to *C*. The longer wire could be wound on two large metal pulleys at *C* and *D*. *A* and *B* were bent tubular insulators through which the longer wire could easily slide. These insulating tubes were fixed on boards about a foot square which ran on wheels along the beam *EF*. They were controlled by an endless cord which enabled them to be moved apart or together after the manner of a stage curtain. The beam

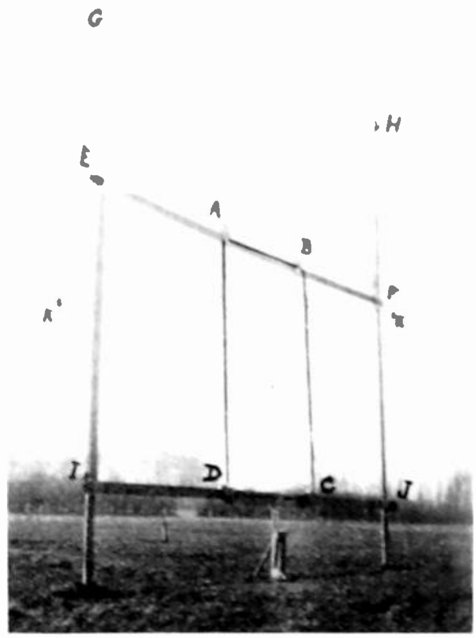

Fig. 6—View of adjustable frame aerial used for receiving.

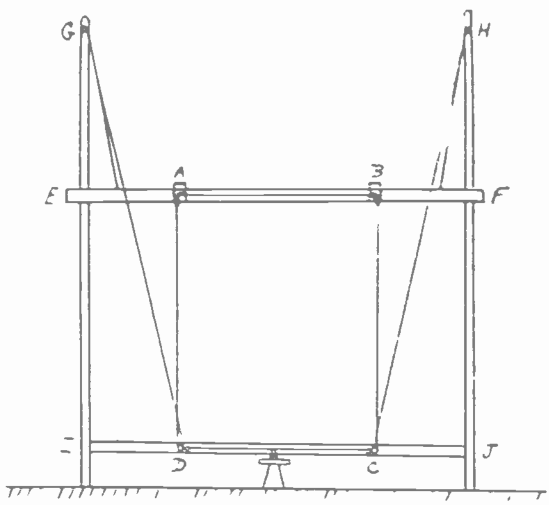

Fig. 7—Mechanical arrangement of adjustable frame aerial.

was supported by ropes running over pulleys at *G* and *H*. The motion of the beam up and down enabled the height of the frame to be altered at will. *D* and *C* could be moved along the fixed beam *IJ* so as to be vertically below *A* and *B*, respectively. In this way the width of the frame was made to vary. In order to adjust the length of wires required for a frame of given dimensions, the metal pulleys D and C were fastened to other wooden pulleys round which a cord was wound in the reverse direction to the wire on the metal pulleys. These cords passed over other pulleys at G and H and heavy lead weights K were suspended from their running ends. (Visible in Fig. 6.) Thus as the frame was expanded, wire unwound from the metal pulleys D and C whilst cord was wound on the wooden pulleys and the lead weights rose. When the frame was contracted the fall of the weights automatically unwound the cord and wound up the unwanted wire. As the wire was bare and the pulleys were of metal the unused wire was automatically shorted. Separate spring pulleys at D and C automatically adjusted the length of the short horizontal wire forming the bottom member of the frame. These spring pulleys were mounted on the same boards as

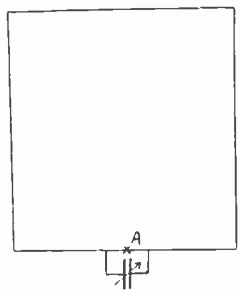

Fig. 8—Electrical circuit of adjustable frame aërial.

those controlling the longer wire and were electrically connected to them by rubbing contacts on the axles.

The scaffold poles and horizontal beams were painted with metric scales from which the frame dimensions could be read.

In order to tune so large an inductance to the short waves it was necessary to shunt the frame with a short piece of copper strip. Across this a condenser, reading up to 0.0005 μf, was connected. To read the current a vacuum thermojunction and microammeter were inserted at A (Fig. 8). This method of tuning ensures that the variation of frame dimensions changes the total circuit inductance by an almost negligible amount, because the inductance of even the smallest frame is very much bigger than that of the copper strip with which it is in parallel.

3. Conditions of Experiment

It was intended to carry out measurements with the plane of the aërial parallel to the direction of propagation. With the previous experiments on Hertzian antennas[2] it was found that this condition was

[2] L. S. Palmer and L. L. Honeyball, "The action of a reflecting antenna." *Jour. I.E.E.* (London), vol. 67, p. 1045; August, (1929).

not easily fulfilled and although the oscillator and antennas were correctly oriented geometrically, the presence of a metal pipe line and a wire fence appreciably deviated the waves. Because of this difficulty, the present site was more carefully selected. A small portable tuned frame with both vacuum thermojunction and ammeter as well as rectifier and headphones was carried round on a pole and the lines of the magnetic component of the radiated field in the neighborhood of the receiving frame were plotted by noticing, with a compass, the orientation of the portable frame for minimum signal strength. In this way the oscillator was so placed that the magnetic component of its field was everywhere perpendicular to the line joining the two scaffold poles.

An attempt was next made to determine the angle γ by using a tilting aërial method similar to that used by Smith-Rose and Barfield.[3] A tuned Hertzian rod oscillator was pivoted about an axis perpendicular to the plane of the frame and rotated until the received current (measured by a vacuum thermojunction and ammeter) was a minimum. This minimum was far from zero in spite of the insertion of a high-frequency rejector circuit and the careful screening of the leads to the ammeter. It was therefore concluded that there was an appreciable horizontal component to the electric vector of the wave. This was to be expected, because, unfortunately, the frame had to be placed fairly close to the ground which is not a perfect conductor. If we take Smith-Rose and McPetrie's recent value[4] for the conductivity of the earth at these frequencies, it is easy to calculate (see below) the ratio of E_H, the horizontal component to E_V, the vertical component, of the electric vector of the wave. Taking the conductivity to be 10^9 e.s.u. and the frequency to be 4.10^7 cycles per second, the value of E_H/E_V is found to be 12.5 per cent. As the exact value is not of immediate moment from the present standpoint, the measurements with the tilting aërial are not being pursued further until the main investigation with the expanding frame is completed. The effect of this elliptical field on the frame aërial measurements is shown in the results given below.

4. Experimental Results

Measurements were made on 7.54, 8.65, and 8.80 meters. The procedure adopted was to raise the horizontal beam to the top of the scaffold poles and to take a series of current readings as the width of

[3] R. L. Smith-Rose and R. H. Barfield, "On the determination of the directions of the forces in wireless waves at the earth's surface," *Proc. Roy. Soc.*, vol. 107A, p. 587, (1925).

[4] R. L. Smith-Rose and J. S. McPetrie, "The attenuation of ultra-short radio waves due to the resistance of the earth," *Proc. Phys. Soc.* (London), vol. 43, pt. 5, p. 592, (1931).

the frame was gradually decreased and then gradually increased. This was then repeated for a smaller height, and so on, till the moving beam was less than a meter above the fixed one. On one or two occasions the width was fixed and the horizontal beam continuously moved. The two methods gave results in complete agreement (See Fig. 9); and hence, as the second method necessitated two assistants, the first method was usually adopted. Operators standing still close to the scaffold poles did not affect the readings. But a six-foot man is a serious absorber of energy at these wavelengths; and, although it was not found necessary to lie down and read the ammeter with a periscopic arrangement of mirrors as was done with the previous work on Hert-

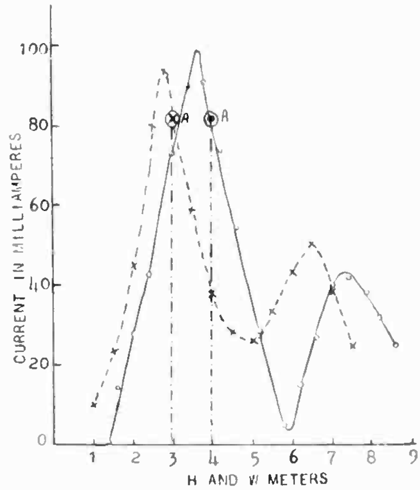

Fig. 9—Variation of received current with dimension of frame aerial (In one case H is kept at 3 meters while W is varied as indicated. In the other case W is kept at 4 meters while H is varied.)
$\lambda = 7.54$ meters
$H = 3$ meters—○—
$W = 4$ meters . . × . .

zian antennas, it was found desirable to remain crouched in one fixed position in order to ensure that bodily movements did not cause current variations. Even with all reasonable precautions the exact values of the actual peak currents are not reliable, but this lack of reliability does not apply to the positions of the peaks with respect to the frame dimensions, and these are the important data from the present point of view.

Current readings were plotted against W for a given value of H. The full line graph in Fig. 9 shows the type of curve obtained by varying the width for a given height of 3 meters, whilst the dotted curve is that obtained by keeping the width constant at 4 meters and gradually changing the height by moving the horizontal beam. The points A on the two curves refer to frames of the same dimensions and it is seen that the two current values are identical. Fig. 10 shows a typical set of curves for values of H varying from 1.0 to 8.0 meters. The chief points to notice from this figure are:

(1) Two peak currents occur, in general, within a width of one wave-length for each value of H;

(2) The positions of the peaks move towards higher values of W as H gets smaller.

(3) the largest peak currents occur for values of W and H which do not differ greatly from each other. That is, the small peak values are

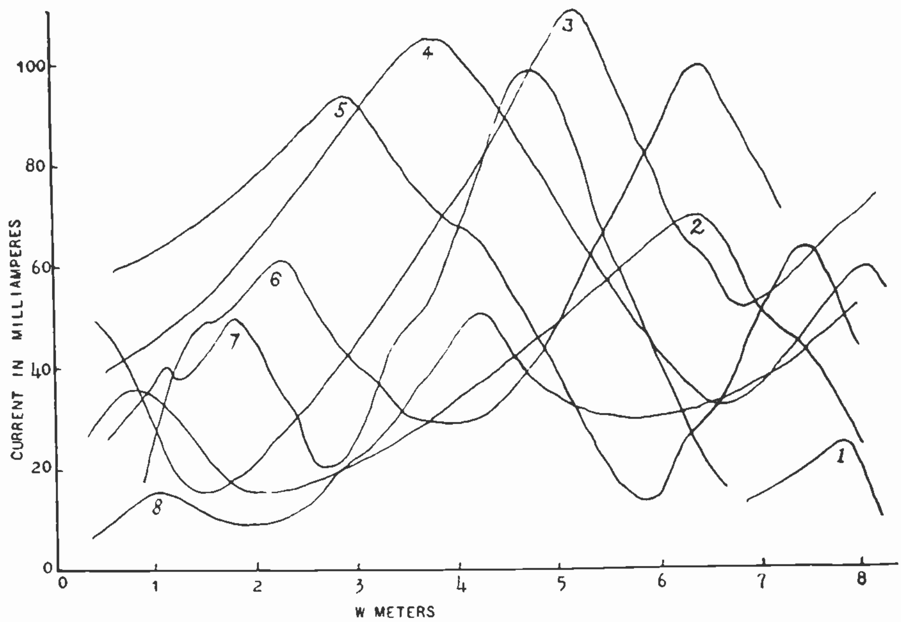

Fig. 10—Variation of received current with width (W) of frame aerial.
$\lambda = 8.65$ meters.
Numbers on curves indicate values of H

associated with either very tall and narrow, or very flat and wide frames.

The first point is a confirmation of the deduction already made from Fig. 3. For example, Fig. 2 depicts two theoretically proportioned frames to which the curve for $H = 5$ meters (or 0.6λ approximately) in Fig. 10 closely corresponds. The theoretical position of the peaks is 0.33λ and 0.85λ, that is about 2.9 and 7.4 meters if $\gamma = 0$. But the theoretical height for these widths is 0.71λ or about 6 meters. This discrepancy in the calculated and observed heights is due to the fact that γ was not equal to zero but was about 8 degrees when the measurements were made. This will be referred to again later, and is also apparent by reference to the points A' on the dotted curves in Fig. 3.

The second point indicates that the positions of the peak currents are varying at least approximately in the manner predicted by the foregoing theory, and it remains to see whether they do actually lie on one or other of the theoretical curves drawn in Fig. 3. By taking the positions of the peaks from graphs such as those in Fig. 10, and plotting the values of H/λ against W/λ the points were all found to lie as shown in Fig. 11, and *not* to coincide with any one of the theoretical curves of Fig. 3. If we now place a trace of Fig. 3 on Fig. 11 we

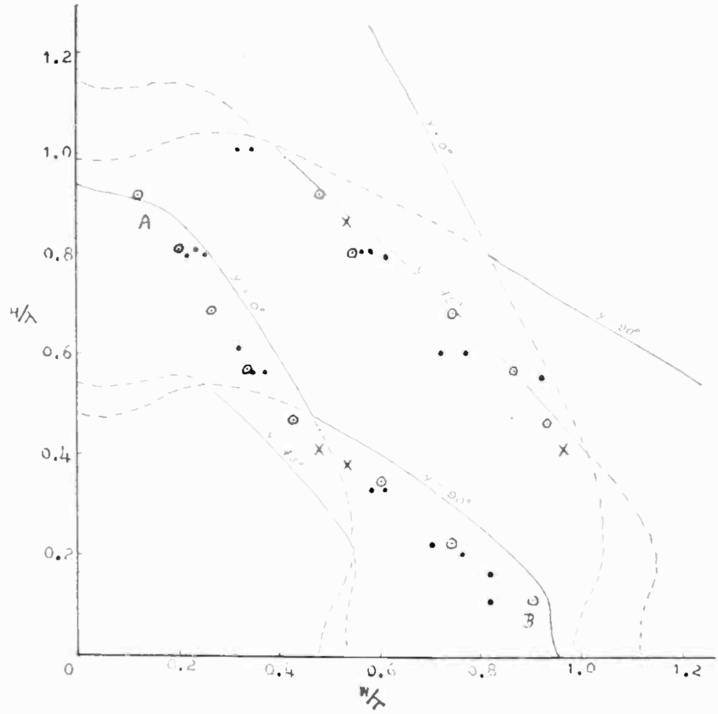

Fig. 11—Comparison of experimental points with calculated curves for various ratios of height (H) and width (W).
o points from Fig. 10. $\lambda = 8.65$ meters.
· points from experiments on $\lambda = 8.80$ meters.
x points from Fig. 9. $\lambda = 7.54$ meters.

find the unexpected result that all the experimental points lie close to the *envelope* of the theoretical curves. To facilitate this comparison the relevant theoretical curves have been redrawn in Fig. 11. The explanation of this seems to be as follows: The wave was found to be elliptically polarized; hence there will always be some component of the electric vector for which $\gamma = 0$ degrees, some component for which $\gamma = 90$ degrees, and some component for all the intermediate values of γ. With a very tall narrow frame as indicated by the point A in Fig. 11, the horizontal component of the electric vector will be practically negligible in its effect on the very short horizontal wires, whilst the vertical component will readily influence the long vertical sides of the frame. Hence point A lies near the graph for $\gamma = 0$ degrees. A similar argument for a flat wide frame leads to the conclusion that the horizontal component is alone operative and hence point B (for which $W/\lambda \gg H/\lambda$) lies near the graph for $\gamma = 90$ degrees.

The frames of more equal dimensions will be affected by both components and hence will have peak current positions on the theoretical graphs for intermediate values of γ.

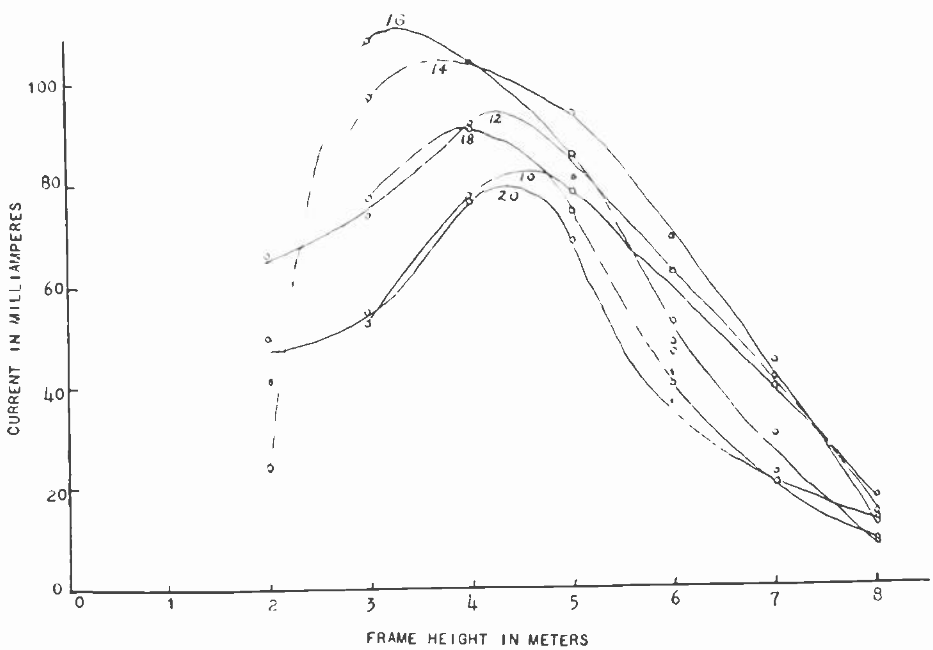

Fig. 12—Variation of received current with frame height for formatised frames of a given area.
Figures on curves indicates values of area in square meters. $\lambda = 8.65$ meters.

In view of the very great difficulties of the experiments it is thought that the greater divergence between the experimental points for the larger frame areas and the theoretical curves is due to some error of measurement peculiar to the larger frames. This error has not been definitely located, but may be due partly to the greater sag in the top horizontal wire when the frame is wide. Correcting this would tend to raise the lower points on the graph, but this defect would not account for the low value of the points for which W/λ is small. The discrepancy does not seem to be of sufficient magnitude to affect the validity of the general theory of the action of a frame aërial which has been discussed in Part I.

The third conclusion (from the graphs of Fig. 10) namely, that the largest currents were produced in those frames which had dimensions not greatly different from each other has already been deduced on page 1351. when discussing Fig. 3. It was there pointed out that a properly formatised frame may have a large or a small area resulting in a large or a small optimum current, and that the large area frames were those with dimensions comparable in magnitude.

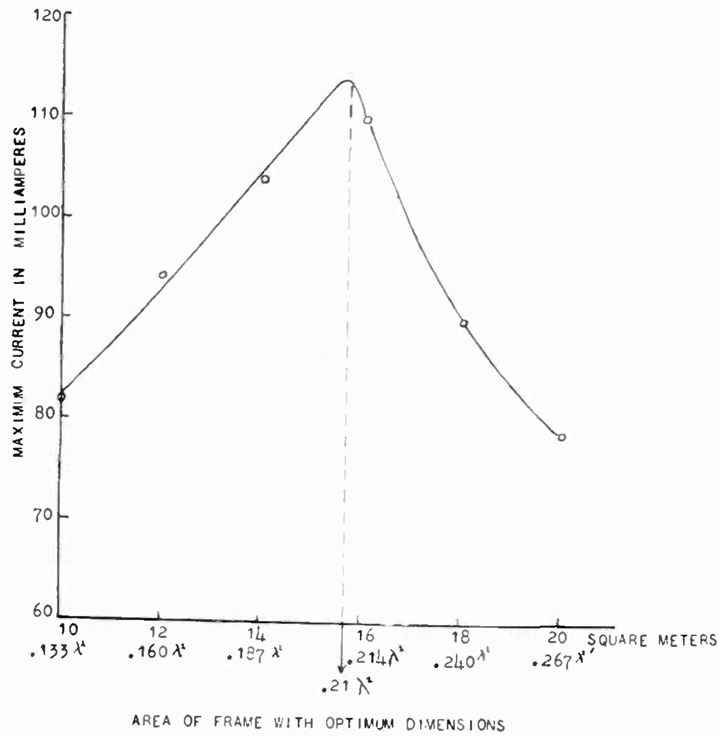

Fig. 13—Variation of received current with area for optimum ratios of width to height.
$\lambda = 8.65$ meters.

A fourth but less obvious conclusion can be deduced from Fig. 10 by tabulating the current values for frames of the same area but of different dimensions. The necessary data can be obtained by considering a given area (say 10, 12, 14 square meters, etc.) and calculating the value of W for the different heights H at which the curves of Fig. 10 were obtained. Then from the graph corresponding to a particular value of H the current value for the appropriate width can be read. This procedure can be repeated for the several selected areas. The data so obtained from the graphs of Fig. 10 are recorded in Table I, and the resulting curves of Fig. 12 show how the current varies as a frame of given fixed area changes its shape. From Fig. 12 we see that there are optimum dimensions for each area and that there is one particular area

TABLE I

Area in Square Meters	10		12		14		16		18		20	
H Meters	W calculated	Current from Fig. 10	W calculated	Current from Fig. 10	W calculated	Current from Fig. 10	W calculated	Current from Fig. 10	W calculated	Current from Fig. 10	W calculated	Current from Fig. 10
1	10.0	—	12.0	—	14.0	—	16.0	—	18.0	—	20.0	—
2	5.0	49	6.0	66	7.0	52	8.0	24	9.0	—	10.0	—
3	3.3	52	4.0	73	4.7	97	5.3	108	6.0	76	6.7	52
4	2.5	77	3.0	91	3.5	103	4.0	103	4.5	90	5.0	76
5	2.0	78	2.4	95	2.8	93	3.2	88	3.6	74	4.0	68
6	1.7	52	2.0	56	2.3	62	2.7	48	3.0	40	3.3	34
7	1.4	39	1.7	48	2.0	44	2.3	32	2.6	20	2.9	22
8	1.25	14	1.5	10	1.75	9	2.0	8	2.25	10	2.5	12

which is better for reception (when of the optimum dimensions) than any smaller or larger area. This is more clearly shown by plotting the maximum possible currents (i.e., the peak values of the curves in Fig. 12) against the corresponding areas. The result is shown in Fig. 13 where the peak occurs for an area if about 0.21 λ^2 (or 15.5 square meters for $\lambda = 8.65$ meters). The foregoing theory predicted a value lying between 0.15 λ^2 and 0.24 λ^2 depending on the value of the angle γ. If the theory be correct then we may conclude that the value of 0.21 λ^2 corresponds to an angle γ of about 8 degrees. This, therefore, would be approximately the slope of the major axis of the elliptical electric field, and from this angle we may calculate the conductivity σ of the ground in the neighborhood of the aërial.

By modifying a formula due to Zenneck, the conductivity $\sigma = f/4 \tan^2\gamma$ where f is the frequency of the wave.[3] For a wavelength of 8.65 meters $\sigma = 3.10^9/4.8\cdot65 \tan^2 8^\circ = 4\cdot4.10^8$ e.s.u. During these experiments the ground was wet and marshy. We may also determine the ratio of the horizontal to the vertical components of the electric vector. This is given approximately by the formula

$$E_H/E_V = \sqrt{f/2\sigma}$$

and for the data given above equals 14 per cent.

These values, from the nature of the very indirect way in which they have been obtained, cannot have any claim to great accuracy. But they agree fairly well with the lowest of the recent values obtained by Smith-Rose and McPetrie, and therefore give further support to the foregoing theory of the action of a frame aërial upon which the above calculation of the slope of the major axis of the elliptical field depends.

The deduction that γ was about 8 degrees when these experiments were performed seems to be confirmed by the appearance of a subsid-

[3] *Loc. cit.*, p. 590.

iary peak on the curve for $H=7$ in Fig. 10. Since $\lambda=8.65$ meters, $H/\lambda=0.81$, and by reference to the dotted portion of Fig. 3 it can be seen that subsidiary peaks may be expected at this value of H/λ if γ lies between 0 and 15 degrees. These subsidiary peaks are indicated by the points C, the second and third of which correspond roughly with the positions of the peaks shown on the curve for $H=7$ in Fig. 10. Unfortunately it was not possible with the frame in use to determine whether a peak corresponding to the first point C was present or not.

In view of the fact that the occurrence of this subsidiary peak is a crucial test of the foregoing theory, a separate series of measurements was undertaken to verify its presence on other wavelengths. These measurements were made in dry summer weather and the peak was again found but was most marked for values of H/λ equal to 0.78 indicating a slightly greater value of γ and a consequent decrease in conductivity.

Conclusion

1. It appears that the resonant current which circulates in a tuned frame aërial under the influence of an electromagnetic wave is not necessarily the maximum current that can be produced by the incident wave. By formatising the frame the current may be increased to a very much larger value.

2. The conditions for formatising a tuned frame lying in the plane of wave propogation are given by the odd solutions of the equations:

$$\tan\left[a(1+\cos\gamma)-\psi\right]=(a^2-1)/a$$
$$\tan\left[a'(1+\sin\gamma)+\psi\right]=(a'^2-1)/a'$$

where

$$a=2\pi W/\lambda,\quad a'=2\pi H/\lambda$$

and ψ is any arbitrary phase angle.

3. The maximum areas of a tuned formatised frame vary from $0.15\lambda^2$ and $0.50\lambda^2$ when $\gamma=45$ degrees to $0.24\lambda^2$ and $0.73\lambda^2$ when $\gamma=0$ degrees or 90 degrees.

Appendix

This appendix is based on a mechanical graph for solving the transcendental equations which determine the formatising conditions.

In order to obtain the curves of Fig. 3 it is necessary to solve equations (5) for different values of ψ. Each value of ψ fixes one point on a curve and each curve is determined by the particular value of γ. The values of a (and hence W/λ) are given by the points of intersection of the graphs:

$$y=(a^2-1)/a \text{ and } y=\tan\left[a(1+\cos\gamma)-\psi\right]$$

whilst the values of a' (and hence H/λ) are given by the points of intersection of

$$y' = (a'^2 - 1)/a' \text{ and } y' = \tan\left[a'(1+\sin\gamma)+\psi\right].$$

To solve these equations the graph $y=(a^2-1)/a$, or the similar graph for a', was plotted on a large sheet of squared paper and the axis of abscissas was raised about one sixteenth of an inch above the surface of the paper by fixing a thin strip of metal along the axis so that it stood perpendicular to the plane of the paper. Along this ledge cardboard templates (e.g., 1, 2, and 3 in Fig. 14) were slid along. These

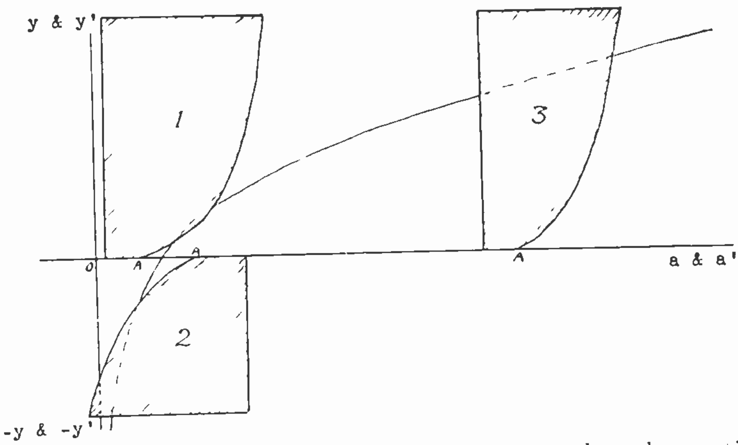

Fig. 14—Mechanical graph for solving transcendental equation.

templates were cut with one edge shaped according to the graphs of the equations $y=\tan a(1+\cos\gamma)$ and $y'=\tan a'(1+\sin\gamma)$ for different values of γ. The points marked A on the raised axis fix the positions of the templates for different values of ψ and these positions were calculated from the equations $a=(n\pi+\psi)/(1+\cos\gamma)$ or $(n\pi-\psi)/(1+\sin\gamma)$. It is readily seen that these equations give the values of a and a' for which y and y' equal zero, respectively.

The required value of a (or a') is then read off from the point where the template cuts the curve drawn on the paper. There may be more than one solution as is shown by the two points where template 1 cuts the curve. Values of a and a' less than unity are obtained by reversing the appropriate template as shown in position 2.

Acknowledgment

We should like to take this opportunity of expressing our thanks to the Department of Scientific and Industrial Research of Great Britain for a grant which enabled us to carry out this investigation.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

WAVELENGTH CHARACTERISTICS OF COUPLED CIRCUITS HAVING DISTRIBUTED CONSTANTS*

By

Ronold King

(Electrical Engineering Department, University of Wisconsin, Madison, Wisconsin)

Summary—*The wavelength characteristics of a simple circuit and of a coupled circuit regenerative oscillator having distributed circuit constants are derived for the case of small resistance and compared with experimental results. A comparison is made with analogous characteristics of circuits with lumped constants and of the electron oscillator.*

THE frequency characteristics of the simpler circuits involving lumped inductances, capacitances, resistances, and negative resistances are well known from both theoretical and experimental investigations. In general, such studies have been limited to frequencies for which interelectrode capacitances of vacuum tubes play no vital part, although more complete researches treating these capacitances have also been made. With the discovery by Barkhausen and Kurz in 1919 of the electron oscillator generating ultra-radio frequencies of wavelengths measured in the small tens of centimeters, an entirely new field was opened. At such tremendously high frequencies ordinary condensers and coils lose their significance as capacitances and inductances, and hence as tuning devices. Here it becomes a question of utilizing the distributed capacitance and inductance of suitably arranged systems of parallel conductors of variable length. With the successful transmission of modulated carrier waves of wavelength in the neighborhood of half a meter accomplished both in this country and abroad, and the interesting possibilities arising from the peculiar properties of such short waves accumulating, it certainly seems in place to investigate theoretically the circuits involved.

In developing new theory, however, it is not sufficient merely to examine certain phenomena; it is also desirable to correlate the results obtained and the methods used with other, similar studies. In this paper it is proposed to analyze theoretically in considerable detail the wavelength characteristics of simple and coupled vacuum tube circuits having distributed constants, and in this way to extend the interpretation of experimental results presented in an earlier paper.[1] The char-

* Decimal classification: R140. Original manuscript received by the Institute, September 4, 1931; revised manuscript received by the Institute, April 29, 1932.

[1] R. King, *Ann. d. Phys.*, Band 5, Folge 7, p. 805, (1930).

acteristics so obtained will be briefly compared with corresponding ones for circuits with lumped constants as described by Ollendorff.[2] The analogy thus exhibited will then be extended to include the electron oscillator in the hope of grouping these three related types of triode oscillatory circuits under the same fundamental theory. With regard to the electron oscillator, reference will be made primarily to the theoretical analysis of Wundt[3] and the experimental data of Hollmann.[4]

Part I. Circuit Analysis

The circuit under consideration is shown in Fig. 1a. It consists of a triode (201-A) and two parallel conductors connected, respectively, to the grid and plate of the tube at one end and bridged by a conducting bridge BB' at any arbitrary point. The length from the bridge BB' to the end EE' is variable. This will be designated as the secondary, while

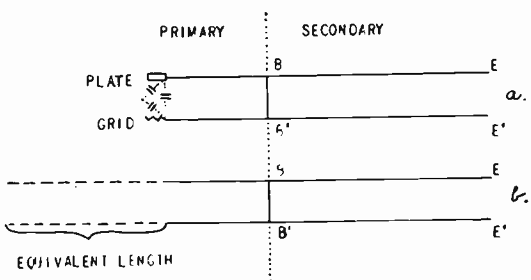

Fig. 1—Actual and idealized triode oscillatory circuit including an open-end variable secondary with distributed constants.

the length from the bridge to and including the triode will be called the primary. The end EE' may be either open or bridged by a conductor. The analysis will be carried through first for the case of an open end and then outlined for the bridged end.

An attempt to set up complete differential equations for the circuit exactly as shown leads to complications on account of the triode included in the primary. In order to simplify the problem it is found convenient to divide the analysis into several parts, as follows: 1. A simple circuit consisting of a triode and two parallel conductors bridged at the end remote from the vacuum tube is first considered. This is essentially the circuit of Fig. 1a, but without the secondary. 2. From simplifications suggested by this analysis, the complete circuit including the secondary is treated under the assumption that the parallel conductors

[2] F. Ollendorff, "Grundlagen der Hochfrequenztechnik," p. 391, published by Julius Springer, (1926).
[3] R. Wundt, *Zeit. für Hochfrequenz*, vol. 36, p. 133; October, (1930).
[4] H. E. Hollmann, *Ann. d. Phys.*, vol. 86, p. 129, (1928); *Zeit. für Hochfrequenz*, vol. 33, p. 128, (1929).

have distributed constants of such relative values that resistance and leakance are both small compared with inductive and capacitive reactance. 3. By graphical methods the wavelength characteristics of the circuit for different lengths of the secondary and a fixed primary are plotted. From values obtained from these characteristics, damping curves are computed. 4. By introducing a negative resistance into the primary, a family of excitation curves is derived. From the three sets of characteristics thus obtained, viz., the wavelength characteristics, the damping curves, and the excitation curves, the behavior of the circuit is completely determined within the limits of the imposed conditions and assumptions. Interpretation of these curves and application to type examples is carried out under 5; the analysis is extended to the case with a bridged end under 6.

1. Wavelength characteristics of the simple regenerative circuit having distributed line constants.

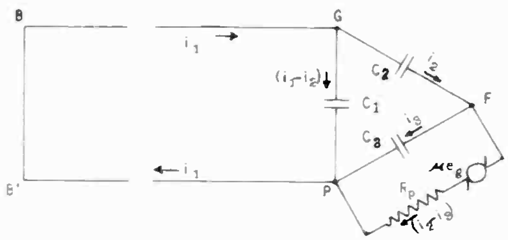

Fig. 2—Schematic circuit diagram of the simple oscillator.

In working with triode amplifiers it is necessary to limit the operating range to the straight-line part of the plate-current grid-voltage characteristic. In the case of oscillators this is not possible since it is precisely the upper and lower curvature in the characteristic which limits the range of operation by decreasing the negative resistance as the amplitude of oscillation becomes sufficiently large. It is, however, possible to treat the characteristics as a family of parallel straight lines extending between horizontally drawn upper and lower limits.[5] With this approximation the negative resistance introduced in the circuit by the tube is, in effect, assumed constant over the entire interval and then considered abruptly reduced to zero at the limits of operation. The approximation is thus inexact only to the extent of neglecting the gradual curving of the characteristics at the upper and lower extremities. Under this assumption it is possible, just as in the case of the amplifier, to regard the characteristics as linear and use only first order terms.

[5] Cf. Gutton, "La Lampe à trois Electrodes," chap. III; p. 67.

The circuit under consideration in this section may be represented schematically as in Fig. 2. BG and $B'P$ are the parallel conductors; BB' is a conducting bridge permanently fixed at their extremities; G, P, and F are, respectively, the triode grid, plate, and filament. The interelectrode capacitances, the internal plate resistance R_p, and the fictitious generator μe_g are represented in the customary way.[6]

Denoting the impedance between GP taken around the distributed circuit by Z', the following equations may be written. For convenience the abbreviation $p = j\omega$ is used.

$$Z'i_1 + (i_1 - i_2)/pC_1 = 0 \tag{1}$$

$$i_2/pC_2 + R_p(i_2 - i_3) - (i_1 - i_2)/pC_1 + \mu e_g = 0 \tag{2}$$

$$i_3/pC_3 - R_p(i_2 - i_3) - \mu e_g = 0 \tag{3}$$

$$e_g = - i_2/pC_2. \tag{4}$$

The following determinant may then be written:

$$D(p) = \begin{vmatrix} Z' + 1/pC_1 & - 1/pC_1 & 0 \\ - 1/pC_1 & 1/pC_2 + R_p + 1/pC_1 - \mu/pC_2 & - R_p \\ 0 & - R_p + \mu/pC_2 & R_p + 1/pC_3 \end{vmatrix} \tag{5}$$

Expanding and equating this determinant to zero gives the simplified expression:

$$p^2 Z'R_pA + p(Z'B + R_pF) + (1 - \mu) = 0 \tag{6}$$

where,

$$A = (C_1C_2 + C_2C_3 + C_1C_3)$$

$$B = C_2 + C_1(1 - \mu)$$

$$F = C_2 + C_3$$

Before attempting to solve this equation it is necessary to evaluate the impedance Z' in terms of the distributed constants of the parallel conductors. Following Cohen,[7] the current at any point x of a parallel line system of length s and bridged at the end $x = s$, is given by:

$$i = \frac{E\sqrt{p + g/c} \cosh K(s - x)}{lv\sqrt{p + r/l} \sinh Ks} \tag{7}$$

where,

$$v^2 = 1/lc$$

$$K^2v^2 = (p + g/c)(p + r/l)$$

$$E \text{ is the voltage impressed at } x = 0$$

[6] *See* for example van der Bijl, "Thermionic Vacuum Tube," McGraw-Hill.
[7] L. Cohen, "Heaviside's Electrical Circuit Theory," McGraw-Hill (1928).

and r, g, l, and c are, respectively, resistance, leakance, inductance, and capacitance per loop unit length of the parallel conductors.

Assuming that $\omega l \gg r$ and $\omega c \gg g$, and remembering that $p = j\omega$, it is clear that $(p+g/c) \doteq (p+r/l) \doteq Kv$. The impedance $Z' = E/i_0$ is, then,

$$Z' = lv \tanh (Ks) \tag{8}$$

or,

$$Z' = lv \tanh (\alpha + j\beta)s \tag{9}$$

where now,

$$K = \alpha + j\beta \text{ where } \alpha = r/lv, \text{ and } \beta = \omega/v.$$

In this preliminary analysis it is sufficiently exact to neglect the resistance of the parallel conductors entirely, thus simplifying (9) to read:

$$Z' = jlv \tan (\omega s/v). \tag{10}$$

Substituting this value of Z' in (6) and writing $j\omega$ for p throughout, the following pair of equations, corresponding to the real and imaginary parts, is obtained:

$$-\omega lvB \tan (\omega s/v) + (1 - \mu) = 0 \tag{11a}$$

$$-j\omega^2 lvR_pA \tan (\omega s/v) + j\omega R_pF = 0. \tag{11b}$$

Either equation yields the same result if the defining relation of μ, viz., $\mu = -C_2/C_3$ is remembered. Solving (11) for ω gives:

$$\omega \tan (\omega s/v) = F/Alv = 1/lvC_t \tag{12}$$

with the quantity F/A, a reciprocal capacity, denoted by $1/C_t$.

In order to obtain a simple picture of the wavelength characteristic of the circuit, it is convenient to consider the two limiting cases: 1. s is so long that the lumped capacitance term C_t is negligibly small compared with the total distributed capacitance sc; 2. s is so short that sc is negligibly small compared with C_t.

Case 1.

Setting $C_t = 0$, since it is to be neglected compared with sc, gives:

$$\tan (\omega s/v) = \infty, \text{ or } \omega s/v = n\pi/2; \quad n = 1, 2, \cdots \tag{12a}$$

then, from the relation $\omega\lambda = 2\pi v$,

$$\lambda = 4s/n. \tag{12b}$$

Case 2.

Taking s very short, $\tan (\omega s/v) \doteq \omega s/v$, so that

$$\omega^2 = 1/C_t ls, \quad \text{or} \quad \lambda^2 = sC_t/4\pi^2 c. \tag{12c}$$

It is thus evident that for short lengths s of the parallel conductors, the square of the wavelength generated is proportional to s, while for extremely long lengths the wavelength varies directly as s. The lower curve of Fig. 3 shows an experimentally determined calibration curve for the circuit under consideration. It clearly agrees with the above derived theory.

If the straight-line part of such a calibration curve carried to still greater values of s be imagined projected back to intersect the negative s-axis ($\lambda=0$) at a point s_1, then the length s_1 may be called the limiting equivalent length of the triode. This length may be directly calculated from the known interelectrode capacitances of the thermionic tube, in this case a 201-A, and the readily evaluated distributed capacitance of the parallel conductors.

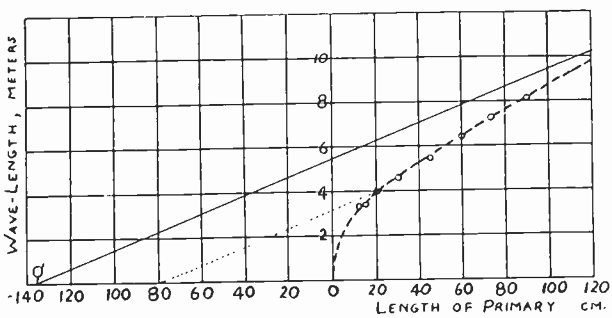

Fig. 3—Actual and idealized calibration curves of the simple oscillator.

Thus, suppose the triode corresponds to a lumped capacitance C_0 connected across the end $s=0$ of a pair of parallel wires. Let the distance from this end to the bridge BB', the current loop, be denoted by s'. The relation between this length and the lumped capacitance C_0 is given by Hund[8] to be:

$$C_0/c = (\cot \beta s')/\beta$$

where c is the capacitance per loop unit length of the parallel conductors. This expression may be written as follows,

$$C_0/c = (\tan \beta s_1)/\beta$$

with $s_1=(\lambda/4-s')$ = the equivalent length of the capacitance C_0. This relation follows at once from the fact that for $C_0=0$, the first equation above reduces to (12a) giving $s_l'=\lambda/4$. Hence $(\lambda/4-s')$ must be the shortening from the ideal extension effected by the capacitance C_0. Since it is desired to obtain the limiting equivalent length of C_0 as the length s' and with it the natural wavelength of the circuit become larger and larger, it is clear that $s_1 \ll \lambda$ so that,

[8] A. Hund, Scientific Paper of the Bureau of Standards, No. 491, page 536.

$$\tan \beta s_1 = \tan (2\pi s_1/\lambda) \doteq \beta s_1$$

and hence,

$$s_1 \doteq C_0/c.$$

This expression, then, gives the limiting equivalent length s_1 of the effective capacitance C_0 of the triode.

An approximate numerical calculation for the circuit constants and triode constants of the oscillator for which the calibration curve of Fig. 3 was plotted, gives[9]:

$$C_0 = C_{pg} + C_{pf}C_{gf}/(C_{pf} + C_{gf}) = 11 \times 10^{-12}\ F.$$

$$c = 0.081 \times 10^{-12}\ F.$$

$$s_1 = 136\ \text{cm}.$$

Inasmuch as the oscillating circuit has one open end and one bridged end, the total length for a purely linear relation between s and λ should be a quarter wavelength, $s_l' = \lambda/4$. Plotting $s_l' = \lambda/4$ from the point $s_1 = -136$ as origin gives the straight-line curve shown in Fig. 3. It is clear that the experimental curve approaches this asymptotically as s becomes larger and larger in accordance with the prediction of the theory.

On the basis of this analysis the assumption will now be made that the primary circuit of Fig. 1a may be idealized into two parallel conductors of length equal to the equivalent length of the triode plus the actual length of parallel conductors attached. This is indicated in Fig. 1b. By equivalent length of the triode is meant that length of parallel conductors which must be added to the obvious primary to make the combined length one-quarter of the wavelength generated. Thus for a generated wavelength of four meters, the calibration curve of Fig. 3 indicates an actual primary length of 20 cm. The quarter wavelength is one meter, so that the equivalent length of the triode must be 80 cm; the ideal calibration curve for this value is shown by the dotted line. In the limit as the length of the actual primary is increased more and more and the generated wavelength becomes greater and greater, the actual calibration curve approaches the limiting asymptote. The limiting value of the equivalent length of the triode is, then, the length $00' = 136$ cm as obtained above.

2. Analysis of the idealized two-circuit oscillator having distributed constants.

[9] Constants for 201-A tube are taken from p. 197, Peters, "Thermionic Vacuum Tube Circuits." The formula for the capacitance of two parallel conductors was taken from Circular 74, Bureau of Standards.

Assuming, on the basis of the foregoing discussion, that the primary may be approximately represented by two parallel conductors of length equal to the equivalent length of the triode plus the length of conductors actually forming the obvious primary, the circuit to be analyzed is that of Fig. 1b.

The general expression for the current at any point a distance x from an applied voltage at $x=0$ in a pair of uniform parallel conductors of total length s and open at the end $x=s$, is

$$i = \frac{E\sqrt{p+g/c}\,\sinh K(s-x)}{lv\sqrt{p+r/l}\,\cosh Ks} \tag{13}$$

with the symbols meaning the same as before. In the idealized circuit under consideration the applied voltage for each end; i.e., for primary and secondary, is the voltage across the bridge BB'; it being tacitly assumed that the two branches react upon each other only through the bridge. Let the bridge BB' be taken as the origin of two systems of coördinates, $x_1=x_2=0$, with the subscripts referring respectively to primary and secondary, and both distances measured in opposite directions from BB'. Let the resistance of the bridge be negligible; let its inductance be M. Then the voltage across the bridge is $E=pM(i_{10}-i_{20})$. The first subscript here refers to the circuit, the second to the value of x at which the current is taken.

Substituting this value of E in (13) written for each circuit with $x_1=x_2=0$, gives, again assuming $(p+g/c)\doteq(p+r/l)\doteq Kv$:

$$i_1 = (pM/lv)(\tanh Ks_1)(i_{10}-i_{20}) \tag{14a}$$

$$i_2 = -(pM/lv)(\tanh Ks_2)(i_{10}-i_{20}). \tag{14b}$$

Eliminating the currents by subtraction leaves

$$lv/pM = \tanh Ks_1 + \tanh Ks_2. \tag{15}$$

Making use of notation already introduced and the relation $M/l=k$, the above equation reads,

$$1/j\beta k = \tanh(\alpha+j\beta)s_1 + \tanh(\alpha+j\beta)s_2. \tag{16}$$

Separating the real and imaginary parts of the hyperbolic tangents gives:

$$\frac{1}{j\beta k} = \frac{\tanh(\alpha s_1)+j\tan(\beta s_1)}{1+j\tanh(\alpha s_1)\tan(\beta s_1)} + \frac{\tanh(\alpha s_2)+j\tan(\beta s_2)}{1+j\tanh(\alpha s_1)\tan(\beta s_2)}. \tag{17}$$

Since the resistance is assumed small, terms of higher order than the first in α may be neglected. Omitting these, separating real and imaginary terms, and simplifying, leaves two expressions as follows:

$$-1/\beta k = \tan\beta s_1 + \tan\beta s_2 \tag{18}$$

$$-\frac{\tanh\alpha s_1}{\tanh\alpha s_2} = \left(\frac{\cos\beta s_1}{\cos\beta s_2}\right)^2. \tag{19}$$

These are the two fundamental equations from which the required characteristics may be evaluated.

3. Wavelength characteristics and damping curves.

Equation (18) involves only the propagation constant β as a function of the length (s_2) of the secondary; the primary being held fixed at a length arbitrarily chosen as unity ($s_1 = 1$). A somewhat laborious

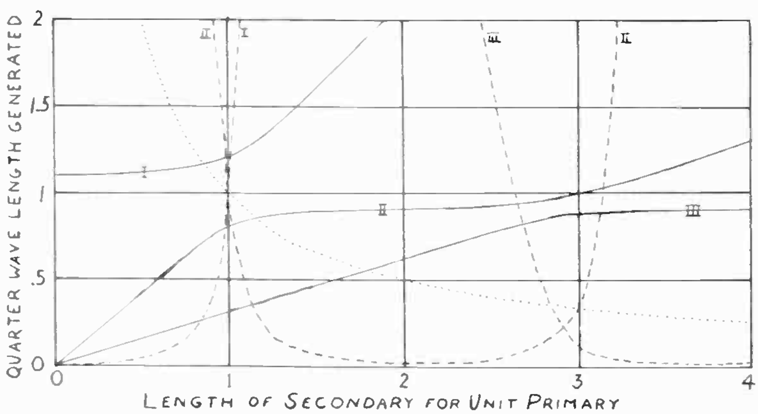

Fig. 4—Theoretical wavelength characteristics (solid lines), damping curves (dashed lines), and excitation curves (dotted lines), of the idealized coupled circuit oscillator with an open-end secondary. The curves are for $k=0.1$. The ordinates for the damping and excitation curves are not given on the figure; they are the same for both sets of curves, and are equal to $(N-r)/rs_2$ in which N, the negative resistance, is the variable parameter.

graphical solution of (18) involves the following steps.[10] Selecting a sufficiently large scale, the curve $y = \tan\beta s_1$ is plotted with β as abcissa. For a given value of the coefficient k the curve $y = -1/\beta k$ is plotted. Then, for a series of suitably chosen values of s_2 a family of curves $y = \tan\beta s_2$ is plotted. The intersections between $-1/\beta k$ and the sum $y = \tan\beta s_1 + \tan\beta s_2$ determine β as a function of s_2. A curve so evaluated and using in addition the relation $\Lambda = 2\pi/\beta$ is shown in Fig. 4 (solid lines). Here $\Lambda/4$ is the ordinate and s_2 the abscissa; k is conveniently chosen as 0.1.

The damping curves are readily evaluated from (19) and the

[10] This is more fully described for the parallel case with a bridged end in conjunction with equation (22a).

values of β obtained from the wavelength characteristic just described. The ratio $-\tanh \alpha s_1/\tanh \alpha s_2$ plotted as ordinate against s_2 (again for $s_1 = 1$) gives the damping curves for the corresponding branches of the wavelength characteristic. They are shown in dashed lines in Fig. 4.

4. The primary negative resitance.

Up to this point the circuit has been analyzed on the assumption that primary and sceondary each consist merely of lengths s_1 and s_2 of parallel conductors coupled by an inductive bridge BB'. No mention has been made of an energy supplying device. This procedure is consistent with the customary practice used in determining the vibration moduli of circuits, and is correct provided the frequency is indeed completely independent of the source of energy. Since the above analysis has taken into account the interelectrode capacities of the triode, which in this case supplies the energy, and since in the regenerative hook-up here used the vacuum tube may be regarded as a pure negative resistance which does not, within the approximations here made at least, affect the frequency prescribed by the circuit constants, the above procedure is justified.[11]

Before the frequency or wavelength characteristic of the circuit can be completely specified, it is necessary to take into account the negative resistance introduced into the primary by the triode. The exact formula for this negative resistance is, for present purposes, immaterial. And since a complete solution of (6) including the distributed resistance of the parallel conductors, (this was neglected in obtaining (11) from (6)), leads to a complicated cubic involving hyperbolic functions, it will not be carried out. A general notion of the type of term involved may be gained, however, by assuming Z' in (6) a pure resistance. By equating the coefficient of the first power term in p to zero, the negative resistance is at once obtained.

The primary has already been assumed idealized into two parallel conductors by the introduction of the concept of equivalent length. It is, therefore, not inconsistent, and mathematically very convenient to assume the negative resistance likewise linearly distributed over this fictitious extension. Moreover, and as has already been stated in conjunction with the limits of operation, the negative resistance will be assumed constant over the entire range. If, therefore, N be defined as the negative resistance introduced by the triode into the primary per loop unit length, the primary damping constant becomes $\alpha_1 = (r - N)/lv$.

[11] It may be noted here that in the Barkhausen-Kurz hook-up discussed in Part III the frequency is no longer a function only of the circuit constants in the sense here used.

The secondary damping constant is unchanged except for the addition of a distinguishing subscript, viz., $\alpha_2 = r/lv$.

The damping curves as plotted in Fig. 4 have as ordinates $y = -\tanh \alpha_1 s_1/\tanh \alpha_2 s_2$ and as abscissas the length of the secondary, s_2; $(s_1 = 1)$. Since α_1 and α_2 are both very small, the hyperbolic tangents may be replaced by their arguments within the range of s_2 here involved. Substituting the above values of α gives the equation:

$$y = -s_1(r - N)/rs_2 = -(r - N)/rs_2; \text{ (for } s_1 = 1\text{)}. \qquad (20)$$

Here $(r-N)/r$ is a constant depending upon the value of N. Selecting several values of N, a family of excitation curves may be plotted with the values of y given by (20) as ordinates and s_2 as abscissas. In Fig. 4 the principal curve of the family, viz., for $N = 2r$, is shown (dotted line). Curves for $N > 2r$ and for $N < 2r$ are easily visualized respectively above and below this reference curve.[12]

5. Interpretation of the theoretically derived characteristics.

A description of the wavelength characteristic of the coupled circuit, distributed constant oscillator is now in order. With the primary fixed at any desired length, here taken as unity, and the oscillator generating a wavelength determined by a calibration curve like that of Fig. 3, it is desired to specify the effect upon this wavelength of the addition of a variable-length secondary. In appearance this secondary is merely an extension beyond the bridge BB' of the parallel conductors forming the primary.

The wavelength characteristic of Fig. 4 shows the theoretical change in wavelength of an oscillating primary of unit length s_1 as the secondary is increased from $s_2 = 0$. The curve consists of several distinct branches with no transitions from one to the other indicated. These transitions, occurring in the neighborhood of $s_2 = 1, 3, 5, \cdots$ are uniquely determined by the intersections of the appropriate branches of the damping curves with an excitation curve specified by the particular value of negative resistance introduced by the triode in question. To each branch of the wavelength characteristic belongs a corresponding branch of the damping curves. The related branches of the two sets of characteristics are numbered alike in Roman numerals. The intersection of an excitation curve with a branch of the damping curve represents a limiting value. Thus, if a given branch of the damping curve crosses the excitation curve in question, it means that for all

[12] It would be graphically more convenient to plot reciprocal ordinates of the damping curves and excitation curves since the latter would then be straight lines. Since the above choice agrees with Ollendorff's curves for lumped constants, it is retained.

values of s_2 at which the damping curve is above the excitation curve, the wavelength corresponding to this branch of the damping curve will be completely damped out. On the other hand, over the entire range of s_2 for which a branch of the damping curves lies below the excitation curve, sufficient energy is available so that the corresponding wavelength is not damped out once the oscillator has started generating it. A few illustrations will help to make this clear.

Consider first the case for which N is less than r. From (20) it follows that the excitation curve for this value of N lies below the axis $y=0$, hence it lies below all branches of the damping curves at all points.

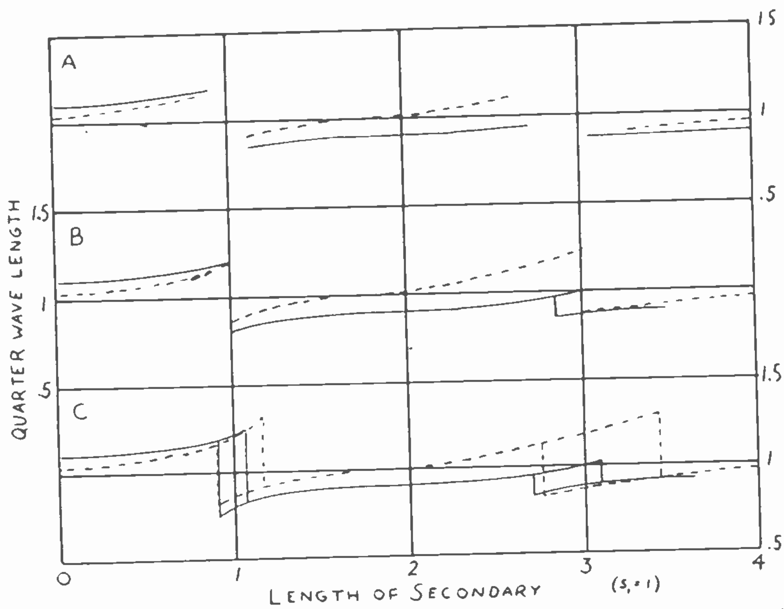

Fig. 5—Typical wavelength characteristics of the idealized coupled-circuit oscillator with an open-end secondary. The negative resistance is taken small for A, near critical for B, large for C.

Therefore all wavelengths are damped out: the oscillator cannot oscillate.

Now let N lie between r and $2r$ so that the excitation curve lies between that drawn in Fig. 4 and the axis of abcissas. Beginning at $s_2=0$, it is clear that the damping curve of branch I of the wavelength characteristic is below the excitation curve. The oscillator will therefore generate a wavelength indicated by this branch. As s_2 is increased, i.e., the secondary lengthened, the wavelength follows branch I of the characteristic until damping curve I crosses the excitation curve. Beyond this point damping curve I rises above the excitation curve and the wavelength given by branch I of the wavelength characteristic is damped out. Since no other branches of the damping curve now lie below the excitation curve for say $N=1.5r$, the oscillator must stop oscillating. This is illustrated in curve *A* (solid line) of Fig. 5. A little beyond $s_2=1$, branch II of the damping curve crosses below the excitation curve, hence the oscillator will resume oscillating but now at a wavelength specified by branch II of the wavelength characteristic. The wavelength generated follows branch II until damping curve II rises above the excitation curve near $s_2=2.75$ where oscillations again break off. They again set in and follow branch III as soon as damping curve III falls below the excitation curve.

For $N=2r$ a similar procedure may be followed. The curve for this value of N is shown dotted in Fig. 4. In this case it is seen that instead of the oscillation breaking off over the range near $s_2=1$, the wavelength changes abruptly but continuously from one branch of the wavelength characteristic to another. For ranges near $s_2=3, 5, \cdots$ the wavelength likewise changes abruptly and continuously, but the transition point is not the same when s_2 is increased over these values as when it is decreased. In other words, there are overlapping ranges within which either of two wavelengths may be generated since both branches of the damping curve lie below the excitation curve. The wavelength actually generated will be the one which happens to have been started. Thus, if s_2 is increased over this overlapping range and then decreased, the wavelength generated will form a so-called loop or Ziehschleife, as the German puts it. Such loops can obviously only occur with values of N for which the excitation curve lies above both branches of the damping curve over a smaller or larger range. Fig. 5b shows a characteristic of this type. It is clear that a value of N can be found for which the transition at $s_2=3$ is single valued, but then the transition at $s_2=1$ is necessarily discontinuous.

For values of N greater than $2r$ the excitation curves lie above both branches of the damping curves at all transition points. Loops will therefore appear at $s_2=1, 3, 5, \cdots$ Fig. 5c shows this situation. The curves indicate that the loops should grow wider; i.e., extend over a greater range of variation in s_2, as the distance of the transition point from $s_2=0$ increases.

There are, then, three possibilities: For relatively low values of negative resistance oscillations break off completely for small ranges in the neighborhood of $s_2=1, 3, 5, \cdots$ For a critical value of N the oscillations do not break off and the wavelength changes abruptly near one of the above values of s_2. For large values of N loops or overlapping ranges in which either of two wavelengths may be generated appear.

The significance of the coefficient k, so far arbitrarily set at 0.1, remains to be discussed. From the definition of k a variation in the value assigned to it is equivalent to changing the inductance of the bridge

BB' common to primary and secondary. To increase k means to increase the inductance M of this bridge. Computed wavelength characteristics over the first transition range are shown in Fig. 6 for three different values of k. The fourth characteristic shown is an experimental curve. Clearly the smaller k is, the less the wavelength departs from the fundamental ordinate $\Lambda/4$, the ideal or theoretical quarter wavelength, and the more nearly it follows the quasi asymptotes $\Lambda/4 = s_2/1$ and $\Lambda/4 = s_2/3$. The wavelength jumps must, therefore, also be corre-

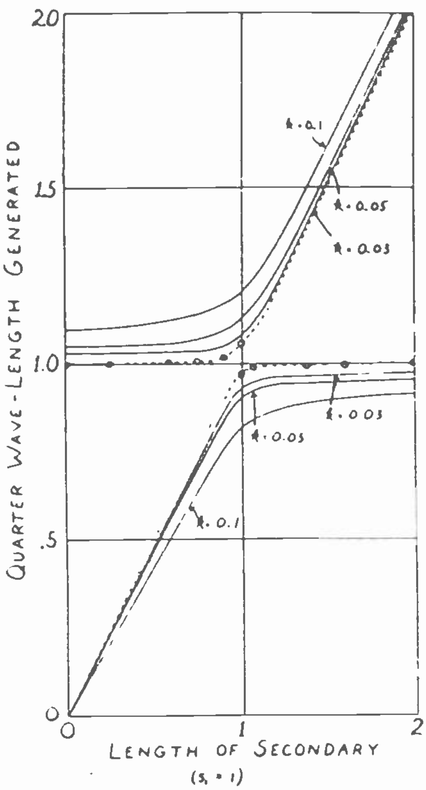

Fig. 6. Theoretical wavelength characteristics of the idealized coupled-circuit oscillator with open-end secondary for three different values of the coefficient k, (solid lines). A typical experimental curve is shown dotted.

spondingly smaller. Likewise, the smaller k is, the steeper the damping curves rise and the narrower possible breaks or loops in the wavelength characteristic must be. These conclusions mean nothing more than that the smaller the inductance M of the bridge BB' the less will be the effect of the secondary upon the frequency generated.

Intimately related with the coefficient k is the meaning of the ideal quarter wavelength used as ordinate in the theoretical wavelength characteristics. Let it be noted, and this was not emphasized above, that the bridge BB' joining the two parallel conductors and serving as the common part of the two coupled circuits is always a part of the primary even when the secondary length is reduced to zero. The significance of this statement lies in the fact that the ideal quarter wavelength is that value which a primary open at one end and bridged by a *perfect* conductor at the other would generate. Now it must be kept in mind that a perfect conductor is one having zero distributed inductance as well as zero resistance. This means nothing less than that the constant $k = M/l$ must vanish if this ideal quarter wavelength is to be generated. Clearly in practice k can never be identically zero, so that the quarter wavelength actually generated even by the primary alone with the secondary completely removed would not be this ideal value, but one depending on the actual value of k as determined by the inductance of the bridge BB'. This point will be taken up again when experimental curves are examined.

6. The analogous case with a bridged end.

The object of the mathematical analysis in this instance is the circuit shown in Fig. 7. Since this differs from the circuit of Fig. 1 only in having a secondary with a bridged end instead of with an open one,

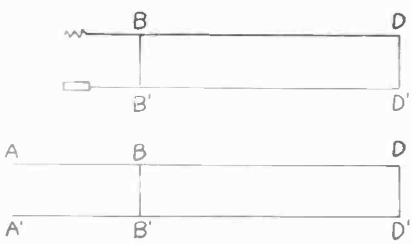

Fig. 7—Schematic diagram of actual and idealized circuits for a bridged-end secondary.

the steps in the analysis are entirely parallel to those of the case just discussed. The primary is clearly entirely unchanged, and the same idealization will be assumed. Following, then, exactly the same argument developed in detail above, the equation corresponding to (16) is:

$$1/j\beta k = \tanh(\alpha + j\beta)s_1 + \coth(\alpha + j\beta)s_2. \tag{21}$$

After separating the real and imaginary parts and neglecting squared terms in α the two equations corresponding to (18) and (19) are:

$$-1/\beta k = \tan \beta s_1 - \cot \beta s_2 \tag{22}$$

$$-\frac{\tanh \alpha_1 s_1}{\tanh \alpha_2 s_2} = \frac{\cos^2 \beta s_1}{\sin^2 \beta s_2}. \tag{23}$$

The graphical solution of (22) is essentially the same as that already carried through, but since it was not described in full above, a small scale reproduction of the solution for the present case is shown in Fig. 8. With the length of the ideal primary again taken as unity and the constant k as before conveniently chosen as 0.1, (22) reads:

$$-10/\beta = \tan\beta - \cot\beta s_2 \qquad (22a)$$

in which s_2, the length of the secondary per unit primary, is to be assigned values ranging from zero to several times the length of the primary. Fig. 8 shows the curve $y = -10/\beta$ plotted against an arbitrary ordinate with β as abscissa. The curve $y = \tan\beta$ is also shown. Sections of a series of curves representing $y = -\cot\beta s_2$ for various values of s_2

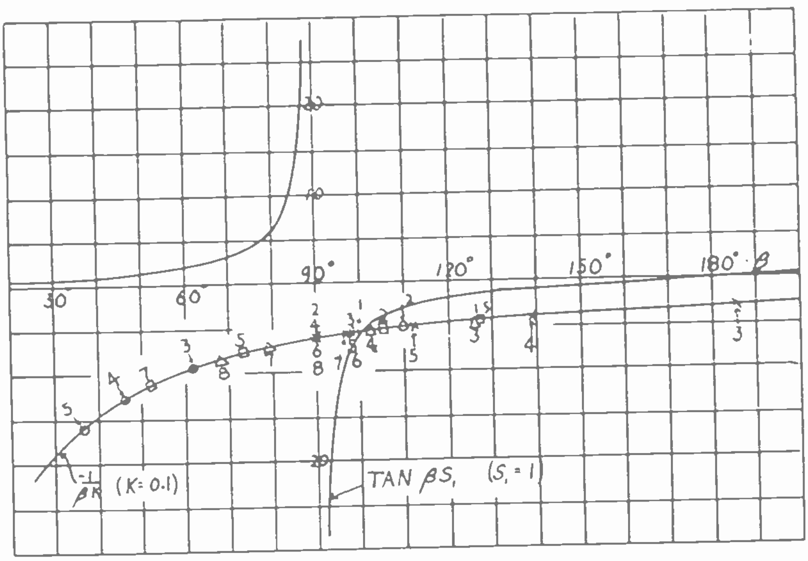

Fig. 8—Graphical solution of (22). Intersections with the first four branches of the cotangent curve are denoted respectively by circles, squares, triangles, and crosses. The numbers written in are the values of s_2 in each case.

(as multiples of s_1) were next drawn and added to the curve $y = \tan\beta$. The points of intersection of these sums with the curve $y = -10/\beta$ gave, along the axis of abscissas, the values of β corresponding to the particular choice of s_2. Since the cotangent curve has an infinite number of branches, an infinity of values of β for each choice of s_2 is a possible solution. However, only the first four of these values were obtained over the limited range of s_2 here required. In Fig. 8 the numerical values of s_2 are written in for the respective points. The family of such points belonging to each of the four branches of the cotangent curve are distinguished in the notation. Plotting the values of β so obtained in terms of the corresponding wavelengths gave the solid curves of Fig. 9. The ordinates are, as in Fig. 4, the ideal quarter wavelength; the abscissas are the values of s_2 for unit s_1.

The damping curves, which are shown in dashed lines in Fig. 9, are readily obtained by substituting the values of β read from the wavelength characteristic just determined into (23). Each family of values of β derived from the individual branches of the cotangent curve and determining one continuous wavelength characteristic, leads to a distinct and likewise continuous damping curve. The ordinates of the damping curves of Fig. 9 are the same as those of Fig. 4. The interpretation of the curves is again carried out by means of excitation curves as given by (20). The excitation curve for $N=2r$ is drawn dotted in Fig. 9.

With the wavelength characteristics, the damping curves, and the

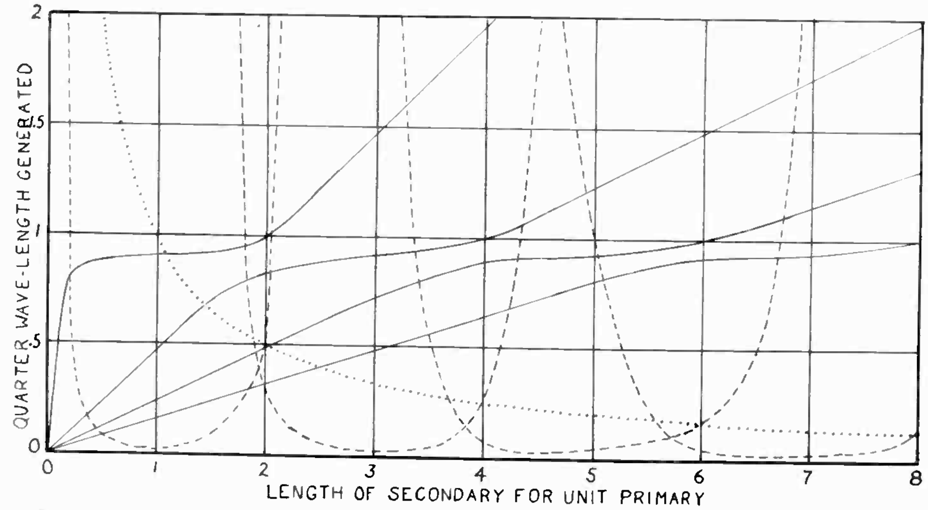

Fig. 9—Theoretical curves for wavelength (solid lines), damping (dashed lines), and excitation for $N=2r$ (dotted line) for the bridged-end secondary. The legend is the same as that of Fig. 4.

excitation curves obtained, the transition from one branch of the wavelength characteristics to another is determined by the intersection of damping and excitation curves precisely as described at length above. In this case the quasi asymptotes are $\Lambda/4=s_2/2$ and $\Lambda/4=s_2/4$.

Part II. Comparison with Experiment

With the theoretical investigation of the circuits of Figs. 1 and 7 completed and the wavelength characteristics determined, a comparison with experimental data is in place. Indeed, in the light of the numerous and certainly not insignificant assumptions and approximations which convenience in mathematical handling imposed, an experimental check should prove highly interesting. Before proceeding to an examination of the available experimental results, however, it will be advantageous to review briefly the salient points of the theoretical analysis with a view to summarize and examine the practical significance of the various assumptions made. It will be convenient to divide the qualifying conditions and approximations made in the course of the analysis into two groups, physical assumptions and analytical assumptions, and to discuss these separately.

Physical assumptions.

The conditions imposed in this group pertain primarily to the constants of the complete circuit of Fig. 1a. With reference to the distributed circuit and the connecting bridge the following may be listed:

(a) Distributed resistance and leakance are small compared with inductive and capacitive reactance per loop unit length.

(b) The resistance of the bridge BB' is small compared with its inductive reactance.

(c) The length of the secondary is not so great that the approximation $\tanh \alpha_2 s_2 = \alpha_2 s_2$ is not justified.

For ultra-radio frequencies and circuits using copper conductors of common size these conditions are automatically satisfied. For example, with $\omega = 4 \times 10^8$ and No. 12 copper wire, the ratio $r/\omega l$ is of the order of magnitude of 10^{-3}, and $\alpha = r/lv$ of the order of 10^{-5}. Clearly to neglect α^2 along the wires, and r in comparison with ωl in the bridge is entirely justified. Since $\tanh x = x - x^3/3 + 2x^5/15$. The secondary would have to become over a kilometer long before the substitution of the argument for the hyperbolic tangent becomes a poor approximation.

Two other physical assumptions are made.

(d) The frequency is determined entirely by the circuit constants and is independent of the negative resistance introduced.

(e) The coupling, or region of energy exchange between primary and secondary, is limited to the common bridge BB'.

It is a matter of common experience that the first of these is in general very nearly true of triode oscillators of the regenerative type. With regard to the second, the only other means of energy exchange between primary and secondary is by radiation. So long as primary and secondary are placed end to end as supposed in the theory, there can be no radiation coupling between them. If the two circuit branches were arranged mutually at right angles, on the other hand, their fields could interact. In this case an extension of the theory to include radiation coupling would be necessary.

Analytical assumptions.

The specializations in this group were made for mathematical convenience. Essentially they all reduce to the one fundamental approximation, namely, the substitution of circuit b for a in Fig. 1. On the basis of the physical assumptions already discussed, circuit b has been quite rigorously analyzed; the question only remains whether the characteristics of circuit 1b do approximate those of 1a. In final analysis this is a question for experiment to decide, but a few significant remarks can be made. In the first place, it is to be noted that to introduce an equivalent length and with it an ideal straight-line calibration characteristic for the primary is in no way to change the physical fact that for a given fixed length the primary alone will generate a definite wavelength. Indeed, so long as the primary and the wavelength are maintained fixed, once an equivalent length has been chosen to represent them analytically, the change from a to b is more in the form of notation than in the physics of the problem. Of course, this is not rigorously true since the response of the idealized circuit to a coupled secondary, a response which does involve wavelength variation, would hardly be expected to be identically that of the actual circuit, even though the two were initially made to have the same natural vibration modulus. Roughly, however, the difference might be described as a second order one. In the same way the introduction of a constant and distributed negative resistance which has no definite physical equivalent, but which, to a first approximation at least, can be made to describe the effect should represent the actual case reasonably well.

In summing up it may be said, then, that physical and analytical assumptions do not seriously restrict the generality of the problem.

Turning now to an experimental verification, the data available are divided into two parts. First, there are the wavelength curves obtained in an earlier paper already mentioned[1] for the case of an open-end secondary. These cover a somewhat limited range of variation in secondary length and do not show the especially interesting phenomenon of overlapping loops (Ziehschleifen). Moreover, and as will be discussed in considerable detail below, the experimental set-up was such as to introduce a complicating interaction between the circuit branches. Since this is in itself of considerable interest, this early data will be considered as an example of the open-end case. Second, and in order to verify accurately the theoretical predictions, a special set of data was taken, this time with a bridged-end secondary. It will be primarily to these latter results that reference will be made for a precise study of the adequacy of the theory as developed.

1. The Open-End Case.

To turn to the open-end case, a study of the circuit used in obtaining the experimental data[13] reveals that all the physical assumptions

[13] The experimental data to be here considered are taken from the paper re-

made in the foregoing analysis apply except only one. In this experiment it was found necessary to arrange the primary and secondary mutually at right angles in order to avoid a progressive increase in the coupling with the measuring system of Lecher wires as the secondary was extended. As has already been mentioned, the assumption that the bridge BB' is the only region of energy exchange between the two circuit branches is no longer justified when these are placed mutually at right angles. Indeed, an experiment with a single circuit oscillator in which this is placed at right angles to a pair of Lecher wires shows that energy is transferred to the wires even when the oscillator is moved 50 cm beyond their end. It is thus clear that with the right-angle setting of primary and secondary used in the experiment, an energy exchange by radiation must take place. In other words, in addition to the inductance of the bridge which is common to both circuits, a mutual reaction distributed along the conductors forming the primary and secondary also plays a part. It is, therefore, necessary to extend the theory to include this additional effect before the experimental curves can be considered.

In order to show in an approximate way the effect of adding a distributed mutual term, it is convenient to reanalyze the circuit of Fig. 1b taking into account only such a mutual term for the coupling between the circuit branches. The actual characteristic should then be obtained roughly by an averaging of the characteristic for bridge coupling with that for radiation coupling. A simple way of approximating such a distributed coupling is to include a distributed mutual factor m (which may be a function of the linear dimension x) with the four distributed constants r, g, l, and c. Taking such a factor into account the familiar partial differential equations, known as the "long-line equations," assume the following form when written for the two coupled branches of the distributed circuit:

$$\begin{aligned} \partial e_1/\partial x_1 &= -\,i_1(r + pl) - i_2pm \\ \partial i_1/\partial x_1 &= -\,e_1(g + pc) \end{aligned} \qquad (21)$$

and a similar pair with subscripts interchanged.

The solution of these two pairs of simultaneous equations is straightforward and quite similar to that carried out by Ollendorff[14] for two coupled circuits with lumped constants. With the assumption that squared terms in α are negligible, the wavelength characteristic

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ferred to in footnote 1. A detailed discussion of apparatus and experimental technique are there given.

[14] Ollendorff, "Grundlagen der Hochfrequenztechnik", p. 391.

and damping curves of Fig. 10 were computed. Excitation curves are in this case horizontal and straight. An examination of this new wavelength characteristic shows that the branches in the range of the first jump are essentially like those previously derived. (The new coefficient $k=m/l$ in this case is assigned a value which will give approximately the same variation in wavelength as the previous choice of $k=M/l=0.1$. This is to facilitate comparison, no attempt being made to estimate relative values in the two widely different cases.) The second jump, on the other hand, is somewhat different. This is best seen in the curves of Fig. 5. Here the solid curves are those from the previous analysis, the dashed ones those estimated using distributed coupling; (the discontinuity between the two parts of branch II of the wavelength charac-

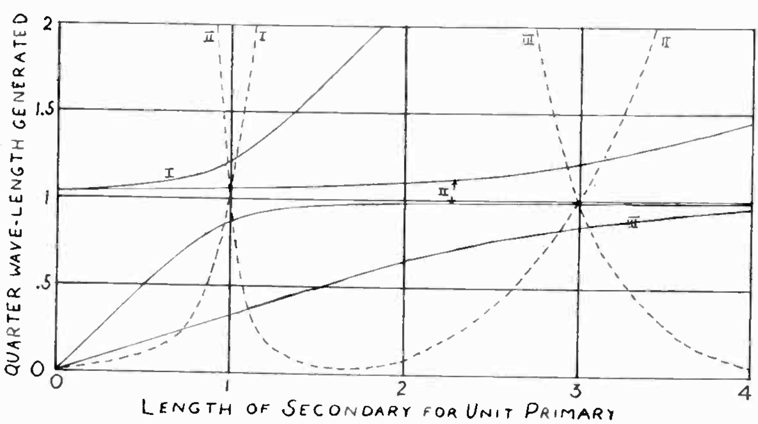


Fig. 10—Theoretical characteristics like those of Fig. 4 but derived assuming mutual distributed coupling instead of coupling through the bridge BB'.

teristic has been smoothed over; it is due to the overidealized hypothesis of exclusive radiation coupling). According to the dashed curves the second transition range involves a wavelength jump corresponding in magnitude to the first one. Without going further into a comparison of the two groups of characteristics, it is clear that with both types of coupling present, the first wavelength jump should be relatively little different from that obtained for pure bridge coupling, whereas the second jump should be smaller than the first one, but not to the extent required by pure bridge coupling.

The experimentally determined wavelength characteristics are shown in Fig. 11; 11b is for an open end secondary like Fig. 1a, while 11a is the corresponding characteristic for a bridged-end secondary. Little need be said about the first wavelength jump since it evidently agrees well with the theory. This is best seen in Fig. 6 where it is clear that a suitable determination of the coefficient k is all that is needed to admit the dotted experimental curve into the theoretical family. A close scrutiny of Fig. 11b and a comparison with Fig. 5b shows that the second wavelength jump does actually fall somewhere between the magnitude suggested by the dotted curve and that indicated by the solid one. The experimental curve is, therefore, in agreement with the theory as modified to include distributed coupling. The fact that the theoretical curve of Fig. 5b shows a loop at the second jump, whereas the experimental curve of Fig. 11b does not, is not necessarily contrary to theory. As seen in Fig. 6 the experimental value of the coefficient k for the first jump is in the neighborhood of 0.01, or one-tenth the value used in the theoretical curve. With such a small value of k a possible loop would be very narrow, so that for a near critical negative resistance a narrow loop at the second jump, or an even narrower break at

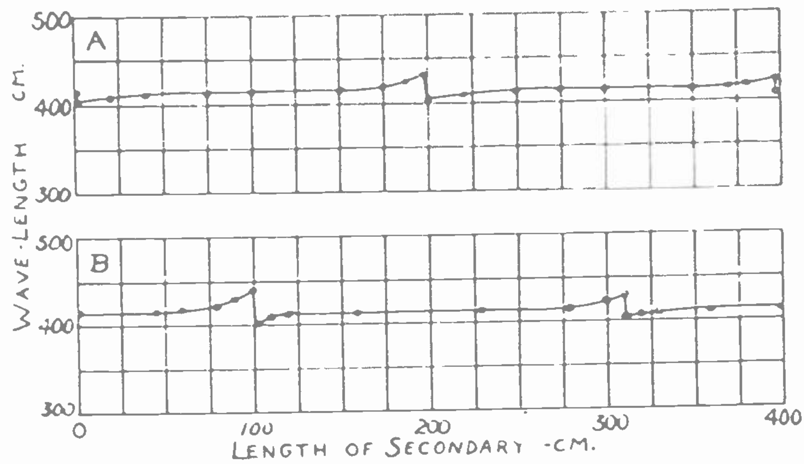


Fig. 11—Experimental wavelength characteristics with the secondary bridged at the end EE' in curve A, open at this end in curve B.

the first jump would not be experimentally observable. This is especially true since the actual transition point of the wavelength, being a point of instability, is difficult to determine with precision. In any case the question of loops is best left to the more complete and more precise data obtained especially for this purpose. This is discussed below.

2. The Bridged-End Case.

The apparatus used to obtain the new set of data now to be described is essentially the same as that described in reference 1 and in a further paper.[15] Nevertheless, both oscillator and resonance indicator differed considerably in design from those used in the previous work. A detailed description of these new pieces of apparatus, which were constructed especially for an investigation of radiation patterns using

[15] R. King, "A screen-grid voltmeter," PROC. I.R.E., vol. 18, p. 1388, (1930).

models, is reserved for a later paper. The general arrangement of the apparatus is shown in Fig. 12. It will be noted that the measuring system and the oscillator are now uniformly coupled regardless of the extension of the oscillator secondary. The parallel conductors which formed the oscillator secondary consisted of flat copper braid of extremely low resistance. Their end was made variable by arranging a brass cylinder, an inch and a half in diameter, in such a way that it could be rolled along the taught parallel system winding up the copper conductor and unwinding a pair of supporting cords. No attempt was made, however, to vary continuously the length of the oscillator secondary in this way. In reference 1 it was demonstrated that a conductor bridge may be moved along a pair of parallel conductors and have the same effect as varying the total length of these provided this total length is properly chosen. Hence, the total length BE (Fig. 4) was adjusted once and for all to be completely out of resonance with the primary, and the secondary BD was varied by moving the bridge DD'.

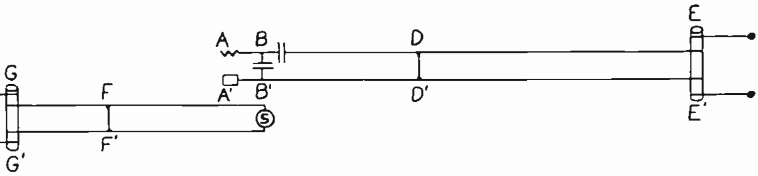


Fig. 12—Schematic diagram of the experimental arrangement for obtaining the wavelength characteristic of the coupled circuit with a bridged-end secondary. AB = primary; BD = secondary; GS = measuring system; S = screen-grid resonance indicator. BB' is the fixed bridge common to primary and secondary; DD' is the movable bridge for varying the secondary length; FF' is the movable bridge of the measuring system. EE' and GG' are the brass cylinders for adjusting the lengths of the copper conductors.

In carrying out the experiment this bridge DD' was moved step by step from an initial position near BB' toward EE'. For each setting, i.e., for each length BD, the wavelength generated by the coupled circuits was measured with great precision on the measuring system loosely coupled to the primary. (This measuring system was adjusted according to the maximum deflection method described in reference 1.) In taking the data for the positions of the resonance peaks on the measuring system, deflections of the indicator were recorded for a range of several centimeters over the peaks. In this way the actual position of the maximum point could be estimated to within a half centimeter or less. The wavelength was thus determined to within a half of one per cent. Care was taken to read the indicator by means of a telescope so placed that body capacity did not affect the readings.

Fig. 13 shows the wavelength curve obtained from a large number of carefully distributed wavelength measurements. The overlapping portions or loops (Ziehschleifen) were obtained by moving the bridge DD' first in one direction until the jump occurred and then in the other until the reverse jump took place. The fact that such pronounced loops were obtained in this instance where the previous data showed none, is due to the circumstance that in this case a 345 power tube with larger negative resistance was used in the oscillator instead of the 201-A previously employed.

It is a simple matter to fit the beautiful experimental curve of Fig 13 into the theory developed. The first essential is to determine the value of the ideal or theoretical quarter wavelength of the oscillator used. As has already been pointed out this is not the quarter wavelength generated by the primary alone, but somewhat larger than this. Referring to the first branch of the wavelength characteristic of Fig. 9,

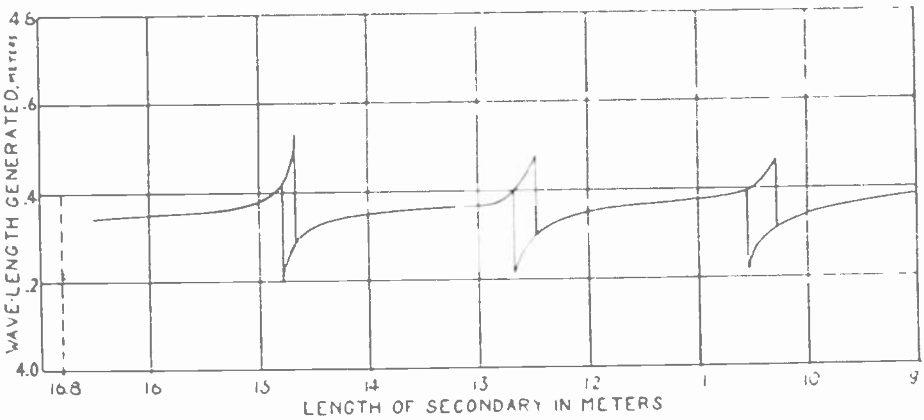


Fig. 13—Experimental curve of the wavelength generated by the bridged-end secondary two-circuit oscillator for different lengths of the secondary.

it is at once clear that for an excitation greater than $X=2r$ as required by the width of the loops, the maximum possible value of the quarter wavelength generated can differ very little from that corresponding to the abscissa $s_2=2$ on account of the steepness of the damping curve. From the ratios of the widths of the three successive loops in the experimental curve, a rough comparison with the widths required by the theoretical damping curves leads to the conclusion that the excitation cannot be very much in excess of that corresponding to $X=2r$. With upper and lower limits thus approximately suggested, and the fact that the ordinate $s_2=2$ corresponds to the ideal half wavelength, it is reasonable to suppose that the maximum wavelength measured at the first jump is slightly larger than the ideal wavelength. In this case the maximum wavelength was 4.53 meters. A choice of 4.50 meters as the ideal wavelength therefore seems reasonable. Selecting this value and reducing the experimental curve to the same relative coördinates used in the theoretical characteristic of Fig. 9, Fig. 14 shows the experimental and theoretical curves plotted together to the same scale. It is clear that the general agreement is excellent except that the magnitude of the jumps is different. Since this depends on the coupling constant k, it is only a question of assigning a suitable value to this constant to fit the experimental curve into the theoretical family. Since the entire experimental curve lies much nearer the ideal quarter wavelength of unity in Fig. 14 than does the theoretical curve for which $k=0.1$, it is clear that k for the experimental case must be smaller than 0.1. Its exact value can be readily computed as follows. By a simple trigonometric transformation (22) can be written in a more convenient form, viz.:

$$k = \frac{\sin \beta s_2 \cos \beta s_1}{\beta \cos \beta(s_1 + s_2)}.$$

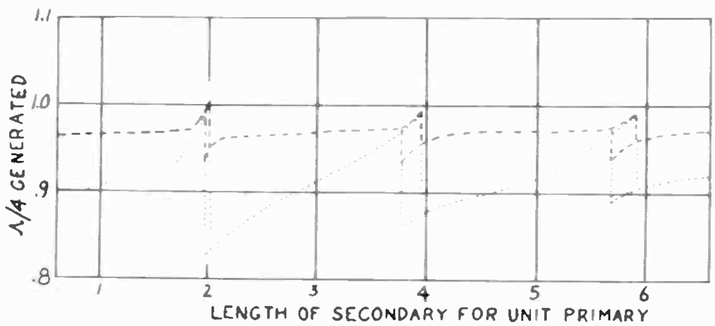


Fig. 14—Combined theoretical and experimental curves.

Substituting the values $s_1 = 1$, $s_2 = 2$, $\beta = 1.68$ as obtained from Fig. 9, k is found to be $k = 0.044$. This is a very reasonable value and a general comparison or actual calculation shows that with this value of k the theory yields curves precisely like the experimental ones.

With the secondary completely removed, the oscillator generated a wavelength measured to be 4.34 meters. This is precisely the value obtained over the long, nearly horizontal parts of the wavelength characteristic between the successive jumps. It is clear that the effect of the secondary is to make the wavelength of the two coupled circuits fluctuate above and below this value near the jumps, but to a lesser degree the longer the secondary becomes. For an infinitely long secondary the wavelength generated would be constant for variations in the secondary and would have the value generated by the primary alone. As the coupling constant k is diminished this value approaches the ideal wavelength represented by the quarter wavelength of unity in the theory.

In concluding the comparison of experimental and theoretical results, it may safely be said that the general agreement obtained with the open-end data, complicated as it was by radiation coupling, is made definitely precise and convincing by the more comprehensive check obtained with the new data using the bridged-end secondary. Certainly in view of the approximations made in the theoretical analysis, the results are most gratifying.

3. Wavelength Measurement.

One further experimental point remains to be discussed. In reference 1 two methods of wavelength measurement are described, a maximum deflection method and a zero or minimum deflection method. The minimum deflection method consists essentially in dividing the measuring system into a primary and secondary by means of a fixed bridge at the first resonance peak from the indicator. The oscillator which is to be standardized is, of course, coupled to the indicator end or primary. With the total length correctly adjusted, a second bridge is now moved beyond the fixed one. At definite intervals the indicator records sharp minima. The distance between these minima is clearly nothing other than the distance between the points at which the wavelength of the coupled circuits jumps. The fact that these minima occur at the same bridge positions regardless of the direction of approach suggests that the excitation must be near the critical value for a continuous, single-valued wavelength characteristic. The question to be answered is this: Is the distance between successive minima a true measure of a half wavelength as stated in reference 1 on purely experimental grounds. Reference to the theoretical curves of Fig. 9 shows that the critical excitation curve intersects the damping curves at

TABLE I

Maximum deflection method		Minimum deflection method	
Position of Peak	Half Wavelength	Position of Minimum	Half Wavelength
13.915 meters		11.810	
	2.170		2.170
11.745		9.640	
	2.175		2.170
9.570		7.470	
	2.170		2.170
7.400		5.300	
	2.170		2.170
5.230		3.130	
	2.165		
3.060			

(The third decimal place is to the nearest 5.)

points which are separated considerably less than the ideal half wavelength, but only very slightly less than the actual half wavelength generated by the primary alone. Since even for $k=0.1$ the error involved in calling this separation a half wavelength is not large, it should for most purposes be entirely negligible for the very small values of l encountered in practice.

In order to demonstrate this experimentally, the wavelength of the oscillator alone without secondary was measured on the long parallel conductors previously used as secondary. A succession of half wavelengths over the full sixteen-meter extension was measured by the same precise method already described, both for maximum and minimum deflection methods. The half wavelengths were determined to the nearest half centimeter. They are recorded in Table I. These results indicate that either method gives the same results.

Part III. Comparison of Three Types of Oscillatory Circuits

In order to do full justice to the interesting and striking analogy existing in the broader lines of comparison, and at the same time stress those differences which characterize the individuality of the three types of oscillatory circuits here in question, many pages would be required. Since a very considerable familiarity with the subject can hardly be presupposed, in view of the fact that the major part of the literature has appeared in foreign periodicals, a compromise will be attempted.

1. Lumped Constants

The circuit with purely lumped constants having a frequency or wavelength characteristic analogous to that derived above for the distributed circuit is well known. It consists of two branch circuits each including resistance, inductance, and capacitance reacting upon each other either through mutual-inductance or by having a part of the self-inductance in common. This circuit is completely solved in Ollendorff.[19] Since the solution is straightforward and probably familiar, it will not be reproduced here. Suffice it to say that the wavelength characteristics and damping curves are essentially the same as those shown in Fig. 11 for the range of the first jump; i.e. up to abscissa 2. In the case of lumped constants there are, of course, no further jumps since there are only two possible coupling frequencies and hence only two damping curves. A set of curves analogous to those of Figs. 4 and 11 are plotted on page 395 of Ollendorff. Damping curves and excitation curves are shown on page 406, and a set of characteristics like those of Fig. 5 is shown on page 407. The essential difference throughout is the presence of only one transition discontinuity, since the repeated resonance points of the distributed circuit are, of course, absent. It may, therefore, be stated that differences in the nature of the constants and of the coupling coefficients lead to corresponding differences in the detail of

[19] *Loc. cit.*, p. 392.

the circuit characteristics in the two cases. But in the broader outline and significant structure the circuit types are fundamentally similar.

2. The Electron Oscillator

Before entering into a discussion of the wavelength characteristics of the electron or space-charge oscillator, it seems desirable to outline briefly the underlying principle of this type of triode high-frequency generator. The circuit used may be identically that shown in Fig. 1a except that the potentials on the electrodes of the triode are reversed. The grid is now raised to a high positive potential instead of the plate, while this latter may be at the filament potential or even negative. Such an arrangement is found to oscillate at a very high frequency as detected by the resonance points observed by moving a bridge containing a current indicator along the parallel conductors, or by the standing waves set up on a suitably coupled Lecher wire system. The first procedure is that followed by Barkhausen and Kurz when they discovered the oscillations; the second was used by Gill and Morrell somewhat later when they were led to contradict the Barkhausen-Kurz assertion that the frequency generated was entirely independent of the circuit external to the triode itself. Theories regarding the finer mechanism of electron oscillations are numerous; but there seems to be complete agreement on the general picture presented in the original paper. This describes the electrons emitted by the filament as accelerated by the highly positive grid to sufficient velocities that the greater number of them fly through the meshes of the grid in a dense cloud. After they have passed through the grid they are retarded by the reversed field on the other side and attracted back again through the grid to the space-charge region of the filament, thus completing a cycle. Disagreement and uncertainty still centers primarily about the mechanism or agency which starts and regulates the motion of such clouds of oscillating electrons.

In this paper, where it is solely a question of wavelength characteristics as determined by circuit constants, the oscillations will simply be taken for granted without further explanation. It is here a question of determining, as in the case of the regenerative oscillator, the effect of a variable secondary upon the frequency generated. Of course, other factors than the length of a secondary are primarily important in determining the frequency. Grid and plate voltages, electrode dimensions and shapes, as well as the degree of evacuation all play a greater or lesser role. The simplest formula for the most idealized case is given as $\lambda^2 E_g = \text{const}$. In the present discussion, however, all factors except the parallel conductors attached to plate and grid will be assumed constant.

It was stated a few lines above that the effect of the secondary upon the frequency generated is to be determined. Clearly the first important question is to decide precisely what constitutes the secondary and what the primary. There appears to be considerable disagreement on this point. Hollmann,[4] who has done pioneer work in investigating the electron oscillator, has followed the seemingly obvious course of using exactly the same division into primary and secondary as shown in Fig. 1a for the regenerative oscillator. In other words, he includes the parallel conductors up to the bridge BB' and the interelectrode capacitances of the triode in the primary. Using this nomenclature he has developed a rather complete theory of distinct frequency ranges in wavelength characteristics obtained by moving the bridge BB' in steps

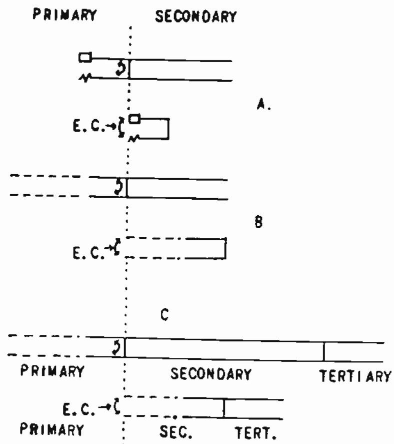


Fig. 15—Electron and regenerative circuits compared. The upper diagram in each pair is for the regenerative, the lower for the electron oscillator. $E.C.$ = electron cloud.

and measuring the wavelength generated on an accessory Lecher wire system. His explanations, complete in themselves, are not well adapted to bring out an analogy between the electron oscillator and the two types already described. For that reason no further reference will be made to them. Instead, a theory paralleling quite completely that given above for the regenerative oscillator will be sketched and the analogy between the two pointed out.

According to the accepted theory for the electron oscillations, the oscillatory circuit consists essentially of a moving cloud of electrons flying back and forth through the grid of the tube thereby causing its potential to fluctuate periodically. Ideally, therefore, it is this cloud of electrons, independent of all electrodes and conductors, which constitutes the primary circuit. The secondary then includes the electrodes as well as all attached conductors. If these are a pair of parallel wires, and if, merely for convenience in making the picture clear, an equivalent length is again introduced to represent the interelectrode capacitances, the circuit may be diagrammatically pictured as in Fig. 15a and b. The regenerative circuit of Fig. 1 is also reproduced for ready comparison. Thus a of Fig. 15 shows the regenerative circuit already discussed, and directly below it the corresponding electron oscillatory circuit. The primary of this latter is the vibrating electron cloud (*E.C.*); the secondary includes interelectrode capacitances and parallel conductors. b of the same figure again shows the two circuits but with the equivalent lengths of the interelectrode capacitances dotted in at the proper points. It will be observed that in each case the secondary has one open and one bridged end.

A very complete theoretical investigation by Wundt,[3] following the general scheme just outlined, leads to results remarkably similar to those obtained for the regenerative case and agreeing well with experimental data. Wundt's analysis is more involved than that carried out for the regenerative circuit since the equations of primary and secondary are no longer the same. In his case the equation for the primary had to be derived from the motion of an electron moving between charged electrodes. Another difference, and one of prime importance in comparing the two types of circuits, is the coupling between primary and secondary. In the preceding analysis it was the current in the bridge BB' which was common to both branches; in the present case it is the potential difference between plate and grid which affects both primary and secondary. Fig. 14 of Wundt's paper shows the theoretically derived wavelength characteristics and damping curves of the electron oscillator. These very much resemble the curves of Fig. 4. A detailed comparison and more precise analysis of differences is here beside the point. It is desired to bring out an analogy, not an identity. Just as was stated in connection with the lumped constant circuit, differences in the circuits, in the coupling involved, must lead to differences in the results; notably, in this case, a much greater variation in wavelength and correspondingly larger jumps. The significant point is that the wavelength characteristics of the electron oscillator with a pair of parallel wires attached to grid and plate, may be explained on the same fundamental basis of coupling frequencies and wavelength jumps as were those of the regenerative oscillator with distributed line constants. And it is consistent with the whole development of physical science to select explanations which unify and simplify interpretation. An outline of the analogy between the electron oscillator and the regenerative oscillator is given in Table II.

One point still may be obscure. If the electron cloud constitutes the primary, and if a pair of parallel wires is the secondary, what interpretation can be given to an electron oscillatory circuit which includes an additional circuit beyond the bridge BB' as pictured in Fig. 1a? The answer is that a third circuit or any number of circuits may be doubled

TABLE II

Comparison of Regenerative and Electron Oscillators Each With Secondary Having Distributed Constants

	Regenerative Oscillator	Electron Oscillator
The primary	Consists of two parallel conductors, the interelectrode capacitances, and the negative resistance of the triode.	Consists of a cloud of electrons moving periodically through the meshes of the highly positive grid of the triode.
The natural frequency of the primary alone.	Is determined by the circuit constants. (*Cf.* the calibration curve of Fig. 3.)	Is purely hypothetical physically, but may be used conveniently to bring out the analogy. The formula $\lambda^2 E_g = \text{const.}$, to which no experimentally observed data exactly correspond, is presumably the mathematical equivalent.
The secondary.	Consists of two parallel conductors which are a continuation of the primary beyond the bridge BB'.	Consists of two parallel conductors attached to grid and plate of the triode and the interelectrode capacitances.
The coupling or reaction between primary and secondary.	Principally the current through the inductance of the bridge BB' which is common to both branches. An interchange of radiated energy may also play a part.	Principally the potential difference between plate and grid of the tube which is common to both primary and secondary.
The natural frequency or vibration modulus of the complete circuit.	Is determined by the circuit constants of primary and secondary.	Is determined by several factors influencing the velocity and distance of motion of the electrons of the primary and by the circuit constants of the secondary.
Effect of the coupling reaction.	Depending upon the relative magnitudes of the negative resistance and the actual resistance in the circuit, the mutual reaction produces discontinuous, continuous, or overlapping changes in frequency as the length of the secondary is varied. (*Cf.* Fig. 5.) Abrupt frequency changes take place in the neighborhood of such secondary lengths as will bring this into resonance with the primary. The process may be described as a mutual building up of amplitude near such resonance points with the frequency changing in such a way as to satisfy energy relations prescribed by resistance and negative resistance.	Abrupt frequency changes take place in the neighborhood of such lengths of the secondary as will bring this into resonance with the primary. Following Hollmann's picture a mutual building up of amplitude takes place as the alternating potentials superposed on the electrodes by the oscillating electron cloud, and the resonating secondary approach each other in frequency. Oscillations generally break off giving rise to discontinuities. The wavelength characteristic resembles the corresponding one for the regenerative case. (*Cf.* Fig. 14, of Wundt's article.)
Repeated wavelength jumps.	Occur, respectively, for secondary open or bridged at the end EE' related to quasi asymptotes which have slopes in the ratio 1:1/3:1/5:, and 1/2:1/4:1/6:. For a quarter wavelength secondary the odd ratios are true, for a half wavelength secondary the even ratios.	As for the regenerative case the odd ratios for a quarter wavelength secondary, the even ratios for a half wavelength secondary. Since the secondary is open at the end coupled to the primary instead of bridged, the end conditions at EE' must be the reverse from those true for the regenerative case.

to the secondary of an electron oscillator, and that the same may be done in the case of a regenerative oscillator. This is illustrated in Fig. 15c. The characteristics of such a triple circuit must naturally be more complicated and should involve additional, but smaller, wavelength variations due to changing the length of the less closely coupled tertiary branch. Such variations were detected by Hollmann and submitted privately to the writer in order to demonstrate that the more intense wavelength jumps obtained by varying the secondary of the electron oscillator are essentially different from the less intense ones observed in the regenerative circuit. Far from disproving the theory outlined above, the existence of such weaker, tertiary jumps fits into it perfectly. But it must be emphasized again that no pretense is made to claim a closer relationship between the two circuits and their characteristics than that of an analogy.

A few words may now be appended with reference to the contradictory opinions voiced by experimenters with regard to the independence of frequency of the electron oscillator to variations in an externally attached system. In the first place, it is to be noted that it is not possible to measure the frequency of the primary, or cloud of electrons, independent of a secondary since this includes the tube electrodes as well as any attached wires. It is therefore futile, from the experimental point of view, to talk about such a hypothetical frequency. Indeed, the only frequency which has any physical significance in the case of the electron oscillator is the frequency generated when a secondary is attached. The situation would be the same for the regenerative oscillator if the entire primary were hermetically sealed in a completely shielded container with only the terminals BB' available. There would, then, be no point, other than that of convenience in interpretation, in talking about the frequency generated in the box since there is no way of measuring it. The instant a pair of wires is attached to the available terminals, the frequency of the complete circuit including the wires is measureable, but this differs from that originally generated by the undisturbed primary. Thus the procedure followed by Barkhausen and Kurz in moving a current indicating bridge along such a pair of wires attached to plate and grid of the tube (which are equivalent to the terminals BB' mentioned above) determined the distance apart of wavelength jumps. This is not necessarily a half wavelength. On the other hand the method of Gill and Morrell and Hollmann of varying the length of the attached wires by moving a bridge[16] and then measuring the wavelength generated on an accessory Lecher system, measures the wavelength of the complete circuit for the particular setting of the bridge. This wavelength, of course, changes if the circuit is varied, since it is determined by the constants of the circuit as a whole.

Objections may be raised by some to calling a purely hypothetical cloud of electrons a primary circuit which admittedly cannot exist independent of a secondary of some sort. This is done for the sole purpose

[16] With the precautions taken by Hollmann with regard to the length of the tertiary circuit beyond the movable bridge, moving a bridge is practically equivalent to varying the length of the wires with the bridge fixed at their extremities. This is discussed in greater detail in reference 1.

of presenting an easily visualized analogy consistent with the basic mathematical analysis. Such pictures are for convenience in understanding, and if they are not convenient they may best be discarded. It is interesting to note in this connection, however, that in the case of the electron oscillator the ideal quarter wavelength is ideal both theoretically and experimentally. With the regenerative oscillator the ideal quarter wavelength is the limiting value of the wavelength actually generated by the primary alone as the coefficient k becomes vanishingly small. In both cases, however, it is convenient to introduce an ideal quarter wavelength for the primary alone, not for physical but for mathematical purposes.

It is hoped that in analyzing and discussing triode oscillatory circuits in an endeavor to develop and correlate their fundamental theory all success has not been wanting. Encouraging is a private communication from Hollmann in which he agrees that in its general principles the alternative theory of the electron oscillator here presented is not inconsistent with his own extensive theoretical and experimental studies.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

A SIMPLIFIED GENERAL METHOD FOR RESISTANCE-CAPACITY COUPLED AMPLIFIER DESIGN*

By

David G. C. Luck

(Department of Physics, Massachusetts Institute of Technology, Cambridge, Massachusetts)

Summary—*The steady state analysis of the general resistance-capacity coupled amplifier stage is thrown into such a form that any amplifier stage is characterized by three easily computed constants which make it possible to read off its complete steady state performance immediately from three perfectly general analytical curves. These curves are shown, as are some applications of the method.*

THE resistance-capacity coupled type of vacuum tube amplifier is very widely used today, and for a great diversity of purposes. It is felt that a wider realization of the possibilities presented by adjustment of shunt capacities would result in an even greater breadth of usefulness. One of the most valuable features of this instrument is its essential simplicity of operation, causing it to lend itself easily to accurate design. However, the designing process often becomes quite tedious, and it is the object of this paper to present what appears to be a great simplification of the work involved.

This method yields only the steady state characteristics of such amplifiers; this is of course not a complete solution of the problem but is of considerable value, not only directly but for the light it throws on the amplifier performance under transients. The only assumptions necessary regarding the amplifier are that it shall be nonregenerative, shall have no appreciable inductive reactances, and shall have tube characteristics linear over the operating range. These assumptions are all normally fulfilled, at least to a sufficient degree, in practice.

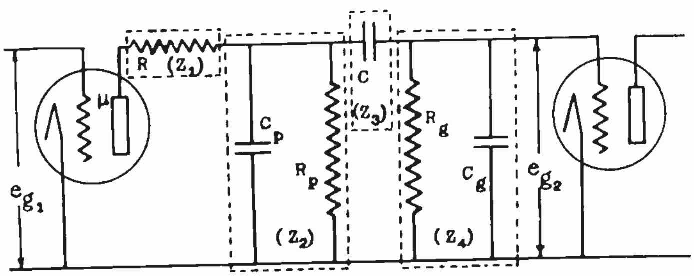


Fig. 1

Fig. 1 shows the most general resistance-capacity coupled amplifier network for one stage; the designations used are standard ones. The ordinary steady state solution of this network yields

* Decimal classification: R363. Original manuscript received by the Institute, April 5, 1932.

$$A = e_{g_2}/e_{g_1} = \mu \cdot \frac{Z_2 Z_4}{Z_1 Z_2 + (Z_1 + Z_2)(Z_3 + Z_4)} \tag{1}$$

where A is the vector amplification per stage.

Writing explicitly the values of the various impedances yields

$$\begin{aligned}\mu/A &= 1 + R/R_g + R/R_p + C_g/C + RC_p/R_gC + RC_g/R_pC \\ &\quad + j(\omega RC_g + \omega RC_p + \omega RC_pC_g/C - 1/\omega R_gC - R/\omega R_gCR_p).\end{aligned} \tag{2}$$

This may be rewritten in the form

$$A_0/A = 1 + j(f/f_0 - f_0/f)/w = 1 + jy/w \tag{3}$$

in which,

$$\left.\begin{aligned} A_0 &= g_m R_s/(1 + C_g/C) \\ w &= (1 + C_g/C)\sqrt{\frac{R_gC}{R_s(C_p + C_g + C_pC_g/C)} \cdot \frac{R_l}{R_s}} \\ \frac{1}{f_0} &= 2\pi\sqrt{R_pR_g(C_pC + C_gC + C_pC_g)}, \end{aligned}\right\} \tag{3a}$$

where

$$\begin{aligned} 1/R_s &= 1/R + 1/R_l + 1/R_g \\ 1/R_l &= 1/R + 1/R_l \\ \mu &= g_m R. \end{aligned}$$

Here $(C_p - C_g)/C$ has been neglected in comparison with unity; this may in all cases be made allowable by proper choice of constants, even when C_p and C_g are separately of the same order of magnitude as C, and C_g/c is therefore not negligible compared to unity. It is apparent that (3) is equivalent to

$$A = |A|\, \epsilon^{-j\phi} = A_0 \cos\phi \cdot \epsilon^{-j\phi}, \tag{4}$$

where ϕ is the phase displacement per stage. From this it may be seen that

$$y = f/f_0 - f_0/f \tag{5}$$

$$\phi = \tan^{-1} y/w \tag{6}$$

$$|A|/A_0 = \cos(\tan^{-1} y/w) \tag{7}$$

represent three completely general analytical curves which completely express the gain-frequency characteristics of any amplifier of this type, both in magnitude and in phase, when the constants given by (3a) are known. The logarithmic performance of the amplifier stage may be represented by

$$D = 20 \log [\cos(\tan^{-1} y/w)] \tag{8}$$

where D is the drop in amplification in decibels per stage from the maximum value. The graph of (5) is shown as Fig. 2, the abscissas being f_0/f when $f<f_0$ and f/f_0 when $f>f_0$. The graphs of (6), (7), and (8) are shown as Fig. 3. It will be noted at once that all points for the graphs of Fig. 3 are obtainable from trigonometric tables, no computation being required.

It is immediately apparent that, with the curves above obtained, the steady state performance of any specific amplifier is completely

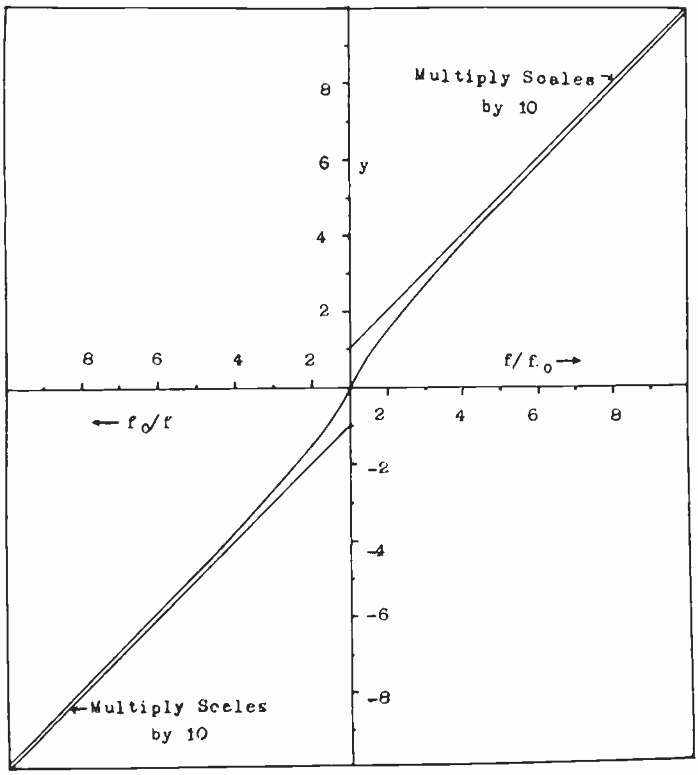


Fig. 2

determined by the three constants given by (3a) which simply fix the scales of the general curves. These constants are: a maximum amplification A_0, which occurs without phase shift at a certain "mean frequency;" f_0, which divides the amplifier gain-frequency characteristic into symmetrical halves; and a "pass band width" w, which multiplies or divides the mean frequency to give the two frequencies at which the gain per stage falls to $A_0/\sqrt{2}$. It is also apparent that there is a simple fixed relation between gain magnitude per stage and phase displacement per stage, the former being proportional to the cosine of the latter. The generality of the curves shows that all resistance-capacity coupled amplifiers must have just the same symmetrical frequency characteristic; the symmetry axis of this characteristic and its frequency scale may however be adjusted within quite wide limits by means of the disposable circuit parameters. Knowing the circuit parameters of any amplifier, its scale constants may be computed from (3a) and its performance read off from Figs. 2 and 3. Conversely, knowing

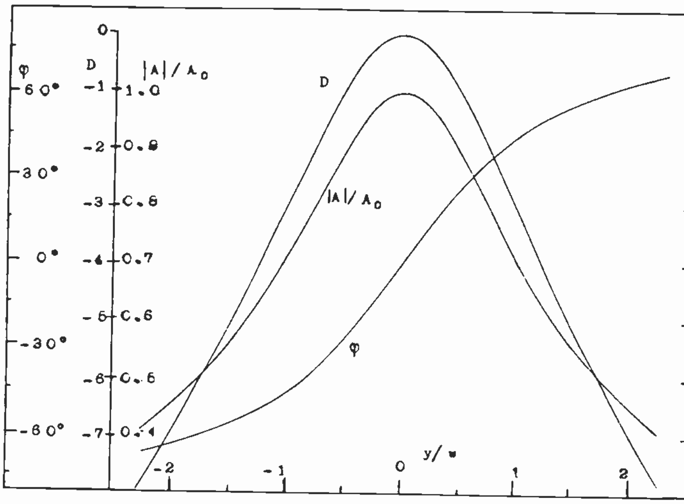


Fig. 3

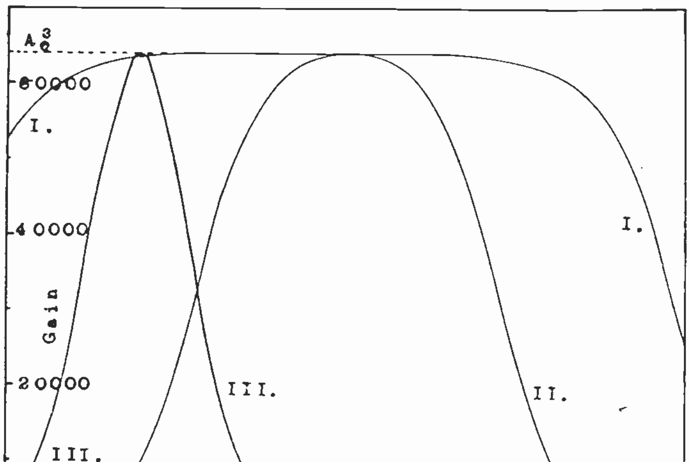


Fig. 4

the performance desired from an amplifier, the necessary values of the scale constants may be determined from the figures, and the disposable circuit parameters adjusted to give these values. Either process may be carried out quite rapidly with a little practice.

As examples of the use of this method, the design of three widely different amplifiers has been performed. Their circuit constants are shown in Table I, and their gain-frequency characteristics in Fig. 4,

TABLE I

Amplifier Type	A_0	w	f_0 c.p.s.	R_g megohms	R_p megohms	C μf	C_p μf	C_g μf	R megohms	g_m ma/v	I_p ma	E_b volts	E_{c_2} volts	E_{c_1} volts
I	40	170	670	2.0	0.05	0.02	0.00002	0.00001	0.75	0.87	1.0	180	45	−1.5
II	40	11.5	1000	2.0	0.10	0.0008	0.00008	0.00008	2.0	0.30	0.4	180	30	−1.5
III	40	2.31	60	0.18	0.24	0.0085	0.0085	0.0085	0.75	0.87	1.0	90	45	−1.5

for three stages of each type. Numerical gain has been plotted, as this is more useful than the usual logarithmic (decibel) plot for almost all nonacoustic uses. The amplifier designated as Type I is a general purpose high gain, high quality amplifier, suitable for example as a photocell amplifier for television and adequate in frequency range for 20 pictures, of 60×72 lines, per second. Type II is intended to amplify harmonics of 60 cycles and suppress the fundamental as much as possible; Fig. 4 shows that the third harmonic is amplified five times as much as the fundamental. The amplifier of Type III is a "tuned" instrument, for suppression of noise in amplification of a 60-cycle signal, but has no inductances to make shielding from stray magnetic disturbances difficult.

In each case the design process was the second one suggested above. First, the necessary scale constants were determined by consideration of the desired amplifier performance. Thus, for Type I, f_0 was the geometric mean of the lowest and highest frequencies it was desired to amplify, and w had to be made as large as possible consistent with sufficient amplification, the limitation being the unavoidable shunt capacity of the tubes; for Type III, f_0 was the frequency which it was desired to amplify and w had to be made as small as possible. Then the desired scale constants were approached as nearly as tube limitations admitted by adjustment of the circuit constants in (3a), giving the final values in Table I. At this stage of the design, the relation

$$A_0 f_0 \omega = \frac{g_m}{2\pi(C_p + C_g + C_p C_g / C)}$$

is very convenient.

These final values of f_0 were then used to determine y for a set of frequencies, from the curves of Fig. 2; these values of y, with the final values of w, sufficed to determine $|A|/A_0$ from the proper curve in Fig. 3, which, when multiplied with A_0 and cubed (for three stages), gave the necessary ordinates for the curves of Fig. 4. Tubes of the '36 type were used in each design, as they have proved exceptionally satisfactory as laboratory voltage amplifiers. The values of g_m and R shown in the table were directly measured; they may be controlled to some extent by variation of screen voltage.

Regeneration might be used to modify the shape of any characteristic, but would seriously complicate the design problem. It may be avoided entirely by adequate shielding and the use of separate batteries for each stage; the low currents and moderate voltages required make the latter procedure economically feasible. However, two stages of less than one hundred gain per stage may usually be safely operated from common batteries.

BOOK REVIEW

The National Physical Laboratory Report for the Year 1931. Published by His Majesty's Stationery Office, London, for the Department of Scientific and Industrial Research. 313 pages, paper cover. Price, 15 s. 0d. Net.

The first thirty-odd pages contain the report of the Executive Committee' being a brief outline of the work carried on by the various departments of the National Physical Laboratory. There are fourteen pages listing the papers published by the Laboratory or members of its staff for the years 1930 and 1931.

Reports on the several departments of the Laboratory follow. In these reports sections are found dealing with the following items: heat, general physics, radiology, sound, optics, electrical standards and measurements, electrotechnics, wireless, photometry, all kinds of measurements and measurement equipment, engineering research, aerodynamics, metallurgy, and various studies in the design of ships.

The radio engineer will be most interested in the report of the Electricity Department, forty-three pages of which are devoted to electrical standards and measurements, and radio. A brief description and some results obtained with a tuning fork standard are given. Another frequency standard briefly described is a quartz ring, with six pairs of electrodes alternately placed near the inner and outer surfaces of the ring.

Other radio subjects treated include international frequency comparisons, standard frequency transmissions, radio-frequency bridge, radio transmission researches, antennas for transmission and reception, propagation of electric waves along earth's surface for frequencies above 30,000 kc, electrical properties of earth's surface at radio frequencies, measurements of plate-grid capacity of screen-grid tubes, measurement of current at radio frequencies, percentage modulation, total harmonic content in an alternating current, and development of a simple constant frequency oscillator.

*E. L. HALL

* 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 and catalogs describing the various products manufactured by the National Carbon Company, Inc., New York, N.Y., are available. Among these, perhaps the most interesting to radio men is the folder describing radio receiving tubes, A, B, and C batteries including the air cell battery, and the 12-page technical bulletin giving characteristics of and suggestions for the operation of Eveready-Raytheon phototubes and glow discharge lamps.

The Pacent Electric Company, 91 Seventh Ave., New York, N.Y., has available a number of leaflets describing their products. Form No. 22 is a catalog of their duo lateral coils and contains an abac showing the relationship between frequency, inductance, and capacitance. Forms 48, 50, 51A, and 56 describe various electric phonograph devices, particularly tone arms. Form No. 55 and a leaflet bearing no form designation describe a series of portable sound motion picture projectors for home entertainment.

The numerous sizes and shapes in which shakeproof lock washers and terminals may be obtained are illustrated in a 20-page catalog issued by the Shakeproof Lock Washer Company, 2501 N. Keeler Ave., Chicago, Ill.

A loose-leaf binder containing a series of technical bulletins gives operating data on the entire line of vacuum tubes manufactured by the Hygrade Sylvania Corporation, Emporium, Pa.

The Yaxley Line of small parts so necessary to the manufacturer, serviceman, or experimenter is illustrated in a 12-page catalog entitled "Approved Radio Products." The Yaxley Manufacturing Company is a division of P. R. Mallory & Company, 3029 E. Washington St., Indianapolis, Ind.

A great number of bulletins illustrating the numerous products manufactured by the Ward Leonard Electric Company, Mount Vernon, N.Y., are available upon request. Although Ward Leonard manufactures resistance devices for a great number of nonradio uses, the following bulletins are of interest primarily to radio men: Circular 507, describing vitrohm resistors and rheostats for radio purposes, Bulletin 80,000 showing vitrohm plaque noninductive, non-capacitive resistors; Sheet No. 1006, showing vitrohm stamped steel rheostats; Data Sheet 81007 describing the type TD2 time delay relay for use with mercury vapor rectifier tubes; Data Sheet 81008 giving details of the type A midget magnetic relays for a-c or d-c operation; Bulletin 69 describing battery charging equipment; Bulletins 79 and 79000 describing faders and volume controls for broadcast stations and public address systems.

Proceedings of the Institute of Radio Engineers
Volume 20, Number 8 *August, 1932*

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.

R000. RADIO (GENERAL)

R030 ×534 Acoustic nomenclature and definitions. *Wireless Engineer & Experimental Wireless* (London), vol. 9, pp. 307–309; June, (1932).

An account of several German units is given.

R100. RADIO PRINCIPLES

R113 M. J. O. Strutt. Zusammenfassender Bericht: Der Einfluss der Erdbodeneigenshaften auf die Ausbreitung elektromagnetischer Wellen. (A comprehensive treatment: The influence of the properties of the earth's surface on the transmission of electromagnetic waves.) *Hochfrequenz. und Elektroakustik*, vol. 39, pp. 177–185; May, (1932).

The treatment is divided into four parts. The first part treats electromagnetic radiation without an earth. Part II is devoted to radiation diagrams under the influence of the earth. Part III treats the field strength on the earth's surface. Part IV takes up the experimental determination of the "Erdbodeneigenshaften."

R113.5 C. G. Abbott. Solar and radio periodicities. *Science*, vol. 75, p. 607; June 10, (1932).

A comparison of data given in "Periodicity in solar radiation" by C. G. Abbott and G. T. Bond (Smithsonian Misc. Coll., vol. 87, No. 9), and "Tables of North Atlantic radio transmission conditions for long-wave daylight signals for the years 1922–1930" (PROC. I.R.E., vol. 20, April, (1932)) show clearly a relation between solar radiation and radio transmission phenomena.

R113.6 F. Ollendorff. Die Adsorption kurzen Wellen in Gebäuden. (The absorption of short waves in buildings). *Elek. Nach. Tech.*, vol. 9, pp. 181–194; May, (1932).

A theoretical treatment is given which is based on the consideration of a simple model. The passage of electromagnetic current through a house wall, theory of series reflection, the transmission law, propagation on a periodically interrupted chain system are topics which receive theoretical consideration.

* This list compiled by Mr. A. H. Hodge and Miss E. M. Zandonini.

R116 P. S. Carter. Circuit relations in radiating systems and applications to antenna problems. Proc. I.R.E., vol. 20, pp. 1004-10041; June, (1932).

Expressions for the self and mutual impedances within a radiating system are developed by the use of the generalized reciprocity theorem. A method for the determination of the field intensities is outlined. Formulas for the self and mutual impedances in several types of directional antennas are given. Questions of practical interest in connection with arrays of half-wave dipoles, long parallel wires, and "V". type radiators are discussed. Different types of reflector systems are considered Curves of the more important relations are shown. The mathematical development is shown in an appendix.

R125 G. L. Davies and W. H. Orton. Graphical determination of polar patterns of directional antenna systems. Research Paper No. 435. *Bureau of Standards Journal of Research*, vol. 8, pp. 555–569; May; (1932).

This paper describes graphical methods for the determination of polar patterns of directional antenna systems.

R126×R326 ×R387.5 H. M. Towne. Lightning arrester grounds. *General Electric Reviews*, vol. 35, pp. 173–77, March; pp. 215–21, April; pp. 280-85, May, (1932).

The general characteristics of grounds, the factors affecting their resistance, the advantages of artificial treatment, maintenance and testing and general considerations with respect to various classes of electric circuits are discussed. Driving ground electrodes and making connections thereto, artificial treatment of soil, energy dissipation in soil, impulse characteristics, and resistance values are treated. Pole-type and station-type arrester grounds, common neutral ground wire, overcoming adverse conditions, railway and signal arrester grounds, inspection and measurement are discussed.

R133 ×R339 W. Dehlinger. On the ultra high-frequency oscillation of the magnetostatic vacuum tube. *Physics*, vol. 2, pp. 432–42; June, (1932).

A physical picture of the phenomena during the electronic oscillation in the mag netostatic oscillator is given. The notion of a critical radius for plate voltages varying between a larger and a smaller value of the critical for a given constant magnetic field is developed. The flying time of the electrons is discussed in its relation to the voltage distribution. The falling angle is expressed in a simple way.

R140 M. Osnos. Eigenshaften eines freischweingenden Kreises der selbstinduktion, Kapazität, und Verlustwiderstand in Reichenshaltung enthält. (Characteristics of an oscillating circuit which contains inductance, capacitance and resistance connected in series). *Hochfrequenz. und Elektroakustik*, vol. 39, pp. 173–177; May, (1932).

A simple series circuit is treated mathematically for free and damped oscillations. It is shown that an oscillating circuit has besides the zero current, three distinct current values (maximum current, "Halbzeitströme," and initial current). The equation for the current is constructed.

R140 N. Howitt. Equivalent electrical networks. Proc. I.R.E., vol. 20, pp. 1042–1051; June, (1932).

The paper shows how to obtain, by a matrix multiplication, networks equivalent at all frequencies to a given network, as well as the networks, with the least number of elements.

R140 M. Reed. The analysis and design of a chain of resonant circuits. *Wireless Engineer & Experimental Wireless* (London), vol. 9, pp. 259–268, May; pp. 320–324, June, (1932).

An analysis is given of a chain of resonant circuits consisting of two, three, and four links respectively. Part II is concerned with the factors which influence the design of the above system. It is shown that the design of chains containing an odd number of links must be treated differently from those which contain an even number of links.

R141.2 ×R361.1 R. H. Langley. The application of permeability tuning to broadcast receivers. *Radio Engineering*, vol. 12, pp. 17–22; May, (1932).

Description of a new method of providing variable inductance tuning for radio receivers. A core of special material is inserted into a coil to provide tuning.

R142.5 L. G. A. Sims. Capacitive output coupling. *Wireless Engineer & Experimental Wireless* (London), vol. 9, pp. 314–319; June, (1932).

A theoretical consideration of the proper capacity to be used in capacitive output coupling.

R148.1 C. L. Farrar. Measurement of class B amplifier distortion. *Electronics*, vol. 4, pp. 196–198; June, (1932).

A method of measuring the per cent distortion of a class B amplifier is given. Representative experimental data are discussed with graphs.

R149 J. R. Nelson. Some notes on grid circuit and diode rectification Proc. I. R. E., vol. 20, pp. 989–1003; June, (1932).

The equivalent input resistance of a grid leak and condenser in parallel and the combination in series with a diode or the grid cathode circuit of a triode are calculated for various combinations by means of the static $I_g - E_g$ characteristics and an extension of the work of Colebrook and Peterson and Llewellyn. Experimental results and conclusions are given.

R165 A. R. Barfield. Dynamic speaker design. *Electronics*, vol. 4, pp. 188–190; June, (1932).

An analytical treatment is given which deals with the design of the magnetic circuit and driving coil which is attached to the diaphragm.

R165 D. N. McLachlan. On the symmetrical modes of vibration of truncated conical shells; with applications to loud speaker diaphragms. *Proc. Phys. Soc.* (London), vol. 44, pp. 408–425; May 1, (1932).

It is shown that in general the stresses in a vibrating conical shell are so complicated that the problem is unsuited to analytical treatment. Experimental work with paper, glass, and aluminum shells is described.

R200. Radio Measurements and Standardization

R210 H. O. Peterson and A. M. Braaten. The precision frequency measuring system of R.C.A. Communications, Inc. Proc. I.R.E., vol. 20, pp. 941–956; June, (1932).

A frequency measuring system is described wherein radio transmitter frequencies are compared with the harmonics of a piezo-electric frequency standard. Discussion of apparatus used, errors and operating aspects are given.

R214 ×R281. G. W. Fox and W. G. Hutton. Experimental study of parallel-cut piezo-electric quartz plates. *Physics*, vol. 2, pp. 443–447; June, (1932).

Y-cut piezo-electric quartz plates were studied with respect to the charges developed on the plates when oscillating near their resonant frequencies. The attractive force between crystal and electrodes was measured. Two methods for determining the piezo-electric voltage developed across the crystal are suggested. It is concluded from a study of crystal breakage that fracture is due to intense mechanical vibration.

R220 J. A. von den Akker. A method for measuring small capacities. *Rev. Sci. Inst.*, vol. 3, pp. 225–229; May, (1932).

A simple method for measuring small capacities is described.

R241.5 L. Behr and A. J. Williams, Jr. The Campbell-Shackelton shielded ratio box. Proc. I.R.E., vol. 20, pp. 969–988; June, (1932).

The Campbell-Shackelton shielded ratio box is intended to serve as a nucleus of a one-to-one bridge for the comparison of impedances. The transformer used in it operates successfully over the range from 50 to 50,000 cycles. The box may be used to form either a bridge with a Wagner earth connection or a bridge in which the junction point of the "standard" and "x" arms is grounded.

R242 H. Schwarz. Strommessung bei sehr hohen Frequenzen. (Current measurement at very high frequencies). *Hochfrequenz. und Elektroakustik*, vol. 39, pp. 160–171; May, (1932).

A comprehensive treatment of the subject is given. A preliminary investigation treats the possibility of comparing current measuring apparatus and the skin effect. Methods of measuring current are then given. Each instrument and its uses are briefly discussed. A method based on the permeability of iron between 2×10^7 and 2×10^8 cycles is given.

R243.1 ×621.313.7 H. Kaden. Über die Frequenzenterrung von Messgeräten MIT Trochengleichrichtern. (On the frequency correction of measuring apparatus with dry rectifiers). *Elekt. Nach. Tech.*, vol. 9, pp. 175–181; May, (1932).

It is shown that by the choice of input transformers and supplementary apparatus one may go a long way toward making measurements with instruments containing rectifiers independent of frequency. For voltage measurement, distinction is to be made between cases with and without previous amplification. Voltage and current measuring apparatus are considered.

R243.1 ×537.7 W. Greenwood. A valve voltmeter method of harmonic analysis. *Wireless Engineer & Experimental Wireless* (London), vol. 9, pp. 310–313; June, (1932).

A method for measuring the harmonics produced by audio-frequency transformers. A voltmeter needle is caused to swing with a beat frequency between the harmonic and a generated frequency. The ratio of the amplitude of swing when the generated frequency is near the harmonic to the amplitude of swing when generated frequency is near the fundamental is a measure of the harmonic amplitude.

R262.3 A. T. Starr. An aperiodic impedance measuring set. *Wireless Engineer & Experimental Wireless* (London), vol. 9, pp. 325–328; June, (1932).

A convenient method of measuring impedance is given. Very simple apparatus and no mathematics is involved.

R270 E. D. McArthur. Determining field distribution by electronic method. *Electronics*, vol. 4, pp. 192–194; June, (1932).

A system is described which is capable of rapidly determining the electric or magnetic field distribution about complicated structures with good accuracy in cases where Laplace's equation applies.

R300. Radio Apparatus and Equipment

R330 A. W. Hull. Electronic devices as aids to research. *Physics*, vol. 2, pp. 409–431; June, (1932).

This article contains information on several uses of vacuum tubes, thyratrons, etc.

R330 Progress in radio tubes. *Radio Engineering*, vol. 12, pp. 31–32; May, (1932).

Characteristics of a group of new tubes.

R330 A new group of receiving tubes. *QST*, vol. 16, pp. 35–36; June, (1932).

Pertinent information on the 56, 57, and 58 type vacuum tubes.

R350 W. R. Lyon. Transformer equipment for large experimental radiotelephone transmitter. *Bell Laboratory Record*, vol. 10, pp. 357–361; June, (1932).

Description of apparatus used in experimental long-wave transoceanic radio transmitter.

R355.5 H. N. Kozanowski. A new circuit for the production of ultra-short-wave oscillations. Proc. I.R.E., vol. 20, pp. 957–968; June, (1932).

This circuit consists of two tubes connected by symmetrical plate and filament Lecher systems instead of the usual plate-grid arrangement. The frequency of the oscillations is determined by the length of the plate Lecher circuit. Tuning of the filament Lecher system governs the amplitude of oscillations.

R355.9 C. A. Culver. An electrostatic alternator. *Physics*, vol. 2, p. 448; June, (1932).

By utilizing a continuously varying capacitance means are developed whereby a pure sine wave, as well as special forms may be produced. Filtering is not required. The theory underlying the production of any desired wave form is outlined. Factors which tend to modify the resultant wave form are pointed out.

R355.9 D. Hale. An audio oscillator of the dynatron type. *Rev. Sci. Inst.*, vol. 3, pp. 230–234; May, (1932).

An oscillator is described which makes use of the fact that a four-element vacuum tube may have a negative internal resistance under certain conditions. The oscillator is capable of frequency variation over a wide range by the variation of this negative internal resistance.

R363 G. Grammar, The A, B and C amplifier classification. *QST*, vol. 16, pp. 25–31; June, (1932).

The identification of amplifiers as class A, B, and C is explained.

R363 S. K. Waldorf. Amplifiers for precise oscillographic measurements. *Jour. Franklin Institute*, vol. 213, pp. 605–622; June, (1932).

A careful study is made of the conditions that an amplifier for use in oscillographic measurements must meet. A suitable amplifier is described. A method of compensating for the steady component of plate current is given. Photographic technique is also considered.

R363 L. L. Ringuet. L'amplificateur a lampes d'une grand sensibilité permet d'éclairer les preblèmes de physique nucléaire: Transmutation rayonnements ultrapénétrants et cosmiques. (The vacuum-tube amplifier of great sensitivity makes possible the explanation of problems of physical nuclei: Transmutation of ultrapenetrating and cosmic radiation.) *L'Onde Electrique*, vol. 11, pp. 157–181; April, (1932).

The use and description of an amplifier of high sensitivity which is used in amplifying small charges is given.

R363.2 C. E. Stromeyer. An improved 120-volt d.c. audio amplifier. *Electronics*, vol. 4, pp. 194–195; June, (1932).

An amplifier which operates from a d-c power line is described. The triple-twin type 291 vacuum tube is used.

R365.2 W. L. Parsons. Loud speakers with independent control added to radio receivers. *Radio Engineering*, vol. 12, p. 29; May, (1932).

Methods of installing several loud speakers on one output are given.

R383 H. L. White and E. A. von Atta. Electrolytic resistors of high resistance. *Rev. Sci. Inst.*, vol. 3, pp. 235–238; May, (1932).

Resistors of from 1×10^8 to 3×10^{10} ohms have been obtained by sealing capillary bridges between calomel electrodes made up of dilute KCl saturated calomel solutions.

R388 H. Neustadt. Thyratron linear time axis for cathode-ray oscillograph. *Electronics*, vol. 4, pp. 198–199; June, (1932).

A condenser placed across plate supply to a thyratron builds up a voltage (linearly with respect to time) which breaks down a kenotron thus discharging. Provision is made for changing frequency of discharge. The circuit may be used at frequencies up to or above 10,000 cycles.

R400. Radio Communication Systems

R470 ×621.319.2 K. O. Thorp. Pacific Gas and Electric extends its carrier system. *Bell Laboratory Record*, vol. 10, pp. 350–356; June, (1932).

Description of a power-line-carrier telephony system.

R500. Applications of Radio

R550 ×R423.5 M. von Ardenne. The transmission and reception of ultra-short waves that are modulated by several modulated high frequencies. Proc. I.R.E., vol. 20, pp. 933–940; June, (1932).

Two modulation connections are shown. Receiving and transmitting apparatus is described.

R583 A. E. Lyle. Sources of light for television. *Radio Engineering*, vol. 12, pp. 16; May, (1932).

Descriptions of modern type, flat plate, and crater lamps for television uses.

R600. Radio Stations

R600 R. L. Davis and V. E. Trouant. Westinghouse radio station at Saxonburg, Pa. Proc. I.R.E., vol. 20, pp. 921–932; June, (1932).

General description of power supply and station equipment.

R600 H. W. Ewen. The new high-power broadcast station at Beromunster, Canton Lucerne. *Marconi Review*, No. 35, pp. 1–14; March–April, (1932).

A new transmitter capable of introducing into the antenna system unmodulated energy of 60 KW and of transmitting on any frequency between the limits of 1000 and 500 kcs is described.

R600 Strong, Mirk, and Gallant. Le grand poste de radio-diffusion de Prague. (The large broadcast transmitting station at Prague.) *L'Onde Electrique*, vol. 11, pp. 182–208; April, (1932).

A general description of the new 200-KW transmitting (broadcast) station at Prague.

R800. Nonradio Subjects

621.314.3 W. Cauer. Ideale Transformatoren und lineare Transformation. (Ideal transformers and linear transformations.) *Elekt. Nach. Tech.*, vol. 9, pp. 157–174; May, (1932).

The equations for a 2n terminal network are set up. An extensive mathematical treatment using matrices is given.

621.314.6 R. Gürtler. Höchstausnutzung von gleichstrombelasteten Eisenkerndrosseln. (The maximum efficiency in iron-cored coils carrying direct current.) *Hochfrequenz. und Elektroakustik*, vol. 39, pp. 171–173; May, (1932).

It is shown that the product of the inductance and the square of the direct current for the highest allowable loss reaches a maximum value, from which the highest efficiency of a coil type is determined.

621.319.2 A. T. Starr. The nonuniform transmission line. Proc. I.R.E., vol. 20, p. 1052–1063; June, (1932).

The problem of transmission of periodic waves along a transmission line whose series impedance and shunt admittance per unit length vary as any powers of the distance from some point is solved. The results can be applied to the cases of a tapered submarine cable, an overhead line with a pronounced sag, and end effects in a high-tension line.

621.374.2 A. Heminway and J. F. McClendon. A. C. Wheatstone bridge for audio and radio-frequency measurements. *Physics*, vol. 2, pp. 396–402; May, (1932).

Description of a Wheatstone bridge suitable for making biological measurements.

621.374.7 W. W. Hansen. A lecture demonstration oscillograph. *Rev. Sci. Inst.*, vol. 3, pp. 305–308; June, (1932).

A cathode-ray oscillograph with a linear time scale is described. The apparatus is designed for demonstration use and will illustrate either 60 cycle or transient phenomena.

621.375.1 R. W. Carson. A grid-glow micrometer. *Electronics*, vol. 4, p. 191; June, (1932).

Needle point contacts cause a grid-glow tube to conduct, thus indicating when contact is made.

CONTRIBUTORS TO THIS ISSUE

Berejkoff, Anatole P.: Born 1893 at Tomsk, Russia. Received E. E. degree and B.S. degree in M.E., Polytechnic Institute of Petrograd and Angola. Technical work for the Russian Government, Departments of War and Means of Communication, 1914–1918. General Electric Test Course, 1920; radio engineering department, General Electric Company, June, 1921 to date. Associate member, Institute of Radio Engineers, 1923.

Berkner, L. V.:

Fick, Clifford G.: Born 1904 at Ida Grove, Iowa. Received B.S. degree in E.E., Iowa State College, 1925. General Electric Test Course, 1925; radio engineering department, General Electric Company, 1927 to date. Associate member, Institute of Radio Engineers, 1925.

Hanson, O. B.: Born 1894 at Huddersfield, England. Attended Royal Masonic Institute, Bushey, Hertfordshire. Chief radio operator, *S. S. Stephano*, 1916. Test department, Marconi Company at Aldene, N. J., 1917–1921; engineer in charge, Station WAAM, 1921–1922; assistant plant manager, radio broadcasting department, American Telephone and Telegraph Company, operating WEAF, 1922-1923; plant manager, 1923–1926. Manager, plant operation and engineering, National Broadcasting Company, 1926 to date. Member, Acoustical Society of America. Associate member, Institute of Radio Engineers, 1918; Member, 1927.

Honeyball, L. L.:

King, Ronold: Born September 19, 1905, at Williamstown, Massachusetts. Received A. B. degree, University of Rochester, 1927; M. S. degree, 1929; Ph. D. degree, University of Wisconsin, 1932. American-German Exchange Student at Munich, 1928–1929; White Fellow in Physics, Cornell University, 1929–1930; University Fellow in Electrical Engineering, University of Wisconsin, 1930–1932. Associate member, Institute of Radio Engineers, 1930.

Luck, David G. C.: Born 1906 at Whittier, California. Received B. S. degree, 1927; Ph. D. degree in physics, 1932, Massachusetts Institute of Technology. Research assistant, Raytheon Manufacturing Company, 1927. Fellow in Physics, Massachusetts Institute of Technology, 1927–1929. Research engineer, Thomas A. Edison, Inc., 1929. Assistant, department of physics, Massachusetts Institute of Technology, 1929–1932. Nonmember, Institute of Radio Engineers.

McNally, J. O.: Received B. S. degree, University of New Brunswick, 1924. Since 1924 with Western Electric Company and Bell Telephone Laboratories, engaged in vacuum tube development and in the study of distortion in vacuum tube output. Associate member, Institute of Radio Engineers, 1926.

Mouromtseff, Ilia Emmanuel: See Proceedings for May, 1932.

Noble, H. V.: Born April 14, 1907, near Noblesville, Indiana. Received B. S. degree in E. E., Purdue University, 1929. Vacuum tube engineering department, Grigsby-Grunow Company, 1929; research laboratories, Westinghouse Electric and Manufacturing Company, 1930 to date. Member, Eta Kappa Nu; associate member, Sigma Xi. Associate member, Institute of Radio Engineers, 1931.

Palmer, Lionel S.: Born August 2, 1891, at London, England. Graduate, Bristol University, 1913. Captain, British Army, 1914–1918. Received M.S. and Ph.D. degrees, 1919–1920. Radio research engineer, H. M. Navy, 1920–1922. Lecturer in Radio Engineering, 1922–1924; Head of Physics Department, Manchester College of Technology, 1924–1926; professor of Physics, University College of Hull, 1927 to date. Member, I.E.E., Fellow, Institute of Physics. Nonmember, Institute of Radio Engineers.

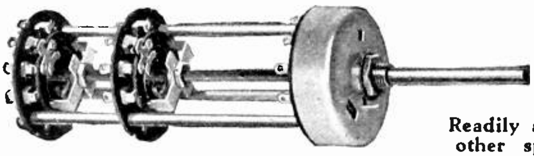
Ritter, E. W.: Born 1902 at Orangeville, Indiana. Received B. S. degree in E. E., Purdue University, 1925. Nela Park Works, General Electric Company, Cleveland, Ohio, 1925–1930; research and development laboratory, RCA Radiotron Company, Inc., 1930 to date. Member, Institute of Radio Engineers, 1930.

Schmit, D. F.: Born 1902 at Port Washington, Wisconsin. Received B. S. degree in E. E., University of Wisconsin, 1923. Research laboratory, General Electric Company, Schenectady, 1923–1926; E. T. Cunningham, Inc., 1926–1931; in charge, application engineering section, research and development laboratory, RCA Radiotron Company, Inc., 1931 to date. Associate member Institute of Radio Engineers, 1925.

von Ardenne, Manfred: See PROCEEDINGS for June, 1932.

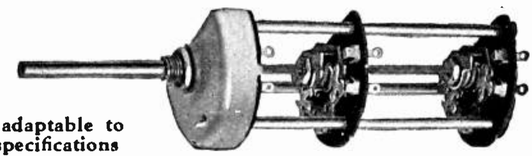
Warner, J. C.: Born February 14, 1896, at Freeport, Indiana. Received B. A. degree, Washburn College; M. A. degree, University of Kansas; M. S. degree in E. E., Union College. Radio work, Signal Corps, U. S. Army, 1917–1919; instructor in physics, University of Kansas, 1919–1920; assistant physicist, Bureau of Standards, 1920; research laboratory, General Electric Company, 1920–1930; vacuum tube engineering department, 1930 to June, 1931; manager, research and development laboratory, RCA Radiotron Company, Inc., 1931 to date. Associate member, Institute of Radio Engineers, 1919; Member, 1926.

SPECIAL ITEMS FOR RADIO MANUFACTURERS

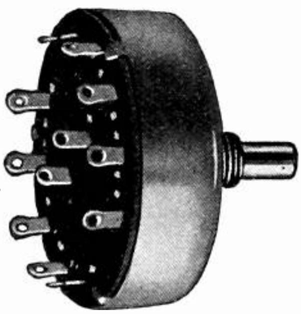


No. 7280A
3 gang, 4 positions
single circuit

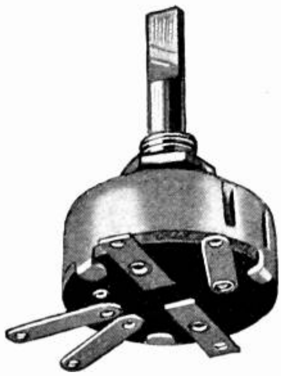
Readily adaptable to
other specifications



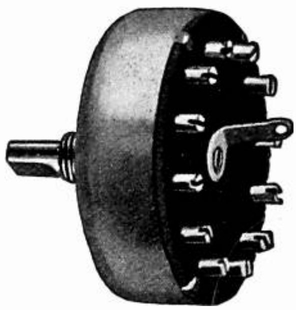
No. 7260A
3 gang, 2 positions
2 circuits



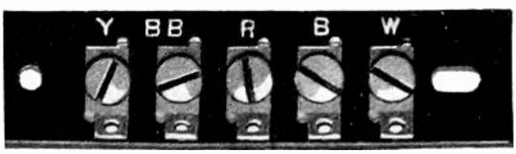
No. 7150A
5 positions, 2 circuits



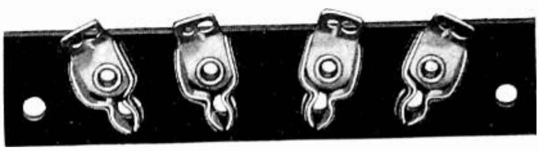
No. 7120A
Combined Acoustic and muting switch 2 or 3 positions



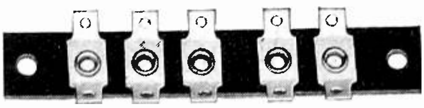
No. 7220A
12 positions, single circuit



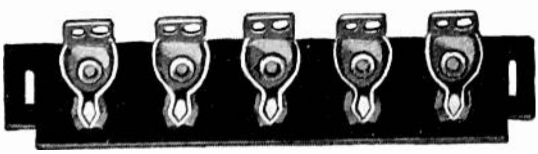
Terminal Strip
5 Lugs No. 6880A, 4 lugs No. 6990A



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.
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"SAY HELLO TO DADDY"

HE MAY be downtown at the office or a hundred miles away . . . yet that happy, eager voice wings across the wires, straight into his heart. It summons up a sudden, tender warmth. It sweeps away cares and worries. It brings sure, comforting knowledge that all is well at home.

Only a small voice, speaking into a telephone. But it can create a moment that colors the whole day.

If you stop to reflect, you will realize how immeasurably the telephone contributes to your family's happiness and welfare. It is a fleet courier . . . bearing messages of love, of friendship. A priceless helper . . . ready to aid in the task of running a household. A vigilant guardian . . . always at hand when emergencies arise.

Security, convenience, contact with all the world—these things the telephone brings to your home. You cannot measure their value in money. You cannot determine the ultimate worth of telephone service.

But consider, for a moment, that your telephone is one of a country-wide system of nineteen million others — a system of many million miles of wire served by hundreds of thousands of employees. Yet you pay only a few cents a day for residential use. And you enjoy the most nearly limitless service the world affords.

AMERICAN TELEPHONE AND TELEGRAPH COMPANY

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)

EMPLOYMENT PAGE

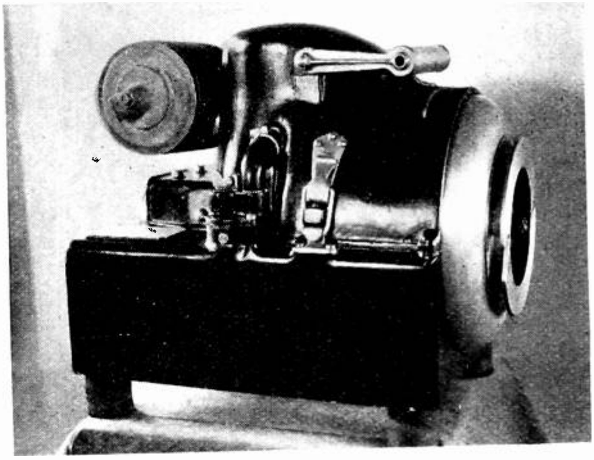
Advertisements on this page are available to members of the Institute of Radio Engineers and to manufacturing concerns who wish to secure trained men for positions. All material for publication on this page is subject to editing from the Institute office and must be sent in by the 15th of the month previous to the month of publication. (August 15th for September 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.

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.

RADIO ENGINEER having three years laboratory work and four and one half years additional responsible position in charge of laboratory of midwest radio manufacturer. Desires position where past training can be most profitably employed. Past experience includes design and production of radio receivers, design of television equipment and commercial operating. College man. Single. Will travel. Age 33. Box 131.

COLLEGE GRADUATE with two years broad factory experience as student engineer, half year field engineer on construction, supervisor, investigation of customer's premises, and two and one half years development of telephone system power plants. Desires immediate position as development, field or plant maintenance engineer. Best of references. E.E. 1927; M.S. in E.E. 1931. Single. Will travel. Age 26. Box 132.



Power *really* Portable!

Heintz and Kaufman Ltd. portable and aircraft equipment was the choice of Rear Admiral Richard E. Byrd, Sir Hubert Wilkins and Captain Charles Kingsford-Smith—it should be your choice!

¶The Type 300 Engine Generator weighs *31 pounds* and occupies a space of but 10″ x 12″ x 14″ over all.

¶There are only *three major moving parts.*

¶There are no revolving windings, sliprings, commutators or brushes.

¶The output is *governed electrically* without moving parts.

¶The output, 300 watts, at 350 cycles may be transformed to any voltage for use directly or through rectifiers and filters. Either single or two-phase can be supplied.

¶In addition to its use for operating radio transmitters, the Type 300 unit will supply power for general lighting, searchlights, moving picture projectors, x-ray tubes, Neon signs—the uses being limited only by the power.

¶Our years of experience in the development of light weight equipment are at your disposal in the sensible solution of your problems calling for portable power for any use.

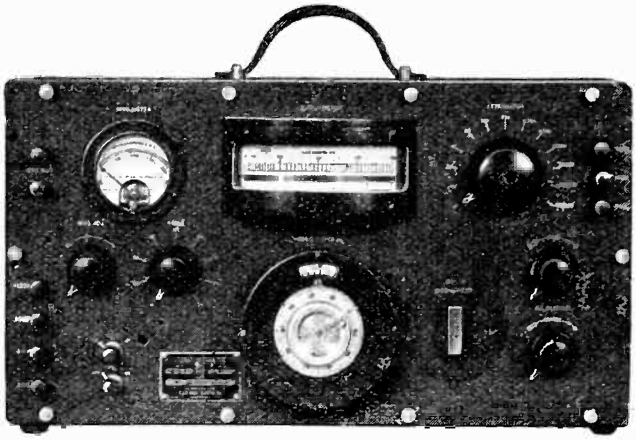
¶Descriptive literature on the Type 300 unit is available.

HEINTZ AND KAUFMAN LTD

311 California St., San Francisco, Cal.

Factory, South San Francisco, Cal.

The *new* type TMV-18C standard Signal Generator



Specifications

Frequency Range: 90 Kc. to 10,000 Kc. using six shielded coils easily removable from the front of the panel. Special coils can be supplied for extending the range from 25 Kc. to 25,000 Kc.

RF Voltage Output: ¼ microvolt to 2 volts continuously variable, output substantially constant over range of any coil.

Modulation: Internal 400 cycle oscillator capable of modulating the output up to 80% in 10% steps. Provision for external modulation. Employs self-calibrating modulation meter.

Output System: Special designed resistance attenuator accurate to within 5% in the broadcast and lower frequency ranges and within 10% above the broadcast range. A precision, long scale, thermo-volt meter is used to measure the voltage input to the attenuator. Meter accuracy 0.25 of 1% at constant temperature.

Frequency Calibration: Accurate to within 0.25 of 1%. Frequency controlled by new precision dial, 60:1 ratio, spreading 270° rotation of condenser over 4500 divisions.

Leakage: Potential differences between any external grounded point, meters or controls, is less than 0.1 microvolt. The stray field is not sufficient to affect the accuracy of measurements within the range of the instrument. The oscillator and attenuator are enclosed in heavy aluminum castings.

Mechanical Features: Enclosed in heavy aluminum shielding case finished with black crackle lacquer. Dimensions approximately 21″ x 12″ x 9″. Weight approximately 40 lbs.

Tubes Required: 1 RCA-56; 1 UX-112A; 1 UX-230.

Price: Complete with six coils (90—10,000 Kc.) and carrying case, $700.00.

INDUSTRIAL PRODUCTS SECTION

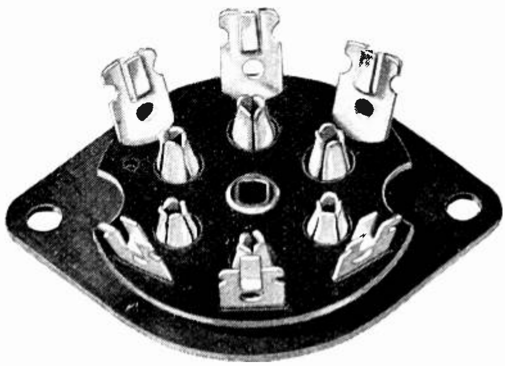
RCA Victor Company, Inc.

A Radio Corporation of America Subsidiary

CAMDEN, N. J.

"RADIO HEADQUARTERS"

New!---the
CINCH RADIO SOCKET WITH
Floating contacts!



Another Cinch achievement! A new radio socket . . . with *floating contacts*. Scientifically designed to eliminate *all* strain on the bakelite . . . preventing warping and loss of tension after tube is inserted. *Floating* principle makes contacts self-aligning assuring a constant, rigid contact on each tube prong. Three styles of contacts. This construction eliminates all holes generally used for riveting contacts to bakelite . . . making the socket considerably *stronger*.

The new Cinch Radio Sockets are made for 4, 5 and 6-prong tubes with $1\frac{27}{32}$", $1\frac{11}{16}$" and $1\frac{1}{2}$" mounting centers.

Write for Blue print and samples . . . yours without obligation.

CINCH PRODUCTS:

Standard and Midget Size Radio Sockets . . . Binding Posts . . . Soldering Lugs . . . Insulated Mounting Strips . . . Tip-Jacks . . . Small Intricate Metal Stampings.

CINCH MANUFACTURING CORP.

2335 W. Van Buren St., Chicago, Ill.

Subsidiary of

UNITED-CARR FASTENER CORP.
CAMBRIDGE, MASS.

Take Your Choice

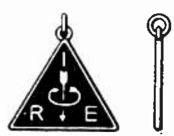
Every member of the Institute of Radio Engineers will want to wear one of these three attractive emblems. All of these emblems are appropriately colored to indicate the various grades of membership and each emblem is approximately the size of the following illustrations.

The Lapel Button is of 14k gold, with background enameled in the membership color, and with the lettering in gold. The screw-back on the back of this button fastens it securely to your coat.

$2.75 postpaid—any grade.

The Pin is also of 14k gold. It is provided with a safety catch and is appropriately colored for the various grades of membership.

$3.00 postpaid—any grade.



The Watch Charm, handsomely finished on both sides, is also of 14k gold. This charm is equipped with a suspension ring for attaching to a watch fob or chain.

$5.00 postpaid—any grade.

Orders, accompanied by checks made payable to the Institute of Radio Engineers, should be addressed to

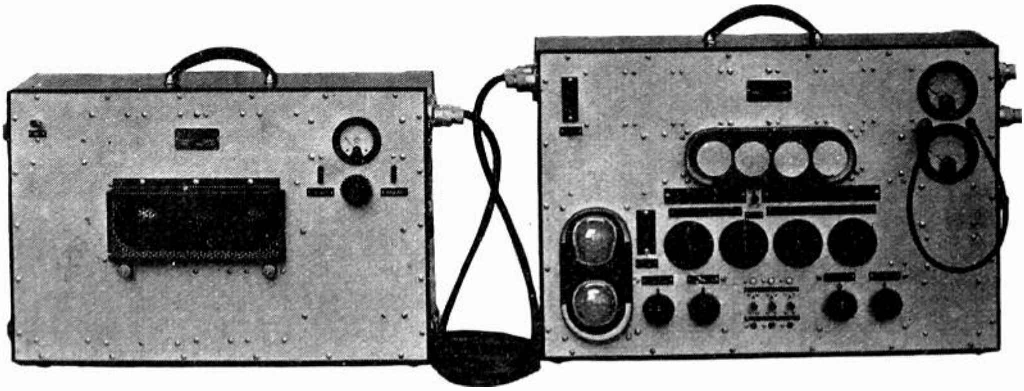
THE INSTITUTE OF RADIO ENGINEERS

33 West 39th Street

NEW YORK CITY, N. Y.

Send in your order now!

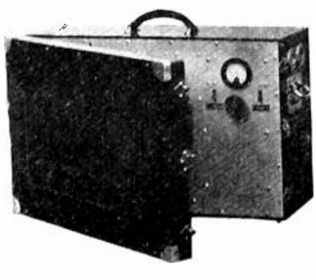
AMERTRAN *Portable*
PUBLIC - ADDRESS SYSTEM



AMERTRAN Type PA-71 is a complete portable public-address system mounted in two units which may be carried about easily. Although of such compact design, frequency characteristics and the efficiency of the circuits equal in performance the standard panel-type AmerTran Sound Systems.

The equipment is mounted in attractive, durable cases of quartered oak with nickel plated hardware, fitted with convenient handles. Plugs with cables facilitate rapid connection of the apparatus. Instruments and controls, with engraved bakelite designation strips are mounted on aluminum panels

The larger unit contains a three-circuit mixer feeding into a four-stage amplifier having an undistorted output of 33dB (12.5 watts). Operating power is obtained from the smaller unit housing a rectifier and filter system. The system may be operated wherever 110 volt, 60 cycle lighting circuits are available, and the source of signal may be microphones, phonograph or radio set.



Overall measurements on Type PA-71 indicate a frequency variation of less than 2dB throughout the band of 40 to 8000 cycles, and a gain which is ample for all requirements, no additional preamplifier being necessary when using microphones.

Unusual freedom from a.c. hum is also obtained by supplying filaments of heater type tubes with rectified and filtered current instead of the usual raw a.c. Furthermore, the equipment may be used with all standard apparatus, as the three input circuits each have 200 ohms impedance and the output has an impedance of 500 ohms.

Write us direct for prices and further information.

AMERICAN TRANSFORMER COMPANY

Main Office and Factory

178 Emmet Street **Newark, N. J.**

Cornell

TRADE MARK REG.

"CUBS"

EARNED LEADERSHIP

It has been our privilege to work with the foremost radio set manufacturers throughout the United States and Canada, to make their condenser problems our own and to meet their particular requirements on a commercial scale, quickly and economically.

Our leadership has been gained through our successful handling of many such cooperative researches throughout the radio industry. If you have any changes in mind, we feel confident you will benefit by consulting us. We can assist you materially. Put your condenser problem up to Cornell.

WRITE FOR CATALOG A

Describing our complete line of Paper Dielectric Capacitors for radio, television and ignition, and Resistors for radio and television.

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

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	

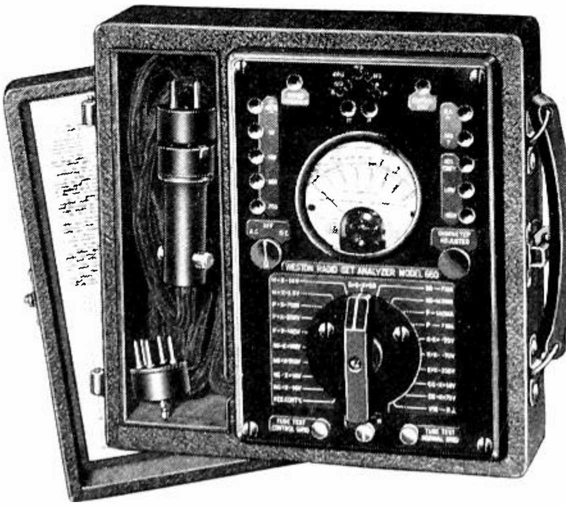
Record may be continued on other sheets of this size if space is insufficient.

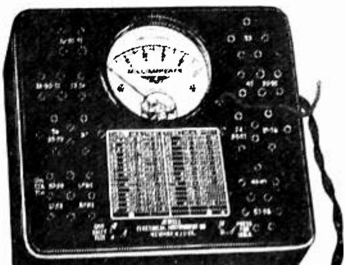
Receipt Acknowledged............ Elected............ Deferred............
Grade .... Advised of Election........ . This Record Filed...........

New ... A Weston Single Meter ANALYZER

Every necessary socket voltage and current measurement, both A.C. and D.C., can be made on all modern receivers with this remarkable Weston Single Meter Analyzer.

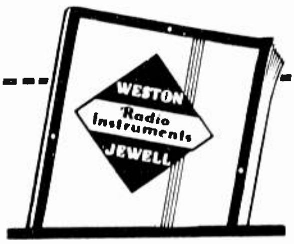
Point-to-point tests are provided for with three resistance and ample voltage and current measuring ranges, available at pin-jacks on the panel. Full description is contained in the new Weston-Jewell catalog.





Pattern 675 Tube-Checker

A new, low-priced instrument that checks all present types of tubes without the use of adapters. Sixteen tube sockets are ingeniously arranged to occupy small space. Test limits for all tubes are etched on the instrument panel. Compact and light in weight, this tube checker is popular for counter and portable use. Write for Weston-Jewell catalog describing this great instrument value!



Weston Electrical Instrument Corp. } 589 Frelinghuysen
Jewell Electrical Instrument Co. } Ave., Newark, N.J.

Please send me a copy of your new catalog listing the complete Weston-Jewell line of radio instruments.

Name ....

Address ....

WESTON-JEWELL

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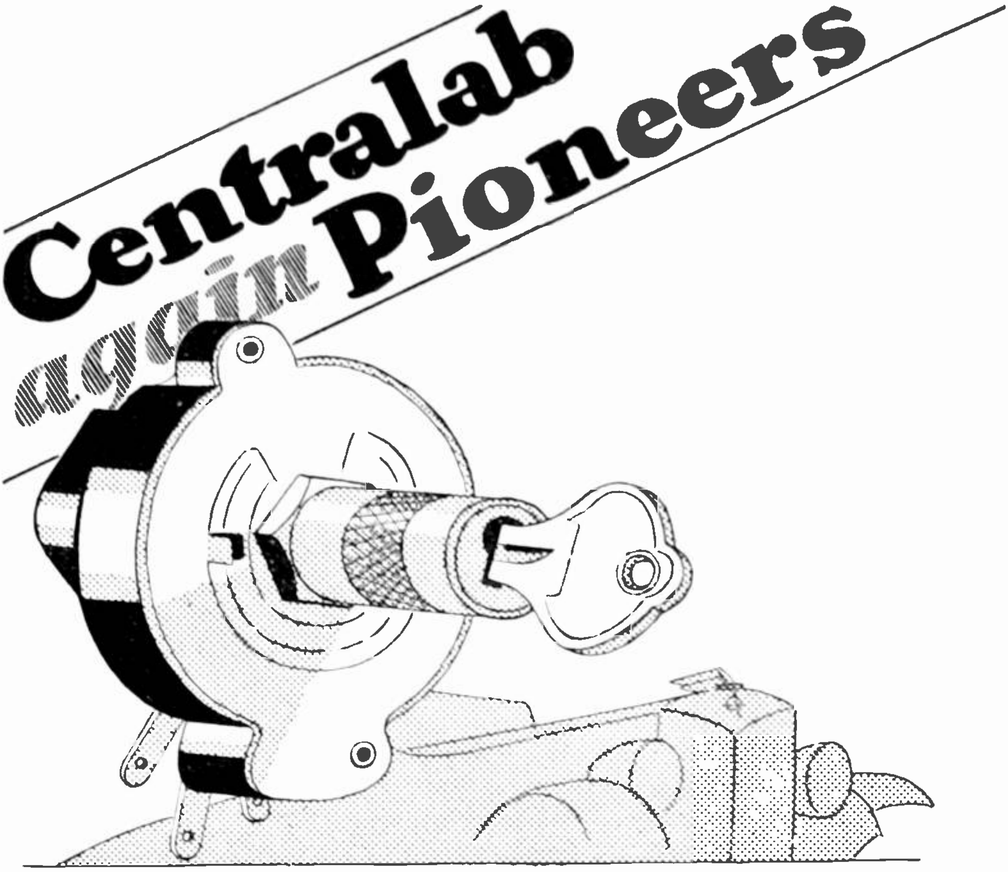

Centralab
again Pioneers

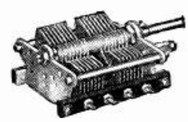
TIME-INSURED QUALITY

CARDWELL variable condensers and CARDWELL-designed and built-to-order apparatus have, for considerably more than a decade, enjoyed a reputation for outstanding excellence won by conscientious adherence to an ideal of fair dealing and a determination to sponsor only *quality* products.

Look around. Somewhere, in the majority of the really fine radio installations and equipments in the country, governmental, broadcast, commercial or amateur, will be found some product of CARDWELL manufacture.

What better proof of the recognized excellence of CARDWELL apparatus could be offered?

The CARDWELL line of condensers includes transmitting condensers for high, medium and low power, and receiving condensers in infinite variety. The CARDWELL Midway Featherweight (all aluminum, none over 7 ounces in weight) for transmitting, receiving and neutralizing, is unequalled for aircraft and portable equipment.



May we send our literature?

CARDWELL CONDENSERS AND MANUFACTURING SERVICE

The Allen D. Cardwell Mfg. Corpn.
93 Prospect Street, Brooklyn, N.Y.

"THE STANDARD OF COMPARISON"

WORLD WIDE SERVICE

FACTORIES in Erie, Pennsylvania, Toronto, Canada and London, England make and distribute Erie Resistors throughout the world. This world wide service has been built on the solid foundation of precision in manufacture plus a product development service of the highest order.

ERIE RESISTORS have been developed constantly to meet the exacting needs of their users. Precision in manufacture has been guarded consistently step by step by laboratory checks. Nothing is left to chance and every detail in the making of ERIE RESISTORS is regarded as a step which must be triple checked and tested to insure successful installation.

Surely, resistors which have established a record for the smallest percentage of rejections ever recorded in the radio industry and with world wide service merit your attention and careful consideration. May we send you samples and prices?

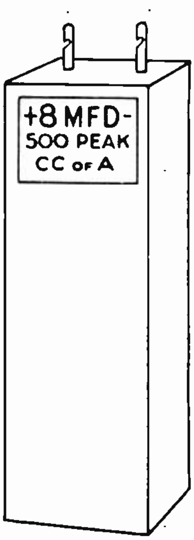
ERIE RESISTORS

ERIE RESISTOR CORPORATION, ERIE, PA.

Factories in Erie, Pa., Toronto, Canada and London, Eng.

If you want dependable condenser performance-specify the new ACRACON SEMI-DRY ELECTROLYTIC CONDENSER

Compare These Characteristics:



Type DC-8
Acracon
Semi-Dry Unit

- peak operating voltage 500
- surge voltage 600
- low initial leakage
- leakage current at 500 volts less than .2 mils per mfd.
- constant capacity; does not decrease with age
- stable power factor; does not increase with age
- non-corrosive connections
- metal or fibre container
- standard and special sizes

Type DE-8
Acracon
Semi-Dry Unit

Write Today For Catalog!

* *Acracon Features are Protected by Patents Pending*

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