VOLUME 19 AUGUST, 1931 NUMBER 8

PROCEEDINGS
of
The Institute of Radio Engineers

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Institute of Radio Engineers
Forthcoming Meetings

PITTSBURGH SECTION
September 22, 1931

SAN FRANCISCO SECTION
September 16, 1931

SEATTLE SECTION
September, 24, 1931

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SUGGESTIONS FOR CONTRIBUTORS TO THE "PROCEEDINGS"

Preparation of Paper

Form—Manuscripts may be submitted by member and nonmember contributors from any country. To be acceptable for publication, manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume, page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer divisions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be 3/16 in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportion.

Abbreviations—Write a.c. and d.c. (a-c and d-c as adjectives), kc, μf, μμf, e.m.f., mh, μh henries, abscissas, antennas. Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a nonspecialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

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Proceedings of the Institute of Radio Engineers
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China	Shanghai, 81 Rue du Consulat	Wang, T. S.

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Proceedings of the Institute of Radio Engineers
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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 August 31, 1931. These applicants will be considered by the Board of Direction at its September 2nd meeting.

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England	Harrow, Middlesex, 29 The Gardens, Vaughan Rd.	Montgomery, A. W.

For Election to the Associate grade

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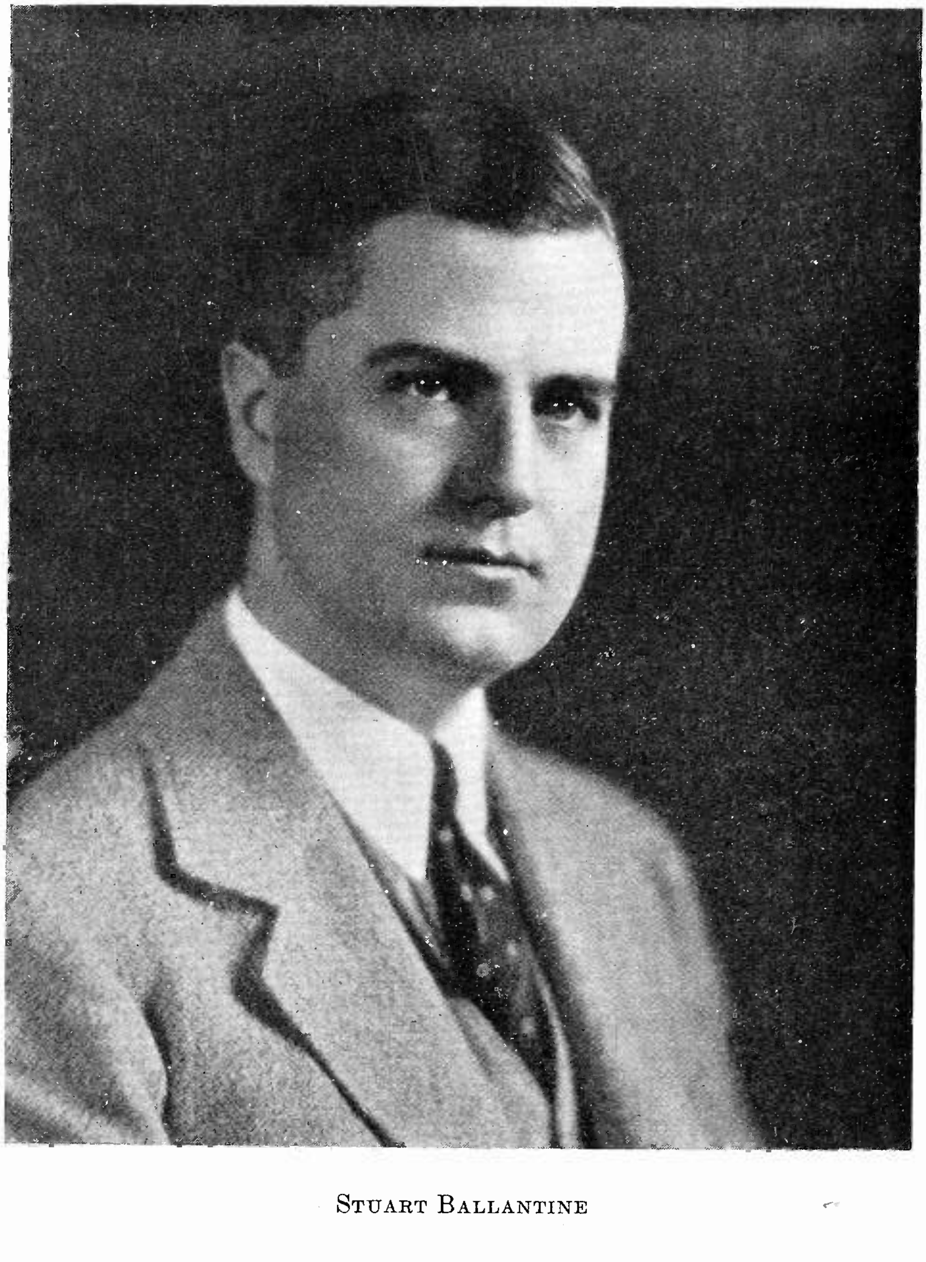

Stuart Ballantine

STUART BALLANTINE

Recipient of Morris Liebmann Memorial Prize for 1931

In view of his outstanding theoretical and experimental investigations of numerous radio and acoustic devices, the Morris Liebmann Memorial Prize for 1931 was awarded to Stuart Ballantine of Boonton, N. J.

Stuart Ballantine was born September 22, 1897, at Germantown, Philadelphia, Pa. He became interested in wireless telegraphy as an amateur in 1908 and served as a ship radio operator during the summers of 1913, 1914, and 1915. He was employed during 1916 by H. K. Mulford Company, bacteriologists, and during 1917 by the Bell Telephone Company of Pennsylvania.

From 1917 to 1920 as Expert Radio Aide at the Philadelphia Navy Yard, he was responsible for the design of the Navy coil-type radio compasses. He discovered the "antenna effect" in coil-type systems and invented the capacity compensator for its control.

During 1920 and 1921 he studied mathematical physics at the Graduate School, Harvard University and spent 1923 and 1924 as a John Tyndall scholar in physics at Harvard.

In 1922 and 1923 while doing development work on broadcast receivers at Radio Frequency Laboratories, Boonton, N.J., he invented a method of stabilizing radio-frequency amplifiers by means of a Wheatstone bridge circuit. Linear detection at high signal levels and automatic volume control for radio receivers were also developed by him at this time.

He was privately engaged in scientific investigations at White Haven, Pa., from 1924 to 1927 and then rejoined Radio Frequency Laboratories in charge of the research division, becoming principally engaged in studies of detection at high signal levels, fluctuation noise in radio receivers and tubes, development of technique for sound measurements of loud speakers and receivers, microphone calibration, and broadcast receiver design.

He collaborated in 1929 with F. M. Huntoon in an investigation of the effects of high pressure on bacteria.

Since June, 1929, he has been President of the Boonton Research Corporation where he developed an electrostethoscope, automatic optical recorder for frequency-response measurements, logarithmic voltmeter, investigation of errors in microphones due to diffraction and cavity resonance, and collaborated with H. A. Snow in development of variable-mu vacuum tubes.

He is the author of one book and thirty-one papers on radio, acoustics, electrical theory, and other subjects.

He organized the Philadelphia Section of the Institute in 1920 acting as its chairman until 1926. He served on the Committee on Membership (1926–1927), Meetings and Papers Committee (1929–1930), Standardization Committee (1930–1931), Awards Committee (1930–1931), Chairman of the Technical Committee on Vacuum Tubes (1930), Chairman of the Technical Committee on Electro-Acoustic Devices (1931), and a member of the Board of Editors (1929–1931).

He is a Fellow of the American Physical Society, The Acoustical Society of America, and a Member of the Franklin Institute. He became an Associate member of the Institute of Radio Engineers in 1916 and a Fellow in 1928.

INSTITUTE NEWS AND RADIO NOTES

Sixth Annual Convention

The Sixth Annual Convention of the Institute was held under the sponsorship of the Chicago Section on June 4, 5, and 6. The headquarters of the convention was the Hotel Sherman in Chicago.

The technical papers of which there were nineteen were presented at four technical sessions, two of which were held on the first day of the convention while the other two were held on the following two days.

A number of inspection trips permitted visits to the manufacturing plant of the Grigsby-Grunow Manufacturing Company, the Long Lines Department of the American Telephone and Telegraph Co., the Exchange of the Illinois Bell Telephone Co., the National Broadcasting Co. Studios, the Hawthorne Works of the Western Electric Co., and the Adler Planetarium and Astronomical Museum. In addition, the ladies' trips included a tea and fashion promenade at Marshall Fields, a theater party, a sight-seeing tour, a luncheon and bridge at Maillard's, the National Broadcasting Co. Studios, and the Adler Planetarium and Astronomical Museum.

On the evening of June 4 a lecture on "Modern Conceptions of the Electron" was presented by Professor A. H. Compton of the Physics Department of the University of Chicago.

During the banquet, which was held on the evening of June 5, announcement was made of the presentation of the Institute Medal of Honor to General G. A. Ferriè of France, who was unable to be present due to the necessity of his being in attendance at the meeting of the International Consulting Committee on Radio in Copenhagen at approximately the same time. The Morris Liebmann Memorial Prize was then awarded to Stuart Ballantine. Almost a dozen golf trophies were presented as the result of the golf tournament played at Calumet Golf Course on the afternoon of the 5th.

On the evening of the 4th, the annual meeting of the Committee on Sections was held and representatives of the Buffalo-Niagara, Chicago, Cincinnati, Cleveland, Connecticut Valley, Detroit, Rochester, and Toronto Sections were present. The discussion which took place covered practically the entire field of section operation and maintenance. A financial report form was agreed upon so that financial reports of sections can hereafter be prepared uniformly which will permit more ready analysis. Although the present rebate system was discussed fully, it was not possible to make any important decisions regarding it in view of the fact that it has been in operation for about six months only.

The Constitution for Sections was reviewed briefly to determine if there were any corrections or additions needed. The necessity of each section's adopting a by-law specifying when the annual meeting of the section would be held was pointed out and as only five of the eight sections represented had adopted such by-laws, officers of all sections were requested to see that such a by-law is adopted.

It was agreed that a meeting of the Sections Committee should be scheduled for Rochester on either November 18 or 19 during the time of the Rochester Fall Meeting. Details of this meeting will be forwarded to all sections as soon as they have been worked out.

The exhibition which was a part of the convention was comprised of almost fifty booths representing practically all of the manufacturers of component parts, measuring and laboratory equipment, and materials and equipment of interest to manufacturers.

The registration for the convention totaled four hundred, the attendance at the banquet being two hundred and twenty-five.

Bound Volumes

The twelve issues of the PROCEEDINGS published during 1930 are now available in blue buckram binding to members of the Institute at nine dollars and fifty cents ($9.50) per volume. The price to nonmembers of the Institute is twelve ($12.00) dollars per volume.

Radio Transmissions of Standard Frequency, August and September, 1931

The Bureau of Standards announces a new schedule of radio transmissions of standard frequencies. This service may be used by broadcast and other stations in adjusting their transmitters to exact frequency, and by the public in calibrating frequency standards and transmitting and receiving apparatus. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to about 1000 miles from Washington, and some of them at all points in the United States. The time schedules are different from those used in transmissions prior to this July.

There are two classes of transmissions provided: one, transmission of the highest accuracy at 5000 kc for two hours afternoon and two hours evening on three Tuesdays in each month; the other, transmissions of a number of frequencies in two-hour periods in the afternoon and evening, one Tuesday a month. The transmissions are by continuous-wave radiotelegraphy. The 5000-kc transmissions consist mainly of a continuous cw transmission, giving a continuous whistle in the receiving phones. The first five minutes of this transmission consist of 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. The transmissions of the other type are also by continuous-wave radiotelegraphy. A complete frequency transmission includes a "general call," "standard frequency signal," and "announcements." The general call is given at the beginning of each 18-minute period and continues for about two minutes. This includes a statement of the frequency. The Standard frequency signal is a series of very long dashes with the call letters (WWV) intervening; this signal continues for about 8 minutes. The announcements follow, and contain a statement of the frequency being transmitted and of the next frequency to be transmitted. There is then a 6-minute interval while the transmitting set is adjusted for the next frequency.

5000-Kilocycle Transmissions
2:00 to 4:00 P.M., and 10:00 P.M., to 12:00 Midnight, Eastern Standard Time

July	August	September
14	11	8
21	18	15
28	25	22
		29

Multifrequency Transmissions

		Frequencies in Kilocycles		
Eastern Standard Time		July 7	August 4	September 1
2:00 P.M.	10:00 P.M.	1600	3600	6400
2:18	10:18	1800	4000	7000
2:36	10:36	2000	4400	7600
2:54	10:54	2400	4800	8200
3:12	11:12	2800	5200	8800
3:30	11:30	3200	5800	9400
3:48	11:48	3600	6400	10000

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 280, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequencies are received (or even only a single one), persons can obtain as complete a frequency meter calibration as desired by the methods of generator harmonics.

The 5000-kc transmissions are from a transmitter of 1 kilowatt power; they occur every Tuesday except the first in each month. The other transmissions are from a transmitter of 1/2 kilowatt power; they are given on the first Tuesday of every month both afternoon and evening.

The frequencies in the 5000-kilocycle transmissions are piezo controlled, and are accurate to much better than a part in a million. The frequencies in the multifrequency transmissions are manually controlled, and are accurate to a part in a hundred thousand.

Since the start of the 5000-kc transmissions the Bureau of Standards has been receiving reports regarding the reception of these transmissions and their use for frequency measurements from nearly all parts of the United States, including the Pacific coast and Alaska. The Bureau is desirous of receiving more reports on these transmissions, especially because radio transmission phenomena change with the season of the year. The data thus far obtained cover the first six months of 1931, and give information regarding approximate field intensity, fading, and the suitability of the transmissions for frequency measurements.

It is suggested that in reporting upon the field intensity of these transmissions, the following designations be used where field intensity measurement apparatus is not at hand: (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 whether slow or rapid, and time between peaks of signal intensity. Statements as to the type of receiving set used in reporting on the transmissions and the type of antenna used are likewise desired. The Bureau would also appreciate reports on the use of the transmissions for purposes of frequency measurement or control.

Reports on the reception of the transmissions should be addressed to Bureau of Standards, Washington, D. C.

Committee Work

COMMITTEE ON BROADCASTING

A meeting of the Committee on Broadcasting was held at 3 P.M. on Thursday, June 18, in the office of the Institute. Those in attendance were L. M. Hull, chairman; Arthur Batcheller, J. B. Coleman, (representing B. R. Cummings), P. A. Greene, J. V. L. Hogan, C. W. Horn, R. H. Marriott, and E. L. Nelson.

The committee continued its consideration of General Order 97 of the Federal Radio Commission, putting their comments thereon in final form to be placed before the Board of Direction at its September 2 meeting for subsequent transmittal to the Commission.

COMMITTEE ON SECTIONS

The annual meeting of the Committee on Sections which is reported with the other activities of the Sixth Annual Convention was attended by the following: Austin Bailey, chairman; R. H. Manson, H. A. Brown, A. B. Buchanan, F. K. Dalton, Samuel Firestone, V. M. Graham, R. A. Hackbusch, G. B. Hammon, J. B. Hoag, L. N. Holland, C. E. Kilgour, J. J. Lamb, J. M. Leslie, B. B. Minnium, B. E. Shackelford, B. Dudley, and H. P. Westman.

STANDARDIZATION

TECHNICAL COMMITTEE ON VACUUM TUBES—IRE

Two meetings of the Technical Committee on Vacuum Tubes of the Institute were held on May 12 and July 7, respectively. The May meeting was attended by B. E. Shackelford, chairman; N. P. Case, H. F. Dart, F. H. Engel, J. F. Hanley, E. A. Lederer, W. M. Perkins, L. M. Price, E. W. Ritter, H. A. Snow, W. M. Tuttle (nonmember), J. C. Warner, K. S. Weaver, P. T. Weeks, and B. Dudley, secretary.

The July meeting was attended by B. E. Shackelford, chairman; R. R. Batcher (nonmember), N. P. Case, F. H. Engel, J. F. Hanley, M. J. Kelly, A. F. Merrill (representing E. A. Lederer), W. N. Tuttle (nonmember), Dayton Ulrey, J. C. Warner, K. S. Weaver, P. T. Weeks, B. Dudley, secretary; and H. P. Westman.

At these meetings the committee reviewed briefly the 1931 Report of the Committee on Standardization as regards that portion of it devoted to vacuum tubes. It was thought desirable to give careful consideration to several of the definitions appearing in this report and a number that might advantageously be added when the next report is issued.

It was decided to establish one or more subcommittees to study certain portions of the report during July and August so that their comments may be available to the committee at its next meeting on September 15.

Institute Meetings

CINCINNATI SECTION

The June 16th meeting of the Cincinnati Section was held at the Hotel Alms, D. D. Israel, chairman, presiding.

The paper of the evening on "New Tubes and their Applications" was presented by R. S. Burnap, commercial engineer of the RCA Radiotron Company. The author described a number of the new tubes which have recently been announced, giving their characteristics and indicating the conditions under which the various tubes were designed to be operated. Discussion of the paper was entered into by Messrs. Israel, Kilgour, and Osterbrock.

After the presentation of the paper, some semiportable loudspeaker measuring apparatus was demonstrated.

The meeting was attended by fifty-two members and guests, twenty-four of whom attended the informal dinner which preceded it.

CLEVELAND SECTION

The April meeting of the Cleveland Section, held at the Case School of Applied Science on the 24th, was presided over by Chairman G. G. Hammon.

The paper of the evening on "Porcelain Insulators for Radio Insulation" was presented by A. O. Austin, chief engineer of the Ohio Insulator Company.

The speaker traced the history and development of porcelain insulators and indicated that the insulation of towers for transmitting stations is receiving more attention in modern installations. Experimental work leading to the production of satisfactory insulators for antennas and towers was described and illustrated with lantern slides.

The attendance at the meeting totaled forty-four.

CONNECTICUT VALLEY SECTION

At the May 27th meeting of the Connecticut Valley Section, held in the Hotel Garde, Hartford, Conn., and presided over by Chairman R. S. Kruse, a paper on "The Development of Station WTIC" was presented by J. Clayton Randall, plant manager of that station.

The speaker displayed motion pictures showing the growth of WTIC from the original 500-watt station to the present 50-kw plant. Following the pictures, Mr. Randall discussed some of the maintenance problems involved in the operation of a large boadcast station with particular reference to economic considerations. A general outline of the equipment, both audio and radio, was given and following the meeting an inspection trip was made to the station which is located at Avon, Conn.

This meeting, which was the last of the spring session, was attended by ninety-nine, thirty-two of whom were present at the dinner which preceded it.

A by-law to the section constitution was made setting the last meeting of the calendar year as the annual meeting of the section. Chairman Kruse appointed W. F. Cotter as Chairman of the Meetings and Papers Committee, H. W. Holt as Chairman of the Membership Committee, and J. F. Furey as Chairman of the Publicity Committee.

DETROIT SECTION

The May 22nd meeting of the Detroit Section was held at the Detroit News Auditorium and presided over by Chairman L. N. Holland. A paper on "Considerations in the Design and Operation of Modulators and Modulating Equipment" was presented by Howard S. Stokes of the Airways Division of the Department of Commerce.

The author discussed the theoretical limitations and the design of circuits using audio amplifiers and audio tubes. Commonly accepted proofs were presented and some of the problems were described. Circuits furnishing unusually large audio power output without undue distortion which had been studied under operating conditions were discussed. Oscillograms of typical modulator tubes were also presented to illustrate the results of various malpractices in design and operation of modulators.

A number of the sixty-five members and guests in attendance entered into the discussion of the paper.

A meeting of the Detroit Section was held at the Detroit News Auditorium on June 26, L. N. Holland, chairman, presiding. The paper of the evening on "Talking Movies" was presented by C. L. Stong of Electrical Research Products, Inc.

The speaker gave a brief history of the development of talking movies describing the experiments of de Forest and others. He then described in detail the two general systems used at the present time which employ a sound track on the film and a phonograph record. The inherent advantages of the film recording were pointed out and a prediction made that within five years most of the talking movies would be of this type. The talk was concluded with a résumé of some of the more recent developments in acoustical treatment of theaters.

During the discussion of the paper, which was entered into by a number of the sixty members and guests in attendance, the speaker pointed out that at the present time the phonograph record type of recording was the only feasible method for use in conjunction with a sixteen-millimeter film. He also described some of the methods used in recording at the motion picture studios.

After the close of the meeting some talking motion picture equipment, which had been employed to show a picture before the presentation of the paper, was demonstrated to those interested.

ROCHESTER SECTION

A meeting or the Rochester Section was held on May 28th at the Sagamore Hotel in Rochester, H. E. Gordon, chairman presiding.

The paper of the evening on "Radio Interference" was presented by H. J. Klumb, director of Associated Gas and Electric Laboratories in Rochester.

The author, who has been closely allied with radio investigation work, as carried on by public utilities for the past nine years, had some very interesting data upon which to base his remarks. He outlined some of the causes and gave an analysis of various kinds of interference, suggesting some of the methods employed in their elimination. As a general conclusion he stated that the most satisfactory cure for interference is to increase the power of the radio transmitter to limits controlled only by the pocketbook of the station owner, with the subsequent decrease in sensitivity of broadcast receivers made possible. In this manner the noise level would be below the sensitivity range of the receiver and not audible while the signal strength of the broadcast station would be sufficiently high to be received satisfactorily.

The paper was discussed by Messrs. Estwick, Graham, Gordon, and Karker of the sixty-seven members and guests in attendance.

As this meeting was the annual meeting of the section, election of officers was held with the following results: Chairman, H. A. Brown; Vice Chairman, A. L. Schoen; Secretary-Treasurer, M. A. Wood.

SAN FRANCISCO SECTION

The May meeting of the San Francisco Section held on the 20th at the Bellevue Hotel in San Francisco was presided over by Walter D. Kellogg, chairman.

A paper on "Patents" was presented by Donald K. Lippincott, a patent attorney. The speaker outlined the patent system in effect in the United States, the form of protection given to an inventor and the precautions to be taken by the inventor in order to protect most fully his invention. The patent situation as it affects the radio art was discussed.

A number of the forty members and guests in attendance entered into the discussion of the paper.

The annual meeting of the San Francisco Section was held on June 17 in the Ball Room of the Bellevue Hotel in San Francisco. Walter D. Kellogg, retiring chairman, opened the meeting and in the election which followed the officers for the ensuing year were named. These were: for Chairman, C. H. Suydam; for Vice Chairman, Ralph M. Heintz; and for Secretary-Treasurer, F. E. Terman.

The retiring chairman then thanked the members of all the committees as well as the other officers for their good work during the past year and upon motion a vote of thanks was given to the chairman and the secretary-treasurer for their successful efforts on behalf of the San Francisco Section during the past year. The new chairman was then introduced.

The paper of the evening on "Some Problems of Sound Reproduction" was presented by P. G. Caldwell who based it upon work done at Stanford University. Studies of sound were made with a dynamic speaker, condenser microphone, and associated apparatus. Lantern slides were projected to show interference patterns observed at different frequencies in the measurement of loud speaker response.

A number of the thirty-nine members and guests in attendance entered into the discussion. The attendance at the informal dinner which preceded the meeting totaled nineteen.

SEATTLE SECTION

The May meeting of the Seattle Section was held on the 28th at Guggenheim Hall, University of Washington, under the chairmanship of A. R. Willson.

The paper by T. M. Libby on "Television" covered the accomplishments to date and also a discussion of the problems to be solved in this field. The paper, which was illustrated by projected diagrams called forth a general discussion which covered many phases of the subject and was entered into by Messrs. Hackett, Lovejoy, Miller, and Willson.

E. W. Lovejoy, local supervisor of radio told of his experience in inspecting television transmitting and receiving equipment while on a trip through the East.

Following the discussion, two reels of motion pictures, "The Earth's Four Corners" and "Man-Made Miracles" were projected. These pictures showed the source of material used in the manufacture of radio tubes and also views of the various machines and instruments used in factory production.

The attendance at the meeting totaled seventy-five.

The June meeting of the Seattle Section was held on the 25th at Guggenheim Hall of the University of Washington, Abner R. Willson, chairman, presiding.

The paper of the evening presented by P. J. Hackett of the Universal High Power Telephone Company was on "Talking Pictures."

The author traced the early developments and general history of the talking pictures. He then with the aid of projected illustrations discussed the complex problems in the design, operation, and maintenance of reproducing equipment. The acoustic problems involved and the methods employed in the measurement of absorption and reverberation were discussed and some of the mathematical aspects covered. The auditorium of Guggenheim Hall served as an illustration of some of the acoustic formulas applied to the problems of acoustic design.

Messrs. Eastman, Libby, and Renfro of the fifty members and guests in attendance participated in the discussion which followed the presentation of the paper.

TORONTO SECTION

The May 7th meeting of the Toronto Section was held in the Electrical Building of the University of Toronto, J. M. Leslie, chairman, presiding.

The paper of the evening, which was presented by Eduard Karplus of the General Radio Company, covered the general subject of the communication applications of short waves. Discussion of the paper was entered into by Messrs. Bayly, Oxley, and Professor Price, of the sixty-five members and guests in attendance.

This meeting was also the annual meeting of the section and the following officers were elected for the forthcoming year: F. K. Dalton, Chairman; R. A. Hackbusch, Vice Chairman; and G. E. Pipe, Secretary-Treasurer.

Personal Mention

Rudolph W. Ackerman, formerly chief engineer of Gold Seal Electric Co. Radio Tube Division has joined the radio engineering staff of the Arcturus Radio Tube Co. of Newark, N.J.

Lieutenant William B. Bailey has been assigned to duty on the U.S.S. Pennsylvania.

Lieutenant Herbert C. Behner has left Yale University for duty on board the U.S.S. Wright.

P. K. Beisel, formerly with the New York Telephone Co., has joined the Engineering Research Staff of Metro-Goldwyn-Mayer in Hollywood, Calif.

Previously with the RCA Victor Co., Richard Bell has become a radio engineer for the Philadelphia Storage Battery Co. of Philadelphia.

Willis H. Beltz has become assistant manager of the Installation and Service Department of RCA Photophone, Inc., previously being in the Broad Street office of the Radio Corporation of America.

Previously on the technical staff of the Bell Telephone Laboratories R. V. Cartwright is now located in the Department of Physics of the University of Colorado at Boulder, Colo.

C. C. Chen previously radio engineer and director of the Chinese Admiralty House, has joined the Electrical Engineering Department of the Kiangnan Dock and Engineering Works at Shanghai, China.

Lieutenant W. P. Cogswell has left the Postgraduate School of the U.S. Naval Academy at Annapolis for duty at the Naval Air Station, Anacostia, D.C.

Previously sound manager of Gaumont-British Picture Corporation of London, Ronald Dixon has become assistant department manager of the Phoenix Telephone and Electric Works of London.

E. E. Eldredge, formerly field engineer for Mackay Radio and Telegraph Company is now with Press Wireless Inc. at their Hicksville, L.I., station.

F. Clifford Estey has become sales engineer for the Illinois Zinc Co., previously having been connected with the Aluminum Company of America.

J. R. Harrison, formerly of Wired Radio, Inc., is now at the Electro-Technical Laboratory of Tufts College, Mass.

Previously with the Westinghouse Electric and Manufacturing Co. at Chicopee Falls, R. A. Henderson has joined the radio engineering staff of the De Forest Radio Co.

Lieutenant Commander W. S. Hogg, Jr. has been transferred from the U.S.S. Schenck, to the Bureau of Engineering, Navy Department, Washington, D.C.

Lieutenant T. R. Horn has left Harvard University and is now stationed at Fort Monmouth, N.J.

Charles E. Huffman, formerly with De Forest Radio Co. is now with the American Television Laboratories at Hollywood, Calif.

F. N. Jacob previously with the RCA Victor Company has become an engineer for the Meissner Manufacturing Co. of Chicago.

R. J. Knouf has left the Steinite Manufacturing Co. to join the staff of the Transformer Corporation of America at Chicago.

Lieutenant O. C. Maier, formerly located at Washington is now at the Georgia School of Technology at Atlanta.

Commander B. V. McCandlish has been transferred from the U.S.S. Maryland to the U.S. Naval War College, Newport, R.I.

R. E. Meyers is now electrical officer on the Wilkins Transpolar Expedition.

E. B. Moullin, formerly lecturer at Cambridge University is now a reader in engineering science at Oxford University, Oxford, England.

Lieutenant L. W. Nuesse has left Yale University to board the U.S.S. Dobbin.

Formerly student engineer at RCA Victor, Ralph J. Orner has become development engineer for the De Forest Radio Company.

Lieutenant W. K. Sherman, previously on the U.S.S. Trenton has been transferred to naval inspection work at the General Electric Co. plant at Schenectady.

H. O. Storm, formerly with the Federal Telegraph Co. has become installation engineer for the All America Cables Co. at Granada, Nicaragua.

H. E. Wallace, previously at the Newark Works of the Westinghouse Electric and Manufacturing Co. has joined the radio engineering staff of Wired Radio, Inc.

Paul E. Watson has been made chief engineer of the Radio Section, Signal Corps Laboratories, Fort Monmouth, N.J.

Harry Wilkie, formerly at the Naval Research Laboratory is now a radio engineer in the Signal Corps at Fort Shafter, Honolulu.

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PART II
TECHNICAL PAPERS

Proceedings of the Institute of Radio Engineers
Volume 19, Number 8 *August, 1931*

APPLICATION OF FREQUENCIES ABOVE 30,000 KILOCYCLES TO COMMUNICATION PROBLEMS*

By

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([1] R. C. A. Communications, New York City; [2] Riverhead, L. I., N. Y.; [3] Rocky Point L. I., N. Y.)

Summary—*The authors briefly describe the results of a number of experiments with frequencies above 30,000 kc covering a period of several years. Since the major interest of radio communication companies has been in long-distance communications, this paper includes some qualitative data covering propagation beyond the optical, or direct vision, range. The authors have found that the altitude of the terminal equipment location has a marked effect on the signal intensity, even beyond the optical range.*

Frequencies below about 43,000 kc appear to be reflected back to earth at relatively great distances in the daytime in north-south directions, but east-west transmission over long distances is extremely erratic.

Frequencies above about 43,000 kc do not appear to return to earth beyond the ground wave range, except at rare intervals, and then for only a few seconds or a few minutes. These frequencies, which do not return to earth, also appear to be free of echoes and multiple path transmission effects. Therefore, they are free from distortion due to selective fading and echoes. The range is also limited to the ground wave range, so these frequencies may be duplicated at many points without interference. As the frequency is raised, the range tends to approach the optical distance as a limit. Experiments with frequencies above 300,000 kc have, so far, indicated that the maximum range is limited to the optical distance. A number of possible applications are suggested, based on the unique properties of these frequencies. A specific application to telephony between the Islands of the Hawaiian group is briefly described.

Soon after the commercial possibilities of high frequencies for long-distance communication became known, investigators found that frequencies above 23,000 kc were apparently less useful than the frequencies below 23,000 kc. Undoubtably, this was partly due to the difficulty of producing considerable amounts of power above 23,000 kc with existing equipment. Aside from this, however, it was quite definite that these higher frequencies were relatively limited in their hours of usefulness and were quite sensitive to magnetic disturbances, particularly on long-distance east-west circuits.

However, in spite of these apparent limitations, many investigators in the communication field continued to make studies of the propagation possibilities of these higher frequencies.

It is the purpose of this paper to describe briefly, a series of investigations made on these ultra-high frequencies by the engineers of the Radio Corporation of America and R.C.A. Communications, Inc.

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* Decimal classification: R423.5 Original manuscript received by the Institute, April 24, 1931. Presented before Boston Section of the Institute, May 15, 1931.

HISTORICAL

Since the major interest of radio communication companies up to the present time, has been in connection with long-distance communication, our early efforts were concentrated on the long-distance propagation possibilities of these ultra-high frequencies. Consequently, this paper includes studies of transmission beyond the optical range, as well as within the optical range.

Several years ago, a simple regenerative receiver was developed which was sensitive enough to pick up 60,000-kc harmonics at Riverhead, L.I., from short-wave transmitters at Rocky Point, L.I., a distance of about fifteen miles. A transmitter using two 150-watt tubes was then set up at Bush Terminal in Brooklyn, N.Y. It was estimated that this transmitter developed about 80 to 100 watts in the antenna at a frequency of 55,500 kc. The antenna was located on one of the towers of the marine station "WNY" as high above all obstructions as possible.

The simple regenerative receiver was placed aboard a ship running from New York to San Juan, Porto Rico. W. I. Matthews was assigned to this ship for making observations.

The signals from this transmitter were strong with the ship at her dock. As the ship sailed the signals gradually became weaker, finally disappearing at a distance of about 40 miles. The Bush Terminal signal was not heard again during the entire round trip. However, on the trip north, harmonics of the Rocky Point transmitters WLL and WTT were heard for a short time at a distance of nearly 900 miles. These signals were between 50,000 kc and 60,000 kc and were, apparently, the third harmonics of the above-mentioned transmitters. They were heard for only a short time and were not heard at lesser distances during the voyage.

The results of the ship tests were encouraging, so it was decided to make further investigations with more powerful transmitters and a more sensitive receiver. A receiver was built having two stages of high-frequency amplification, using special UX-222 tubes. This receiver was mounted in a Ford sedan. A water-cooled tube was used at the transmitter and was adjusted to take an input of about 7 kw at all frequencies between 50,000 and 30,000 kc. The average power in the antenna was probably around two or three kilowatts, being rather lower on the longer waves due to losses in the chokes. This transmitter was located in a special building at Rocky Point, and was used with a vertical doublet, a horizontal doublet and a beam antenna, the latter directed towards Havana.

With this improved receiver mounted in the Ford car, tests were first made around Long Island listening to Rocky Point transmitting on 50,000 kc. The signals were heard at Montauk Point, a distance of about 60 miles. It was found that the signals were weak with the receiver on low ground, but were strong with the receiver on hilltops. This demonstrated that the 50,000-kc signals could be heard beyond the distance of direct vision, and that elevation was apparently important.

After completing these preliminary tests, Matthews started south with the test car, listening to Rocky Point transmitting on 50,000 kc with the water-cooled tube and radiating on the vertical doublet. This signal was heard well at Atlantic Highlands, N.J., 265 feet above sea level, but it was very weak at sea level. The distance was about 67 miles, indicating that the normal level ground range would be around 40 miles. The signal was also heard at Shark River Hills, near Belmar, N.J., a distance of about 90 miles. The signal was fairly strong there, but fading was quite violent. A conservative rating for good reception would be about 80 miles, with the receiver on a moderately high hill.

From Belmar, Matthews went on to Ocean City, N.J., then Cape May, N.J., then Ocean City, Md. Nothing was heard for days, excepting on one occasion at Ocean City, Md. This occasion is very well summed up in Matthew's log, which reads as follows:

"11:52 A.M.—As I had been listening for the last two days and nothing was heard, and this morning tuning continuously since 9:30 o'clock in the vicinity of 2XT, I set the dials on his wave and left them there. While I was thinking about what moves I should make, and where to try next, the signal suddenly burst in. It did not drift in, the intensity coming up slowly, but came in R8 and stayed in for about 30 seconds, the same strength. It went out the same as it appeared, suddenly. I was so surprised that I did not touch the dials for at least five minutes. Then I tuned around for the signal, thinking it might have drifted. Only once did I hear it, at 12:01, a very faint dash was heard."

The distance from Rocky Point to Ocean City is about 238 miles and the transmitter was on the vertical doublet at the time. Matthews requested that the transmitter be put on the beam antenna at alternate hours on the following day. He listened all day, but heard absolutely nothing.

From this point he moved on to Virginia Beach, Va., Cape Henry, Va., Kitty Hawk, N.C., Sumpter, S.C., Titusville, Fla., and finally Palm Beach, Fla. Nothing was heard at any of these places, although very careful observations were made. This group of tests was finally terminated at Palm Beach, Florida.

It appeared quite conclusive that it was impossible to duplicate the results on the ship when the signal was heard at a distance of over 800 miles. The experience at Ocean City, Md., was a very strong indication that all of the long-distance signals on 50,000 and 60,000 kc were caused by some temporary abnormal condition. Certainly the long-distance signals were not normal phenomena for these frequencies.

In the meantime the old transmitter at Bush Terminal was removed and placed on an Eagle boat at New London for tests. A receiver was placed on a similar boat. It is estimated that the transmitter put about 100 watts into the antenna at 55,000 kc. G. S. Wickizer, who made the tests, reported that the range was about 12 miles with this equipment. He also noted certain directional effects, that is, it was easier to transmit and receive towards the bow of the boat, due to less shielding from the superstructure. Considerable trouble was experienced due to vibration at both receiver and transmitter. Doubtess, the range could be increased somewhat with higher antennas, and equipment designed to withstand vibration.

The second group of tests observed by Matthews was started from Riverhead in the Ford car, shortly after his return from the Florida tests. The Rocky Point transmitter "2XT" was tuned to 42,000 kc, using a horizontal doublet. The signal was heard at Atlantic Highlands, N.J., but apparently no better than the 50,000-kc signal. The signal was better further on at Shark River Hills. The signal decreased towards sea level just as in the case of 50,000 kc, but not as markedly. At Cape May Lighthouse, 184 miles from Rocky Point, the 42,000-kc signal was fairly steady, using an antenna on a 50-foot tower. The signal was also heard at Ocean City, Md., 238 miles away, and was last heard at Kitty Hawk, N.C., 390 miles away. At Kitty Hawk, the signal was heard for only a few seconds at a time and was entirely unreliable. At Ocean City the signal was also rather unreliable due to violent fading, but it was better at Bethany Beach, Del., a few miles north, where a 35-foot pole was available for the antenna. The transmitter used a horizontal doublet during all of these tests.

While Matthews was at Kitty Hawk, the Rocky Point transmitter was tuned to 36,600 kc. The signal was received very poorly, but considerably better than the 42,000-kc signal.

The 36,600-kc signal was not heard beyond Kitty Hawk, indicating that under the existing conditions, 390 miles was near the limit of the ground wave range for this frequency and was at the very outer fringe of the ground wave range for 42,000 kc. It was evident that the ground wave range on 36,600 kc and 42,000 kc was greatly in excess of the ground wave range on 50,000 and 60,000 kc.

While at Kitty Hawk, signals were picked up from a harmonic of "XDA" Mexico City. This harmonic was very close to 2XT on 36,600 kc, and was much stronger than 2XT, although the distance to Mexico City was around 1600 miles.

Proceeding further south, listening tests were made at several points, but the 36,600-kc signal from Rocky Point was not heard again.

While listening at Brunswick, Ga., Matthews heard harmonics of several stations, these harmonics ranging from 33,000 to 43,000 kc. Among the stations heard were LP1, LP2, LP3 (Buenos Aires), and KMM (Bolinas, California). These signals were heard with excellent intensity at Wilmington, N.C., Kitty Hawk, N.C., Cape Charles, Va., and Asbury Park, N.J. Some other harmonics from stations to the south were heard, but no stations to the east were intercepted. In all cases, these long-distance signals were heard only during all daylight conditions.

Since the ground wave range of the ultra-high frequencies was considerably greater with the receiver located on hill tops, it was considered highly desirable to make some tests with the receiver located at a considerable altitude.

Accordingly, the Rocky Point transmitter was again tuned to 50,000 kc, radiating on a horizontal doublet. The receiver was installed on a dirigible using a small horizontal doublet for the antenna. With Matthews aboard as observer, the ship was first flown toward Rocky Point until the 50,000-kc signals were picked up. Then the ship was navigated at different altitudes up to 2500 feet, and at different distances up to 150 miles from Rocky Point. At this distance on the ground, or on a hilltop, no signals had ever been heard on 50,000 kc, aside from the single occasion of a few seconds at Ocean City, Md. On the dirigible, the effect of altitude was found to be very marked and, at a distance of 150 miles, the signal could not be heard below an altitude of 1000 feet, but was strong at 2000 feet.

The receiving antenna was found to be quite directive, and was also found to be shielded by the metal framework of the dirigible. The maximum signal was received when the horizontal doublet was broadside to Rocky Point with the ship turned so that the doublet was on the side of the ship nearest the transmitter. This directive effect rather complicated the analysis of the results, yet the effects noted were qualitatively very definite. There was no question about the marked increase in range with the receiver at a high elevation.

For the next group of tests, the Rocky Point transmitter was tuned to 33,000 kc. Matthews started towards Montreal with the receiver in

Fig. 2, is an attempt to show qualitatively, the transmission characteristics of 37,500 kc and 43,000 kc. The reliable gound wave range lies above "R5," so the "R5" range checks approximately with the ground wave range indicated in Fig. 1. For example, the 37,500 kc curve shows that the signals were fairly reliable at 150 miles, but were heard as far away as 400 miles, although they were extremely erratic at the greater distance, as indicated by the tests at Kitty Hawk, N.C. Beyond the 400-mile distance, the signals appear to "skip" up to about 1600 miles, at which point the signals begin to come in again during "all daylight" conditions. The signals become fairly good again in the range from 2500 to 5000 miles, as shown by Fig. 2. This sky wave curve is made up of all the harmonics observed between 33,000 kc and 43,000 kc. It should be noted that these harmonics were received mainly from the south and west under all daylight conditions, whereas, the ground wave range is practically independent of time of day.

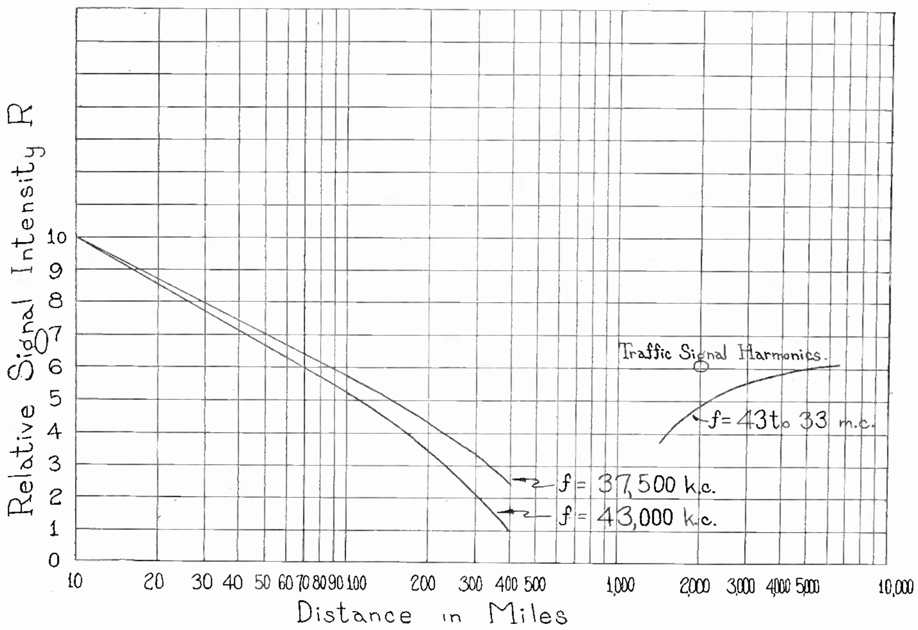

Fig. 2—Curves showing variation of signal intensity with distance.

Some of the long-distance signals below 43,000 kc were received well enough to indicate a possible value for long-distance communication purposes, although these frequencies did not appear to be nearly as reliable as the lower frequencies, and also appeared sensitive to magnetic disturbance. It was noted that certain transmitters appear to radiate strongly on these high frequencies, while similar transmitters near the same frequency at the same location, cannot be heard at all.

Fig. 3, is an attempt to show the effect of the altitude of the receiver on the reception of the ground wave on 50,000 kc. The data for these curves were taken partly from the dirigible observations, and partly from other observations. It is evident that the reliable ground wave range of 50,000 kc might be extended from 30 miles to 150 miles, or more, if the receiver or transmitter, or both, can be suitably located at high elevations. At a distance of 100 miles, with the transmitter at sea level, the receiver would have to be located at an elevation of about 6700 feet to obtain optical vision to the transmitting point. Since the signals on the dirigible came in strong at 2000 feet, it is evident that there must have been some bending of the wave at 50,000 kc.

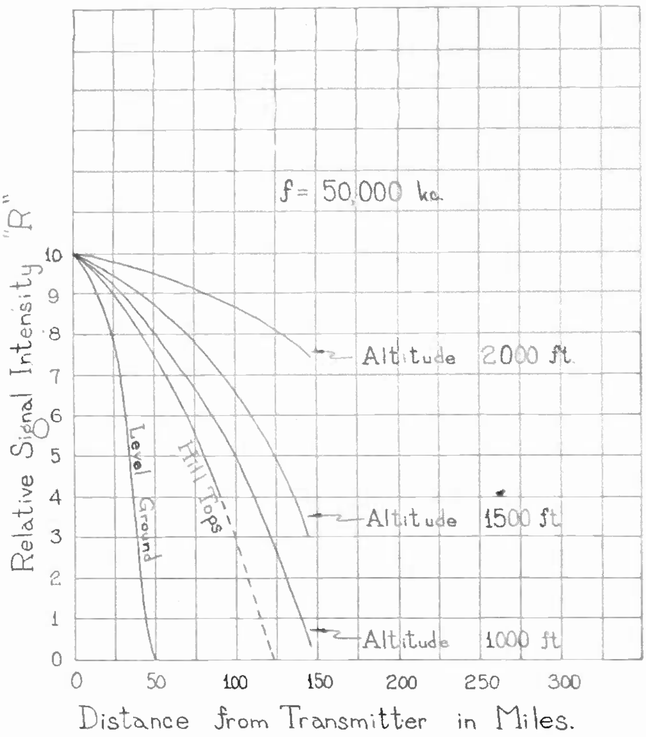

Fig. 3—Curves showing variation of signal intensity with distance and altitude.

HAWAIIAN ISLAND TESTS

It seemed obvious from the results obtained from the long-distance tests, that circuits of commercial quality should be quite practicable on these ultra-high frequencies if one or both of the terminals were sufficiently elevated. A proposition where our conclusions might be put to a practical test came to our attention while we were conducting these investigations. The Mutual Telephone Company of Honolulu, had for some years, been studying the possibilities of interconnecting the various Islands of the Hawaiian group for telephone service. Fig. 4 is a map of the Hawaiian Islands showing the telephone circuits desired. The rough seas and depth of water between these Islands made the use of submarine cable impracticable. Tests between the Islands with radiotelephony on frequencies between 1500 kc and 5000 kc were none too encouraging. Selective fading and strong atmospherics were experienced on those frequencies. To overcome these obstacles would call for an investment unwarranted by the probable earning power of the circuits. The ultra-high-frequency spectrum seemed to offer a possible solution.

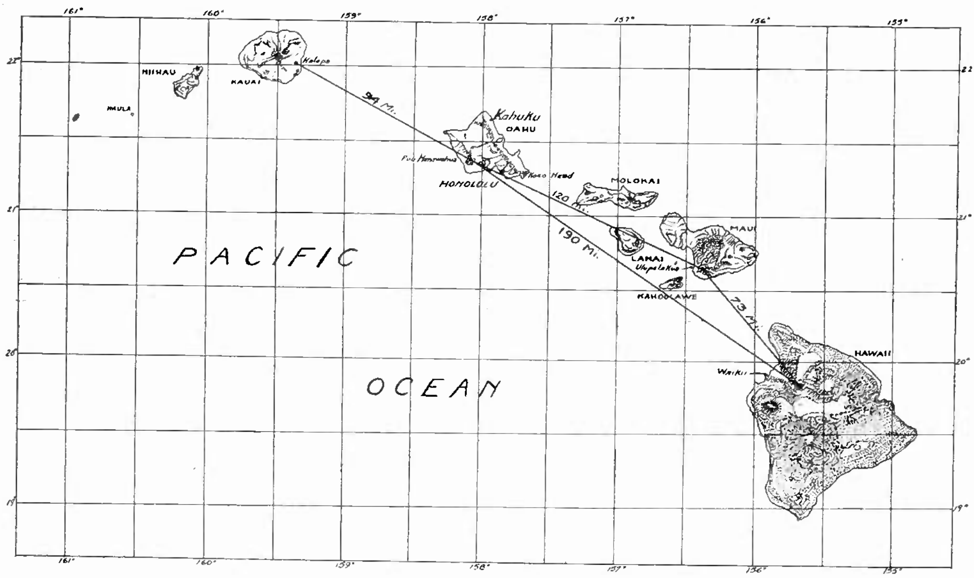

Fig. 4—Map of the Hawaiian Islands showing proposed radio links for interconnecting the telephone systems on the major islands of the group.

W. I. Harrington of the Mutual Telephone Company, had conducted a series of tests on 60,000 kc, obtaining a range of up to 80 miles. The transmitter was situated at an elevation of about 100 feet in a suburb of Honolulu. Receiving tests were made at various locations in the vicinity of Koko Head, Oahu, and on the Islands of Molokai and Maui, at elevations up to 500 feet. These tests gave consistent results and indicated that better results might be expected if sufficient altitude could be obtained at the terminals. It was decided that a more extensive study should be made, and J. A. Balch, President of the Mutual Telephone Company, arranged for the coöperation of the Radio Corporation of America.

Accordingly, W. I. Matthews and S. H. Fifield, were sent to Honolulu with two complete transmitters and receivers, specially prepared to make the desired telephone tests. The transmitters were of the modulated power oscillator type, with long line frequency control.[1] The oscillator consisted of two UX-852 tubes operated push-pull. An output of about 75 watts could be obtained at frequencies between 37,000 kc and 60,000 kc. Fig. 5 shows one of the transmitters used. The plate supply rectifier and the modulator system were built into the unit, making the apparatus convenient for survey work.

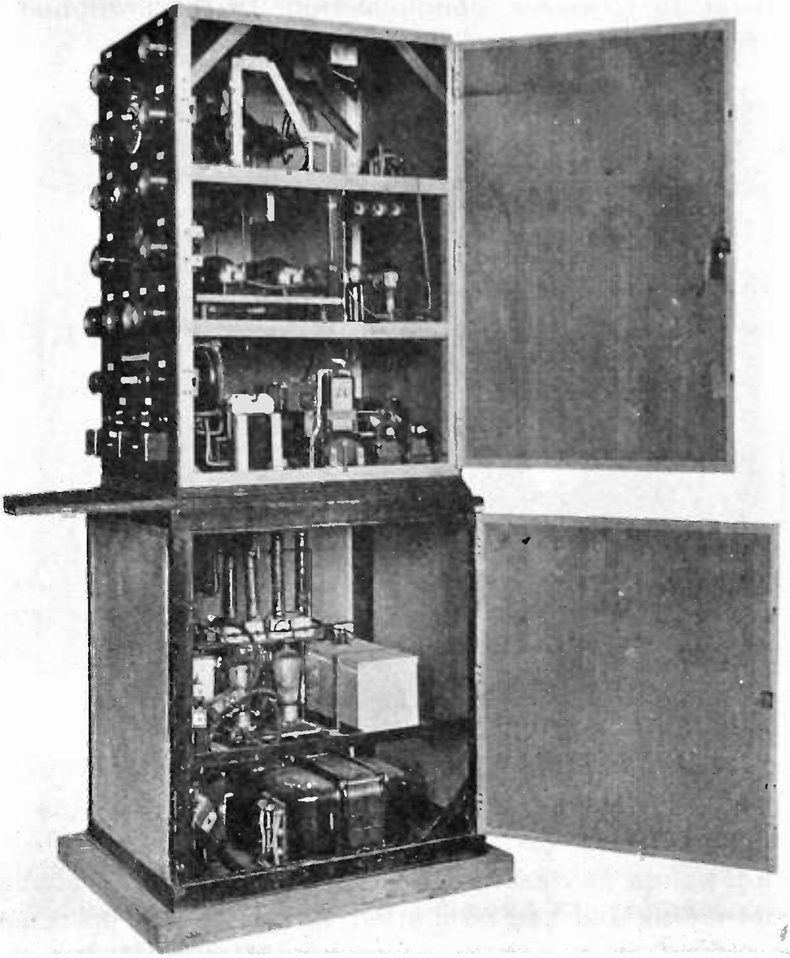

Fig. 5

The receivers were of the superheterodyne type. Fig. 6, shows one of the sets as used on this project. The first detector was preceded by two stages of tuned r-f amplification using a special type of screen-grid tube. A tuning range of from 30,000 to 65,000 kc was provided for. A standard AR-1286 aircraft beacon receiver was used as intermediate-

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[1] Conklin, Finch, and Hansell, "New methods of control employing long lines," to be published in a forthcoming issue of the PROC. I. R. E.

frequency amplifier and second detector. Any intermediate frequency between 250 and 500 kc could be used. As a rule, something on the order of 500 kc was used. Quite satisfactory performance was obtained with this combination of transmitting and receiving components.

One transmitter was set up on the Island of Oahu at a point known as Puu Manawahua. This site was at an altitude of 1700 feet and commanded unobstructed lines of sight to the other Islands with which it was desired to establish communication. In the beginning a simple doublet was used at the transmitter.

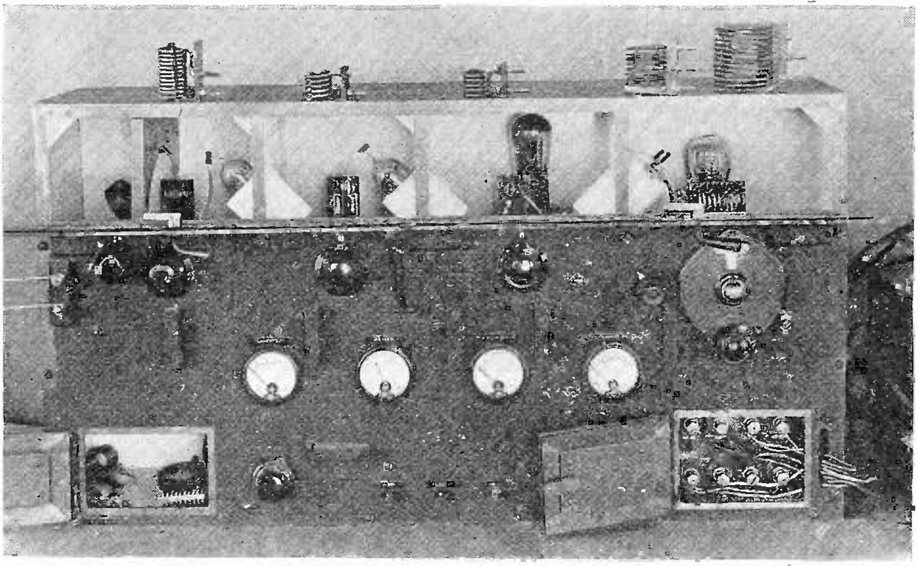

Fig. 6

The receiver was first taken to the Island of Maui, about 100 miles away. Reception tests were made at two different sites, at elevations of 1000 feet and 2700 feet, respectively. The telephone transmissions from Puu Manawahua were heard perfectly at both locations. The signal was free from fading and had no diurnal variations. The operation of this circuit was so easy that the receiving expedition decided to proceed at once to the Island of Hawaii, about 190 miles away, where more difficulty might be expected because of the greater distance involved.

The initial receiving tests on Hawaii, made at an elevation of 6200 feet on the slope of Mauna Loa, were successful. An intelligibility of 85 per cent was obtained. A carrier frequency of 37,500 kc was used during this period. A site at Waikii at 4800 feet elevation was later selected for more extensive tests. From a consideration of the distances of direct visibility for the elevations of 4800 feet at the receiver and 1700 feet at the transmitter, it was calculated that the range of the signal must have been greater than the range of vision by a distance of 56 miles out of the total of 190 miles traveled.

Successful transmission was also obtained over this circuit on 36,600 kc and on 42,600 kc. However, the signal strength on the latter frequency was appreciably weaker.

A series of polarization tests were conducted on 40,000 kc. A vertical and a horizontal doublet were made available at the transmitter on Oahu and schedules of alternately vertical and horizontal radiations were arranged. The receiving site likewise had both horizontal and vertical doublets for observational purposes. It was definitely determined that a signal radiated by a horizontal doublet was received best on a horizontal doublet and that a signal radiated by a vertical antenna was received best on a vertical antenna. This relationship was most fixed on the shorter circuit between Oahu and a location on Maui, a distance of 127 miles. At Waikii, on the Island of Hawaii, there was evidence that the polarization of the Oahu signal had a tendency to become twisted at times. This tendency appeared in the form of fading. The periodicity of this fading was very slow and the resulting signal variations were easily eliminated with automatic volume control applied at the receiver. It was finally determined that vertical polarization was most adaptable at these locations. This decision was influenced to quite an

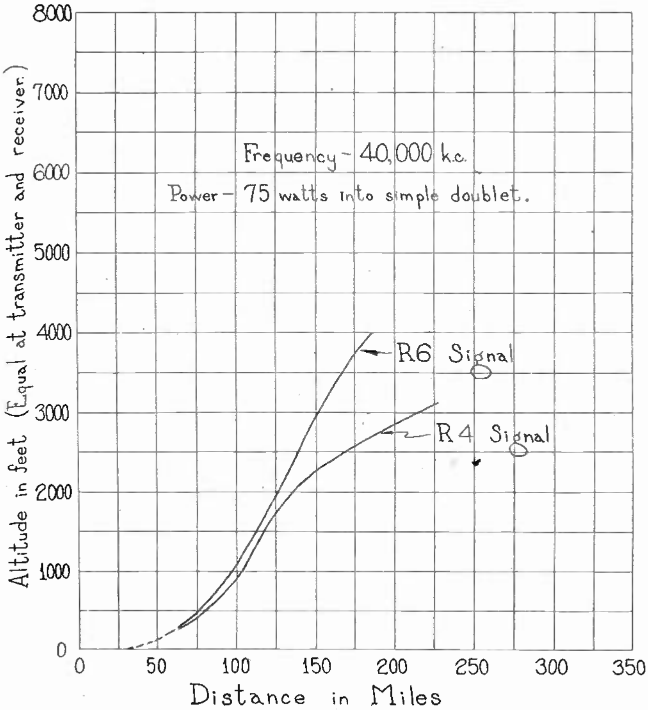

Fig. 7—Curves showing variation of signal intensity with altitude and distance.

extent by the sloping nature of the terrain. The R.C.A. Model "B" directive antenna which radiates vertically polarized waves, had been chosen as most suitable for this work. This antenna consists of a number of wires pointed in the direction of the circuit and inclined to the horizontal at an angle of about 17 degrees.[2]

The directive antennas were next erected to establish a two-way test circuit between Puu Manawahua (Oahu) and Waikii (Hawaii). These antennas resulted in a very decided improvement in the circuit. It was determined that conversation could be held over the circuit with as little as 12 watts input to the transmitter anodes, when using these antennas. An extended period of two-way testing demonstrated the commercial feasibility of this circuit.

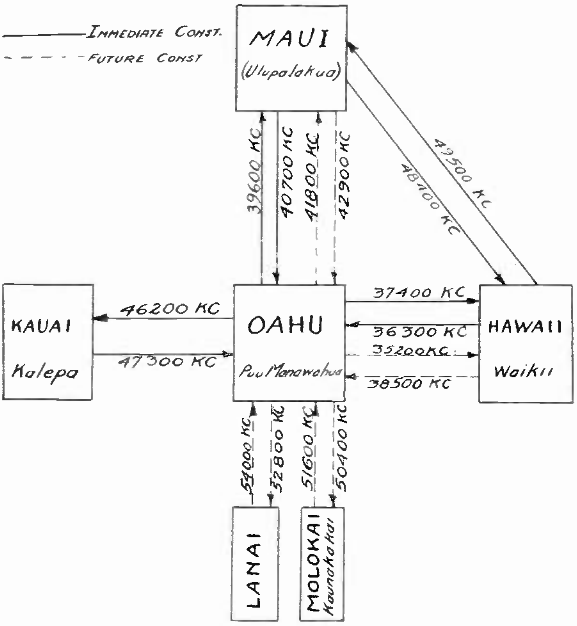

Fig. 8—Frequency assignments for Hawaiian interisland telephone system.

Receiving tests (40,000 kc) were later made on the Island of Kauai, 91 miles distant from Puu Manawahua. Strong signals from Oahu were received at an elevation of 700 feet and a usable signal level was observed at sea level. While at Kauai, it was found possible to receive at sea level, signals from Waikii (Hawaii), 280 miles distant. This unprecedented range of the ground wave was probably due to the use of directive antennas and to the favorable conditions of over-water transmission.

Fig. 7 shows some relationships between altitude and distance derived from these experiments in the Islands. These curves are plotted

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[2] P. S. Carter, C. W. Hansell, and N. Lindenblad, "Development of directive transmitting antennas by R. C. A. Communications, Inc.," to be published in a forthcoming issue of the PROC. I. R. E.

on the assumption that transmitter and receiver are both at equal elevation. The antenna input is 75 watts and the frequency 40,000 kc. The R4 signal indicated for simple doublets, would make a commercial circuit possible with directive antennas. The change in slope of the curve suggests that for weaker signal strengths, the signal would be heard at great distances with moderate elevations. The experience of hearing Hawaii at Kauai supports this conclusion.

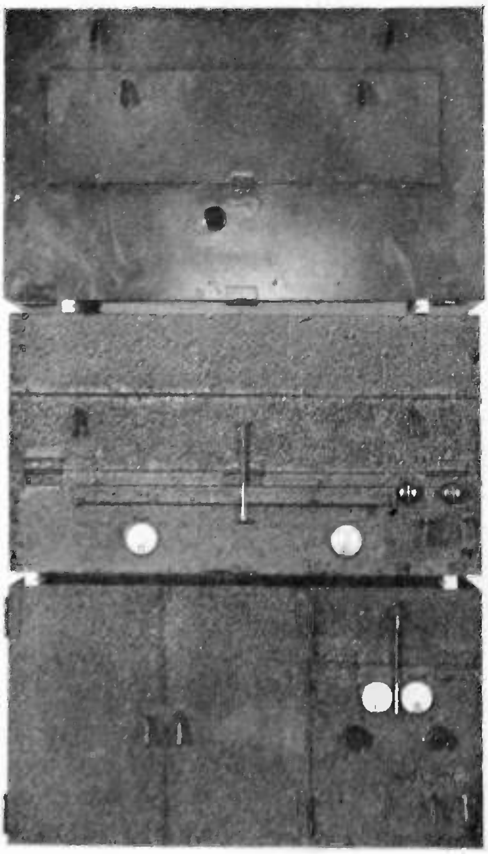

Fig. 9

The extended tests indicated a remarkable freedom from atmospherics and selective fading. Conditions of cloudy weather had no effect on the circuits and no differences were observed between daylight and darkness. On the strength of these findings, a telephone network using these frequencies has been planned for the Islands and the facilities are in the process of preparation. Fig. 8 shows a diagram of the channels contemplated. These circuits together with the facilities now being installed by R.C.A. Communications in Hawaii to connect the Mutual Telephone Company with the telephone system in San Francisco, will soon link the important Islands of Hawaii, with many countries through the world-wide telephone system of the American Telephone and Telegraph Company.

To reduce operating costs to a minimum, receiving and transmitting apparatus will be installed in the same building. The apparatus is of such design as to require a minimum of attention. Fig. 9 shows the type of receiver that will be used on these circuits. The upper box contains the high-frequency unit. To prevent drift of oscillator frequency with change of temperature, the entire unit is enclosed in a temperature controlled, heat insulated box. Fig. 10 shows this unit with doors open. It is expected to have these circuits in commerical operation before the end of 1931.

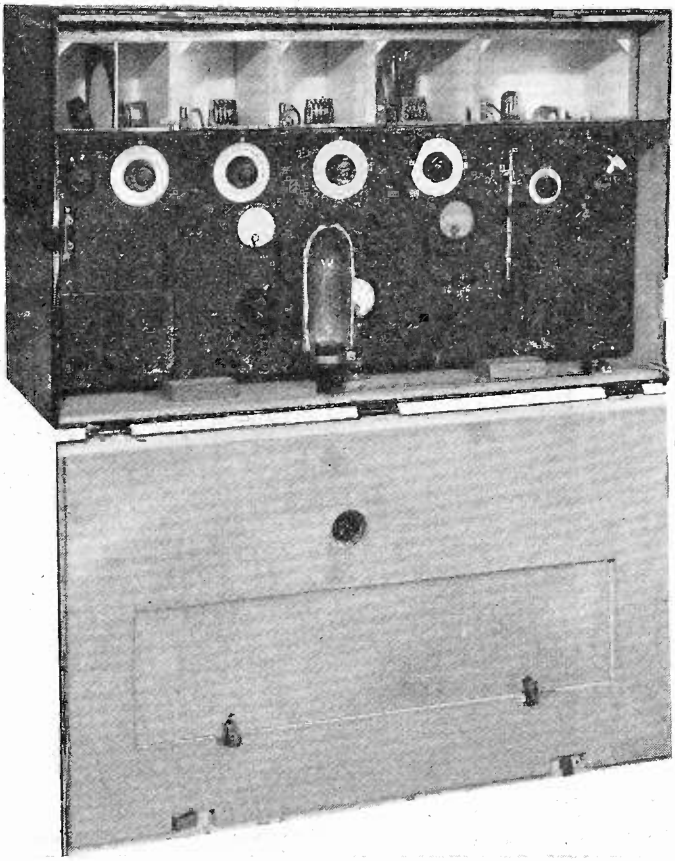

Fig. 10

TESTS WITH AIRCRAFT

Because of the advantages of altitude already indicated, it was thought probable that these frequencies might be of some service in aircraft radio applications. A UX-210 oscillator delivering about one watt of energy to a horizontal doublet was mounted in an airplane and some range tests were conducted, transmitting to a fixed receiving site situated at Riverhead, L.I., about 80 feet above sea level. The plane was heard at distances up to about 100 miles during these experiments. These tests were conducted on a frequency of 60,000 kc. The same apparatus gave a maximum range of only three miles over flat country with the transmitter located in an automobile.

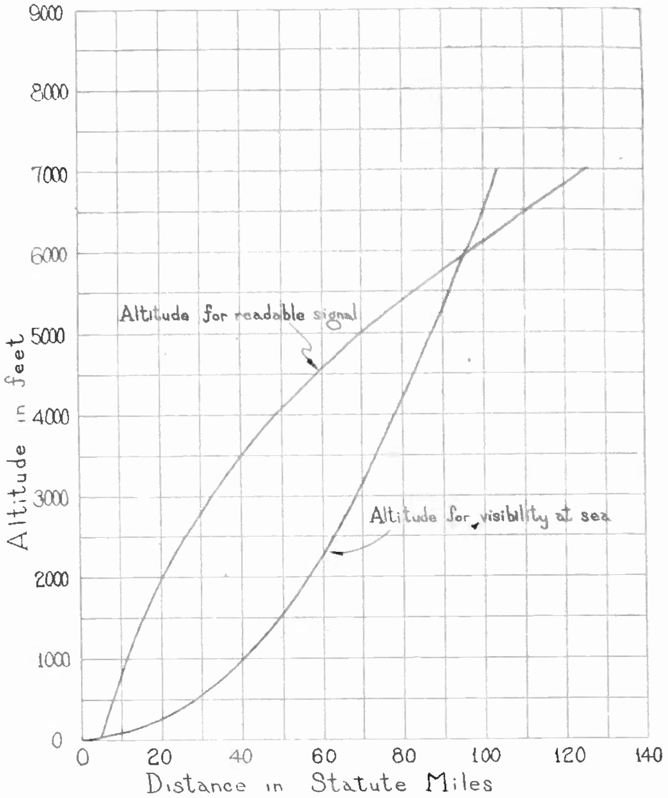

Fig. 11—Curves showing height required at a given distance for communication from ground to plane on 47,000 kc.

It was observed that at all times, the plane had to reach sufficient altitude to command an unobstructed path to the receiving site. It was not possible to communicate over circuits requiring the signal to follow the curvature of the earth part of the way. This reluctance of the signal to follow the curvature of the earth might have been partly due to the higher frequency used (60,000 kc). It might also have been partly due to the small amount of power.

At a later date, range tests were conducted with a receiver mounted in a plane and a transmitter situated near the west side of New York City about 50 feet above sea level. The transmitter was a crystal controlled telegraph set delivering about 100 watts to the antenna at a frequency of 47,000 kc. The receiver was a superheterodyne of the same type as used on the Hawaiian Islands survey. A vertical antenna was used at the transmitter and a trailing wire was used for receiving on the plane. This trailing wire was reeled out to a length equal to four wavelengths. Since the tests were conducted with the plane flying away from the transmitter, the directive properties of this sloping wire trailing out beneath the airplane were correct for best reception.

The data obtained in these tests from ground to plane are summarized in the curve Fig. 11. The curve showing the altitudes required for direct line visibility of various distances is shown for comparison. It should be noted that the altitude required was, in general, greater than that required for direct visibility. This checks with the results previously discussed for the case of transmission from plane to ground. One explanation for this high altitude requirement, especially for the distances within 50 miles of the transmitter, might be the fact that the signal had to travel over New York City. Since the transmitter was situated at a comparatively low altitude, the signal was probably seriously obstructed at the start. Another handicap in these tests was the ignition noise level on the plane. The ignition system was completely shielded in the conventional manner found effective on lower frequencies, but considerable noise level was experienced on 47,000 kc. A special type of shielding might be necessary for the ultra-high frequencies.

It would seem that these high frequencies have properties that will make them useful for certain types of aircraft work. The small size of the antenna systems required and the possibilities for reducing weight of equipment are desirable features for this class of service.

DIRECTION FINDING

During some of the work between plane and ground, a horizontal doublet which could be rotated about a vertical axis, was available. This antenna was found to be a fairly effective direction finder. The signal strength was zero or minimum when the doublet was oriented so as to point at the transmitter. The horizontal doublet on the plane was likewise found to be quite directive. Rotating directive beacons using these ultra-high frequencies might be quite useful in the navigation of aircraft in thick weather.

A possible application for high-frequency direction finding as an aid to navigation, presented itself in the form of the New York Harbor ferryboat situation. These boats make scheduled trips back and forth across the river in all kinds of weather. It is sometimes so foggy that visibility is limited to a few hundred feet or less. If, in making a crossing, a detour has to be made to avoid another vessel, it sometimes becomes quite difficult to find the slip on the other side. Radio direction finding on the usual frequencies would be objectionable because of the interference it might cause to other services. The comparatively great number of channels available in the ultra-high frequency part of the spectrum, encouraged a consideration of using 60,000 kc for this work.

Accordingly, a series of experiments were conducted on one of the New York Central Railroad Company ferry boats operating between Weehawken, N.J., and Manhatten Island. The Hudson River is about a mile wide at this point. The 60,000-kc transmitter was mounted in the loft of the Weehawken ferry slip. The antenna was well above the roof at an altitude of about 100 feet above the river. The carrier was modulated with a characteristic tone and the set left running continuously.

A simple tuned r-f receiver was used on the ferry. A horizontal doublet was mounted at the upper end of a brass tube extending up through the roof of the wheel house. This doublet could thereby be rotated about the vertical axis. The energy from the doublet was carried down inside the vertical tube on a two-wire transmission line. The lower end of the tube fitted into the top of the receiver, completing the shielding.

Quite sharp bearings were obtained and it was demonstrated that it should be quite practicable to find the slip in foggy weather by these means. It also became apparent that the same facilities might be used to advantage as a medium of communication between the central office and the various boats en route. Such facilities should be very useful in certain emergencies.

It was observed that the signal intensity varied up and down at regular intervals as the ferry moved across the river. This condition was most likely due to an interference pattern resulting from the combination of the energy coming drectly from the transmitting antenna with that which was reflected from the water somewhere between the transmitter and receiver.

A few tests were conducted to determine the shielding effect of city buildings. A receiver was mounted in a test car and continuous observations of the Weehawken signal were conducted while driving through various streets on the Manhattan side of the river. It was found that the signal could be heard in streets four or five city blocks back from the water front. Driving along a street parallel to, and several blocks back from, the river, it was found that the signal strength increased greatly whenever one of the streets perpendicular to the river was crossed. This suggests that buildings may serve as fairly effective reflectors. Thus it might be possible to obtain quite effective broadcast service to all parts of a city if the transmitter is situated at the top of one of the tallest buildings. Due to the obstructions in city areas, however, the service range will be much less than is indicated in Fig. 1, particularly for high-grade entertainment broadcasting and television.

CONCLUSIONS

This paper has narrated in a more or less chronological order, a number of experiences in the operation of ultra-high-frequency radio equipment. Certain definite propagational properties have been indicated, some of them semiquantitatively. It is felt that more research should be carried on along similar lines to establish more quantitative relationships that might be used in the engineering of circuits. It might be a fitting conclusion to this paper, to enumerate some of the uses for which this part of the spectrum might be especially well suited.

As might be concluded from the work that has been described, frequencies above 45,000 kc do not seem to be reflected back to earth by the Kenelly-Heaviside layer, and, furthermore, as the frequency is increased, the maximum range tends to approach the optical range as a limit. This limitation should in many cases, prove an advantage. It should eliminate the selective fading effects and frequency distortion now so troublesome on the lower frequencies. It should furthermore, be possible to use the same frequency channels over and over at geographically separated points on the earth. It is suggested that the ultra-high frequencies hold forth promise for many applications, a few of which are enumerated herewith:

(1) Point-to-point communication up to 300 miles between mountains.

(2) Ground-to-aircraft communication up to at least 100 miles and communication between aircraft.

(3) Point-to-point communication between high buildings or towers up to 50 miles or more.

(4) City police alarm distribution up to a few miles. Portable receivers may be carried by scout cars.

(5) Possible application to high speed visual image distribution over local areas.

(6) Local audio, facsimile, or ticker distribution.

(7) Communication and direction finding for ferryboats, tugs, and harbor craft.

(8) Marker beacons for air and water craft.

Some of these short distance services have been suggested by other investigators, particularly by engineers of C. Lorenz-Aktiengesellschaft of Germany,[3] who have carried out some very interesting studies, especially in the optical range. R.C.A. Communications' engineers have also made some experiments using frequencies above 300,000 kc over distances up to the limits of the optical range. They have had a two-way circuit in experimental operation on 460,000 kc (65 centimeters) for an extended period. It is hoped that this work may be published in the I.R.E. PROCEEDINGS at some future date. Their experiences with frequencies above 300,000 kc check very well with the results reported by Eduard Karplus in a paper presented before the Institute of Radio Engineers in New York on April 1, 1931.[4] Similar results have recently been reported in the press by the International Telephone and Telegraph Company, and their associates, in connection with a demonstration of telephony and facsimile across the English Channel, using frequencies above 300,000 kc.

ACKNOWLEDGMENT

The experiments between the Islands of the Hawaiian group were carried out jointly by the Mutual Telephone Company of Honolulu and the Radio Corporation of America. Without the coöperation of the Mutual Telephone Company in providing sites, building antennas, telephone lines, power lines, and active participation in the tests, these extensive experiments could not have been undertaken. The active interest and coöperation of J. A. Balch and W. I. Harrington of the Mutual Telephone Company, were particularly helpful and encouraging.

The especially fine work of W. I. Matthews in making a great number of the receiving observations described should be acknowledged. The coöperation of F. H. Kroger, S. H. Fifield, G. S. Wickizer, and D. R. Goddard in various phases of the work, as well as the active support of C. H. Taylor, is also acknowledged.

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[3] Esau and Hahnemann, "Report on experiments with electric waves of about 3 meters," PROC. I. R. E., 19, 471; March, 1930.

[4] Eduard Karplus, "Communication with quasi optical waves," delivered before the I. R. E. in New York, April 1, 1931.

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AUTOMATIC COLOR ORGAN*

BY

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Summary—*The automatic color organ, a by-product of radio, produces colors by means of music and synchronizes colors with music. Acoustic power on the order of microwatts controls lighting power of hundreds to millions of watts, which is varied in accordance with rapid fluctuations of the input.*

INTRODUCTION

WHILE the primary purpose of this paper is to describe an instrument which automatically produces colors by means of music and synchronizes colors with music, there are a number of important factors to be considered as far as its entertainment value is concerned.

The device itself makes commercially possible the control of an ouput ranging from hundreds to millions of watts, by a few microwatts of acoustic power from music or voice. With a lighting load, it then becomes what is termed a color organ and thus enters the field of color music.

Color music, as an art, has been under consideration for several centuries, and it is only natural that experimenters have been attracted by its vast possibilities. Many have treated color as an art totally divorced from music by appealing to the sense of sight by colors in the same manner as music to the sense of hearing. Others have endeavored to correlate color and music, thereby simultaneously appealing to both senses with consequent greater emotional and æsthetic values.

MUSIC AS AN ART

To understand the problem more definitely, a brief résumé of musical history may be of interest.

Occidental music is traced from the Assyrians and Egyptians by the wall paintings and bas reliefs, which have been preserved to us. Egypt had a definite musical science before 3000 B. C., and during the Golden Age music was not only employed as a social diversion, but also for religious services. The Hebrews and Greeks obtained their knowledge from the Egyptians, and our present day music, as distinct from the tom tom drummings of the savages, may be traced from them.

It is interesting to note that music, while related to the development of civilization, is always the last art to be developed seriously.

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Painting and architecture, followed by literature and drama, come first, and music is last. Yet with the primitive man, music is first.[1]

By referring to Helmholtz[2] the answer seems plain. He states music was forced to select artistically and then shape for itself the material with which it works. Painting and sculpture find their materials in nature itself. Likewise poetry finds its material in words of language. While architecture must create its forms, these are forced upon it by technical and not purely artistical considerations. Music, on the other hand, has a greater and more absolute freedom in the use of material than any of the other arts. Hence without any guides or external landmarks a difficult task is faced, which must be undertaken slowly. If this is true for music, how much greater are the obstacles in the way of obtaining color music. Surely freedom is available in this field.

SIGHT AND HEARING

The senses of sight and hearing are superior to our other senses. They are removed from the necessary life functions. Colors and tones are rather semblances or distant signs of material realities than the realities themselves. Therefore, the senses are particularly fitted for æsthetic perception with its calm and dreamlike detachment. In addition, by these senses a number of persons may join in a common act of æsthetic contemplation, and in doing so, it permits the impressive conviction that æsthetic experience is a common possession and a form of social enjoyment. Furthermore these two senses most readily increase perceptual enjoyment through their resonant effects of sympathy.[3]

Music deals with longitudinal vibrations in air, having a frequency range from approximately sixteen to forty thousand cycles, with a speed about one-fifth mile per second. Although the frequency range impresses eleven octaves, only seven are used musically.[4] Pitch determination is only possible from twenty to twenty thousand cycles.[5]

Color, on the other hand, is due to transverse vibrations in the so-called ether, embracing an approximate wavelength range from red at 0.0007 millimeters to violet at 0.0004 millimeters for good visibility, or less than one octave. Expressed in frequency the approximate range is from (red) 430,000,000,000,000 to (violet) 750,000,000,000,000 cycles. The eye responds to radiation between indefinite limits from 300 to $1000\mu\mu$ (millionths millimeters).[6] The speed of light is roughly 900,000 times that of sound.

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[1] Faulkner, What We Hear in Music.

[2] Helmholtz, Sensations of Tone.

[3] Encyclopaedia Britannica, (Aesthetics).

[4] Glazebrook Dic. App. Physics, (Sound).

[5] Fletcher, Speech and Hearing.

[6] Nutting, *Bull. Bur. Stand.*, 5, 2, 1908–1909.

It seems that to correlate sound and color is at once impossible of solution. In spite of this, pleasing results can be obtained, for æsthetic enjoyment is not based on formula.

With color, the eye perceives three factors—hue, degree of saturation, and brightness. The hue is the wavelength of the monochromatic radiation, and saturation depends upon the amount of white light in the total.

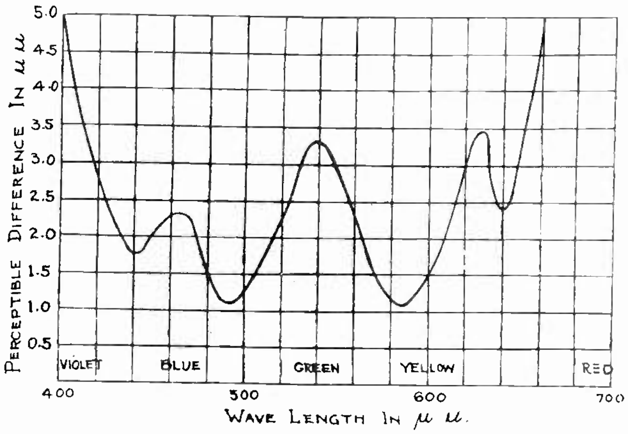

Fig. 1

The perception of hue and the rate at which it changes with wavelength is very irregular. With two adjacent areas illuminated with monochromatic light of different hues or wavelengths, it is possible to determine the extent by which they must differ in order to be perceptible. This is shown in Fig. 1.[7]

It will be seen that a great difference is necessary at the extreme red and violet ends of the vision range, and thus the eye, in this particular respect, performs as does the ear at the extreme ends of the audio spectrum where variations in pitch are difficult to determine within narrow limits.[5]

Of particular importance from a lighting standpoint is the chart in Fig. 2, which records results made with a number of observers on the visibility of various colors.[8] Red at 0.70 μ and violet at 0.40 μ to be seen with the same intensity as yellow-green at 0.56μ must have many times the energy flux of yellow-green. Consequently in arranging for lighting loads this factor must be carefully considered in order to secure color balance. Here again the eye and the ear have a part in common, as the ear has a sensitivity curve somewhat similar.[5]

With the foregoing relations in mind, we are now in a better position to discuss the color music and its harmony.

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[7] Glazebrook Dic. App. Physics, (Eye); Jones, *Jour. Opt. Soc. Amer.*, i, 63, 1917; Nutting, *Bull. Bur. Stand.*, 6, 1, 1909–1910.

[8] Coblentz and Emerson, *Bull. Bur. Stand.*, **14**, 1918–1919.

COLOR MUSIC

Klein[9] has very ably covered its history, the color organs devised for color projection, and the color scales. The control of most of the instruments, which have found favor in the past, has usually consisted of a keyboard similar to that of a piano or a series of levers for manual operation. Notable devices of this type have been built by Castel, Remington, Greenewalt, Wilfred, Hector, Luckiesh, Klein, and others.

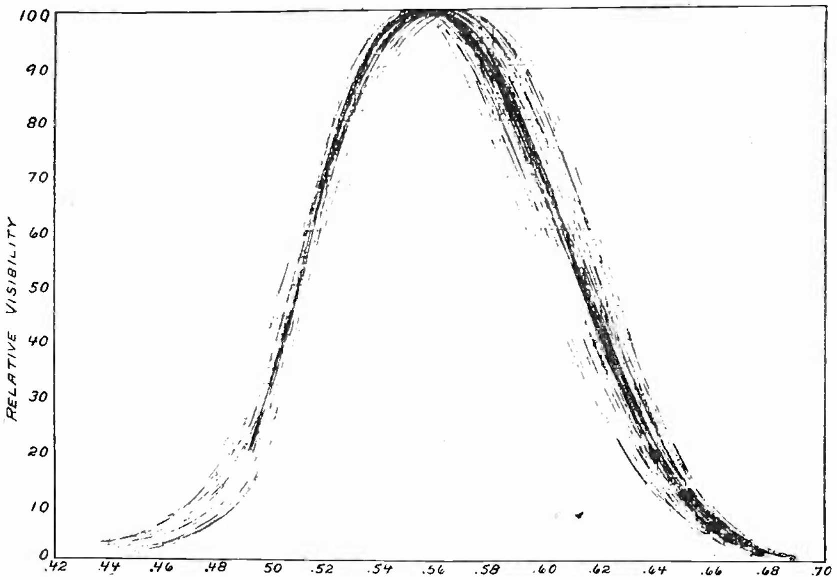

Fig. 2—Composite visibility curve of 125 persons.

Wilfred, in addition to the manual control, has also developed a mechanically driven Clavilux requiring no keyboard. Colors are obtained by the projection of white light through a painted disk. With this predetermined selection, the colors may be repeated at will. Other controls have been added to provide for patterns and background lights. His designs are very beautiful and have a dreamlike character, which aids in the appreciation of the exhibit. As far as it is known, the only devices to appear commercially are those operated by hand or mechanical drive.

While a knowledge of apparatus employed in the past and present is of importance, the color scales or harmony systems are of prime interest. It is to be expected that color schemes analogous to musical scales have received considerable attention. According to Klein, Aristotle can be credited with the first thought along this line and he has been followed by a host of others.

A typical scale is Remington's, which he used with his keyboard

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[9] Klein, Color Music—The Art of Light.

color organ with and without musical accompaniment. He assigned to the note C—deep red, D—orange-crimson, E—yellow, etc., through the octave to violet. For the next higher octave he assigned the same colors but with less saturation. Scriabine also had a scale wherein colors were assigned positions with respect to the musical frequencies.[10]

While we can appreciate the very natural tendency to associate color with music in this manner, with some thought, it does seem other systems would be less artificial.

It is generally recognized that colors do exert, over the majority of us, a profound influence. The following table by Luckiesh[11] gives a series of colors with the commonly associated reactions:

Red—warm, exciting, passionate
Orange—warm, exciting, suffocating, flowing, lively
Yellow—warm, exciting, joyous, gay, merry
Yellow-green—cheerful
Green—neutral, tranquil, peaceful, soothing
Blue-green—sober, sedate
Blue—cold, grave, tranquil, serene
Violet—solemn, melancholy, neutral, depressing
Purple—neutral, solemn, stately, pompous, impressive

By studying the frequency content and the general mood of the music played, it is possible to employ colors which aid in the interpretation of the mood. At least the goal should be the correlation of the moods even though the accomplishment may not be perfect.

A method of cuing the color to music is to assume that the bass notes of the drum indicate an effort, on the part of the composer, to create a stirring effect and hence a red color, In practice red may usually be assigned this position. The other colors, however, represent more of a problem.

DESIGN

With the discussion of the important factors of color music, we are in a better position to design the instrument.

It is first of all necessary to segregate certain bands of audio frequencies which will serve to actuate lights of various colors.

Because of the variation of the audio content, the mood expressed in the music, and of the natural tastes of the audience, a switching means should be provided to permit easy interconnections between the audio bands and the colors.

A choice of the colors to be projected must also be given considera-

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[10] Luckiesh, Color and Its Applications.
[11] Luckiesh, The Lighting Art.

tion. There is a difference between the mixing of colored lights and the mixing of pigments, as the results are diametrically opposite. The effect of mixing colored lights is to produce a color which is more nearly white than its components, whereas the mixing of pigments as for a filter is to produce a lower transmission than either of the two components.[12]

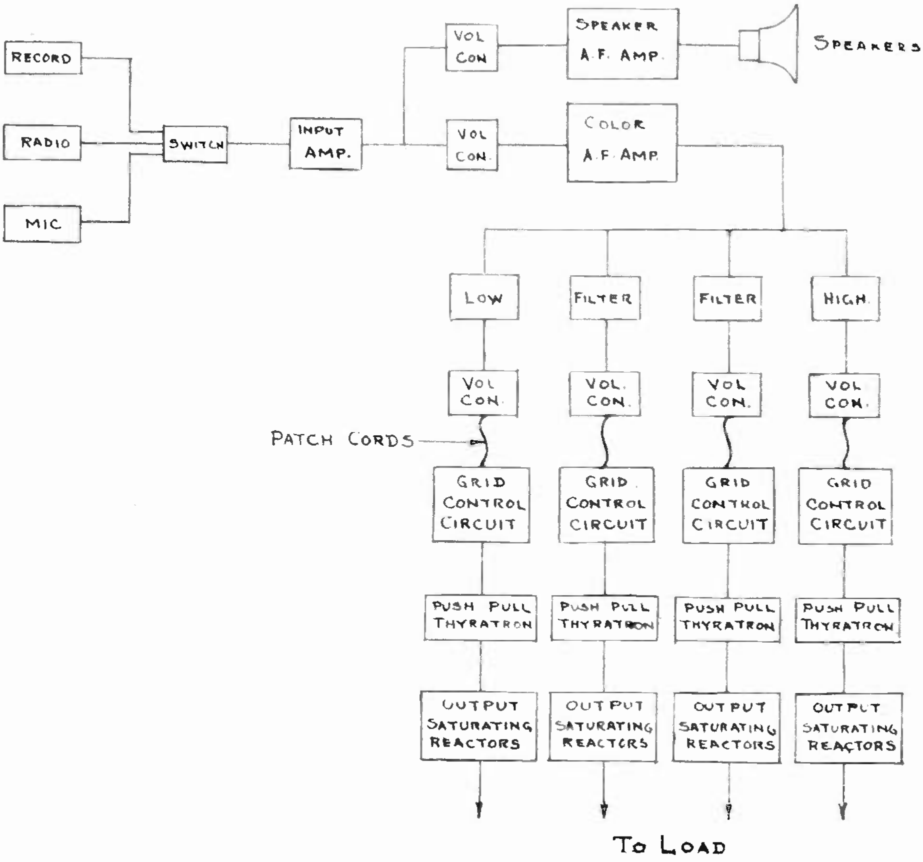

Fig. 3—Automatic color organ.

Since it is possible to match every known color by mixing correctly colored lights of red, green and blue,* it seems desirable to employ these colors and in addition clear light, which permits variation of saturation.

The color organ must provide means to amplify the music picked up by a microphone or from a record or radio in order to attain sufficient amplitude for control purposes. Means must also be provided to control a light source in accordance with the rather rapid pulsations from the music.

[12] Jolley, Waldram, and Wilson, Theory and Design of Illuminating Eng. Equip.

* Luckiesh suggests red with dominant hue near 0.66μ, green near 0.54μ, and blue near 0.45μ for this purpose, although other sets may be employed.[11]

The general layout of the equipment to meet the requirements is shown in Fig. 3, and the general plan is much like the original working model designed some years ago. At that time thyratrons were not available and rectified audio frequency was employed to close a series of sensitive relays which, in turn, controlled the lighting load.

At the left of the diagram are the various inputs, such as radio, record and microphone. Next is the switching device to connect them to the input amplifier. Here a standard amplifier with a '27 for the first stage is transformer coupled to two '45 tubes in push-pull. A standard last stage of two '50 tubes in push-pull drives the color circuit. Several types of push-pull last stages may be used for the loud speakers depending upon requirements.

From the color amplifier, the audio frequency passes through a series of filters. Actual practice has shown four will usually be sufficient and they are represented in the diagram. The audio frequency from these filters is connected to rectox rectifiers and the direct current thus obtained serves as the control for the type FG-27 thyratrons, one circuit of which is shown in Fig. 8 and described later.

Provision is made for individual color volume controls, as these are of advantage in balancing the colors or causing certain colors to predominate. Patch cords connect the various filter circuits to the various thyratrons and their colors and permit any audio band to control any color.

Other arrangements are, of course, possible but the equipment represented in Fig. 3 permits the use of standard parts without recourse to new designs.

The thyratrons are connected in push-pull to avoid the possibility of flickering on account of the sixty-cycle power which is applied to them.

While the power obtained from the thyratrons is on the order of five hundred watts for each set of tubes, this is not sufficient for commercial purposes. Hence output saturating reactors are introduced and with these it is possible to control 50 kw of load for each set or 200 kw for eight tubes. Because of the color balance required, the entire amount cannot be realized or else the clear or white light would greatly exceed the other colors. However, thyratrons may be connected in parallel for the weaker colors, thereby controlling more power and thus balance is maintained economically.

THYRATRON CONTROL

Of particular interest is the thyratron control, without which the enormous amplification of microwatts to millions of watts would not be economically feasible.

Hull[13] describes the thyratron as consisting of a cathode, grid, anode, and a small amount of inert gas. It is possible to start current through the tube as an arc by the grid. However, after starting, the grid exerts no further influence. To stop the arc, the anode voltage must be removed. Because of this characteristic the tube lends itself to a-c operation as the anode voltage necessarily passes through zero periodically and stops the current which may then be restarted by the grid.

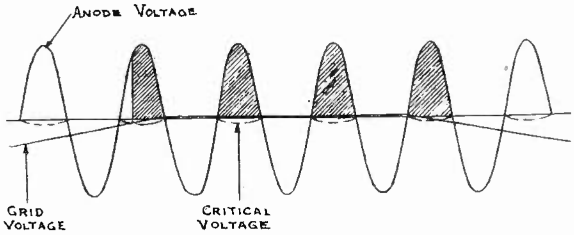

Fig. 4

Fig. 4 shows the simplest method of operation. The grid-voltage critical value (indicated by the dotted line) is where the current will just start. With the grid voltage more negative than this value no current flows. As the grid becomes less negative than the critical value current flows for the remainder of the half cycle.

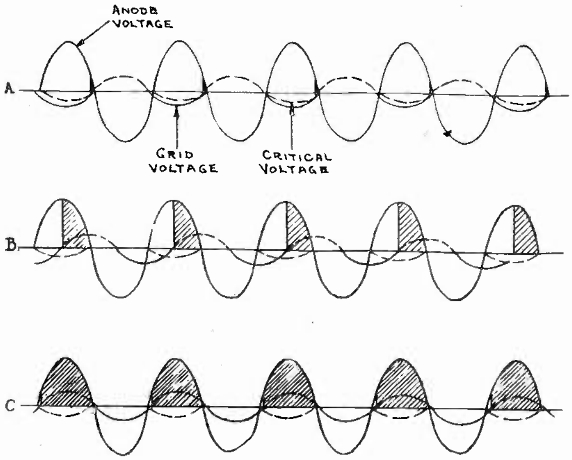

Fig. 5

Hull[13] describes another method of control illustrated in Fig. 5. Here alternating voltage is impressed on both the grid and anode and the phase of the grid is varied with respect to the anode. The grid, almost out of phase with the anode voltage, is shown at (A) and no current flows except at the end of the cycle. With the voltages more nearly in phase the current starts in the middle as at (B), and when in

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[13] Hull, *Gen. Elec. Rev.*, **32**, No. 7.

phase current starts at the beginning of the cycle and continues, as shown at (C).

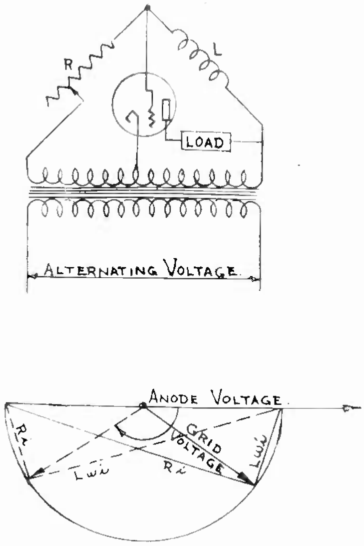

Fig. 6

The phase variation may be accomplished by combinations of resistance, inductance, and capacity. Fig. 6 shows a resistance control

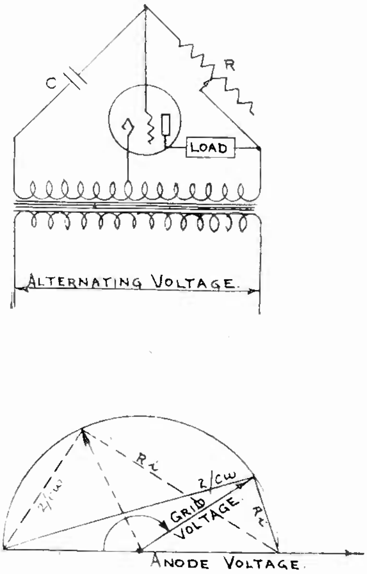

Fig. 7

with inductance where the phase is retarded through 180 degrees, as the resistance is decreased from a high to a low value. Fig. 7 shows a capacity-resistance method.

In Fig. 8 is a control circuit, which has been found particularly useful, as it permits control of the thyratron by rectified audio fre-

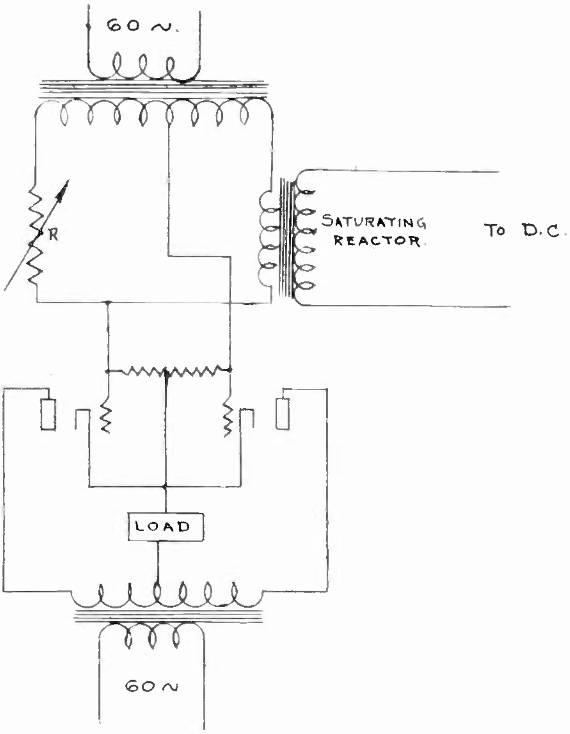

Fig. 8

quency. Variation in the sensitivity of the circuit to the direct current is obtained by adjustments of resistor (R).

The thyratron tube cabinet is shown in Fig. 9. The transformers

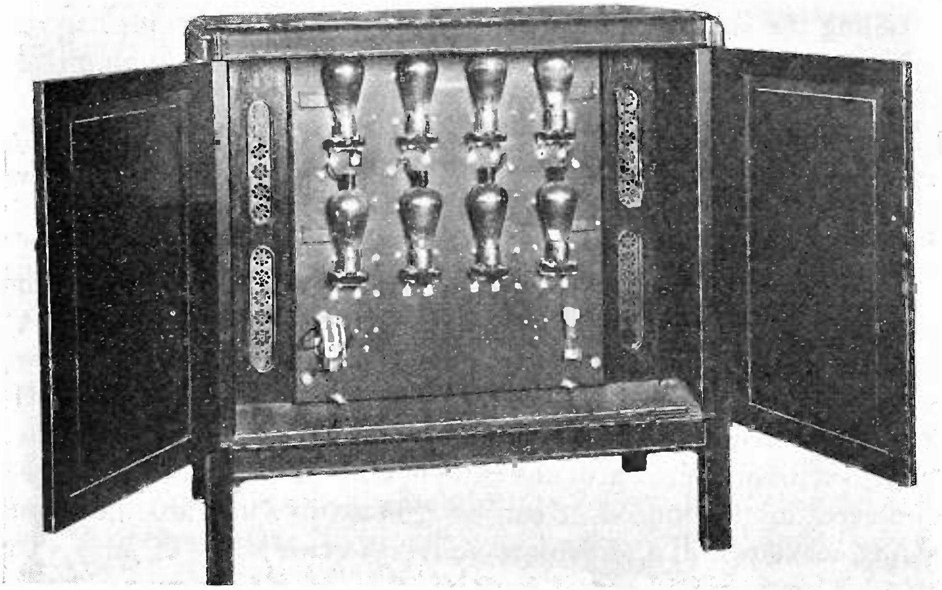

Fig. 9

and incidental equipment are mounted in back of the panel. A second cabinet of similar size contains the audio-frequency amplifiers.

Fig. 10 shows the control cabinet. In the upper right-hand corner are controls for varying the intensity of the various colors. Patch cords are provided to interconnect the jacks from the filters to the thyratrons controlling the color output. The top row of jacks is provided to read saturating current of the saturating input reactors. The knob directly under the meter is the master volume control for the light. The filters are located in the bottom of the cabinet. An automatic record changing device is on the left-hand side. Auxiliary input controls are not shown.

Fig. 10

Operation

In operation, the musical selection is played to determine its audio-frequency content and its mood. With patch cords, connections are made from the various audio-frequency bands to the lights. In many cases the arrangement will consist of red at low, blue and green for the intermediate frequencies, and amber or white for the high frequencies. If there is a preponderance of one band of frequencies, which may upset the musical interpretation, it can be reduced by the individual band controls, as shown at the cabinet above the two strips of jacks. The particular arrangement of the equipment does not require any adjustment for the average recording or musical selection, as there appears to be sufficient flexibility to take care of pianissimo and fortissimo passages.

It is possible, with special arrangements, to obtain a most sensitive control, the colors following practically every change in the music. However, violent fluctuations tend to become objectionable. Where extremely rapid changes are required, incandescent lamp filaments should not be too heavy on account of the time delay in heating and cooling.

While we have roughly determined the color of the lights to be employed, the success of the presentation depends greatly upon the manner of light projection and also on the introduction of some moving patterns, which serve to relieve the possibility of monotony.

There are a number of effect machines to produce clouds, waterfalls, etc. These, for the most part, consist of a revolving or painted disk in front of a spot light. These spots may be directed on a curtain, and used in conjunction with ordinary border- and footlights as will be found in a theater.

Very elaborate lighting schemes are coming into prominence where bare walls are painted by color patterns and projected pictures. These systems serve to focus public attention on the lighting art, and lend themselves to the easy adaption of color music.

Another application of the color device will be to fountains, where, in addition to the color, it will also be possible to control water valves where high amplitudes of music cause increased amounts of water flow. In other words, if mechanical motion is required, it can be obtained as the color organ is not confined to a lighting load of incandescent lamps.

A method now being investigated to produce variations in pattern is along the lines of Miller's Phonodeik.[14] Here audio-frequency vibrations by means of a vibrating mirror, cause a spot of light to fall on a revolving mirror, which, in turn, projects a beam of light of constant intensity on a screen. This can be used with a color light, the intensity of which varies in accordance with the music. In addition, a distorting prism or crystal can be introduced in the path of the light to secure a distorted wave form, as this apparently is more pleasing to the eye because of its added variation.

In the creation of patterns, moving or still, care must be exercised in avoiding too definite a structure. The imagination is important in giving æsthetic enjoyment which cannot be realized to the fullest extent when the pattern is too concrete in form, even though it may be very beautiful in design.

[14] D. C. Miller, The Science of Musical Sounds.

Where a microphone is employed with an orchestra some very excellent presentations can be obtained. The instrument playing a solo part can be assigned a definite color and the microphone or microphones can be placed in such a position that a very strong signal can be received from the solo instrument. The band-pass filters are then employed, which embrace the frequency range of the instrument in question. This permits the solo part to stand out very definitely in relation to the background colors produced by the other instruments.

Last but not least, education plays an important part in the appreciation of musical themes. More enjoyment is usually obtained in listening to a symphony if we understand what the composer is attempting to interpret. The same holds true for color, and explanations should usually accompany presentations when it is at all possible to do so. In this way the audience, if it is at all receptive, is in a better position to understand the color technique.

CONCLUSION

It is difficult to predict the future of color music. However, it is believed the radio by-product described in this paper will be a tool in advancing the struggling art. Certainly the main task to be undertaken now is in the creation of projection apparatus to permit the rendering of color in a form as appealing to the eye as a symphony is to the ear. The artistically inclined lighting expert should find this a fertile field.

Success should attend those who nurture the art if a sane and reasonable attitude is pursued. Helmholtz[2] adequately covers the case when he writes, "Beauty is subject to laws and rules dependent on the nature of human intelligence. Art works with design, but the work of art should have the appearance of being undesigned and must be judged on that ground. Art creates as imagination pictures, regularly without conscious law, designedly without conscious aim."

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Proceedings of the Institute of Radio Engineers
Volume 19, Number 8 *August, 1931*

SOME DEVELOPMENTS IN COMMON FREQUENCY BROADCASTING*

By
G. D. Gillett
(Bell Telephone Laboratories, New York City)

Summary—*This paper describes the results of the simultaneous operation of radio stations WHO and WOC broadcasting the same program on a common frequency using independent crystal controlled oscillators. These stations had previously been compelled to share time on 1000 kc and each is now able to render full time service.*

The exceptional stability of the crystal controlled oscillators used at each station is described. Since even these oscillators require occasional readjustment to maintain them in isochronism, a monitoring receiver was established midway between the stations and the resultant program is sent back by wire line to WOC to provide an indication for readjusting its frequency to exact isochronism with WHO. An audio oscillator used to modulate the carriers in the monitoring receiver provides a tone independent of the program for the guidance of the operator. Curves are presented showing the quality impairment caused by different degrees of isochronism and signal strength ratios.

The improvement in distance reception with simultaneous operation is reported and an explanation given. The impaired reception in the area midway between the stations and outside their normal service range is shown to be a function of the degree of modulation of each transmitter, of the field strength ratio and of the audio phase angle and independent of the carrier phase at the transmitters. It is pointed out that reception equal to that from either station alone may still be obtained in this area by the use of a simple directive antenna.

The marked increase in the service rendered by these stations through simultaneous operation is indicative of the improved service that can be rendered to urban areas by common frequency broadcasting. Although it is probable that the high powered station on a cleared channel will remain the best means of affording a high-grade service to a metropolitan area while also rendering an acceptable service to large rural areas, common frequency broadcasting now appears to offer definite means by which to provide an improved coverage to a number of noncontiguous communities.

THE development of chain broadcasting and the congestion in the broadcast frequency range has naturally led to a consideration of the possibilities of operating a group of stations on a single frequency.[1] The possible usefulness of such a system has resulted in a number of attempts to secure the additional coverage offered by the simultaneous operation of two or more stations broadcasting the same program on a common frequency. This problem has been attacked in two different ways.

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* Decimal classification: R550. Original manuscript received by the Institute, May 15, 1931. Presented before Sixth Annual Convention, June 4, Chicago, Illinois.

[1] De Loss K. Martin, Glenn D. Gillett, and Isabel S. Bemis, "Some possibilities and limitations in common frequency broadcasting," Proc. I.R.E., 15, 213–223; March, 1927.

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In one case, a control frequency has been transmitted either by wire line or radio to each station and a frequency multiplier used to develop directly the carrier frequency which was to be transmitted from the station. This method has met with some success both here and abroad. It was used in this country for the commercial operation of WBZ-WBZA[2] and in Germany the Postal authorities have operated several stations experimentally with equipment developed by the Telefunken G.m.b.H. and the C. Lorenz A.G.[3,4] Both the WBZ-WBZA and the Telefunken systems used a high control frequency which was particularly suitable for transmission over open wire lines while the Lorenz system used a lower control frequency which was suitable for transmission over cable circuits as well. Three stations located at Berlin, Stettin, and Madgeburg, respectively, are now in commercial operation on a common frequency using control equipment manufactured by the Lorenz firm.[5] In Sweden the postal authorities have developed a similar system of frequency control capable of using either a high or low standard frequency interchangeably. This system was used in placing the broadcast stations at Malmo and Halsingborg in commercial operation on a common frequency in the latter part of 1929.[6] Intensive development work on similar systems is under way in the United States. The National Broadcasting Company has in operation in their network, two groups of two stations each which are being operated synchronously using a standard reference frequency transmitted between stations over telephone circuits. The Bell System has developed a common frequency broadcast system using a standard reference frequency suitable for transmission over telephone circuits. This system has been given a practical test in coöperation with the Columbia Broadcasting System. It will shortly be commercially available.

The other method of attack has been to derive the carrier frequency at each station from an independent oscillator. In England,[7,8] electrically driven tuning forks have been used to supply an audio frequency of high stability from which the carrier frequency has been derived by means of frequency multipliers. With this equipment it has been possible to maintain the derived carrier within a few cycles per second of

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[2] Frank B. Falknor, "A history of synchronization," *Citizens Radio Call Book Magazine and Technical Review*, **12**, 38–40; March, 1931.

[3] W. Hahn, *Funk*, 35, 247–248, 1928.

[4] W. Hahn, *Die Sendung*, 5, 430–432, 1928.

[5] F. Gerth, "A German common frequency broadcasting system," Proc. I.R.E., **18**, 510–512; March, 1930.

[6] Erik Esting, *Elektroteknik*, pp. 109–112, June 7, 1930.

[7] P. P. Eckersley, "The operation of several broadcasting stations on the same wavelength," *Jour. I.E.E.*, 1929.

[8] P. P. Eckersley, "The simultaneous operation of different broadcast stations on the same channel," Proc. I.R.E., **19**, 175–194; February, 1931.

isochronism[9] and this has been sufficient to permit a satisfactory service to be rendered to the territories immediately adjacent to each station. As will be shown in detail later there is a substantial difference between the service range of a station operating in almost perfect isochronism with the other stations in the common frequency broadcast system and that of a station which is more than a small fraction of a cycle per second out of isochronism. In this country "matched crystals" and other means of independent frequency control have been tried but the frequency stability of the best equipment available in the past has fallen far short of that required for the satisfactory operation of the stations on a common frequency.

In the spring of 1930 the Central Broadcasting Company of Iowa found themselves in the possession of a concrete example of the need for the simultaneous operation of two stations on a common frequency in that their stations WHO and WOC had been compelled to divide time equally on 1000 kc so that the Davenport and Des Moines areas each received service from their local station but half the time. These stations are 153 miles apart and either could be depended upon to render a high-grade service only within a radius of about fifty miles of the station. It was felt that with the simultaneous operation of both stations, each of these areas would receive full time service from its local station.

The Central Broadcasting Company presented their problem and asked for equipment capable of maintaining the carriers of these two stations within the limits of isochronism required for their simultaneous operation. Bell Telephone Laboratories, therefore, undertook the necessary development work.

The degree of isochronism required for the various conditions existing under the different types of common frequency broadcast systems is in fact a fundamental question that must be answered before any logical delineation of the problem can be attempted. Unfortunately there exists no similar condition in ordinary human experience from which a valid analogy can be drawn, so that the *a priori* assumptions which have been used in the preliminary theoretical discussions of the various phases of this problem have of necessity been based primarily upon personal opinion and the resultant conclusions have quite naturally varied between extremely wide limits.

The problem had been studied intensively during the preliminary field tests of common frequency broadcasting which were made in the

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[9] The term "isochronous" has been used instead of "synchronous" in order to exclude the concept of identity of phase which is usually included in the meaning of the latter together with the meaning of identity of frequency which is common to both words.

fall of 1929 in coöperation with the Columbia Broadcasting System using stations WABC and WCAU. It proved to be very difficult to get accurate and consistent data from such field observations without a very extensive series of tests because the fortuitous variations in the transmission medium continually altered the test conditions. These were especially troublesome since the frequency difference between the carriers is but one of the two independent variables of primary importance which affect the quality of the program received at any given point, the other being the ratio of field strength received from the two stations at the point in question.

It was, therefore, necessary to set up in the laboratory apparatus which would simulate as closely as possible the conditions existing in the field but with all the variables under definite control. Two identical miniature transmitters were modulated by the same program. The modulated carriers were then attenuated through independent transmission paths and received by a high-quality detector. The layout of the apparatus is shown schematically in the block diagram of Fig. 1. It will be seen that with this equipment the signal strength received at the detector from either station may be varied so that any desired signal strength ratio may be obtained. The frequency difference, Δf, was fixed directly by the adjustment of the carrier frequencies of the two transmitters to the require degree of isochronism. These transmitters, operating at a frequency of approximately 50 kc were quite stable and capable of accurate adjustment.

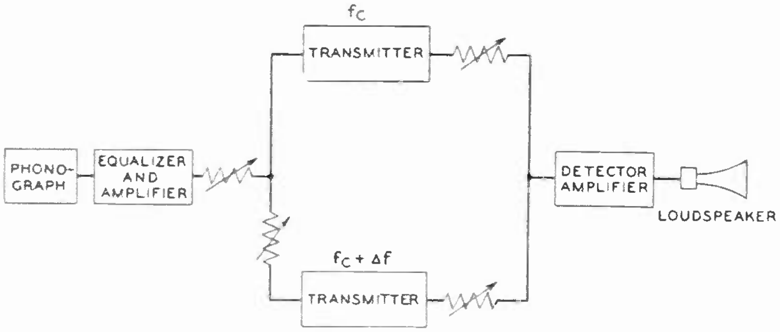

Fig. 1—Block diagram of apparatus set-up for determinations of quality impairment with different degrees of isochronism and field strength ratios.

The over-all audio-frequency transmission characteristic of the whole system was even better than is available in the better commercial radio receivers. The observers were engineers well acquainted with the effects to be expected and whose judgment was extremely critical. Tests were made with material consisting of both musical and talking programs and, while the effects are more noticeable with musical programs due to the presence of sustained tones, the difference was not marked. The observers compared the quality of the program received from the two stations with varying field strength ratios and degrees of carrier isochronism with that received from one of the stations transmitting alone. The change from the test condition to the reference condition could be made at will and the gains of the various circuit elements were adjusted so that the apparent program level was the same under the two conditions. Each test covered a considerable period of time and the curves shown in Fig. 2 mark the field strength ratios at which the observers could not distinguish between the test and reference conditions. The data shown are, therefore, believed to be distinctly conservative and to represent a criterion much more severe than any which will be encountered in commercial operation. These results are also in agreement with the experimental data that were obtained from our field tests and also check closely the data obtained by the engineers of the British Broadcasting Company in similar field tests in England.

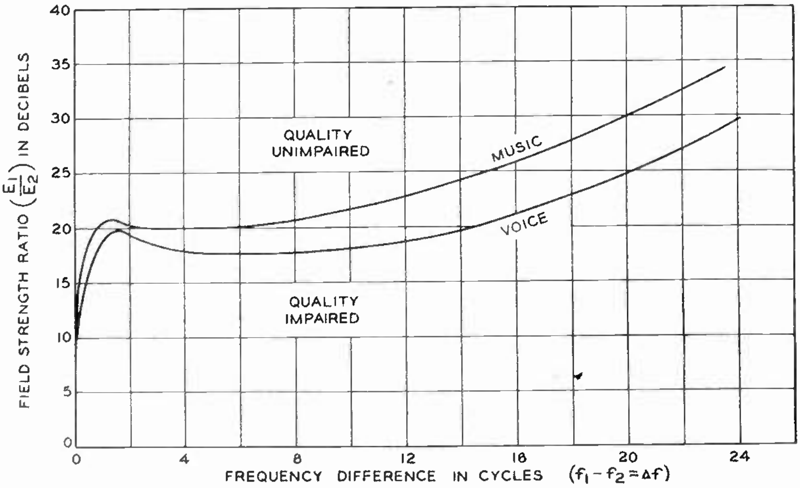

Fig. 2—Quality impairment vs. the frequency difference f_1-f_2 and field strength ratio E_1/E_2. The curves mark points where the quality impairment was just perceptible.

It will be noted that when the frequency difference is very small, closely approaching isochronism, unimpaired reception is assured provided the field strength ratio is at least 10 db, but that as soon as the frequency difference is at all appreciable the required field strength ratio for ordinary programs rises sharply to about 20 db and is approximately constant within the range from 1 to 10 cycles per second.

Our field strength distribution surveys and studies have shown that, for 5-kw stations separated by two or three hundred miles, a field strength ratio of 20 db is obtained only at points well within the normal service area of the station. On the other hand the limits of the 10-db ratio lie for the most part outside the normal service range of the station. Thus if such a station is to be operated on a common frequency chain, the carriers must be maintained approximately in isochronism if a large portion of the listeners within the normal service range of the station are not to receive a seriously impaired program. If approximate isochronism is maintained, the service area of each of these stations should not differ materially from that which selective fading and interference would establish for that station transmitting alone.

In order to maintain unimpaired reception in the region where the field strength ratio is between 10 and 20 db, it is necessary that the stations be operated so that their carriers are not permitted to differ in frequency by more than one cycle in 10 seconds and this demands a frequency stability of an entirely new order of magnitude for commercially available independent oscillators. However, at the time that this development was undertaken for the Central Broadcasting Company, previous tests had shown that a newly developed crystal controlled oscillator unit designated as the No. D-90684 oscillator-amplifier possessed an exceptional frequency stability for commercial equipment and that minor modifications would give it the stability required for the simultaneous operation of a small group of stations on a common frequency.

It was therefore planned to replace the existing crystal control equipment by one of these new units located at each station and supplemented by a monitoring receiver located midway between the transmitters.

The No. D-90684 oscillator-amplifier is a relay rack mounted assembly consisting of a shielded unit containing a constant temperature oven and a crystal oscillator, an amplifier having a maximum power output of thirty watts, and the necessary power control equipment. The amplifier tubes, instruments, and controls are mounted on the front of the panels as is shown in Fig. 3 and all other apparatus is mounted in the back and enclosed by a metal locker. The assembly of the various components inside the locker is shown in Fig. 4. The power equipment is placed in the lower part, the constant temperature oven and crystal oscillator unit is mounted on slides in the middle compartment, while the upper shielded section isolates the buffer and output stages from the rest of the transmitter. The door of the locker is fitted with safety switches which automatically disconnect all high voltages from the equipment before the door can be opened. It was a simple matter to install one of these compact, self-contained units adjacent to each transmitter to replace the existing crystal control equipment as the source of the carrier frequency. A corner of the operating room at station WOC is shown in Fig. 5, with a part of the radio transmitter at the extreme right and the oscillator-amplifier mounted adjacent to it. The author is holding the crystal oscillator and constant temperature oven, and over his head to the left is the loud speaker on which the program from the monitoring receiver is received.

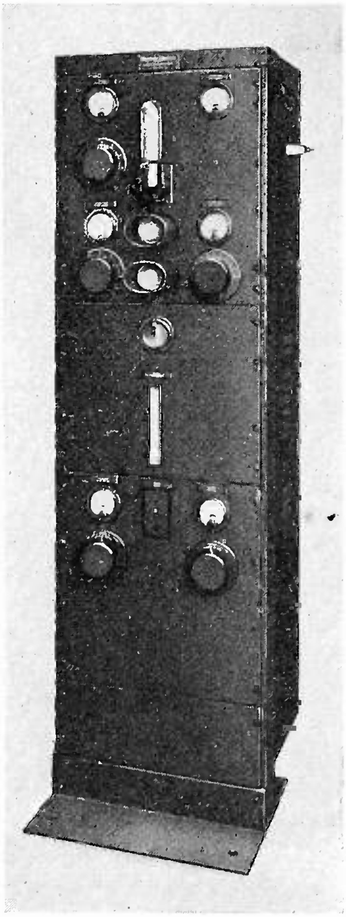

Fig. 3—Front view of crystal controlled oscillator-amplifier unit.

The extraordinary frequency stability of these units has not been obtained through any radical change in design but has come rather as a result of the refinement of all the component elements to form a coördinated unit. A clamped crystal has been used in an improved type of holder, designed to maintain a constant pressure on the crystal and at the same time to prevent any lateral movement which would cause a change in the crystal frequency. The crystal and its holder are mounted in an oven fitted with an improved thermostat capable of maintaining the temperature of the crystal constant within extremely narrow limits. This constant temperature oven is built as an integral part of the oscillator, which has been designed to work the crystal under the conditions of optimum stability.

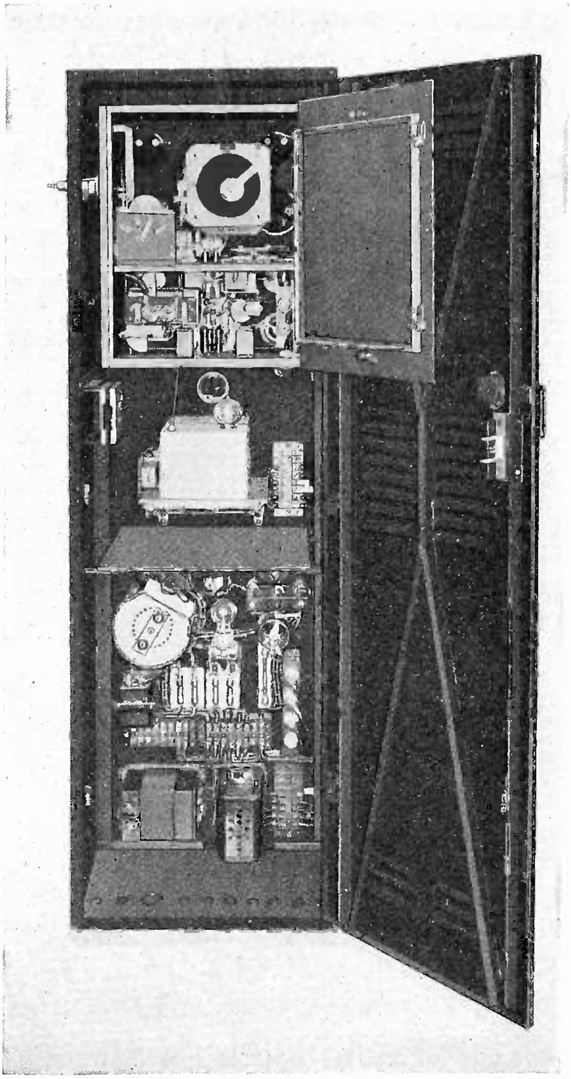

Fig. 4—Rear view showing interior of oscillator-amplifier unit.

The oscillator and crystal are carefully shielded and isolated from the output stage by several buffer stages in order to prevent any change in the load conditions from being reflected back to the oscillator and thereby changing its frequency. Careful tests in the laboratory have shown that the output power could be varied from zero to full load without affecting the frequency within the limits of observation, which were about one part in a hundred million. It is relatively insensitive to changes in filament current, though this is maintained constant within narrow limits by a ballast lamp. Since a change of one per cent in the plate voltage causes an immediate change in the frequency of about one part in fifteen million and an ultimate change of about one part in two million, the crystal oscillator is now being operated from batteries.

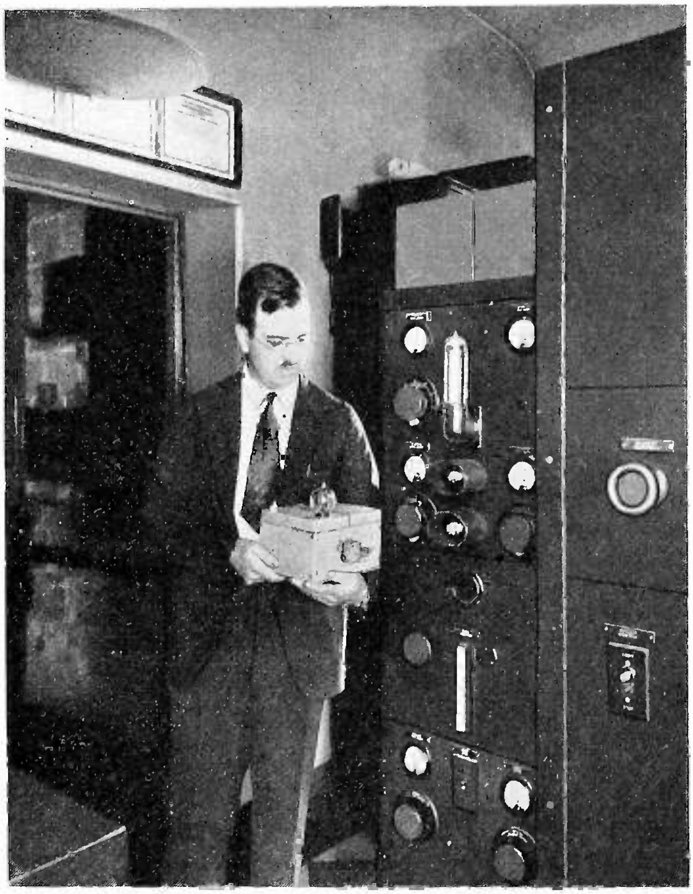

Fig. 5—A corner of the operating room at station WOC.

Since even with these oscillators absolute isochronism cannot be maintained indefinitely without readjustment, WHO was chosen as the reference frequency station and WOC was provided with means by which its carrier frequency could be brought into exact isochronism with that of WHO. In order that the operator of WOC could easily determine the degree of isochronism, a monitoring receiver was set up at a point midway between the stations and the program received there was transmitted back to station WOC by wire line. A departure of the two stations from isochronism is shown by a slow variation in the level of the program received and the operator can then make the adjustment necessary to restore the stations to isochronism. The nicety of this adjustment can best be appreciated by the fact that a complete revolution of the control dial varies the carrier frequency at WOC by but one part in a million.

It was found difficult at times to determine the beat frequency resulting from a lack of isochronism on account of the masking effect of the rhythm or beat of a musical program. The receiver was therefore equipped with a small audio-frequency oscillator which was arranged to modulate completely the incoming carriers received from the two stations. These combined modulated carriers are detected in the usual way and the output from the receiver is then an audio tone, the level of which is directly proportional to the resultant of the combined carriers. This tone overrides the program and is transmitted back to the station where even very slow changes in its level are easily detected by the operator. This also has the advantage that the degree of isochronism can be determined before any program is broadcast and that any necessary readjustment to restore isochronism can be made during silent periods in the program. This tone is required only at the

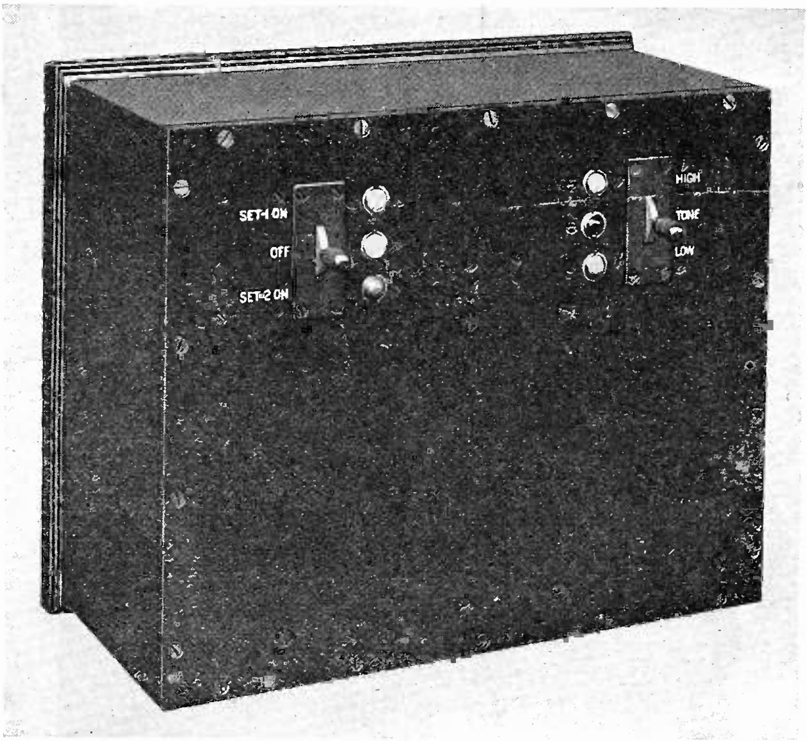

Fig. 6—Control panel for remote control of the monitoring point receivers.

time of adjustment, and relays at the monitoring point have been arranged for remote control from station WOC by which the audio oscillator is turned on whenever desired. These relays also permit the operation of either of two receivers and permit the setting of the gain of either receiver at the proper level for day- or nighttime reception. The control panel at the station, shown in Fig. 6, is equipped with supervisory signal lamps which indicate the position of these relays.

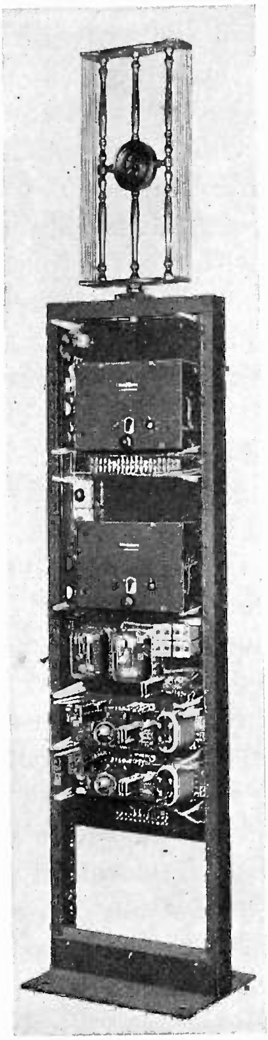

Fig. 7—Remote controlled monitoring point receivers.

Fig. 8—Can covers removed from rear of monitoring point receivers to show the location of the receivers and remote control relays.

The equipment at the monitoring point, shown in Figs. 7 and 8, is mounted on a single relay rack and includes the loop antenna which has been made sufficiently unidirectional to permit the obtaining of an exact balance between the signal strengths received from the two stations. The two radio receivers with their associated audio oscillators and the relay control panel complete the equipment at the monitoring point. The rack is arranged for the complete enclosure of all the equipment by means of dustproof can covers which also serve to prevent any accidental disturbance of the settings and adjustments.

With this equipment in commercial operation, a checking of the frequency every ten minutes in connection with the regular routine inspection of the transmitter has been sufficient to maintain the carriers within an average of two cycles per minute of absolute isochronism. Departures from isochronism of this order of magnitude are not detectable within the normal service area of either station.

While with an installation of this type one is primarily concerned with frequency stability rather than permanence of calibration, the Laboratories have measured the frequency of these stations periodically. It was found at the time of the installation, after the reassembly of the equipment subsequent to its shipment from New York, that the frequency was about two cycles per second different from that measured before shipment. Measurements since that time have shown that the frequency has varied over a period of time between seven cycles above the assigned frequency and seventeen cycles below it. It is known, however, that these variations were primarily due to the variations in plate voltage at WHO, which were permitted in the interest of economy, since the ensuing variations of the two carriers were still far within the requirements of the Federal Radio Commission.

More recent measurements made on a similar unit installed in a broadcast station where precautions have been taken to insure the application of a constant plate voltage to the oscillator have shown a frequency variation of less than five cycles per second from the assigned value over a period of several weeks.

Before approval was sought from the Federal Radio Commission for the full time simultaneous operation of these stations on a common frequency, careful surveys of the areas served were made by the engineers of the Federal Radio Commission, the Department of Commerce, the Central Broadcasting Company, and Bell Telephone Laboratories during their simultaneous operation on an experimental basis during the early morning hours in order to determine the nature of the service being rendered. Nearly three thousand miles were covered by the radio test cars during these tests. Upon completion of these surveys the Federal Radio Commission immediately granted permission for the simultaneous operation of WHO and WOC during regular broadcast hours.

These surveys showed that the service rendered by the simultaneous operation of these two stations was substantially twice as great as the service given on a shared time basis. The normal service area of each station was maintained and the nighttime reception at points over a hundred miles distant from either station was improved by the partial elimination of rapid and selective fading as well as by an increase in the average field strength received.

This improvement in distant reception was confirmed by the letters received in response to requests, made during the tests, for reports as to the quality of reception. In making these requests, the nature of the distortion that might be experienced was carefully described and it was especially emphasized that mere reports of reception would be of no value and that the information desired concerned the quality of the program received during the simultaneous operation of the stations as compared to that from either one alone. Several hundred replies were received from outside the State of Iowa beyond the normal service range of either station. These were almost unanimous in reporting better reception with simultaneous operation. The reports received from distant points during the first year's commercial operation are in full accord with these test data. This improvement apparently occurs wherever marked selective and general fading is experienced in the reception of either station alone.

It has been generally accepted that fading, commonly experienced in the nighttime reception of programs from a distant station, is due to the arrival of the signals along at least two different paths. In the mathematical analysis of this problem it will be convenient to represent each portion of the carrier which arrives at the receiving point via an independent path as a vector of constant amplitude and random phase variation. It will then be possible to represent the fading signal received from a single station as the sum of at least two such vectors. It is then logical to assume that the signal received from two distant stations operating on approximately the same frequency is the summation of at least four of these vectors of constant amplitude and random phase relation. This assumption of random phase relation is valid for any of the common frequency broadcast systems now being developed commercially either here or abroad. If the carriers are derived directly from a reference frequency transmitted via wire line circuits to the several stations, the slight phase variations caused by temperature and humidity changes are sufficient to cause the phases of the derived carriers of the different stations to vary in a fortuitous manner. Furthermore, even if the carriers of two stations were held exactly in phase at their respective antennas, or at some point midway between the transmitters, the variations in the path lengths of the waves arriving at any given distant point would be sufficient to cause a random phase variation. It is helpful in the mathematical analysis to assume also that these vectors are of equal amplitude. While this is not strictly true in all cases, our field observations have shown that it is the limit which tends to be approached as the distance from the stations is increased.

With these assumptions it can be shown mathematically[10] that the probability P_2, that the ratio of the sum of two vectors to their absolute sum will be less at any instant than a given value λ, is given exactly by the expression

$$P_2 = \frac{2}{\Pi} \sin^{-1} \lambda.$$

For larger values of "n," the exact expression is difficult to evaluate but a close approximation to the probability P_n for "n" vectors is afforded by the expression given below:

$$P_n = \frac{12n^2 - 6n + 1}{12n} \lambda^2 - \frac{12n^2 - 18n + 13}{24} \lambda^4$$

$$+ \frac{12n^3 - 36n^2 + 55n}{72} \lambda^6 - \frac{12n^4 - 60n^3 + 155n^2}{288} \lambda^8$$

$$+ \frac{12n^5 - 90n^4 + 350n^3}{1440} \lambda^{10} - \frac{12n^6 - 126n^5 + 646n^4}{8640} \lambda^{12}$$

This probability of the sum of "n" vectors being less than any given percentage of the absolute sum of "n" vectors has been computed by means of these expressions for the cases corresponding to the distant reception of 1, 2, 3, and 5 stations. The results of these computations have been plotted in Fig. 9.

There are two aspects of these curves which are of especial interest in connection with this problem. First it will be noticed that as the number of stations is increased the percentage of time that the signal fades below a small value such as 5 per cent of its maximum should be decreased. Thus the percentage of time that bad quality will be received due to the elimination of the carrier should be noticeably reduced as the number of stations is increased. Also it will be noted that a rapid reduction in the percentage of time that the signal approaches the max-

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[10] See Lord Rayleigh's, "Scientific Papers," Vol. 6, section on "Flights in 1, 2, and 3 Dimensions," and also section on "Random Unit Vibrations."

imum should occur as the number of stations is increased. This serves to emphasize the second aspect of the problem, i.e., the level of the signal received should remain near the mean for a much larger percentage of the time as the number of stations is increased. Thus a distant listener can set his receiver so that a normal level should be obtained for a much larger proportion of the time as the number of stations is increased. As an example, let us consider the proportion of the time that the level of the received program should lie between the limits of 25 and 50 per cent of the maximum signal, for one station it should be but 17.5 per cent while for two stations it should be 32.5 per cent, for three 45 per cent, and for five 55 per cent of the total time. A further development of the probability integral given above has shown that not only should the proportion of the time that a normal program is received increase, but that the instantaneous rate of fading should also decrease as the number of transmitting stations is increased. This is important because the sensitory reaction to fading, within ordinary limits, apparently depends more upon the rate of change of program level than it does upon the absolute total volume change. Since the same arguments apply equally well to each of the individual frequencies comprising the side bands, it can be seen that the general tendency of increasing the number of isochronously operated stations is to improve markedly the satisfactoriness of the program received at a point distant from all the stations of the chain.

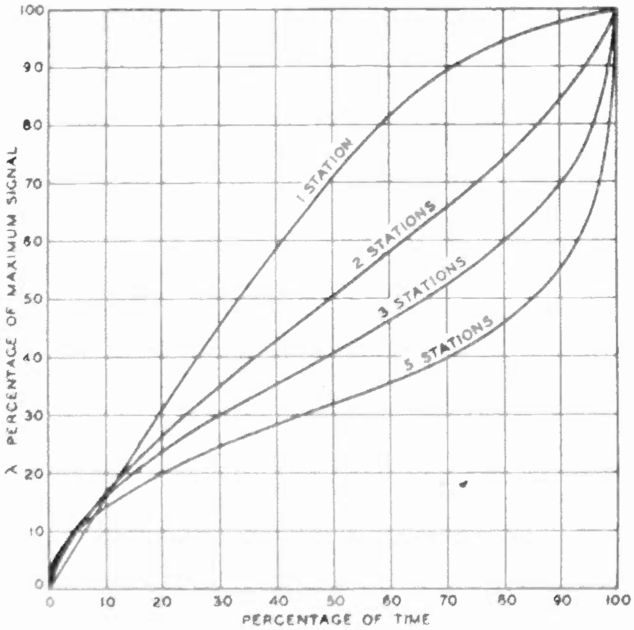

Fig. 9—Curves showing the probability that the instantaneous sum of the signal from a number of distant stations will be less than any given percentage, λ, of the absolute sum.

On the other hand in a small area midway between the stations, which received but a mediocre service originally since it lay outside the normal service area of either station, the reception with simultaneous operation was somewhat further impaired. The conditions that exist in this no-man's land between any two stations operating on a common frequency seem worthy of a detailed consideration, especially since wide publicity has been given to the misconception that the maintenance of the carries in perfect synchronism at the transmitters would entirely eliminate this area of impaired reception. It will be shown below that fundamentally the degree of isochronism merely determines the rate at which alternate strips of bad and good quality reception are swept across this territory. The attainment of exact isochronism would only mean that these strips would tend to be fixed in space and that a certain proportion of the listeners would then receive bad quality all the time instead of getting their share of the good with the bad.

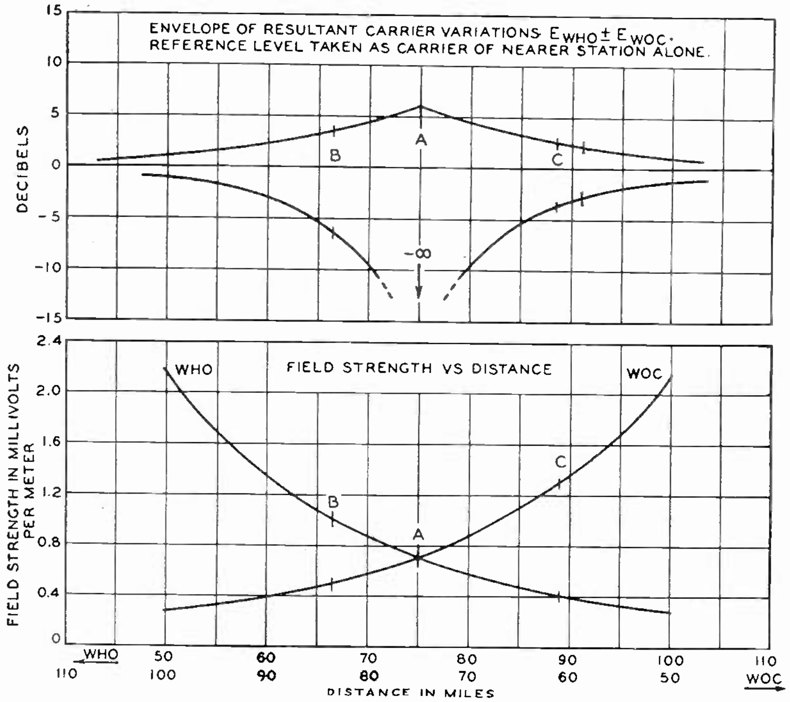

Fig. 10—Smoothed curves of field strength vs. distance for stations WOC and WHO, $fc=1000$ kc and the envelope of resultant carrier variations for $E_{\text{WOC}} \pm E_{\text{WHO}}$.

A smoothed curve of the daytime field strengths from WOC and WHO existing in the middle area on a line between the stations is shown in the lower part of Fig. 10.[11] The range of variation that the resultant carrier level will undergo as the carriers pass in and out of phase is shown in the upper part for the corresponding points along this line. In Fig. 11 enlarged sections showing in detail the variations of the resultant carrier level with distance are given for the point of equal signal strength *A* and for the points *B* and *C* where the field strength ratio is 6 and 10 db, respectively. Any departure from isochronism will have the effect of making these points of maxima and minima move along this line in space at the rate of one-half wavelength per cycle difference in frequency.

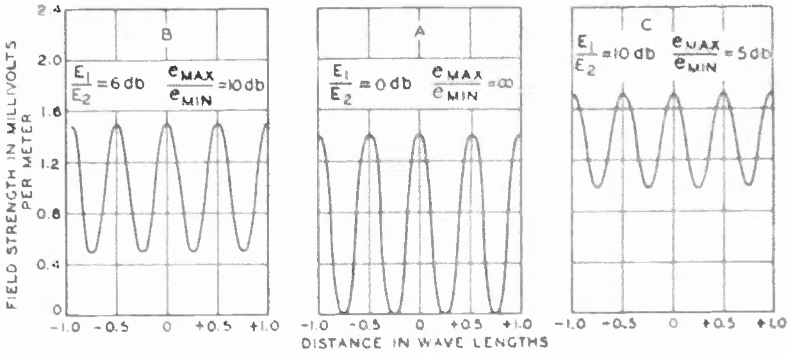

Fig. 11—Enlarged sections of the points *A*, *B*, and *C* of Fig. 10 showing in detail the resultant carrier variations with distance.

Now since the side-band frequencies must perforce differ in wavelength from the carrier, they will arrive at any given point out of phase with the carrier, the amount depending on the distance from the transmitter and the side-band frequency. Thus the side bands will not for the most part be in phase opposition at the same points in space as are the carriers, and distortion will result from the elimination of the carrier while strong side band components are present. The magnitude of this distortion is primarily a function of the existing field strength ratio between carriers and, while the distortion occurs for only a small proportion of the fading cycle, it is extremely objectionable where the field strengths approach equality. Here the carrier is almost entirely eliminated momentarily and the resultant program consists mainly of second harmonics and other distortion products.

It is entirely outside the scope of this paper to attempt to present a complete analysis of this problem but an effort has been made to indicate the quantitative results that may be expected by selecting a few typical examples. The signal being detected has been assumed in all cases to consist of the ordinary carrier and double side band transmission. The theoretical work which follows has been based upon the use of a square-law detector as being representative of the majority of the

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[11] These curves are based on field strength measurements made by Radio Inspectors J. M. Sherman and H. T. Gallaher of the Department of Commerce, and furnished through the courtesy of Mr. H. D. Hayes, U. S. Supervisor of Radio, Chicago, Ill.

existing receivers. In order to avoid undue complexity the curves have been computed for a single frequency audio signal.

In a square-law detector distortion appears primarily in the form of second harmonics and the ratio of these to the fundamental has been taken as a measure of the distortion present under the varying conditions of reception that may exist in the middle area between the stations. There are so many variables concerned in this problem that it is necessary to hold first one and then another fixed while different aspects of the situation are studied.

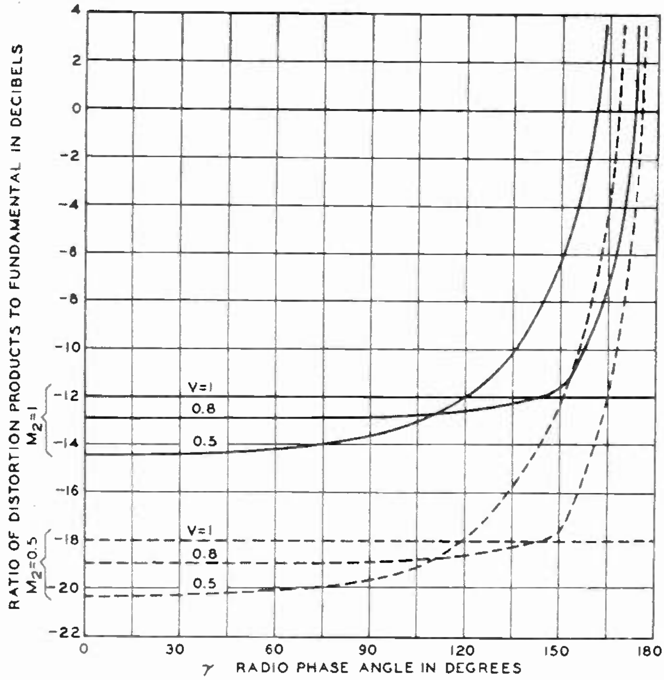

Fig. 12—Ratio of distortion products to fundamental for the point directly between and equidistant from the two stations, with varying carrier phase angle, γ, and different degrees of modulation of the two carriers, ($V = M_1/M_2$) where the audio phase angle $\beta = 0$ and the field strength ratio $E_1/E_2 = 1$.

The first set of curves, Figs. 12, 13, and 14, shows the conditions which exist at the point directly between and equidistant from the two stations when the audio signal supplied to the two transmitters is exactly synchronized, i.e., the relative audio phase angle $\beta = 0$. With this variable fixed the curves in each successive figure of the series have been plotted for successively decreasing signal strength ratios in order to show the effect of the varying radio phase angle with different degrees of modulation. It will be seen from these curves that making the modulation of the two carriers equal effects a tremendous reduction in the amount of distortion present in the hollows of the fading cycle that occur when the carriers approach phase opposition, i.e., $\gamma = 180$ degrees.

Also it will be seen from a comparison of the family of curves for 100 per cent and 50 per cent modulation ($M_2 = 1$ and $M_2 = 0.5$), that a reduction in the degree of modulation of both carriers by 6 db effects an equal reduction in the amount of distortion. Furthermore, a comparison of the curves in the successive figures will show how rapidly the maximum distortion due to the unequal degrees of modulation of the two carriers is reduced as the field strength ratios diverge from unity.

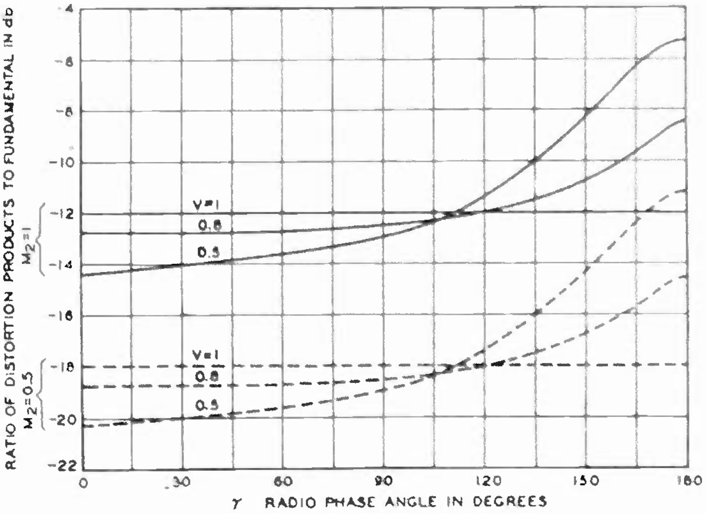

Fig. 13—Same as for Fig. 12 except that the field strength ratio $E_1/E_2 = 0.707$.

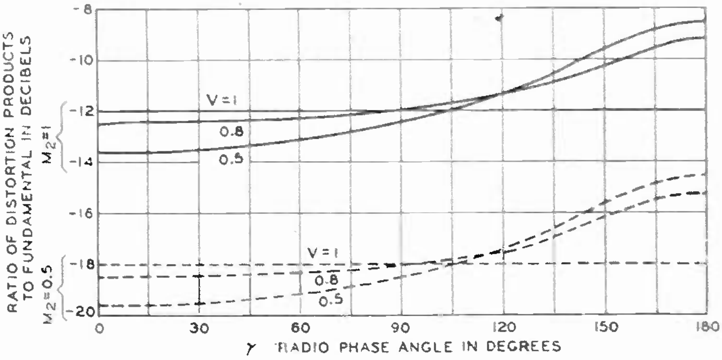

Fig. 14—Same as for Fig. 12 except that the field strength ratio $E_1/E_2 = 0.5$.

In the second series of curves, Figs. 15, 16, and 17, the effect of varying the carrier phase angle is shown for different representative values of audio phase angle while the degree of modulation of the carriers is fixed and equal and the field strength ratio is given a different value in each successive figure. Here the marked increase in the amount of distortion present as the audio components depart from synchronism and the carriers approach phase opposition is most striking. Also the rapid decrease in the amount of distortion present as the field strength ratio diverges from unity is noteworthy where $\beta \neq 0$.

The distortion for values of $\beta = 0$ and equal degrees of modulation ($V = 1$) remains constant in both these series of curves because this is the limiting case and is the distortion that would result from the reception of a similar program from but a single station. This fact affords a basis for comparison in considering the additional distortion that results under certain conditions from the simultaneous operation of two stations. While the distortion products loom large in proportion to the fundamental at times, these are also the times when the fundamental is fading out and the actual magnitude of the distortion products is not large.

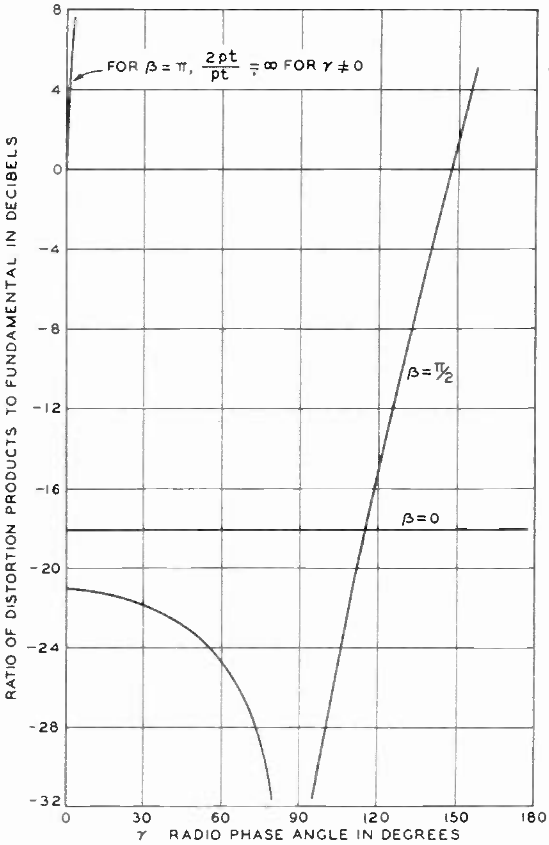

Fig. 15—Ratio of distortion products to fundamental with 50 per cent modulation of each carrier for varying audio and radio phase angles, β and γ respectively, where the field strength ratio $E_1/E_2 = 1$.

In practice, broadcast programs do not consist of a single frequency but represent instead a complex frequency distribution. The distortion resulting from the reception of such a program, therefore, represents a general average of all the different conditions shown in the curves given above. These are averaged in the final analysis by the listener, whose ear is far from linear in its response and whose judgment is affected by his personal opinions and past experience. We have, therefore, considered it futile and perhaps misleading to attempt to present any graphical summation of the effective distortion present in the reception of an ordinary broadcast program under such conditions. On the other hand these theoretical studies were un-

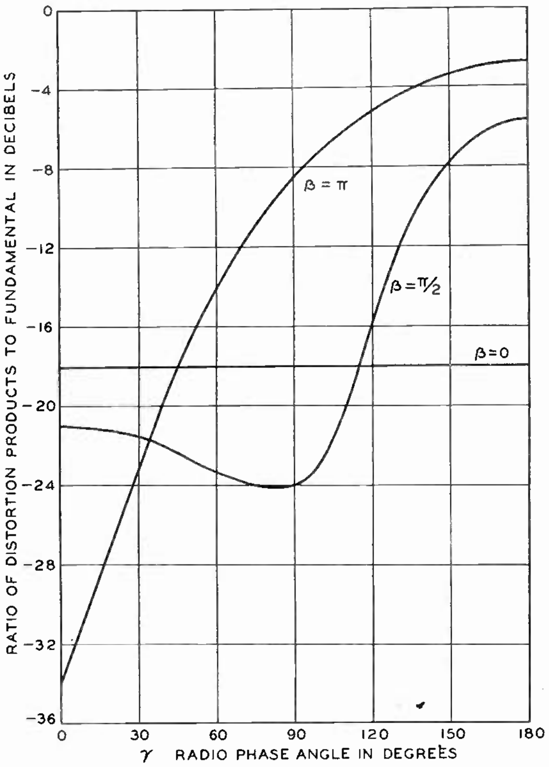

Fig. 16—Same as in **Fig.** 15 except that $E_1/E_2=0.707$.

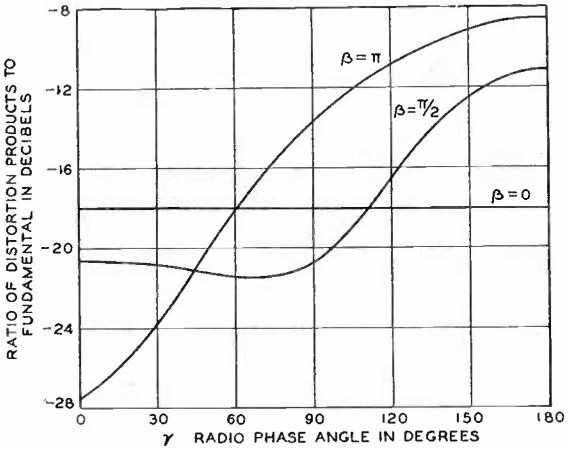

Fig. 17—Same as in Fig. 15 except that $E_1/E_2=0.5$.

dertaken in order to explain certain phenomena observed during the preliminary field tests as the degree of modulation and the field strength ratio were varied. The actual results have been quite closely corroborated by the conclusions reached from a study of these single frequency curves which have been of great value in obtaining a physical picture of the conditions that exist in this middle area.

These curves have been limited, for the sake of simplicity, to the consideration of the conditions at the points in the middle area equidistant from the two stations. It will be seen from these curves for this limited case that if, with equal degrees of modulation, each individual frequency component of the program could be synchronized at the two transmitters no additional distortion would be caused at these points by the isochronous operation of the two stations. But for points not equidistant the audio phase angle will not be zero even though it is maintained so at the transmitter and the magnitude of this divergence from synchronism will be different for each audio frequency and for each separate point in space. The magnitude of this divergence increases rapidly as the distance to the respective stations becomes more unequal. Furthermore, the problem of maintaining in synchronism every component of the broad frequency spectrum required for program transmission appears to offer tremendous technical difficulties.

It seems especially questionable whether such synchronism is necessary since tests in this area have shown that the use of a simple directive antenna capable of moderate discrimination against the weaker of the stations at the point in question is sufficient to render the reception at least comparable to that from either station alone. A loop antenna grounded at one side instead of the center was found to be very effective.

Population studies made in connection with the field surveys show clearly how marked is the improvement in the service rendered by these particular stations under simultaneous operation as compared with operation on a shared time basis. On a shared time basis a population of approximately 1,000,000 received adequate service from these stations half the time, the value of which was greatly impaired by its intermittent character. With simultaneous operation the service area of each station receives full time service. No accurate estimate can be made of the number of people in the middle area, and outside the normal service range of either station alone, whose reception has been further impaired by simultaneous operation. The importance of this effect can, however, be estimated from the fact that but 60 complaints of impaired receptions were received by these stations in the first 35 days of simultaneous operation during the regular hours and that the total for the first year is less than one hundred.

The marked increase in the service rendered by these stations through simultaneous operation is an indication of the possibilities of the improved service that can be made available to urban areas by the use of isochronized transmitters for the broadcasting of a common program. Although it is probable that the high powered station on a cleared channel will remain the best means of affording a high-grade service to a metropolitan area while also rendering an acceptable service to large rural areas, common frequency broadcasting now appears to offer a definitely useful means by which to provide an improved coverage to a number of noncontiguous communities.

In conclusion, the author wishes to acknowledge his especial indebtedness to the following members of Bell Telephone Laboratories; to Mr. G. R. Stiblitz for the development of the probability curves, and to Mr. C. B. Aiken and Mr. R. J. Jones for their preparation of the distortion curves as a part of their general mathematical study of the problem.

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THE DETERMINATION OF POWER IN THE ANTENNA AT HIGH FREQUENCIES*

By

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Summary—*Following a brief review of general methods of measuring radio-frequency power in the antenna, a series of tests on a particular transmitter operating from 4000 to 26,000 kc is described. These tests indicate that it is reasonable for such a transmitter continuously variable in frequency and of fairly modern design to be expected to put not less than 50 per cent of its input plate power of the last stage into the antenna as radio frequency from 4000 kc to 8000 kc and not less than 32 per cent at 24,000 kc. Intermediate efficiencies are indicated for the intervening frequencies. Coil losses and stray losses have been separated and the indications are that with modern insulation a moderate increase in efficiency can be expected in a transmitter designed for as high L/C ratio as possible, which is particularly feasible in the case of a transmitter operating on only one frequency. Marked increases in efficiency in the upper end of the spectrum can only be expected by the use of tubes better adapted to the upper frequencies.*

THE determination of radio-frequency power has always been a matter of some difficulty even at moderately low frequencies, while at high frequencies the difficulty of measurement is very much greater. The direct measurement of power would ordinarily be carried out either by some satisfactory determination of current through a low resistance of known value or would involve the measurement of voltage, current, and power factor simultaneously. At commercial frequencies wattmeters are of course available which will automatically carry out this measurement and give the results directly in units of power. It seems likely that instruments somewhat similar to commercial wattmeters might be evolved for very low radio frequencies but for even moderately high radio frequencies the difficulties due to distributed capacity, inductance, and resonance effects would be very great indeed.

The method of measuring the current into a known resistance can be carried out at fairly high frequencies but even here there is a limitation because of the great difficulty in constructing suitable resistances to handle any considerable amount of power at very high frequencies and at the same time to know the value of these resistances within a desirable degree of accuracy. The method of using artificial loads, such as incandescent bulbs, is very useful. The temperature of the bulb on the high-frequency load and on a measurable direct current or commercial

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frequency load can be made the same by the use of an optical pyrometer, and if the filaments operate in both cases at the same temperature, it is fair to assume that the power dissipation is the same in both cases but even here there are limitations as to frequency.

At very high frequencies it would be a serious mistake to assume that the filaments were noninductive. Moreover, the measurement only shows what the transmitter in question can do under certain power conditions—it does not tell us what power is actually being put into a definite antenna array. At low and intermediate frequencies much may be done by using a dummy antenna made up to have approximately the same capacity, inductance, and resistance as the actual antenna upon which the transmitter is to be used. However, this latter method is very unsatisfactory for high-frequency transmitters because under certain service conditions they have to work into antennas having tremendous ranges of variation of all their constants and, indeed, in many cases such transmitters have to work into antennas which are not tuned to resonance. The dummy antenna method, therefore, has found little favor with engineers desiring to study transmitter performance on frequencies higher than 2000 kc.

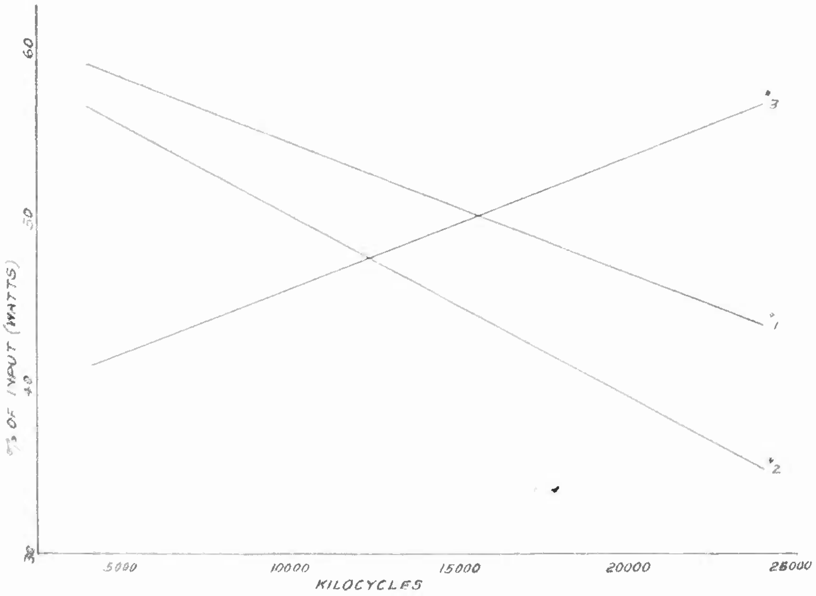

Fig. 1—These curves are only approximate for intermediate points between 4000 and 24,000 kc.
Curve No. 1—(W) radio frequency generated.
Curve No. 2—(W) radio frequency in antenna.
Curve No. 3—(W) plate dissipation.

With the advent of water-cooled tubes a new avenue of approach was opened up. By measuring the amount of heat carried off into the cooling water the dissipation of energy at the plate of the tube could be determined and by subtracting this from the plate power input it was at least possible to say how much of the input power was converted into radio frequency. It was still difficult to say how much of this power actually left the transmitter terminals and went into the antenna system with its associated feed lines. The determination of the total radio-frequency energy developed by a power amplifier was, however, a distinct gain and some years ago the Naval Research Laboratory undertook experiments looking to the extension of this information to air-cooled tubes. It is understood that several other laboratories made similar experiments. One would naturally think of using the optical pyrometer to determine the temperature of the plate of the tube but modern tubes are ordinarily not designed to operate at a high enough plate temperature to give efficient action of the optical pyrometer.

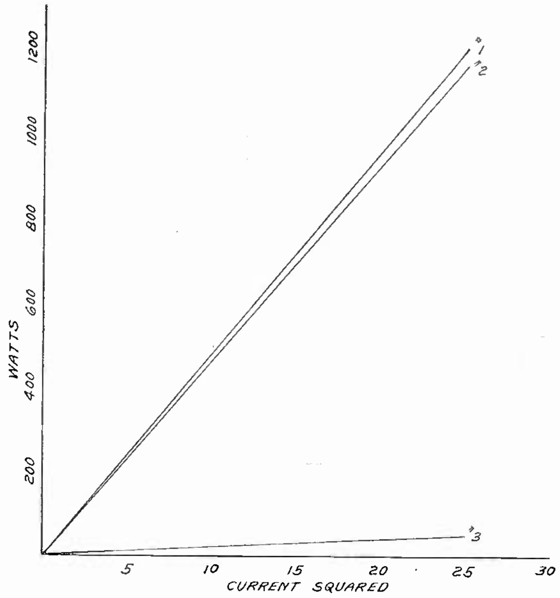

Fig. 2—Distribution of losses at 4017 kc—12-turn coil.
59 per cent of total (W) = radio frequency generated
56.6 per cent of total (W) = radio frequency in antenna
Without antenna
Curve No. 1—19.5 per cent of input watts generated into radio frequency = 100 per cent r-f loss.
Curve No. 2—18.58 per cent of input watts generated into radio frequency and lost in coil = 95.25 per cent r-f loss in coil.
Curve No. 3—0.92 per cent of input watts generated into radio frequency and lost in condenser, shield, etc. (stray loss) = 4.75 per cent r-f loss.

The experiments at the Naval Research Laboratory resulted finally in the use of a small and not very sensitive thermocouple which is strapped on to the wall of the tube to be tested forming a metallically shielded blister which, attached to a suitable galvanometer, will give an indication depending on the temperature of the interior of the tube; namely, the plate. This device could be calibrated by dissipating known amounts of plate power without oscillations developing, that is, with the oscillating circuit open. This permitted plotting of the curve of dissipated plate power against the reading of the indicating instrument connected to the thermocouple. Naturally considerable care had to be taken to shield not only the thermocouple but associated leads from the influence of high-frequency current, but when properly carried out the procedure, although somewhat laborious, gave consistent and fairly satisfactory results. This method is quite comparable to a method used with water-cooled tubes and both can be extended so as to measure

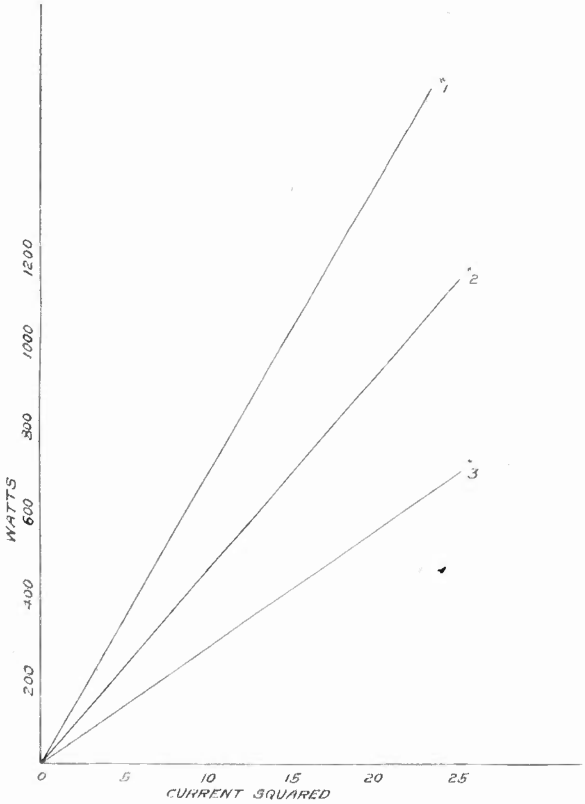

Fig. 3—Distribution of losses at 8035 kc—8-turn coil.
60.8 per cent of total (W) = radio frequency generated
56.6 per cent of total (W) = radio frequency in antenna
Without antenna
Curve No. 1—25.3 per cent of input watts generated into radio frequency = 100 per cent r-f loss.
Curve No. 2—15.85 per cent of input watts generated into radio frequency and lost in coil = 62.6 per cent r-f loss in coil.
Curve No. 3—9.45 per cent of input watts generated into radio frequency and lost in condenser, shield, etc. (stray loss) = 37.4 per cent r-f loss.

quite approximately the radio-frequency power in any antenna system regardless of the nature of its impedance.

It seems certain that the method herein described, while perhaps not essentially new in principle, will be of interest in so far as the practical results with a transmitter of modern design are concerned, and may lead to improvements in future transmitter design and to the introduction of instruments on the transmitter panel from which the antenna power can be readily determined.

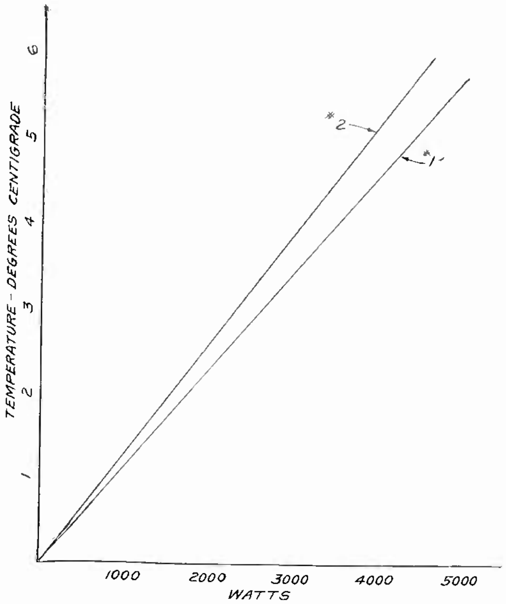

Fig. 4—Temperature rise versus plate dissipation. Nonoscillating condition.

	Plate dissipation
Curve No. 1—Tube No. 1	850 W. per 1 degree C
Curve No. 2—Tube No. 2	765 W. per 1 degree C

The method used depends briefly upon the following: Assuming that the antenna system is coupled in some way to the output circuit of the last power amplifier in the transmitter, it is obvious that in order to transfer the power into the antenna a certain amount of current has to be set up in the main power amplifier circuit, which we shall call the tank circuit. A meter, either directly in this circuit or coupled to this circuit, can be used to give an indication of the amount of current in this tank circuit. It is not necessary to have the absolute value of the current but only to have an indication which will be proportional to it. If we find under normal operation what appears to be a most satisfactory antenna coupling this tank circuit meter reads a certain amount and then if we disconnect the antenna and reëstablish this same current in the tank, obviously with the aid of much less power, we have at hand a method of determining the actual amount of power in the antenna system. For instance, if with 10-kw plate power input we find 4 kw dissipated by the plates of the tubes during normal operation we know that 6 kw have been converted into radio-frequency power. During this normal operation our tank circuit meter has indicated a certain amount. Upon disconnecting the antenna we are able to establish the same amount in the tank circuit by the application of 1-kw plate power input, finding at the same time that 500 watts are dissipated in the plate. Obviously then, the other 500 watts was required to maintain the tank current in question, which means that under normal operation with 10-kw input, 6 kw are converted into radio frequency and 5.5-kw are actually put into the antenna.

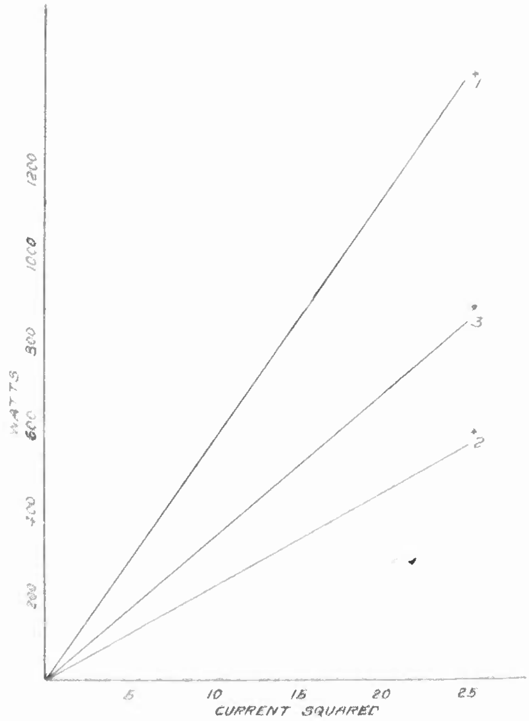

Fig. 5—Distribution of losses at 16,000 kc—6-turn coil.
C' set at minimum
52.5 per cent of total (W) = radio frequency generated
48.2 per cent of total (W) = radio frequency in antenna
Without antenna
Curve No. 1—26.4 per cent of input watts generated into radio frequency = 100 per cent r-f loss.
Curve No. 2—10.5 per cent of input watts generated into radio frequency and lost in coil = 39.8 per cent r-f loss in coil.
Curve No. 3—15.9 per cent of input watts generated into radio frequency and lost in condenser, shield, etc. (stray loss) = 60.2 per cent r-f loss.

The figures actually given would hold for a good design of transmitter in the 4000–6000-kc band used for telegraphic purposes only. The method is obviously subject to one error but a careful study of this error has shown that under practical operating conditions it is not large enough to invalidate the method. This error is introduced by the fact that the removal of the antenna generally necessitates a slight retuning of the tank circuit to keep it in resonance. Unless this retuning is a very large amount indeed (which is ordinarily not the case) the error does not seem to be in excess of 1 per cent or 2 per cent. It is possible, after having determined the percentage of plate power which has arrived in the antenna and the amount that has been dissipated within the set itself as radio-frequency losses, to segregate somewhat these losses if the transmitter is of the type using a water-cooled output coil system. The rate of flow and temperature change in the water through the tank coil system can be checked and will give a determination of the radio-frequency losses within the coil. The remaining losses may be ascribed to faulty insulators, condensers, and to eddy currents induced in the shielding or the framework of the set.

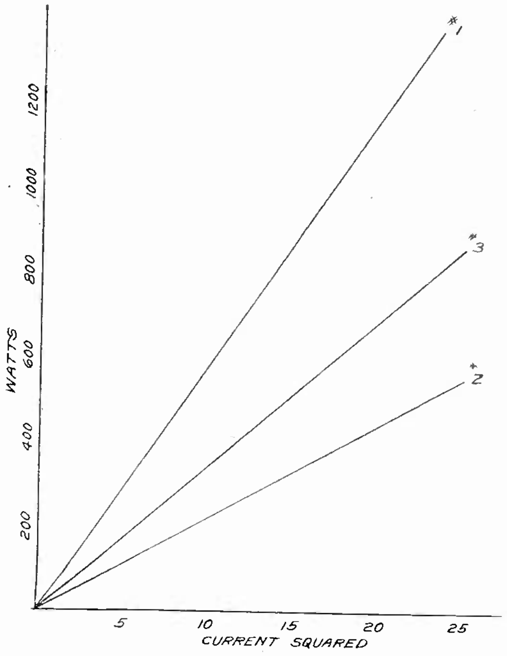

Fig. 6—Distribution of losses at 16,000 kc—2-turn coil. *C* set at maximum

50 per cent of total (W) = radio frequency generated
40.25 per cent of total (W) = radio frequency in antenna
Without antenna
Curve No. 1—42.8 per cent of input watts generated into radio frequency = 100 per cent r-f loss.
Curve No. 2—16.7 per cent of input watts generated into radio frequency and lost in coil = 39 per cent r-f loss in coil.
Curve No. 3—26.1 per cent of input watts generated into radio frequency and lost in condenser, shield, etc. (stray loss) = 61 per cent r-f loss.

It might be well to point out at this time that the most desirable tank circuit would be one that offered a very high impedance to the plate circuit and with the antenna so coupled to it that the load would probably match the impedance of the tubes. The tank circuit losses would be reduced to minimum provided the insulation of the system was fairly adequate. This type of circuit design is easy enough in a transmitter of a single frequency but where the transmitter has to be continuously variable over a very wide range it is, from a design point of view, easier to use a variable condenser in parallel with the output inductance of the tank circuit to extend the tuning range of the same. However, much can be done with a continuously variable coil system, especially in those cases where the power is not so high that it is necessary to resort to water-cooling of the tank circuit.

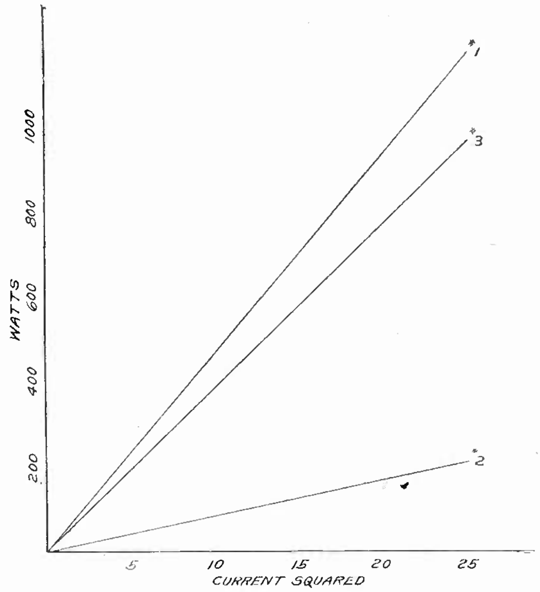

Fig. 7—Distribution of losses at 24,102 kc—2-turn coil.
43.5 per cent of total (W) = radio frequency generated
35 per cent of total (W) = radio frequency in antenna
Without antenna
Curve No. 1—33.3 per cent of input watts generated into r-f loss.
Curve No. 2—5.8 per cent of watts generated into radio frequency and lost in coil = 17.4 per cent r-f loss in coil.
Curve No. 3—27.5 per cent input watts generated into radio frequency and lost in condenser, shield, etc. (stray loss) = 82.6 per cent r-f loss.

This report covers the performance of a transmitter of fairly modern design which covers the frequency range from 4000 to 26,000 kc. The transmitter can be used either with or without crystal control and consists of the following units:

(a) Master oscillator circuit used with or without crystal control, containing one 50-watt tube operated at low power. An alternative arrangement uses a $7\frac{1}{2}$-watt tube for this stage.

(b) A 75-watt shield-grid power amplifier which may be used as a frequency multiplier also.

(c) A third stage contaning two 75-watt tubes (shield grid) used as amplifiers or frequency multipliers.

(d) Two 500-watt tubes, (shield grids) tunable up to 30,000 kc as amplifiers without frequency multiplication.

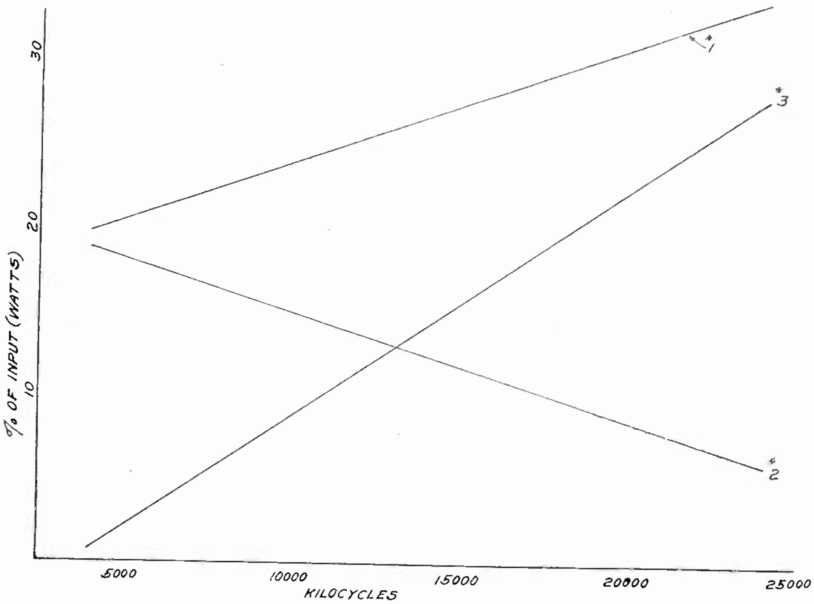

Fig. 8—Coil and stray losses versus frequency, without antenna.
These curves show an average value of r-f loss with increase of frequency (without antenna).
Curve No. 1—per cent of radio frequency generated without antenna.
Curve No. 2—per cent of radio frequency generated lost in coil.
Curve No. 3—per cent of radio frequency generated (stray loss).

(e) A high power stage containing two water-cooled 10-kw tubes arranged push-pull, tunable from 4000 to 26,000 kc by means of a set of changeable ouput coils operated in parallel with a plate circuit tuning condenser running from a minimum capacity of 125 $\mu\mu$f to a maximum capacity of 284 $\mu\mu$f.

In order to insure overlap five different coil systems are employed in the output stage. These are made of copper tubing which can be water-cooled. Due to the changing L/C ratio resulting from this method of tuning the output circuit, the efficiency with respect to power delivered to the antenna will not follow a regular curve over the entire frequency band. In general, the tendency of course is for the losses to increase with the higher frequencies, with a corresponding reduction in percentage of power put into the antenna system, but certain exceptions in this general order of results may occur due to the fact that it is, for instance, possible to get a better efficiency at, say 16,000 kc than at 12,000 kc due to the fact that the 16,000 kc happens to be put out with a more favorable L/C ratio. The results, however, indicated that if the L/C ratio could be maintained we might expect a linear decrease of circuit efficiency with increasing frequency. The same would hold true of the percentage of power put into the antenna.

This particular transmitter was designed to operate its last stage from a d-c generator delivering 10-kw at 7500 volts and most of the efficiencies were therefore determined at approximately 10-kw input in

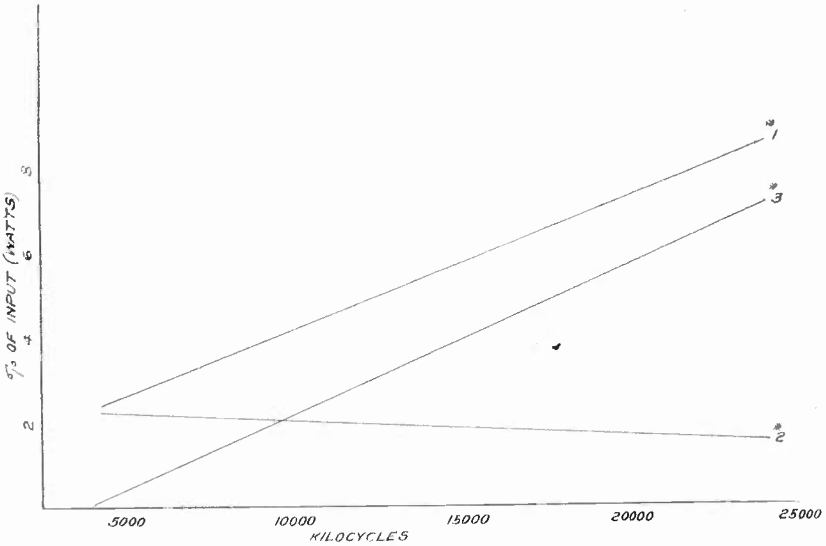

Fig. 9—Distribution of losses, antenna converted.
These curves show an average value of r-f loss with increase in frequency (with antenna).
Curve No. 1—per cent of total (W) lost in r-f circuit.
Curve No. 2—per cent of total (W) lost in r-f output coil.
Curve No. 3—per cent of total (W) lost in r-f stray loss.

the last stage. The antenna system was of the doublet type fed through a trunk 35 feet long and 8 inches in diameter. This antenna system was so adjusted as to absorb the power satisfactorily at any point in the frequency range of 4000 to 26,000 kc. The transmitter was coupled to this antenna system through a pair of variable series antenna condensers, one in each leg of the line. These condensers were connected directly one to the plate of each of the high power push-pull amplifiers excepting that for frequencies in the neighborhood of 4000 kc the condensers were connected into taps on the coil systems, two turns each side of the center tap.

The measurements of input power were made with direct-current voltmeters and ammeters, an ammeter being placed in series with each tube so that the power level in the two tubes could be kept in fairly close agreement as it was found that the highest efficiency could be obtained when the two tubes were drawing equal power. Under these conditions, of course, the transmitter could be worked to maximum power far in excess of the rated power of this particular transmitter. A thermometer calibrated to 0.1 of 1 degree C was used in measuring the input and output temperatures of the water-cooling system, as well as the input and output temperatures of the water circulated through the output coil system. All thermometers were so placed as to be at the same potential as the frame of the set both for direct current and high frequency. To correct for possible errors in heat radiation before the water reaches the thermometers in passing through the hose systems, the input water temperature was raised through the entire working range of the thermometers and the output thermometers checked against the input temperature so that the three output thermometers were thus calibrated in terms of the input termperatures. This procedure also prevented creeping of temperatures during the process of taking readings. Only one set of readings was usually necessary. Knowing the normal position of the output thermometers with a given input water temperature and then reading the temperature with power on, it was easy to find the rise in temperature by a simple subtraction. A water pressure gage was used in the input water circuit to make sure that the water remained constant during the period of calibration and measurement. The flow through the tube was not measured but plate power without oscillations was developed up to normal maximum dissipation and the corresponding thermometer readings were taken, it being assumed that if the pressure remained constant the flow would remain steady.

In the case of the determination of the radio-frequency circuit losses in the output coil, a radio-frequency ammeter with a small pick-up loop was coupled in a fixed position to the output circuit. By taking a series of readings at different watts input without the antenna coupled to the circuit, data were obtained for a curve which showed a linear relationship between the square of this meter reading and the total radio-frequency losses in the output circuit. With the antenna disconnected it was also possible to calibrate the thermometers in the discharge water of the radio-frequency output coil, which in turn can be checked by measuring the rate of water flow in this system.

The preliminary calibration without oscillations is made at a fairly high plate voltage by lowering the negative *C* until the desired amount of plate power is dissipated. Three points will generally determine the straight line relation between temperature differences for outlet and inlet water and power lost in plates. The transmitter is then tuned to a given frequency with the radio-frequency ammeter loosely coupled to the plate circuit in a fixed position. This ammeter had to be calibrated, of course, whenever the frequency was changed over a wide range, especially when coil systems were shifted. It is not desired to calibrate this ammeter to read directly the current of the coil system but rather as a check on the energy of the coil system for it is actually calibrated in terms of the energy necessary to maintain a given reading. The procedure then is to tune up the transmitter for normal operation to what appears to be the best coupling to the antenna system. Note plate input power. Subtracting the power indicated by the temperature of the cooling water we have left the total radio frequency generated. Note the reading of the radio-frequency ammeter loosely coupled to the *tank* and reëstablish this same reading with a lesser amount of applied power or whatever may be necessary when the antenna is disconnected. Of course, it is not necessary to establish exactly the same reading since our calibration has already given us a curve connecting the square of this meter reading with the circuit losses. The circuit loss is thus determined by subtracting from the total of radio frequency, giving us a remainder which has left the terminals of the transmitter and gone into the antenna system. It is, of course, necessary to note the plate power dissipated during the operation with antenna removed because the efficiency of the transmitter will change during this operation. This change makes no difference in the results because due allowance is made for it.

The transmitter losses may be further separated by determining from the temperatures of the input and outlet water for the coil system together with its rate of flow, how much energy has been dissipated in the main tuning inductance. In the particular arrangement used here the rise in temperature of 1 degree produced by the heat in the coil system itself meant a loss of 29 watts. Since it was thought that there might be some error in this particular measurement due to heat radiation from the coil itself, a check was made upon the coil losses by supplying a heavy direct current at low voltage to the coil system. This measurement, however, gave a very satisfactory check showing that the 29 watts per degree was substantially correct. After having run the check on the efficiency of the type described, it was then practicable to make other runs with various degrees of coupling since the percentage of power into the antenna is quite markedly affected by the coupling. It was generally found that if the set were tuned up originally with that degree of coupling which made the set operate with good stability and with normal power input without crowding the plate voltage up or the negative *C* voltage down, the coupling thus instinctively chosen was pretty close to the correct coupling for optimum energy into the antenna.

The results of the measurements are shown in the appended table which is the principal point of interest in this report, as showing what may be reasonably expected in different frequency bands with a reasonably good design. It will be noted that the total radio frequency generated drops from 59 per cent of the input power at 4000 kc to $43\frac{1}{2}$ per cent at 24,000 kc. The percentage of the input power actually delivered to the antenna varies from 56.6 per cent at 4000 kc to 35 per cent at 24,000 kc. The total radio-frequency losses which involve losses in the coil system, in insulators and condensers and losses due to eddy currents in the shields and framework of the set vary from 2.4 per cent of the input power at 4000 kc to 12.75 per cent at 12,000 kc. The irregularity in this last figure is plainly due to changing *L/C* ratios as this point is carefully covered by separate experiments. The general trend for approximately constant *L/C* ratios is for a steady increase in the total loss as the frequency is raised. The irregularity in coil losses ranging from 1.48 per cent to 6.69 per cent is traceable to the same source; namely, variable *L/C* ratios but in general it may be said that the coil losses alone do not by any means show a general tendency to increase with frequency. On the other hand, the stray losses do show a strong tendency to increase with the frequency. The particular reading at 12,000 kc had to be taken with considerable tuning condenser in use. This naturally meant a rather large circulating current which gave excessive coil loss and also excessive stray loss.

TABLE I

Frequency	Per Cent of Radio Frequency					Per Cent of Direct Current	
	Generated	Antenna	Total Loss	Coil Loss	Stray Loss	Plate Loss	Size of Coil
4000	59%	56.6%	2.4%	2.28%	0.114%	41%	12 turns
8000	60%	56.6%	4.2%	2.63%	1.57%	39.2%	8 turns
12000	55%	42.25%	12.75%	6.69%	6.06%	45%	6 turns
16000	52.5%	48.25%	4.25%	1.69%	2.56%	47.5%	6 turns
16000	50%	40.25%	9.75%	3.8%	5.95%	50%	2 turns
24000	43.5%	35%	8.5%	1.48%	7.02%	56.5%	2 turns

A perusal of this table indicates that one should safely be able to specify delivery of power into the antenna on frequencies below 8000 kc to the extent of 50 per cent or better. Between 8000–16,000 kc one should be able to specify a delivery into the antenna of 40 per cent or better, while at 24,000 kc it would probably be unsafe to require more than 32 per cent in the antenna.

It will be noted that there are two sets of readings in the table for 16,000 kc for two coils of the same diameter, one of which had 6 turns and the other 2. It so happens that the tuning condenser in the high power stage reached 16,000 kc for one of these coils at approximately maximum and for the other at approximately minimum. It is interesting to compare the efficiencies for these two cases. The 6-turn coil shows a net gain of 8 per cent increase in antenna energy over the 2-turn coil. The total losses in the set are $9\frac{3}{4}$ per cent for the 2-turn coil and $4\frac{1}{4}$ per cent for the 6-turn coil. Both coil losses and stray losses are more than double for the 2-turn coil than what they are for the 6-turn coil. Without extra high grade insulation this would hardly be possible but every indication obtained from these tests indicates extremely low losses in insulation, which throughout the set consisted of the very best grade of insulators, wherever these insulators were in a strong radio-frequency field. The indications are that by dropping off the tuning condenser altogether and making the tuning adjustment by means of variable inductance, the efficiencies would be still higher. However, it is believed that for general performance of a set of this character, flexibility and rapidity of adjustment in covering such a wide range of frequencies, the efficiencies are all that can be expected. After all, even at 24,000 kc the total losses are not exceptionally high. Failure to produce higher efficiency in the upper bands must be due to the properties of the tubes themselves which are admittedly not ideal for the upper frequencies. The values of negative C used during these tests were fairly high, running generally between 850 and 1100 volts. The plate voltage usually ranges between 6000 and 7500. We may conclude then by expressing the opinion that a moderate increase in efficiency is to be expected over the entire frequency scale by the use of high L/C ratio but that any large increase in efficiency will depend upon the use of improved tubes better adapted to these upper frequencies.

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A THERMIONIC TYPE FREQUENCY METER FOR USE UP TO 15 KC*

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Summary—*A new type of frequency meter is described which is adapted to the measurement of low and intermediate frequencies. The instrument absorbs a negligible amount of power from the circuit being tested, has a linear calibration curve and a sensitivity of about eight microamperes for one per cent change in frequency. An experimental model is described in detail. The input to this model is between five and ten volts. Other models may require different inputs.*

Introduction

The methods now available for the measurement of frequencies within the audible range are few and inconvenient. The two in most general use are (1) comparison with a calibrated oscillator and (2) some sort of an alternating-current bridge arrangement. The first requires a calibrated oscillator and the second has limitations usual to bridge arrangements, namely, the detection of balance becomes difficult at high and low audio frequencies and, secondly, appreciable power must, in general, be absorbed from the circuit being tested in order to obtain adequate sensitivity. The purpose of this paper is to describe a new type of frequency meter which is convenient to operate and circumvents the obstacles listed above.

Theory

In a recent Institute paper[1] a phase-shifting or inverted bridge was described which has been found to be useful for the present purpose also. This is essentially a reactance bridge in which the positions of two adjacent, unlike arms are interchanged. In the previous application use was made of the phase-shifting characteristics of this bridge; in the present application use is made of the fact that at one particular frequency the voltage is the same across all arms, and equal to $0.7E$, where E is the voltage applied to the bridge. If points A and B, Fig. 1, are connected to the grids of two thermionic tubes when this condition holds, and the tubes are adjusted to operate as a balanced detector with C connected to the common grid return, then a d-c microammeter

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* Decimal classification: R211.1. Original manuscript received by the Institute, January 28, 1931. Revised manuscript received by the Institute, March 20, 1931.

[1] Turner and McNamara, Proc. I.R.E., 18, 1746; October, 1930.

shunted across the plates will read zero. Under all conditions other than the critical one ($f=1/2\pi RC$) the instrument reading will depart from zero, in one direction when the frequency is below the critical value and in the opposite direction when it is above. Consequently the 'zero' reading is preferably adjusted to some convenient point on the instrument scale and this arbitrary 'zero' will be indicated by the use of quotation marks in what follows.

The above discussion is based on the presumption that matched tubes are used. To overcome this limitation the 'zero' is chosen with normal voltage applied to the bridge and the two grids connected to the point O, Fig. 3. The two resistors across the input circuit are so chosen that the voltage from O to C is $0.7E$. Hence the 'zero' is set with the same voltage applied to the two grids as will be applied at the critical frequency. This makes the device essentially a null instrument with its well-known advantages. Moreover, with reasonably similar tubes, slight variations in the input voltage should cause a negligible change in the 'zero' or critical settings.

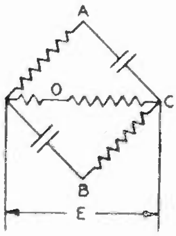

Fig. 1—Inverted bridge.

Analysis shows that in such a circuit the change in plate current for a departure from the critical frequency is given by

$$I=K\ \Delta f/f \text{ (approximately).}$$

This indicates that the meter reading is directly proportional to the percentage change in frequency and that the departure from 'zero' is linear, or the sensitivity constant, for small changes in frequency. Both of these points are illustrated by Fig. 2 while Fig. 3 shows a schematic diagram of the arrangement used. In obtaining the data for Fig. 2 the 'zero' current was set at twenty-one microamperes by adjusting the slider on the voltage divider in the plate circuit. This adjustment is made with both grids tied to the point O, Fig. 3. The two grids were then transferred to points A and B, respectively, as shown, and the oscillator frequency varied on both sides of the critical value. When the frequency was below the critical value the current decreased and when it was above the current increased. To bring out the symmetry the decreases below the 'zero' current have been reversed in sign and are shown by the full line curve for values below 1000 cycles per second. It will be noted that the sensitivity is about eight microamperes for one per cent change in frequency, this being a function of the voltage applied to the bridge. Fig. 4 shows the variation of sensitivity with various values of E which lie within the allowable range of the 201-A tubes used in this experiment. It is apparent that this sensitivity constant makes it possible to evaluate the departure of the test frequency from the critical value. The instrument is calibrated for a given value of input which may be indicated by a milliammeter in the common plate lead if this is desirable. It has been found that with approximately bal-

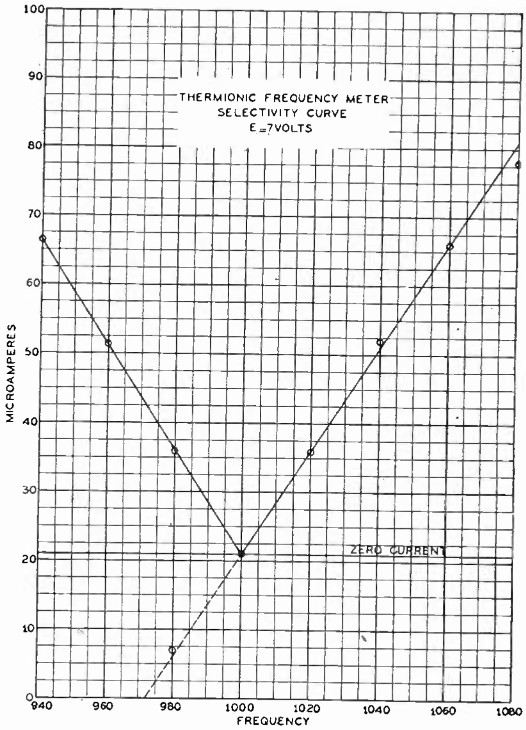

Fig. 2—Selectivity curve.

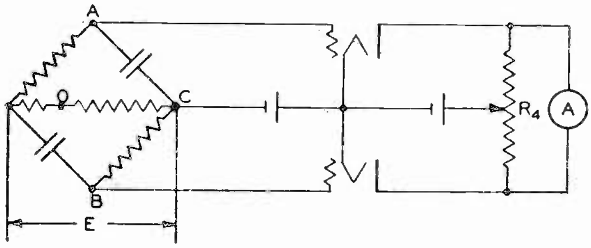

Fig. 3—Schematic diagram.

anced tubes a 15 to 20 per cent deviation from this test value produces the same deflection as a 1 per cent change in frequency.

It has been shown that the bridge is balanced when $f = 1/2\pi RC$. In order to have a linear calibration curve it is necessary that either R or C vary so that its reciprocal will be proportional to the angular displacement of the dial, the other remaining constant. The latter expedient was the simpler in our case, therefore, a pair of old s.l.f. condensers were cut down[2] to make them s.l.x. (i.e., having reactance at a given frequency directly proportional to the dial setting). This decreased the value of capacitance so that it was necessary to use large values of resistance to obtain a balance at the low frequencies. The resistors used were simple grid leaks. These resistors are apparently hygroscopic because an error of as much as 5 per cent was obtained on damp days; when wire-wound resistors were substituted in the same holders no such effect was observed. Analysis shows that if the ratio $C_{\max}/C_{\min} = M$; and n pairs of resistors are used, it should be possible to cover a frequency range of M^n. Thus, if $M = 10$ and $n = 3$ the total range might be from 50 to 50,000 cycles per second. It is apparent that

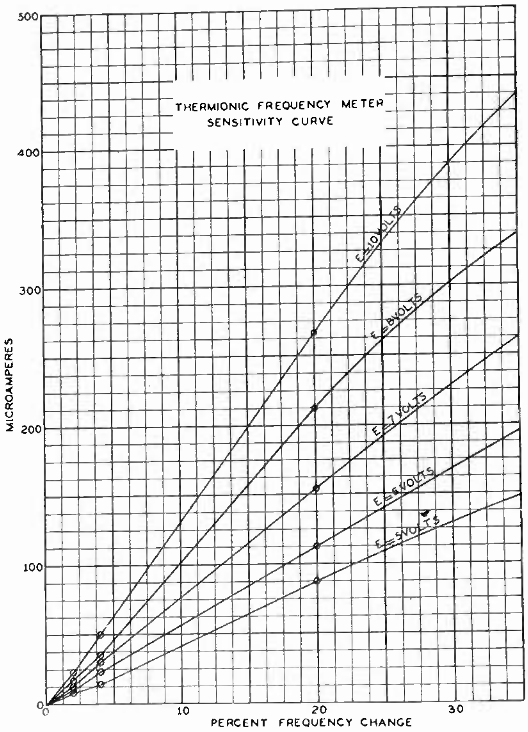

Fig. 4—Sensitivity curve.

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[2] O.C. Roos, "Simplified s.l.f. and s.l.w. design," Proc. I.R.E., December, 1926.

if the resistors chosen are equal and the pairs are exact multiples of each other a single calibration curve with appropriate multipliers may be used throughout.

EXPERIMENTAL MODEL

Fig. 5 is a photograph of an experimental model for which the calibration curves are shown in Fig. 6 and the complete wiring diagram in Fig. 7. The condensers used in this experimental model had a value of *M* of about seven but to allow for considerable overlapping from one frequency range to the next the effective range was made about five to one. When using four pairs of resistors with nominal values of 10, 5, 1, and 0.25 megohms the meter covered the range from 60 to 14,000 cycles per second, which is the range in which we were particularly interested. Other models have been operated up to 50,000 cycles per second. The plate meter is a Weston, Model 301, having a full scale reading of 500 microamperes; it is planned to replace this meter with a similar one having a full scale reading of 200 microamperes. The two binding posts on the upper left of Fig. 5 are normally short-circuited but may be used to insert a more sensitive meter such as the Weston, Model 322. Switching is accomplished by means of an anticapacity switch. The plate voltage is removed while switching since the grid circuits are open during this interval. The middle blade of the 3-pole switch effects this.

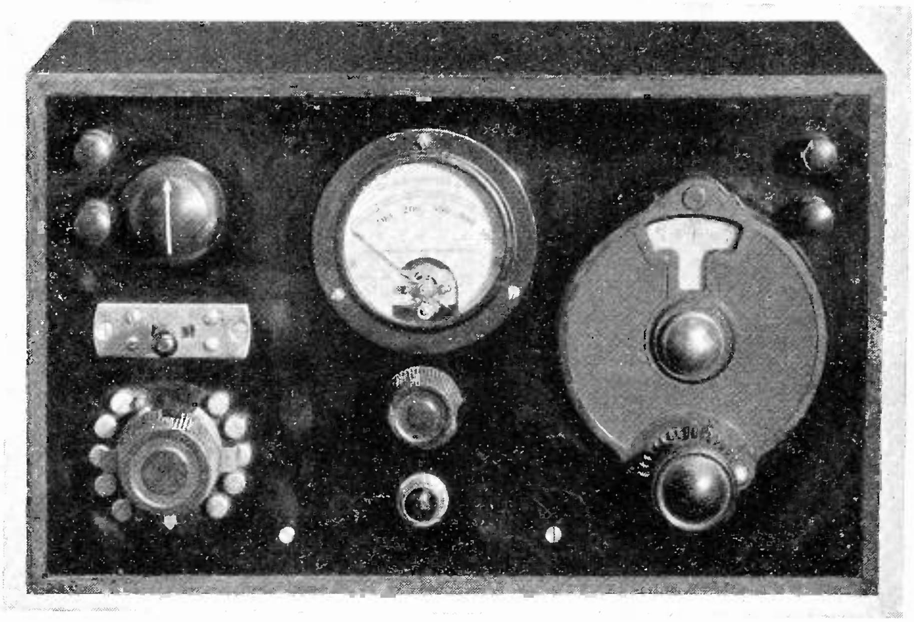

Fig. 5

CONCLUSION

As indicated above variations in voltage have only a small effect on the balance reading; similarly, harmonics in the supply line, since they combine as if in quadrature with the fundamental and with each other, produce little effect; their effect may be eliminated by using a reversing switch so that each tube is connected alternately across a resistance and a capacitance and taking the mean of the readings found under these two conditions. The 'zero' must, of course, be set at the value at which the meter is calibrated and good quality batteries must be em-

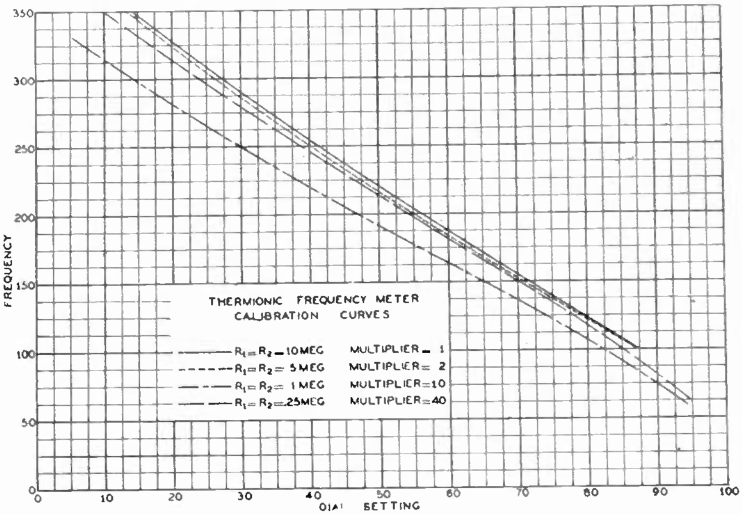

Fig. 6—Calibration curves.

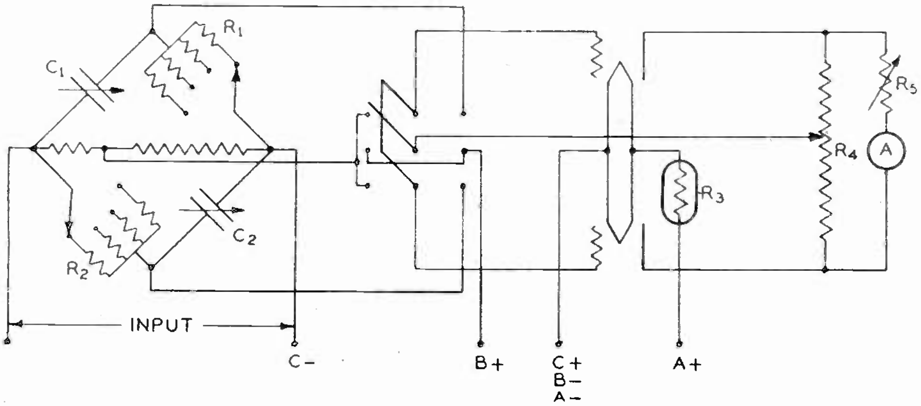

Fig. 7—Complete wiring diagram:
R_1 and R_2 = bridge resistors.
R_3 = amperite.
R_4 = 2000 ohms.
R_5 = 50,000 ohms.

ployed so that the operating point will not shift and so change the rectification characteristic. The 60-cycle point makes a check calibration relatively simple.

In conclusion it may be pointed out that this instrument may be used as a detector of frequency modulated waves if the carrier frequency is not too high.

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PERFORMANCE OF OUTPUT PENTODES*

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Summary—*The comparison of power output, distortion, power sensitivity, and a-c/d-c power economy of a group of experimental pentodes is made with corresponding triodes. The apparatus and method of measuring are described.*

The pentodes' a-c/d-c economy and power sensitivity is considerably higher than that of the corresponding triodes. The harmonic distortion is found to be generally worse with the pentodes. The variation in power output with changes in load resistance, arbitrarily called "output distortion," is shown to be about the same for both classes of tubes.

The need of a large capacity shunting the bias resistor in a self-biased pentode amplifier is shown. Its effect on power output and power sensitivity is discussed.

In the conclusion, five types of distortion occurring in triode and pentode operation are compared. The principal use for the pentode appears to be with battery and 110-volt d-c types of receivers.

MUCH has already been published on the relative merits of various pentodes singly but no attempt to correlate the various results has come to the attention of the author. Samples were received from various manufacturers and measurements undertaken to make such a comparison. It is hoped that the results obtained will be of assistance to those debating the merits of the triode and pentode, and will aid in obtaining satisfactory operation from pentodes now available.

FACTORS TO BE DETERMINED

In determining output performance, four principal factors are to be considered, namely; power output, power sensitivity, a-c/d-c economy, and fidelity. The first three mentioned are relatively easy to determine and require little time. However, the fourth, fidelity, requires a determination of the harmonic content and interelectrode capacity. Both of these require special procedure and apparatus compared with the other measurements. For simplicity only pure resistive loads are considered in this investigation.

The power sensitivity, as defined by Ballantine and Cobb,[1] has been obtained by dividing the square root of the output watts by the

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* Decimal classification: R335×R262. Original manuscript received by the Institute October 6, 1930. Revised manuscript received by the Institute March 30, 1931.

[1] S. Ballantine and H. L. Cobb, "Power output characteristics of the pentode," PROC. I.R.E., 18, 452; March, 1930.

root-mean-square input voltage. The power sensitivity in this case is proportional to sound pressure and allows a direct comparison between tubes.

The a-c/d-c economy has been taken as the ratio of a-c output to d-c input watts. Filament power has not been included because of the ease with which it may be obtained in most sets.

It is also desirable to consider how the power output varies with load resistance, since with most available loud speakers the load impedance is a function of frequency. In order to compare this variation, a procedure has been developed in which an arbitrary factor of "output distortion" is calculated. This factor is determined by dividing the minimum by the maximum load impedance at which eighty per cent of the maximum power output is obtained. The output is most constant when this factor is smallest.

METHODS AND APPARATUS

The following are some of the methods available for harmonic analysis of a complex electric wave such as that found in the output of a vacuum tube.

(*a*) *Graphical and mathematical method.*

This method requires a determination of the static characteristics, from which a cycle of the output wave is plotted and then analyzed by means of Fourier's analysis. This is an accurate method and can be adapted to various conditions of load, operating voltage, and input, but is quite tedious.

(*b*) *Tuned filter method.*

This method ordinarily consists in successively reading the harmonic voltage through a tuned filter which passes the frequency being determined. It requires an elaborate filter system and generally affects the output circuit of the tube under measurement, introducing different impedances to the various harmonic frequencies. A modification of this method was used by Ballantine and Cobb[2] in which the harmonics were separated from the fundamental and the total harmonic content determined.

(*c*) *Oscillograph method and analysis of resultant curve.*

This method does not adapt itself to the measurement of small harmonics and is quite tedious, in addition to requiring elaborate and expensive equipment as well as a photographic processing set-up.

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[2] S. Ballantine and H. L. Cobb, "Power output characteristics of the pentode," PROC. I.R.E., **18**, 461; March, 1930.

(*d*) *Method of beats.*

In this method[3] the sum of a portion of the output voltage and the voltage from a local oscillator are impressed on the input of a vacuum tube voltmeter which reads true root-mean-square voltages. By adjusting the local oscillator to within a few beats per minute of the harmonic frequency, the voltmeter follows the amplitude of the beat. From the limits of the swing the actual magnitude of the harmonic may be calculated. This method is very sensitive and harmonics as low in amplitude as 0.1 per cent of the total voltage may be detected, although it is rather a slow procedure where many determinations are to be made.

After a serious consideration of the various methods, the last one, the method of beats, was decided upon as being best suited to our requirements. It is faster than the graphical method and possesses all the sensitivity and accuracy required. In addition it has great flexibility as to measuring the harmonics of a source of almost any fundamental audio frequency and does not introduce any appreciable load on the circuit under measurement.

Some difficulty was experienced in obtaining a tube for the voltmeter which operated along a true quadratic characteristic but a very close approximation was finally had by using a 171-A type with low B voltage and keeping the input below four volts peak value. In order to read small amounts of second and third harmonics it was necessary to keep the complex input voltage low with respect to that from the local oscillator, a ratio of four to one being actually used. It was not necessary to have a local oscillator of absolutely pure output. A General Radio Type 377 low-frequency oscillator was used without any filter and gave less than 5 per cent of any harmonic. This oscillator had a frequency range from 60 to 100,000 cycles per second.

In making all of the dynamic measurements it was necessary to have a source of constant frequency containing less than 0.2 per cent of any harmonic and giving over 35 volts output. To meet these requirements a vacuum tube oscillator was built containing one 112-A type tube as a 400 cycle oscillator and another of the same type as an amplifier. The output of the amplifier passed through a low-pass filter of three "T" sections to an attenuator feeding the grid circuit of the triode or pentode whose performance was being determined.

The output load was coupled to the tube under test by a choke-condenser combination of circuit. The load was composed of two resistance

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[3] C. G. Suits, "Harmonic analysis of electrical waves," PROC. I.R.E., **18**, 178; January, 1930.

boxes in series, with the portion of the output to be analyzed being taken from one box, in this way forming an attenuator for the voltmeter. By means of a switching arrangement the same vacuum tube voltmeter was used to measure input, output, and harmonics.

In measuring the interelectrode tube capacities a rough method was used which adapted itself to the equipment on hand. Two elements in the triodes were connected together and the capacity measured to the third element. This gave the sum of each pair of interelectrode capacities, from which it was possible to calculate the individual capacities. With the pentodes, all grids except the control grid were connected to the cathode and the tubes treated as triodes.

In determining performance, the maximum input to the tube under test was arrived at by gradually increasing the a-c input until the grid current reached zero, as indicated on a d-c microammeter. Approximately at this point the electron grid current starts to flow, and if appreciable would cause much rectification and distortion when used with a high impedance input circuit. A very complete article on grid current flow has been published by Nottingham.[4]

EXPERIMENTAL RESULTS

In the material which follows data are presented which show the performance of a series of pentodes and the corresponding triodes.

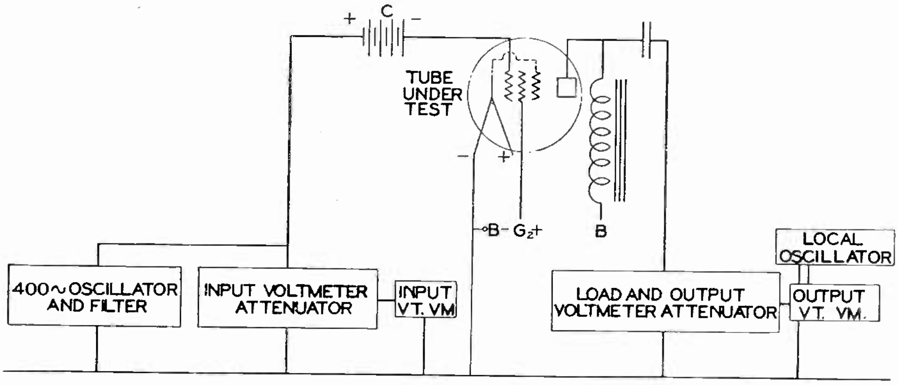

Fig. 1—Schematic diagram of circuit used in determining performance and distortion of triode and pentode tubes.

The samples used in obtaining the pentode data were picked from lots of one to four tubes of each manufacturer. For the triode data, tubes were chosen which had the published values of plate current and mutual conductance. One tube of each type was measured. Fig. 1 shows the schematic circuit arrangement.

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[4] W. B. Nottingham, "Measurement of small d.c. potentials," *Jour. Frank. Inst.*, **209**, 290; March, 1930.

The three tables which follow contain the results of the actual measurements made on the various tubes. Table I contains the d-c data for no input. The UX-112-A, pentode No. 5, and Mullard Pentode PM-24-A were measured at two conditions of d-c voltage.

TABLE I

D-C OPERATING CONDITIONS

TUBE	VOLTAGES				CURRENTS		
	Filament	Control Grid	Accelerating Grid	Plate	Filament	Accelerating Grid	Plate
1. UX-112-A	5	−9	—	135	0.25	—	6
2. UX-112-A	5	−13.5	—	180	0.25	—	7.5
3. UX-171-A	5	−40.5	—	180	0.25	—	19
4. UX-245	2.5	−50	—	250	1.5	—	32
5. Pentode 1	5	−9	135	135	0.25	1.5	6.5
6. Pentode 2	5	−16	200	300	2.0	5	29
7. Pentode 3	5	−20	160	285	2.0	2	33.5
8. Pentode 4	2.5	−20	170	300	1.5	1	21
9. Pentode 5	5	−15	250	250	2.0	13	44
10. Pentode 5	5	−18.5	275	275	2.0	12	42
11. Mullard PM-24-A	4	−25	200	300	0.275	5	17
12. Mullard PM-24-A	4	−33	200	300	0.275	2.3	6.7

Table No. II shows the dynamic performance when maximum input is applied. In this table the "a-c/d-c power" includes d-c power for self-biasing.

TABLE II

DYNAMIC PERFORMANCE WITH MAXIMUM INPUT

TUBE	Max. Input	Max. Power Output for Max. Input	Power Sensitivity	a-c/d-c Power	Load Resistance	R_{min}/R_{max} for 80% of Max. Output*	d-c Power
	Volts	Watts	$\sqrt{W}/E_{gAC}$	Per Cent	Ohms	Per Cent	Watts
1. UX-112-A	6.3	0.135	0.058	13	5000	14.5	1.04
2. UX-112-A	9.5	0.320	0.060	17	6000	15	1.88
3. UX-171-A	28.4	1.06	0.036	19	2000	23	5.6
4. UX-245	35.5	2.07	0.041	17	2000	16	12.2
5. Pentode 1	6.5	0.417	0.100	28.5	15000	17.5	1.46
6. Pentode 2	10.8	2.50	0.146	21.5	10000	22	11.6
7. Pentode 3	13.9	5.50	0.169	42.5	7000	20	13.0
8. Pentode 4	14.4	3.23	0.125	44	15000	21	7.4
9. Pentode 5	9.9	2.90	0.172	19.5	15000	18	14.9
10. Pentode 5	12.6	4.25	0.164	25	10000	19	17.0
11. Mullard PM-24-A	18.15	3.26	0.100	37	12000	21	8.8
12. Mullard PM-24-A	24.5	2.96	0.070	40	10000	24	7.4

* This column represents "output distortion," as previously defined.

It may be seen here that the pentodes require higher load resistance than the triodes to obtain maximum output. Also, the pentodes' a-c/d-c power economy and power sensitivity is considerably higher and the "output distortion" about the same.

Table III lists the harmonic and interelectrode capacities of the tubes measured. For the triodes the values are given for 5 per cent second harmonic since the harmonic distortion continues to decrease as the load increases.

TABLE III

Harmonics and Interelectrode Capacities

Tube	LOAD FOR MINIMUM 2ND HARMONIC			INTERELECTRODE CAPACITIES		
	Input R-M-S Volts	Load Ohms	3rd Harmonic Per cent Total Volts	G-P $\mu\mu f$	G-F $\mu\mu f$	P-F $\mu\mu f$
1. UX-112-A	6.3	13300	0.46	9.9	4.4	3.3
2. UX-112-A	9.5	15900	0.55	9.9	4.4	3.3
3. UX-171-A	28.4	6900	0.45	8.1	4.7	3.9
4. UX-245	35.5	6000	0.30	8.8	6.0	4.8
5. Pentode 1	6.5	17600	13.1	7.8	4.4	24.6
6. Pentode 2	10.8	9200	5.8	9.5	9.5	28.5
7. Pentode 3	13.9	8400	14.0	10.0	11.3	28.5
8. Pentode 4	14.4	14200	14.7	9.2	9.9	27.3
9. Pentode 5	9.9	6600	3.5	1.2	9.5	14.8
10. Pentode 5	12.6	7400	5.7	1.2	9.5	14.8
11. Mullard PM-24-A	18.15	18000	16.5	1.6	10.6	13.8
12. Mullard PM-24-A	24.5	28000	20.0	1.6	10.6	13.8

This table has been included principally to show that the third harmonic of the pentodes is quite high when minimum second harmonic is obtained. This is contrary to that which might be expected, for in the case of the triode both the second and third harmonics tend to decrease continuously with increasing load resistance. This will be more readily seen in some of the curves which follow.

The lines numbered the same in the various tables correspond as far as operating conditions are concerned. The tube marked "Pentode 1," tabulated on line 5 of the tables, was a battery type of tube with an output comparable to that of the UX-112-A listed on line 2. Pentodes No. 2, 4, and 5, in addition to the Mullard PM-24-A are nearest comparable in output to the UX-245; while Pentode 3 is of a higher output order, more nearly classed with the UX-250. Pentode 5 had cylindrical construction and a heater type of cathode while all the others had ribbon filaments.

Performance and Distortion Curves

Figs. 2 and 3 contain the output and harmonic distortion curves for the UX-112-A and Pentode 1 tubes when operated under the fairly comparable conditions given on lines 2 and 5 of the preceding tables. The curves are plotted to the same vertical scales.

It will be immediately apparent that the shape of the output curves are very nearly the same, the principal difference being in their respective amplitudes. However, there is a very marked difference in the curves representing the harmonic distortion. With the triode the second harmonic gradually decreases with increasing load resistance, while with the pentode it may be seen to dip rapidly to zero at a load resistance approximately equal to that giving maximum power output,

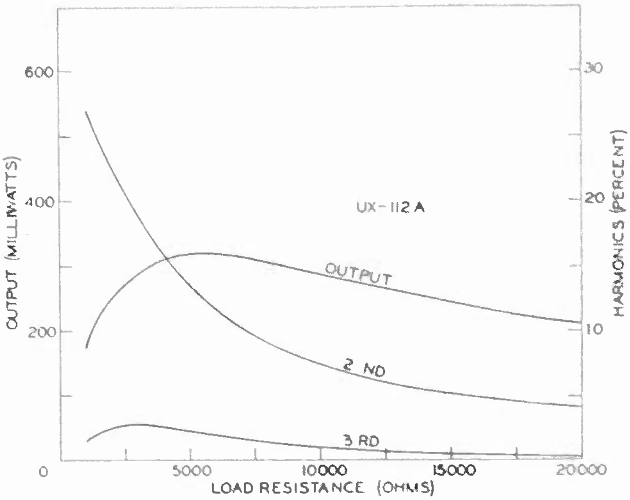

Fig. 2—Power output and harmonic distortion of UX-112-A.

then increasing again at a high rate. The third harmonic is small and much lower than the second for all load resistances with the triode, while with the pentode it rises to a very considerable maximum at a

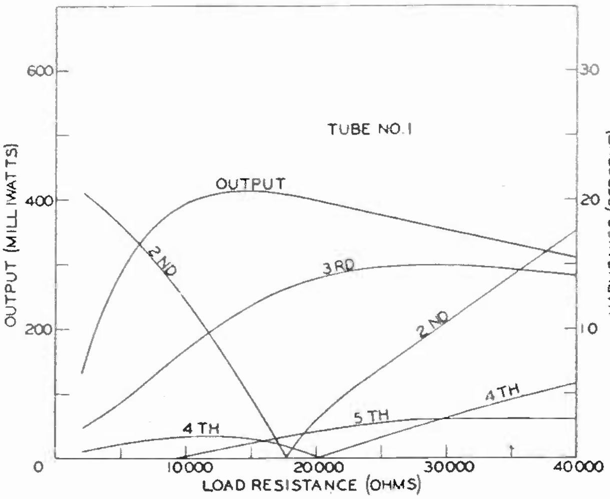

Fig. 3—Power output and harmonic distortion of pentode No. 1.

load approximately fifty per cent greater than that at which minimum second harmonic was obtained. Thus, it may be seen that the third harmonic of a pentode does not follow the characteristics of the second harmonic in any way.

The fourth and fifth harmonics of the triode are negligible. For the pentode they are included in Fig. 3, being of sufficient amplitude to warrant consideration. From graphical analyses, the dip to zero of the second and fourth harmonics was found to indicate a phase reversal with respect to the fundamental.

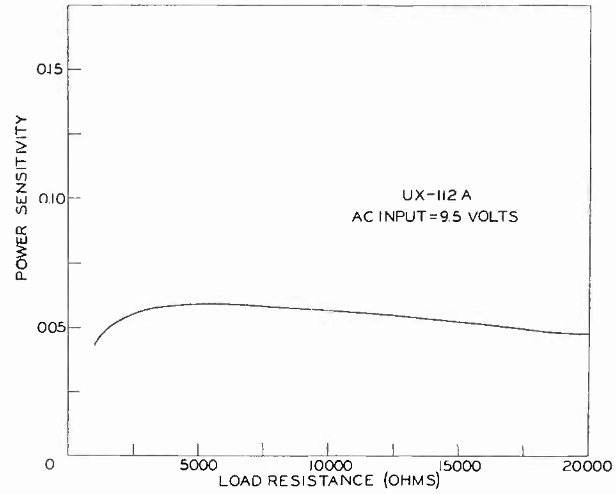

Fig. 4—Power sensitivity of UX-112A.

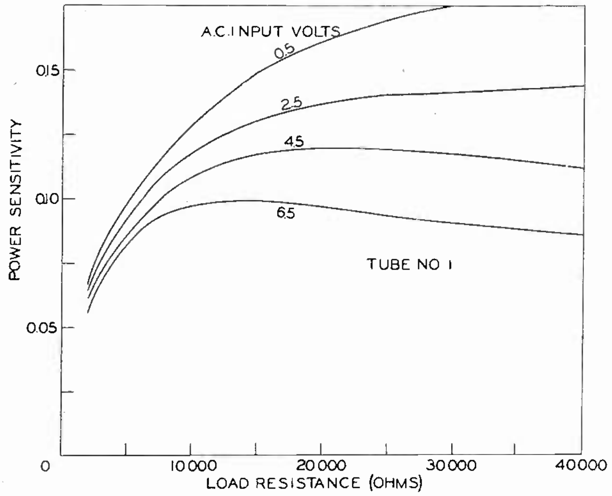

Fig. 5—Power sensitivity of Pentode No. 1.

With both the triode and pentode the harmonic distortion gradually decreased as the input was lowered, showing that the distortion was not caused by grid current.

Figs. 4 and 5 contain the power sensitivity curves for the UX-112-A and Pentode No. 1 under the same operating conditions as those just discussed. Both curves are plotted to the same vertical scale. The power sensitivity curves for various inputs of the triode are about the same so only one such curve is shown, while with the pentode the power sensitivity decreases for any given load resistance as the input increases. This is shown by the spreading of the curves in Fig. 5. Since these two figures are drawn to the same scale, an idea of the relative loud speaker volume per volt input at any given frequency may be had, it being proportional to the ordinates of these curves.

These four sets of curves, which have just been discussed, for the UX-112-A and Pentode No. 1 are typical triode and pentode curves. Also, the discussion applies to practically all tubes of these two types.

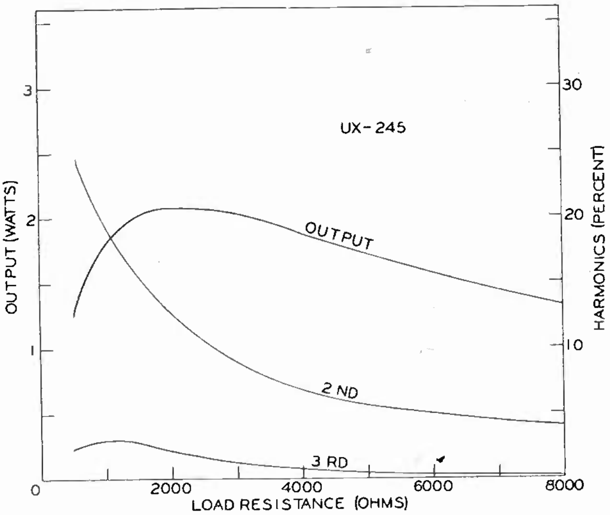

Fig. 6—Power output and harmonic distortion of UX-245.

Figs. 7, 8, 9, 10, and 11 are the output and harmonic distortion curves of the larger pentodes listed in Tables I, II, and III. They are all drawn to be compared directly with the UX-245, whose characteristics are shown in Fig. 6, with the exception of Fig. 9 where it was necessary to plot the output scale double that for the UX-245. Figs. 8 and 9 are both for the same tube, Pentode No. 5, with different operating conditions. Fig. 8 is for the condition shown on line 8 and Fig. 9 for that on line 9 in the preceding tables. The latter condition is one of higher operating voltages. Likewise Figs. 10 and 11 are for two conditions of input and bias voltages with the Mullard Pentode. Fig. 10 is for the manufacturer's specified bias of 33 volts, shown on line 12; and Fig. 11 for a lower bias, shown on line 11 of the tables. This comparison readily shows the improvement in harmonic distortion and output obtained by operating with the proper, rather than too high, a bias. The saving in d-c power required by the tube with the high bias is not enough to recommend such practice generally. It is evident that these curves are all quite similar to those shown in Figs. 2 and 3.

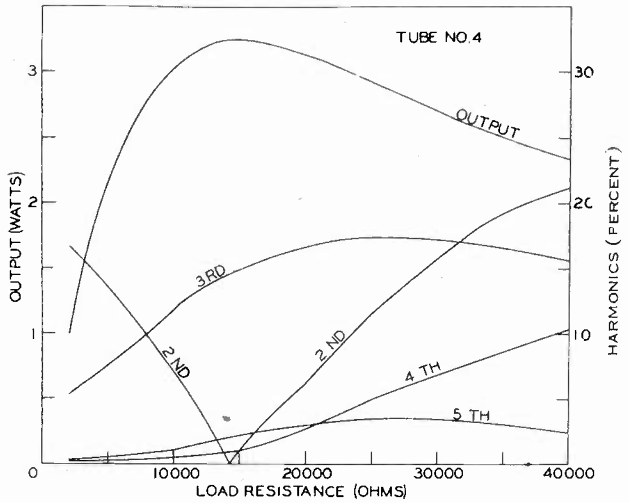

Fig. 7—Power output and harmonic distortion of Pentode No. 4.

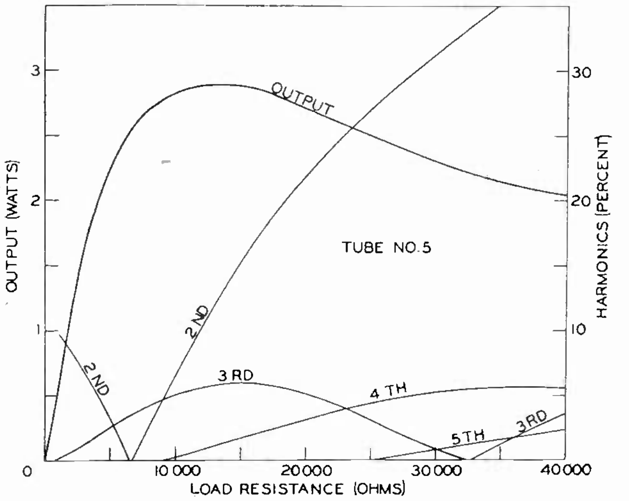

Fig. 8—Power output and harmonic distortion of Pentode No. 5.

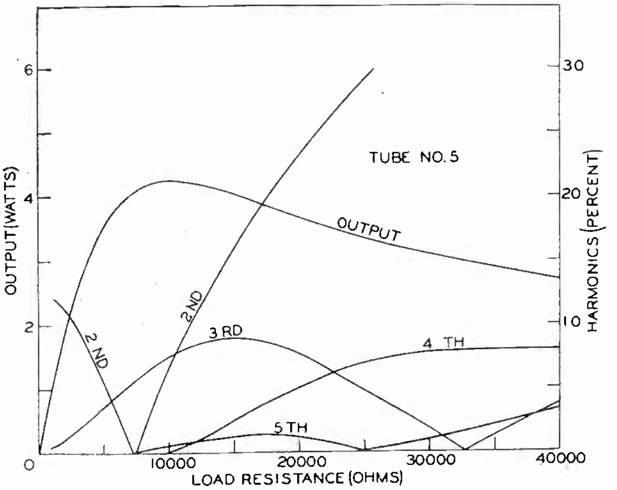

Fig. 9—Power output and harmonic distortion of Pentode No. 5ı

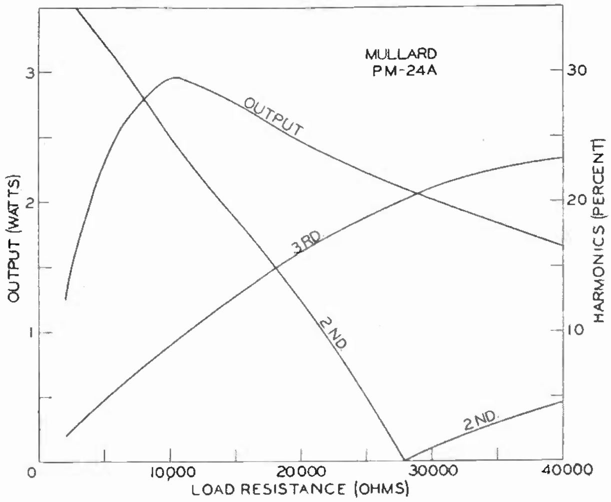

Fig. 10—Power output and harmonic distortion of Mullard Pentode PM-24A.

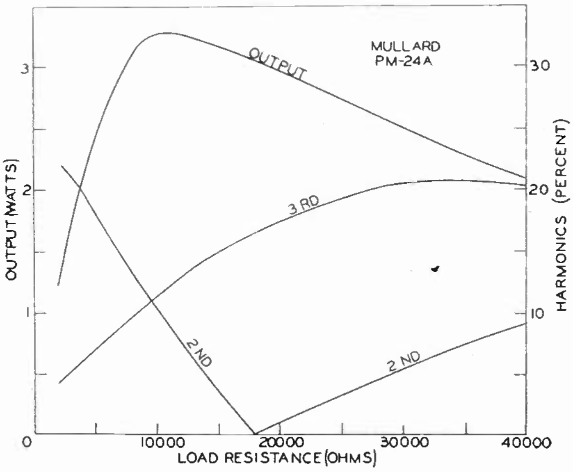

Fig. 11—Power output and harmonic distortion of Mullard Pentode PM-24A.

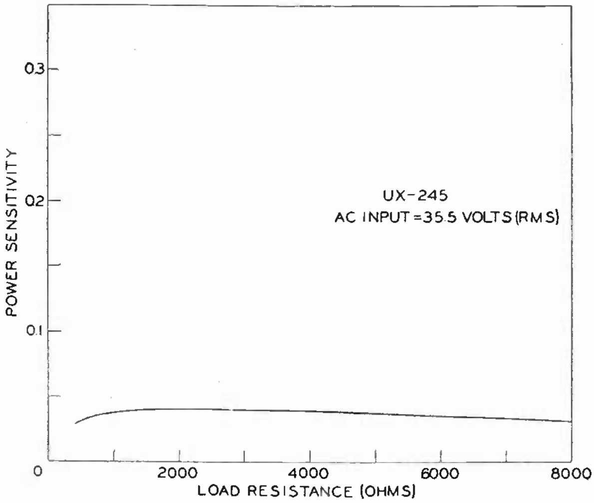

Fig. 12—Power sensitivity of UX-245.

Figs. 12, 13, and 14 are the corresponding power sensitivity curves to accompany the set of curves just discussed. They are all plotted to the same power sensitivity scale and will be seen to follow the same general trend of those shown in Figs. 4 and 5. Fig. 14 contains the curves for both bias conditions of the Mullard Pentode.

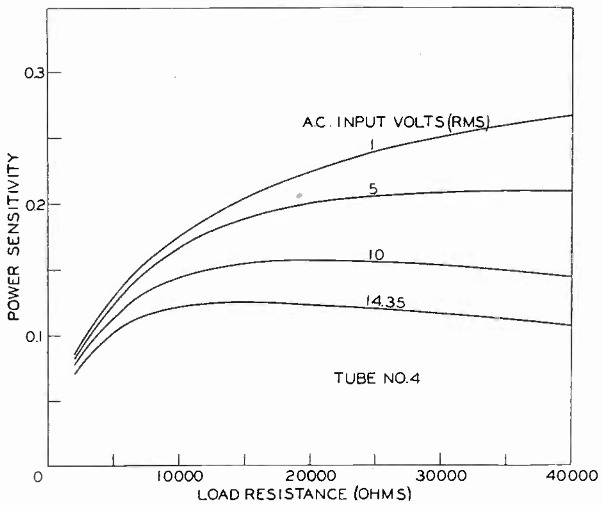

Fig. 13—Power sensitivity of Pentode No. 4.

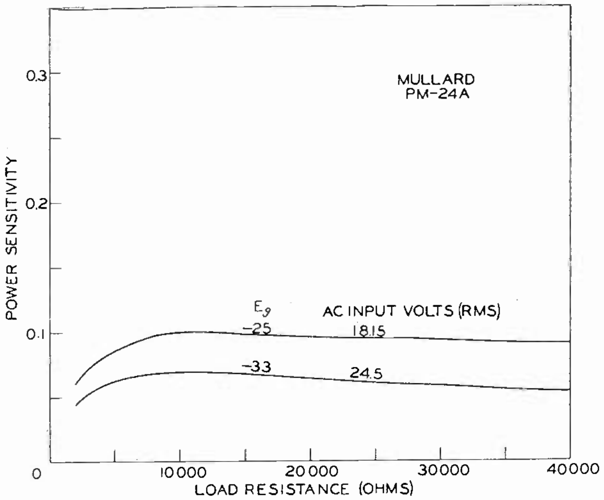

Fig. 14—Power sensitivity of Mullard Pentode PM-24A.

Figs. 15 and 16 compare the performance of the tubes discussed. They represent data taken with the load resistance which gave maximum power output. Fig. 15 shows that the pentodes lie in a group together with the exception of Tube No. 1 which is not very different from the UX-112-A. The steeper these curves are, the better the performance. The curve for Pentode No. 5 is about the same for both operating voltage conditions, with the exception of the maximum output. This tube gives 2.9 watts for the low voltage condition and 4.2 for the high voltage condition. Fig. 16 shows how the pentodes drop off in power sensitivity as the input is increased. This is, of course, undesirable since it causes a separate form of distortion distinct from harmonic distortion.

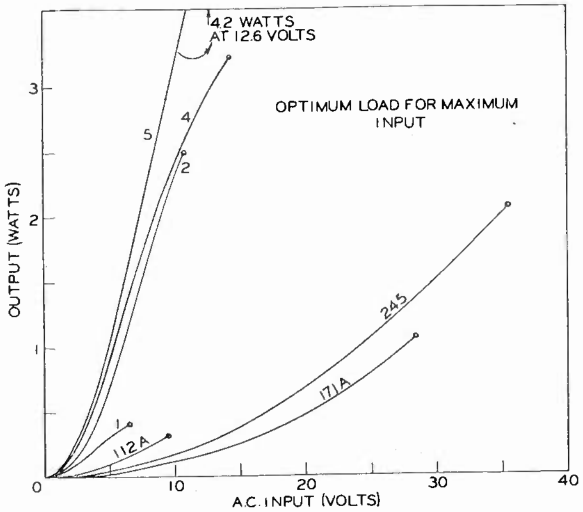

Fig. 15—Power output versus input of triodes and pentodes.

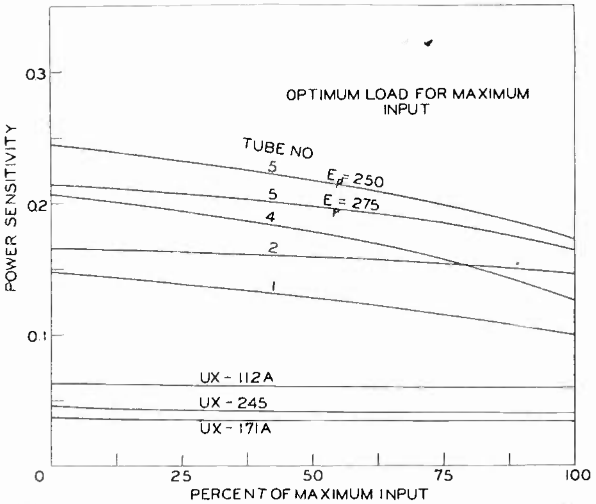

Fig. 16—Power sensitivity versus input of triodes and pentodes.

While operating pentodes as self-biased amplifiers, the by-pass capacity needed for the bias resistor was found to be quite large to give a good output. Data were therefore taken on Pentode No. 5, which is shown in Fig. 17, using 400 cycles. It may be seen from this figure that at least six microfarads were necessary. The bias resistor had a resistance of 330 ohms. The output under the same conditions with a battery bias was 4.2 watts, which is slightly higher than the output with the eight microfarads shown in the figure. The effect of this capacity drops off directly with frequency so that with eight microfarads at 50 cycles the effect is only that shown for one microfarad.

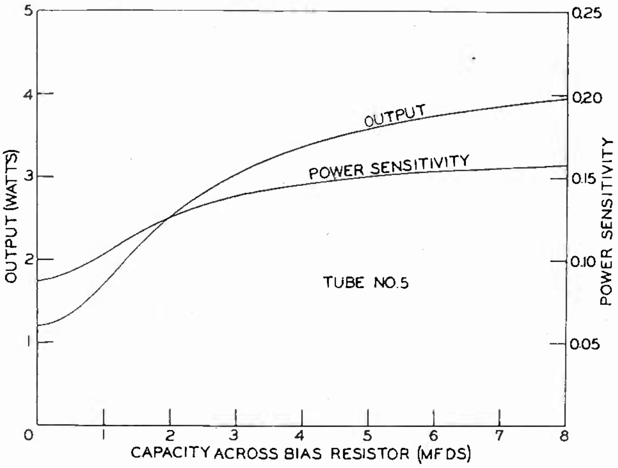

Fig. 17—Power output of self-biased pentode amplifier. Capacity across bias resistor of 330 ohms.

CONCLUSION

In making a final comparison between the pentode and triode it is necessary to consider the various types of distortion present.

The principal type to be considered is that of harmonic generation caused by nonlinearity of the plate current-grid voltage curves. This is more serious with the pentode since much third harmonic is generated and, of course, cannot be removed by push-pull. With the triode the harmonics are generally lower at the working conditions and the principal one, the second harmonic, can be practically removed by using push-pull.

A second type of distortion, which also introduces harmonics, is caused by grid current. This is not likely to be appreciably different in the pentode than in the triode.

A third type is due to the effect of input tube capacity shunting the higher frequencies. This effect is generally greater in the pentode than in the triode due to the high amplification, which reflects a greater amount of grid-plate capacity. For example, the input capacity of the UX-245, calculated from data in Tables II and III, was 31 micromicrofarads while that of Pentode 4 was 160. This large input capacity is not necessarily inherent with pentodes as is evidenced in the case of Pentode 5. The use of small cylindrical elements in this tube with the grid lead brought out of the top reduced the grid-plate geometric capacity to

1.2 micromicrofarads and resulted in an effective input capacity of only 30 micromicrofarads which is about equal to that of the UX-245.

A fourth type of distortion is caused by the large changes in loud speaker impedance at various applied audio frequencies.[5] These result in load variations which cause changes in tube output and harmonic distortion. The tabulation of R_{min}/R_{max} for 80 per cent of maximum output shows this change in output of the pentode and triode to be approximately the same. The curves of harmonic distortion plotted against load resistance show a somewhat more adverse condition with the pentodes than with the triodes for the same per cent load changes.

Another type of output distortion is caused by input changes, where the output variation is not proportional to the input. This is much worse with the pentodes than with the triodes as was shown by the power sensitivity curves.

The last distortion to be mentioned is that which occurs in a self-biased amplifier when the bias resistor is not sufficiently by-passed. This effect on output was illustrated for a pentode by Fig. 17, in which it was seen to be quite appreciable for low frequencies and undoubtedly is one cause of the highs predominating with pentodes. Insufficient by-passing did not increase the harmonic distortion.

In conclusion, the advantages of the pentode may be summed up in two ways. First, they permit more economical operation for the same output; and second, they have higher amplification than the triode.

Offsetting these advantages, the pentodes generally introduce considerable harmonic distortion and require a fairly high load impedance, which in turn makes it necessary to use a more expensive impedance matching transformer for coupling the loud speaker than is required with triodes. Also, hum due to a-c filament operation is likely to be more serious due to the high voltage amplification.

The interest in the pentodes for use in small a-c receivers appears to be due largely to the fact that greater over-all receiver sensitivity and power output can be obtained from a single audio-frequency amplifier tube. More economical d-c power consumption is not a vital consideration with a-c receiver operation.

The use of the pentode with battery and automobile receivers appears to offer a decided advantage from the d-c power standpoint, where economical operation is of extreme importance.

Another possible application of the pentode is with 110-volt d-c receivers where an output tube is needed that will operate satisfactorily on low voltages and yet deliver a high output.

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[5] E. A. Uehling, "The electrodynamic loudspeaker," *Radio Broadcast*, 113, December, 1929.

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DEVELOPMENTS IN SHORT-WAVE DIRECTIVE ANTENNAS*

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Summary—*Part 1 of this paper discusses the relative importance of the factors which limit the intelligibility of short-wave radio telephone communication. The more important of these factors are inherent set noise, external noise (static, etc.), and signal fading. The possibility of counteracting these limitations through antenna directivity is indicated.*

Part 2 describes an antenna system which maintains a desirable degree of directivity throughout a broad continuous range of frequencies. The cost of this antenna is more favorable than that of many types of fixed frequency antennas of equal effectiveness.

INTRODUCTION

BEFORE discussing specific antenna systems, it appears desirable to review the general problems of short-wave communication and to observe wherein antenna design can assist in overcoming existing circuit limitations. Accordingly, this paper is divided into two parts; the first will outline the requirements in the problem, and the second will be a description of an antenna system which has proved effective, despite its low cost of construction.

The writer's experience with antenna systems has been largely confined to the standpoint of reception, therefore, the following discussion will be largely on this basis. It will be apparent to the reader, however, that many of the features are likewise applicable to transmitting antenna installations.

PART 1. THE SHORT-WAVE PROBLEM

RADIOTELEPHONE CIRCUIT LIMITATIONS

An analysis of the factors limiting the excellence of the output quality of a receiver governs the design of the entire radio circuit and associated equipment. Assuming well-designed apparatus throughout, we still encounter difficulties, especially at times of low signal strength, the more important of which are enumerated as follows:

(a) Inherent receiver noise.
(b) External noise (static, man-made noises, etc.)
(c) Signal fading.

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The design of the receiving antenna system has an important bearing upon all three of these factors, brief explanations of which are given below.

(a) Receivers of very high gain characteristics are troubled with an inherent noise adequately described as a "hissing" sound. This may be due to several[1] causes such as shot-effect, etc. Much of this noise can be minimized through proper design, the methods of which are beyond the scope of this paper. Finally, however, an apparently irreducible minimum of noise is encountered, commonly referred to as[2] "Johnson" or circuit noise. This noise, under conditions of matched impedances, is so related to the circuit signal efficiency that the ratio of noise to signal cannot be appreciably altered except through somewhat impractical expedients such as lowering the absolute temperature of the circuit. All this tends to show that the designer of receivers must eventually rely upon his being able to increase the signal outputs from antennas to override the residual receiver noise difficulties on low field strength signals.

(b) Unpublished work, by a member of our laboratories,[3] has indicated that on many occasions short-wave static is highly directional. Interfering signals and electrical noises of human making are, of course, directional. It is quite evident that where the desired signal direction differs from that of the interference, receiving antenna directional discrimination is of immense importance.

(c) At times, remarkable reductions in short-wave fading have been achieved through extremely sharp directional characteristics of the receiving antenna. On the basis that certain types of fading are due to phase interference between multiple path signals of varying path length, it is reasonable to believe that where an angular difference exists between these paths, fading can be reduced by directivity which accepts only one of the paths. This, of course, assumes that the accepted path is stable in its direction. When this is not true, the reduction of fading through directivity becomes difficult.

THE RELATIVE IMPORTANCE OF THE VARIOUS CIRCUIT LIMITATIONS

The most serious hindrance to reliable, long-distance, short-wave communication is the great loss in signal fields which accompanies magnetic storms. Maintaining service under such conditions, develops into a battle against set noise and static. It is during these periods

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[1] F. B. Llewellyn, "A study of noise in vacuum tubes and attached circuits," PROC. I.R.E., February, 1930.

[2] J. B. Johnson, "Thermal agitation of electricity in conductors," *Phys. Rev.*, 32,, 97, 1928.

[3] K. G. Jansky, Bell Telephone Laboratories.

that effective receiving antennas are the most appreciated. The research worker on receiving antenna systems always welcomes such periods for his experimental work, since he knows well that under conditions of strong signals, a simple antenna appears to perform as well as one considerably more elaborate and expensive.

Fig. 1 will assist in comparing the relative importance of set noise and static interference. The figure is not intended to be strictly accurate as to numerical values but will convey the idea of the principles involved. There is plotted as a function of wavelength, for an arbitrary location and season, the average static voltage level delivered to the first tube of a receiver by a half-wave, vertical antenna through its coupling circuits. Likewise, we have plotted the circuit noise delivered to this same tube as a function of wavelength. The fact that these curves intersect is of importance.

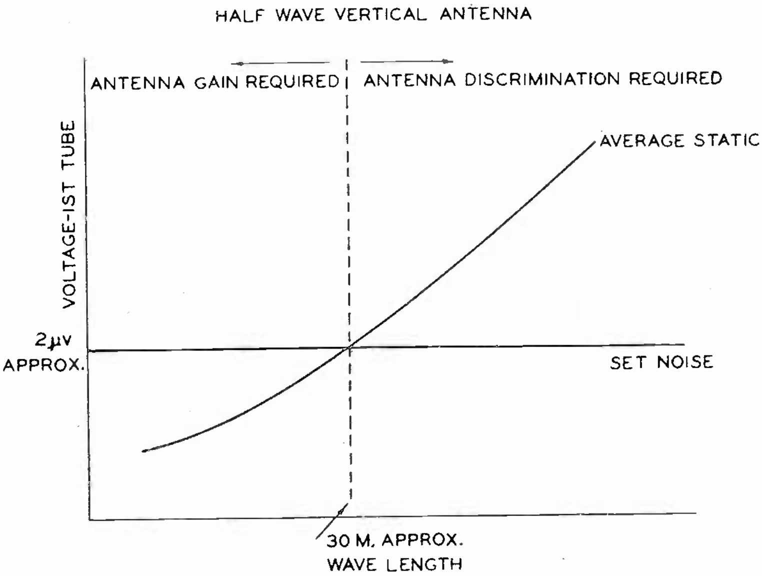

Fig. 1—Relative distribution of static and set noise with wavelength.

At wavelengths considerably below the point of intersection, a weak signal falls into the level of the set noise. Increased signal output from the antenna is desirable to override this noise. It is evident that static reduction through directional discrimination is of little use in this region, therefore an antenna having directional properties but possessing no marked gain in output over a simple nondirectional antenna has no merit. At wavelengths considerably above the point of intersection, static reduction through directivity is of utmost importance, while a gain in antenna output would be of little value if it meant a gain in static as well as in signal. It is interesting to observe, however, that a sufficient reduction of static through directivity would lower the whole static curve until it lay below the set noise curve. Such being the case, signal gain would again be required.

The above arguments are intended to show that, at the shorter wavelengths, receiving antennas should be designed for a gain in signal output. At the long wavelengths, directive discrimination in reception is the major requirement. In contrast to this, a transmitting antenna has no such wavelength eccentricities. Its purpose is always to lay down at the receiving point as great a field as possible. We must not forget, however, that the time is near when more attention should be paid to marked directive discrimination in transmitting antennas as a means of reducing interference between congested communication channels.

While set noise and static are at times important factors in limiting successful short-wave communication, fading practically always presents varying degrees of annoyance. It is really surprising how much fading can be tolerated without radically affecting speech intelligibility, but for services such as high-grade program transmission where naturalness is also important vast improvements are required; consequently much attention has been, and is being, paid to this phase of the problem.

INCREASING THE SIGNAL OUTPUT OF RECEIVING ANTENNAS

Under conditions of optimum output impedances, the magnitude of signal developed at the receiving antenna load is simply a function of the ratio of the effective induced voltage to the effective antenna resistance. The term effective induced voltage is used, as attention must be directed toward proper phasing, where the antenna dimensions are an appreciable part of a wavelength or more. Usually at short waves, the effective resistance is almost entirely the resistance equivalent of the reradiation losses. This resistance can be lowered through directivity, a simple example of which can be illustrated with the aid of Fig. 2.

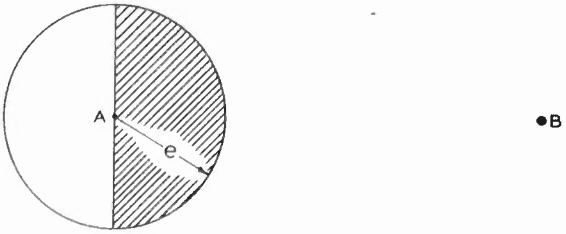

Fig. 2—Effects of antenna directivity.

If we can conceive of a point source of radiation at A, equipotential radiation surfaces would be spherical in shape and symmetrically disposed around A. The field intensity at point B would be unaffected if we had some means of avoiding radiation through the unshaded half of the sphere, with a consequent saving of half of the radiated energy. If instead of saving this energy we added it to the shaded side, the energy available at B would be doubled. This is a simple explanation of the effect of directivity in the transmitting case. The receiving case is quite similar.

If the transmitter is at B, the energy available at A is diminished by reradiation losses. If we avoid reradiation through the unshaded half of the sphere, the radiation equivalent resistance is halved and the load energy will be doubled, after rematching the load to the antenna impedance.

With this knowledge of the usefulness of sharpened directivity, the designer is tempted to carry it to an extreme. The degree of directivity that may be beneficially attempted is, of course, limited by the variation in the apparent direction of wave arrival. For transatlantic, 16-meter signals over a daylight path, the horizontal plane angular variation, at New York has been[4] measured, by observing phase differences between spaced antennas, to be some 5 degrees or less, but apparently random throughout this range. Over a combination path of darkness and daylight, a horizontal angle variation considerably greater than this magnitude is frequently observed.

In the vertical plane, the variations in the apparent directions of arrival are considerable and also random. On rare occasions, angles as high as sixty degrees from the horizontal have been recorded. A sharp low angle antenna may well be expected to decrease in output as the angle of the wave direction becomes high.

Knowing that the interpretation of wave directions, by means of observed phase differences between spaced antennas, might be complicated if multiple waves of varying angles were present, two vertically polarized test antennas were built having optimum response at 27 degrees and at 6 degrees from the horizontal, respectively as shown in Fig. 3-A. These angles were experimentally obtained from airplane measurements. Fig. 3-B, which has been smoothed out for publication, is characteristic of about 80 per cent of the comparative data obtained on these two antennas, as measured by automatic signal recorders. Examination will show that, very frequently, the high angle antenna increases in output as the low angle antenna loses, or vice versa, indicating that the waves are varying in their vertical angle. Similar methods

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[4] H. T. Friis, "Direction of propagation and fading of short waves," Proc. I.R.E., May, 1928.

have also cross-checked the horizontal plane movements previously mentioned.

Where it is planned to design a single fixed antenna for a particular service, the antenna should be sufficiently broad in its directivity to include most of the directional variations in signal arrival that may be encountered. In such cases, we have adopted the policy of simultaneously comparing the signal outputs of various size antennas through the measurements of automatic signal recorders over long periods. A photograph of one such signal recorder is shown in Fig. 4.

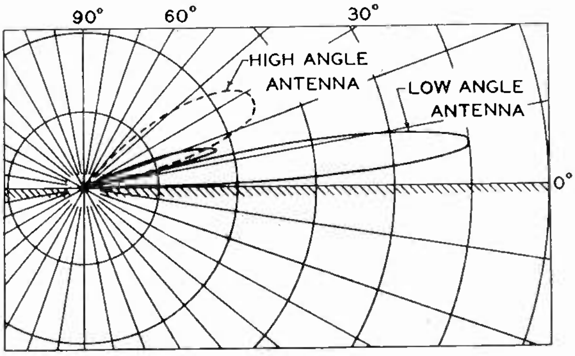

Fig. 3a—Comparative directive diagrams of a high and a low angle antenna.

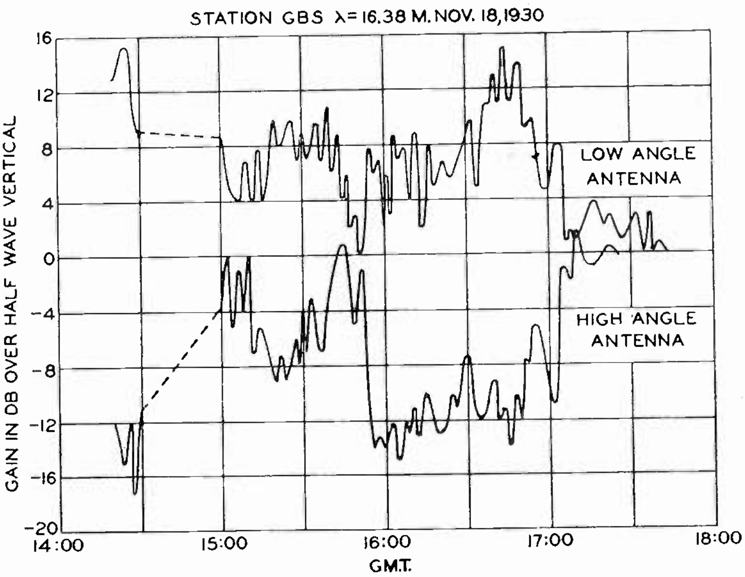

Fig. 3b—Comparison of signal outputs of a high and a low angle antenna.

Several of our test antennas have proved to be too sharp. On occasions, their output exceeded that of any of the smaller, less directive antennas, but when averaged over long intervals of time, they proved to be deficient. At first, we tried to avoid putting too much weight on gain data obtained when signals were normally very strong but long experience seems to show that wave direction variation has little correlation with the field strength of signals.

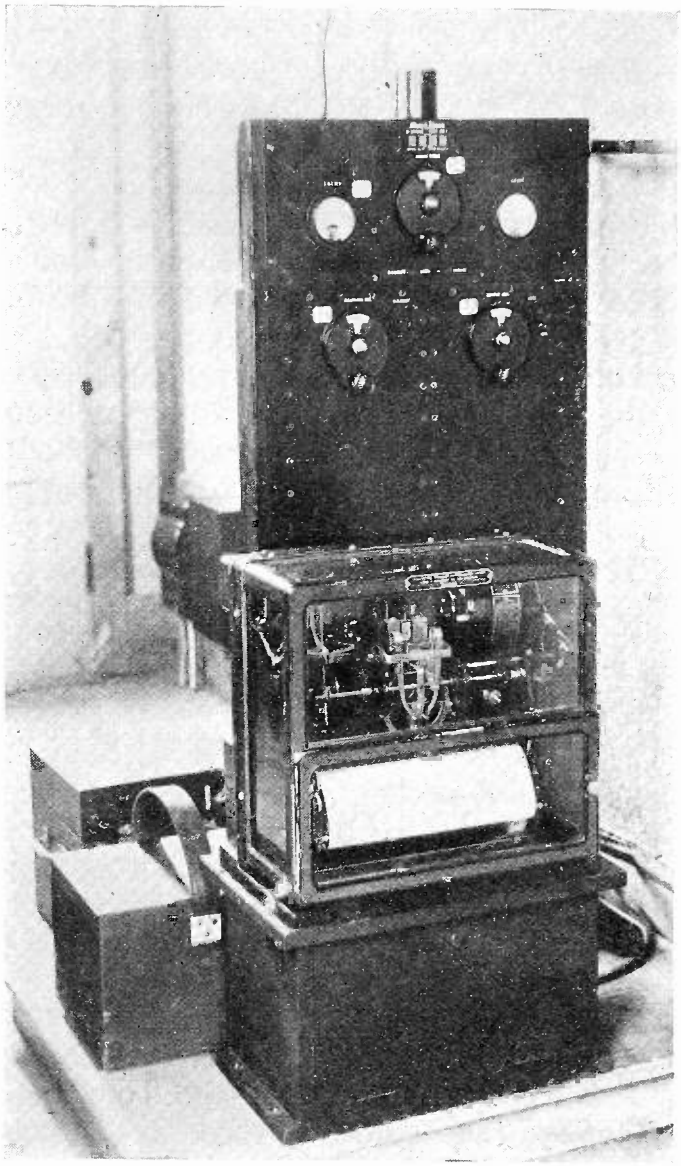

Fig. 4—An automatic signal recorder.

STATIC REDUCTION

Referring again to Fig. 2, assume that point A, receiving from B, is surrounded in all directions by static of uniform intensity. If A is made responsive only in the shaded directions, half of the static appears, at first, to be eliminated, but we must remember that, by previous arguments, the static output from the shaded region is doubled; thus the over-all static output is the same. For uniform distribution, the static output level is independent of the degree of directivity, provided that impedance matching between the load and the antenna is always maintained. We see, therefore, that the improvement in signal-to-static ratio in this case is the same as the signal improvement alone.

If static were always uniformly distributed about an antenna, the problems of signal gain and improvement in the signal-to-static ratio would be synonymous. The fact that short-wave static is usually highly directional puts an entirely different aspect on the problem. If, in Fig. 2, the static came from a direction included in the unshaded portion of the characteristic, the improvement in the signal-to-static ratio would be infinite. In a receiving antenna, therefore, emphasis must be placed on the deep suppression of response in other than the favored direction.

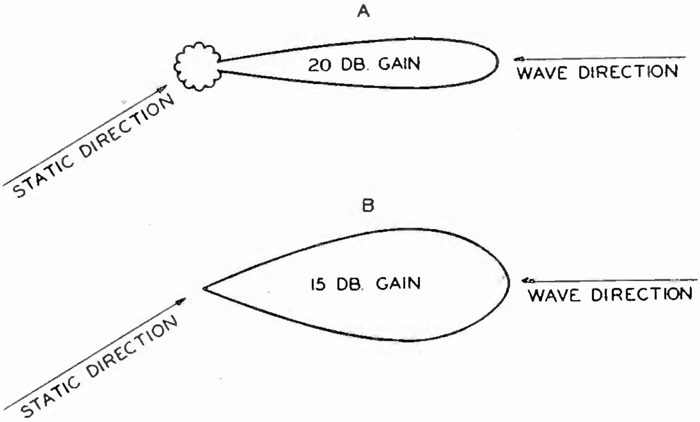

Fig. 5—A comparison of directive diagrams.

Fig. 5 is intended to illustrate the case described. The antenna characteristic 5A, having a signal gain of 20 decibels over a nondirectional antenna, does not accomplish deep rejection in other directions. It follows, therefore, that the better discriminating characteristic 5B would give a vastly better signal-to-static ratio, in spite of a smaller signal gain.

Fading Reduction

Many schemes for counteracting fading are in use and have been suggested. These include compensation for fading through automatic control of the receiver gain, the automatic selection of the best of several antennas, single side band with an unvarying locally supplied carrier, etc. All of these systems have merit, but are not a complete cure for the very prevalent selective type of fading, where several depressions may exist within a frequency band width of speech magnitude.

Under certain conditions, selective fading can be combatted through antenna directivity, but it is not without its difficulties in attainment. This is a direct attack on the multiple path source of the evil, eliminating a cause which makes fading selective with frequency. At times, very marked fading reduction has been obtained by this means.

ECONOMICS OF RECEIVING ANTENNAS

We have indicated briefly that the receiving antenna system has an important bearing upon all the major factors which are limitations in the present short-wave art. As long as these improvements can be effected in the receiving antenna system at a cost less than, for instance, a corresponding increase in transmitter power, concentration on the development of antenna design is well warranted.

One often hears the question whether one type of directive antenna is better than some other type. The answer usually depends on an economic comparison rather than an electrical one. The sharpness of directivity, the gain, etc., are determined by existing conditions. Numerous types of antennas can be designed to meet these specifications, therefore it is evident that the final selection is often based on over-all costs.

In Part 2 of this paper an antenna system will be discussed which is the result of an attempt to produce an effective antenna at a cost more favorable than the types we have been accustomed to use up to the present time.

PART 2—LONG WIRE ANTENNAS

TYPES OF DIRECTIVE ANTENNAS

Directive methods, employing a finite number of spaced elements of specific phase and amplitude relations, have been known for a long time. Most of the more recent innovations, in this form of antenna, have pertained to the methods whereby, in their practical applications, these phases and amplitudes have been achieved. Considerable use has been made to date of such antennas, but they are quite expensive in their larger sizes and often their frequency range is very limited. As a result of these frequency restrictions, the radiotelephone receiving station at Netcong, N. J., employs ten[5] antennas, all differing in their design frequency but having the same favored direction toward England.

For some time, it has been appreciated that if it were possible to

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[5] A. A. Oswald, "Transoceanic telephone service—short wave equipment," *Bell Sys. Tech. Jour.*, April, 1930.

substitute a single directive antenna, having frequency characteristics sufficiently broad as to cover the above mentioned ten channels, a very large economic saving could be effected. Development work was undertaken which has not only resulted in an antenna of considerable frequency latitude, but this new antenna structure is actually less expensive than a single, equally effective unit of the previous type. The remainder of this paper will be devoted to a discussion of various applications of this form of antenna.

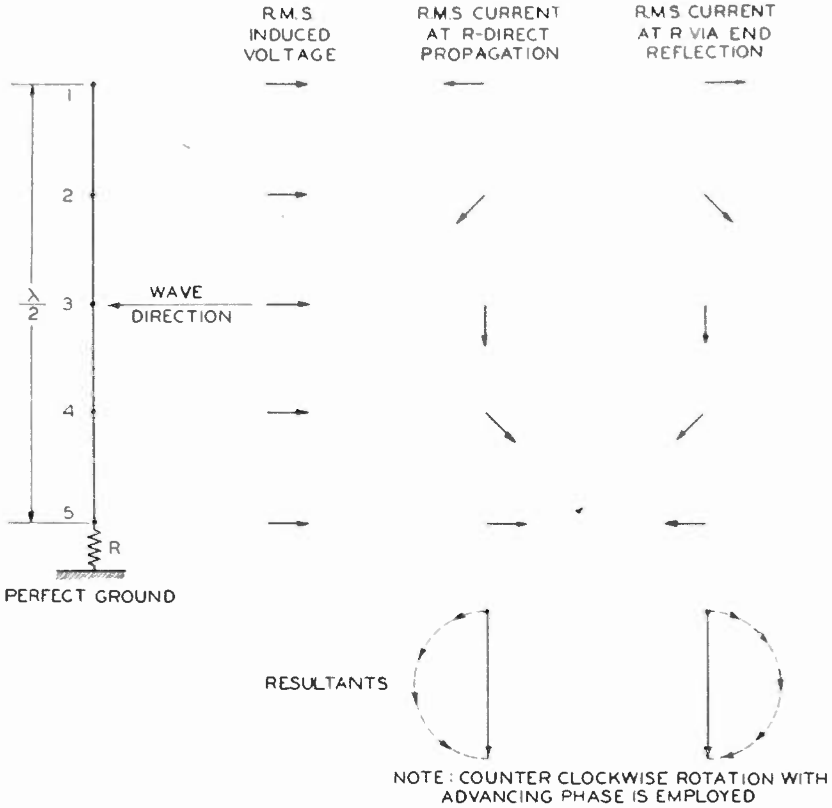

Fig. 6—Vector relations in a half-wave vertical antenna.

PRINCIPLES OF "TILTED" WIRE ANTENNAS

The elementary principles underlying "tilted" wires can be explained more readily by presenting a physical picture, through the use of r-m-s vector representation, rather than through a more or less cumbersome mathematical treatment. The vector representations that follow are not rigorous but they serve to convey quickly the ideas under consideration and give results which are in sufficiently good accord with the complete mathematic analysis.

As we increase the length of a simple vertical antenna exposed to horizontally propagated waves, always rematching impedances by varying the load at its base, we obtain increases in the load power up to the point where the antenna wire length reaches one-half wavelength. The vector representation of this one-half wavelength case constitutes Fig. 6.

The first column of vectors represents the phase of the induced voltages, assumed to be lumped at points 1 to 5. The second column of vectors indicates the phase of the directly propagated currents arriving at R and due to each lumped voltage. The phase changes are due to the varying intervals of time required to traverse the intervening path.

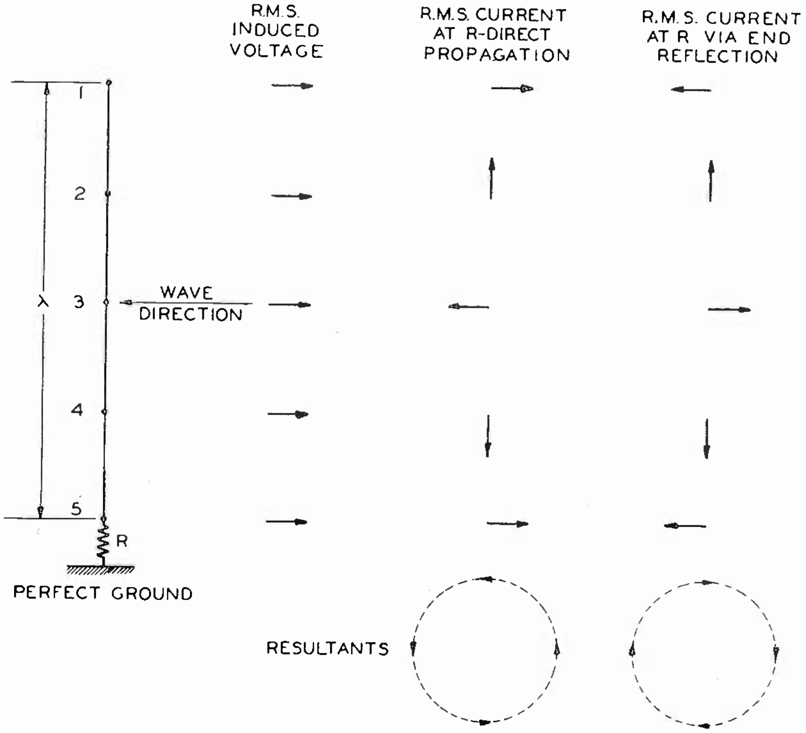

Fig. 7—Vector relations in a one-wave vertical antenna.

Likewise, the third column represents the current reaching R by way of the open-end reflection where a 180-degree phase change occurs. Summing up either column of current vectors, we trace a semicircumference and the resultant is a diameter. Had the antenna wire been slightly longer, the circumference would have been further closed and the resultant smaller. Fig. 7 illustrates an extreme case where the currents in R are zero for the vertical antenna length of one wavelength. Analyzing these vectors, we establish an important principle, as follows:

The length of a straight antenna wire is an optimum value, for currents directly propagated to the load, when the elementary currents due to voltages induced in small lengths at the two wire extremities are opposite in phase at the load, provided that this does not also occur for intermediate points. This statement has been restricted to the directly propagated currents since, in what follows, we shall, practically always, dissipate the currents propagated to the far end in appropriate terminating impedances. In many of the diagrams, the load currents which would arrive from open-end reflections have been included merely as of general interest.

The above stated principle permits us to remedy the null situation of Fig. 7 by tilting the wire as shown in Fig. 8. Notice that point 1 has been advanced into the wave propagation so that, at any given instant, point 1 is later in phase than for instance, point 5. The directly propagated currents of Fig. 8 trace a semicircumference and, therefore, the wire length[6] appears to be an optimum for the tilt selected.

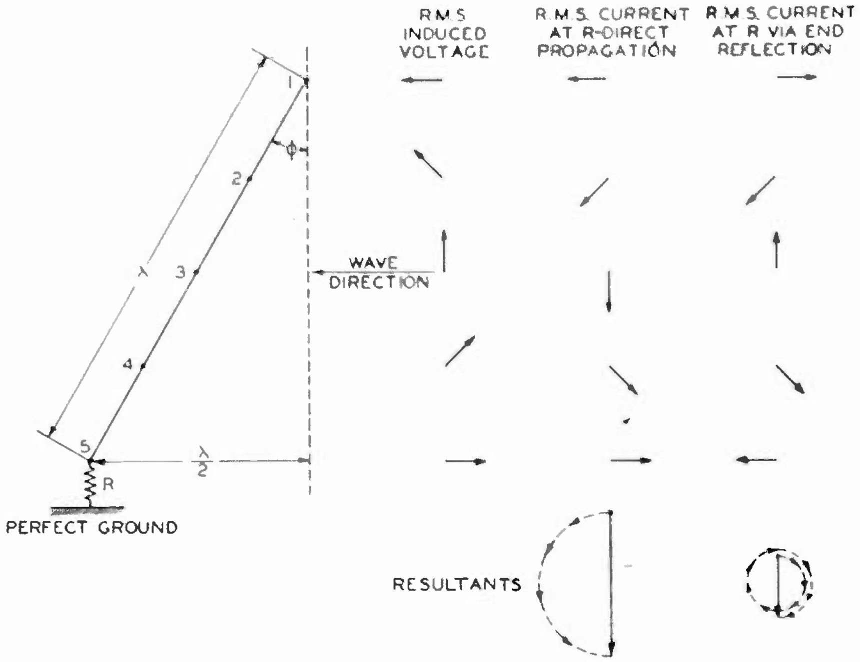

Fig. 8—Vector relations in a tilted wire antenna.

For any wire tilt angle, there exists a wire length which will trace a semicircumference similar to the above. This occurs when the tilt is such that the wire length is one-half wavelength longer than its projection upon the wave direction of propagation. Using appropriate tilt

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[6] For rigid accuracy in determining optimum dimensions, a small correction must be applied to these rules. This correction occurs in cases where, upon changing the wire tilt angle, the rate of change of induced voltage is comparable to the rate of change of load current as described above.

angles, as the wire length increases, output gains are achieved through increased effective induced voltage in the wire. Still further gain in output is available through the increasing directivity that is bound to result from the increasing dimensions.

One of the chief features of the tilted wire antenna is that in its longer lengths it is effective over a broad range of frequencies. This is illustrated by Fig. 9 which is a plot of the wire length versus the tilt angle utilizing the above mentioned rules. For example, if the antenna were designed for a frequency such that the wire was ten wavelengths long but it was used at another frequency where the wire length was only eight wavelengths, Fig. 9 shows that the inaccuracy of tilt angle would be only about two degrees, which in most cases is inappreciable. As we shall see later, even this inaccuracy can be compensated by another wire in combination having an opposite trend.

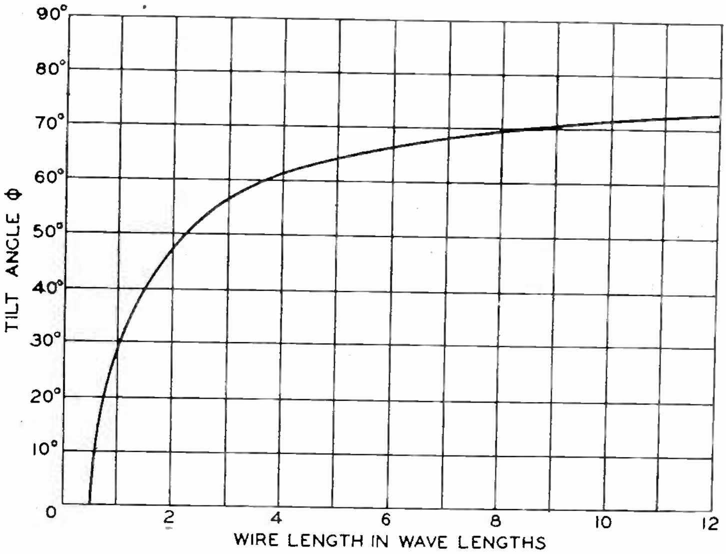


Fig. 9—Optimum tilt angle for long wires.

Broad Frequency Range in Arrays

As is true for any antenna, the tilted wire may be used as an element in all the usual forms of arrays. Successful experimental antennas have been constructed consisting of a succession of tilted wires disposed in broadside relation, in the line of transmission and also stacked one above another. Some of these arrangements confine the effectiveness of the resulting antenna to a single frequency. Appreciating that one of the principle features of the tilted wire was its effectiveness over a broad frequency range, we have particularly stressed the development of those combinations of tilted wires which would not place restrictions on this frequency range. One such combination is discussed in the following section.

THE INVERTED V

The combination of two tilted wires to form the inverted V is shown in Fig. 10. The directional characteristics are appreciably improved with a consequent increase in signal output; also, the far end of the antenna becomes accessible for termination purposes, near the ground. These terminations will be discussed later. The inverted V requires no more supporting structure than the tilted wire, therefore its additional cost is very small where the land is available. Fig. 10 is a vector picture indicating that the two elements of the inverted V add in proper phase relation.

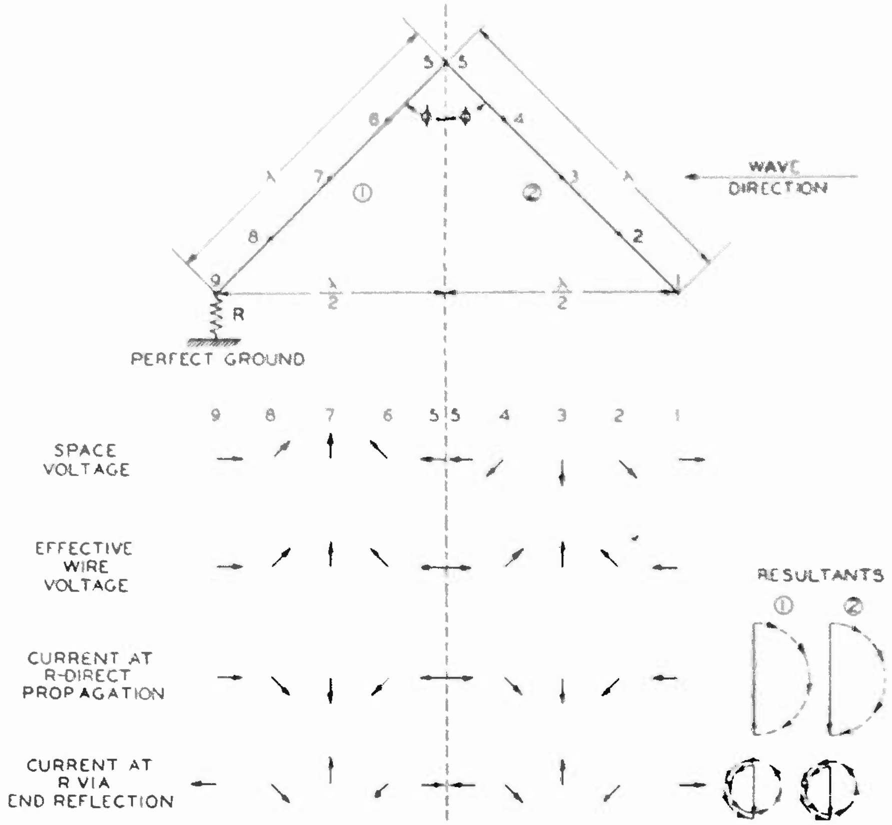


Fig. 10—Vector relations in an inverted V antenna.

In connection with Fig. 9, it has been mentioned that the small inaccuracies in tilt angle, due to departures from the design frequencies, can be counteracted by another wire in combination having an opposite trend. The inverted V of Fig. 10, is an example of one such possible arrangement. Since the tilt angle error is opposite in direction for each leg of the V, in combination, their optimum direction of response will remain unaltered. This will be illustrated by calculated directive diagrams which will be given later.

Asymmetrical Directivity Through Far End Terminations

Where it is desired to make an antenna responsive to signals in a given direction but to discriminate against signals in the opposite direction, reflector systems are often employed. These reflectors may be parasitic or they may be directly connected to the receiver through apparatus controlling their phase and amplitude relations. Our experience has shown that reflectors may be employed in connection with the type of antenna under consideration for the purpose of obtaining unilateral directivity. However, the use of reflectors restricts the possible frequency range, as they only function efficiently at specific spacings in

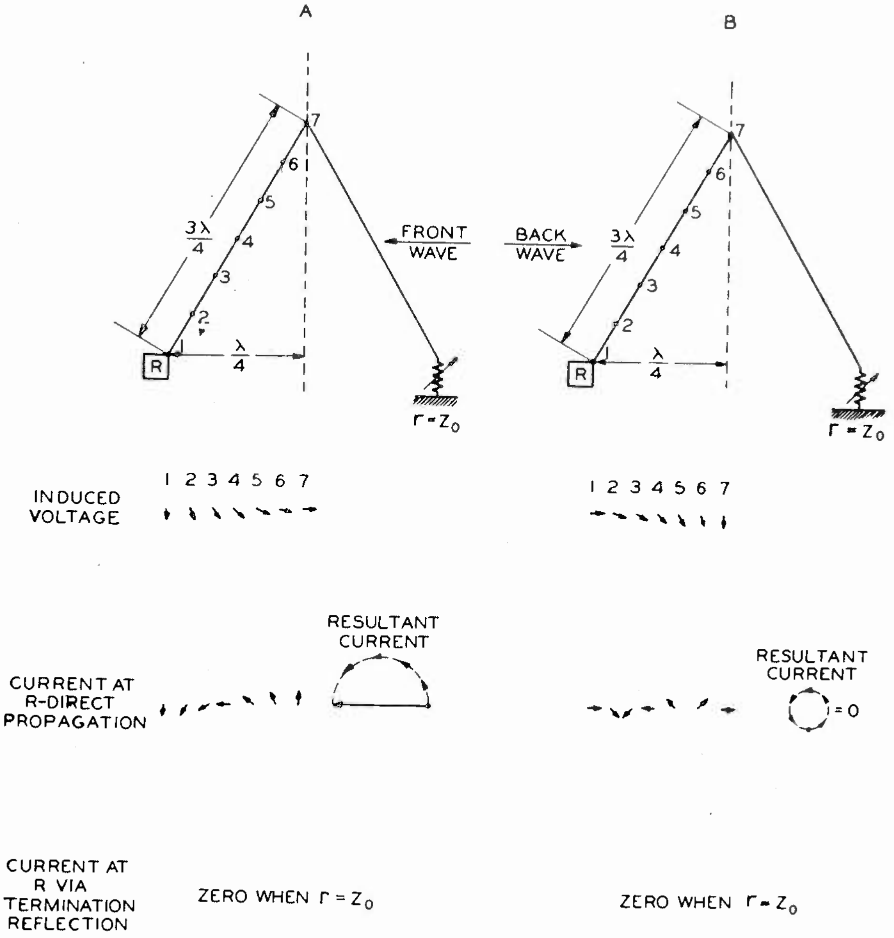


Fig. 11—Vector relations in an inverted V antenna—asymmetrical directivity.

relation to the wavelength used. For this reason, reflectors will not be discussed in this paper, although they are employed where a broad frequency range is not essential.

Tilted wire antennas and their combinations are particularly adapted to obtaining directional asymmetry through proper terminations of the end remote from the receiver. A simple example is illustrated in Fig. 11.

The end of the inverted V remote from the receiver *R*, in Fig. 11, is so terminated as to absorb signals without reflections. In other words, a termination equal to the antenna characteristic impedance is employed. Only the vectors for one leg of each of the inverted V's have been drawn, as the second leg is simply a reproduction of the first, and add directly thereto, after all phase relations have been determined.

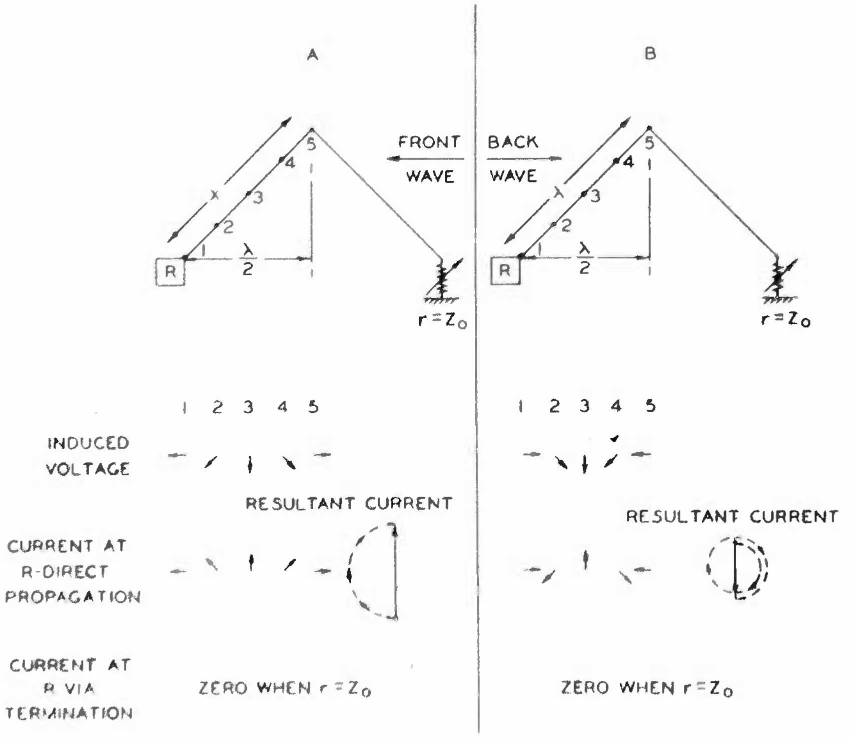


Fig. 12—Vector relations in an inverted V antenna—asymmetrical directivity.

In Fig. 11A, a wave from the right produces elementary load currents which trace a semicircle, as previously discussed. Note that when the wave arrives from the left as in Fig. 11B, the phase change is more rapid and a closed circle is traced making the resultant zero, thus we have achieved an infinite front-to-back ratio. It can be shown that this advantageous condition exists for tilted wires where the wire length of each element is an odd integral multiple, greater than one, of one-quarter wavelength, provided that the previously mentioned optimum tilt, in relation to the wave direction, is maintained.

At first glance, it might appear that the frequency range is restricted, since the above rule is limited to certain wire lengths expressed in wavelengths. The most disadvantageous case exists when the wire length is an even integral multiple, greater than two, of one-quarter wavelength. Fig. 12 illustrates one such case, the wire being one wavelength long and at optimum tilt. It will be observed that the front-to-back ratio is not infinite but there still exists some directional discrimination, due to the fact that the back wave has resulted in the elementary currents tracing one and one-half rotations, thus obtaining partial cancellation. It is important to notice that longer wires would result in an increasing number of rotations and the resultant current of the back wave would become smaller and smaller as compared with the resultant of the front wave. This is a further argument for the use of long tilted wires. The calculated front-to-back ratios obtained with characteristic impedance terminations for various lengths of wires at optimum tilt are plotted in Fig. 13.

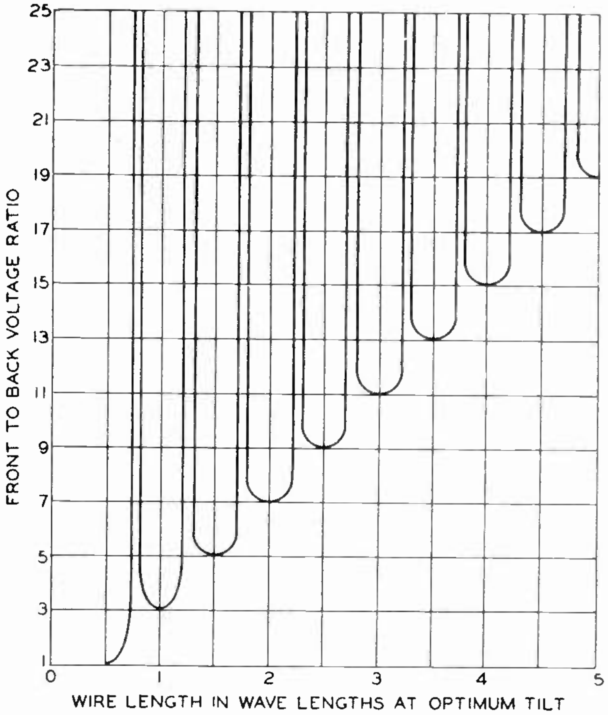


Fig. 13—Front-to-back ratios for characteristic impedance termination.

A very interesting feature about terminations is that, provided we are willing to make slight readjustments in their value, it is possible to obtain infinite front-to-back ratios at all frequencies within range. This is accomplished by cancelling the residue of back signal by means of a small reflection from the end termination obtained by departing slightly from the characteristic impedance adjustment. It can be shown that this results, for wires which are in length an even multiple, greater than two, of one-quarter wavelength, when the termination is the characteristic impedance times the cosine of the angle made by the wire with the direction of wave propagation.

For long wires, the above readjustment is very small. As an example, a ten-wavelength wire is properly tilted when it makes an angle with the direction of wave propagation whose cosine is 0.950. Thus, only a five per cent reduction in the termination from the characteristic impedance value will give an infinite front-to-back ratio.

In practice, we usually adjust a termination to a value which is a compromise between the above value and the characteristic impedance This gives very favorable front-to-back ratios at any frequency within the range of the antenna, particularly in the case of long wires.

Theoretically infinite front-to-back ratios have been mentioned several times in the preceding discussion. It is an experimental fact that where very minute adjustments can be made in both the resistive and reactive components of the termination impedance, the front-to-back signal voltage ratio is only limited by the rigidity of the antenna elements in space. Voltage ratios in excess of 1000 to 1 are readily obtained, although such extremes are seldom warranted in practice. This deep depression can be "steered" through a considerable range of directions largely through changes in the reactive component of the termination impedance, the resistance alteration required being small. This permits a high degree of discrimination against many specific cases of interference in the rear quadrant of the antenna.

THE DIAMOND-SHAPED ANTENNA

In terminating inverted V antennas to ground, trouble has been experienced due to the instability of the ground contact resistance during varying weather conditions. In addition, the signal "pick-up" in the connecting leads was not always small compared with the antenna signal response in directions of antenna minima. These difficulties were avoided by terminating to the center point of a straight wire, substantially a half wavelength in total length, lying perpendicular to the favored wave direction.

As is well known, a quarter wavelength open-ended element appears to be a very low resistance when measured between its terminal and ground or another similar element. Two such low resistance quarter wavelength elements are effectively in parallel in the above arrangement and the center-tapped symmetry substantially balances out the effect of voltages induced in these elements.

Variations of the above type of artificial ground have been used in connection with inverted V antennas but, with few exceptions, they have required readjustments as the frequency was altered. A more satisfactory arrangement from several points of view is the double-V or diamond-shaped antenna shown in Fig. 14. This provides a balanced arrangement eliminating the necessity of a "ground" connection; furthermore, it does not place any frequency limitation upon the system.

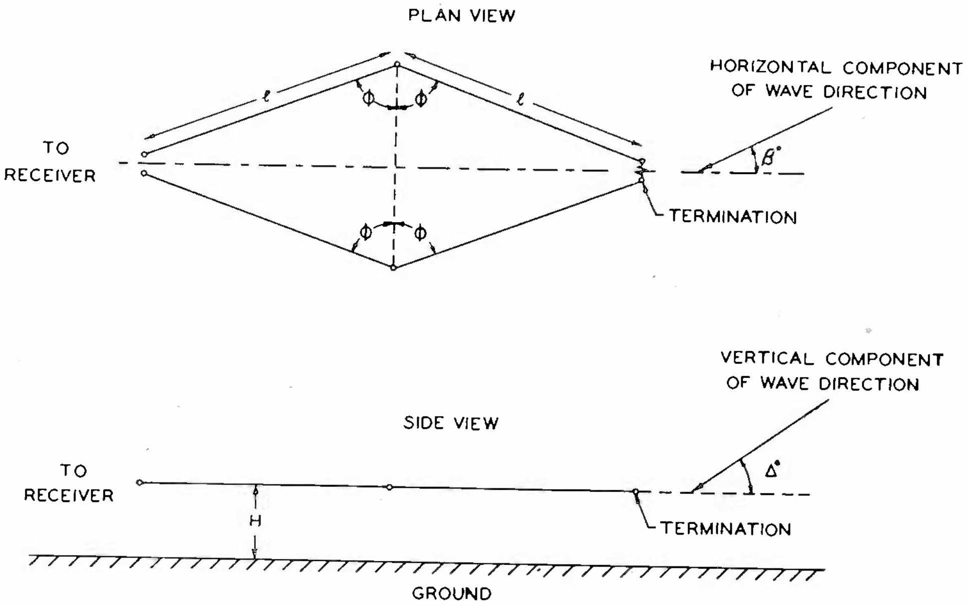


Fig. 14—The horizontal diamond-shaped antenna.

The antenna in Fig. 14 may be used with its plane either vertical or horizontal, being responsive, respectively, to vertically or horizontally polarized waves. It has found its greatest application in its horizontal form, however, due to reasons enumerated below.

(a) The supporting structure in its horizontal form is less costly, since only four relatively short poles are required.

(b) The inherent high angle directive characteristics of horizontal antennas discriminate against ignition, power, and other noises originating near the ground.

(c) The solid directive diagram of the diamond-shaped antenna is sharpest in the plane of the antenna. Since the direction of wave propagation is more stable in the horizontal plane, it is desirable to have the plane of the antenna horizontal.

(d) The directivity of the horizontal diamond-shaped antenna can be aimed, to some extent, at the most desirable vertical angle merely by altering the "tilt" angle ϕ of the antenna.

(e) The performance of the horizontal antenna is stable with varying weather conditions, since horizontally polarized waves are less affected than are the vertical by varying ground constants.

The use of the antenna horizontally, in the usual short-wave range, assumes that the strength of horizontally polarized waves are at least as great as are the vertically polarized components. Several observers have reported them more so, but the experience of the writer has been that there is little choice where horizontal and vertical antennas, having the same degree of directivity and optimum direction, are compared.

Up to this point in this paper, the attempt has been made to present simply a broad picture of some of the applications of long tilted wires to antenna design. It now seems worth while to give in somewhat more detail a sample of the design methods employed and the performance measurements on one typical form of antenna; accordingly a medium size horizontal diamond-shaped antenna has been selected.

THE HORIZONTAL DIAMOND-SHAPED ANTENNA

In calculating the directive diagrams of the horizontal diamond-shaped antenna, the antenna wires have been assumed to be without resistance. As long as we are contented in knowing only the relative shape of the directive diagrams, this approximation is quite accurate and results in a tremendous simplification of the problem.

In all of the calculations, a perfect ground has been assumed. Fortunately, for horizontally polarized waves, variation in the ground constants do not radically affect either the amplitude or phase of the ground reflections, so that the following equations can be used as rough approximations even where imperfect ground conditions are encountered.

Vertical Plane Directivity

The vertical plane directivity of the horizontal diamond-shaped antenna is determined by three factors, i.e., the length of each leg, the "tilt angle" and the height above ground.

For the cases where the element length is an integral multiple of a half wavelength and where the far end termination is the characteristic impedance multiplied by the sine of ϕ (see Fig. 14), the equation for the vertical plane directivity over perfect ground has been calculated to be,

$$I_R = k\left[1 - e^{-j4\pi H \sin\Delta/\lambda}\right]\left[\frac{1+\cos\Delta}{1-\sin^2\phi\cos^2\Delta}\right]\left[1 \pm e^{-j2\pi l \sin\phi\cos\Delta/\lambda}\right]^2$$

where, as shown in Fig. 14,

H = height above perfect ground in wavelengths.
Δ = wave angle from horizontal in the vertical plane
ϕ = tilt angle of elements.
l = element length in wavelengths.
k = proportionality factor.
I_R = receiver current.

Note: In the third bracketed quantity use, in the ± sign, − when l is an even integral multiple of $\lambda/2$ and + when l is an odd integral multiple of $\lambda/2$.

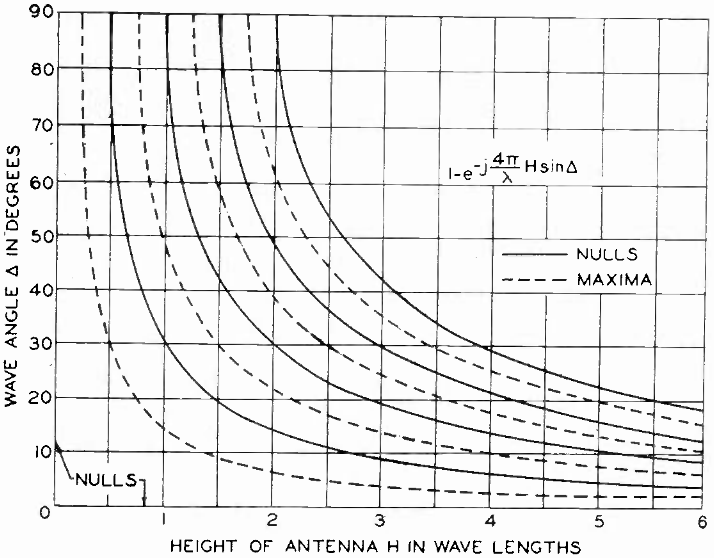


Fig. 15—Vertical plane design chart.

It will be noted that neither the length nor the tilt angle appears in the first bracketed term. It can be shown that this factor appears as a multiplier for nearly any type of horizontal antenna, accordingly the location of nulls and maxima for this factor are separately plotted in Fig. 15.

In the same manner the nulls and maxima of the product of the second and third bracketed terms have been plotted in Fig. 16 for an element length of four wavelengths.

The curves of Figs. 15 and 16 are design curves and their use can be illustrated by the following example: Measurements on the directions of wave arrival have indicated that the most usual directions are from 10 to 15 degrees above the horizontal. It is desired to construct a horizontal diamond-shaped antenna for this reception, employing four-wavelength elements. Fig. 15 indicates that the most economical pole height for 15 degrees is approximately one wavelength. Now referring to Fig. 16, we see that the largest tilt angle, to accomplish this, is about 65 degrees. It is always desirable to use the largest possible angle of tilt to obtain the use of the largest lobe of the directive diagram.

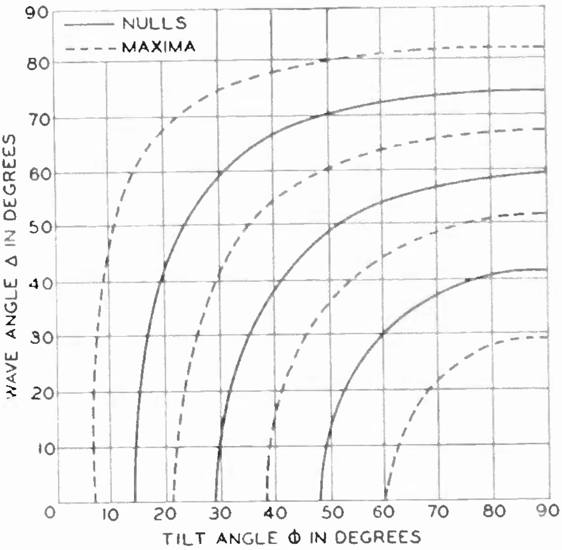


Fig. 16—Vertical plane design chart.

Figs. 15 and 16 likewise give us the null points. These are seen to be 0, 30, and 90 degrees in Fig. 15 and 34, 57, 74, and 90 degrees in Fig. 16.

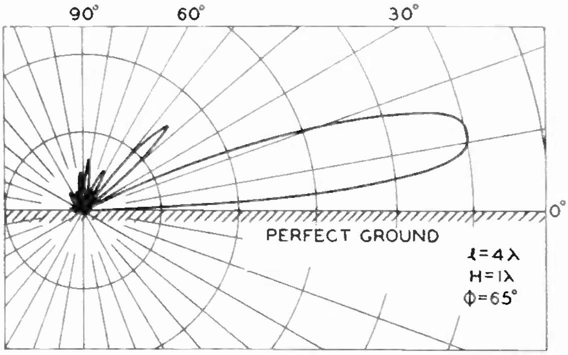


Fig. 17—Vertical plane directive diagram.

Using the above determined dimensions, the complete directive diagrams are calculated to determine whether a satisfactory result has been accomplished. Fig. 17 is the complete vertical plane diagram as calculated from the previously given equation. Should some undesirably large minor lobe be present, it is often possible to suppress it by slightly changing one of the variables. A knowledge of the location of the null points, as given by Figs. 15 and 16, is a valuable guide in this accomplishment.

Horizontal Plane Directivity

Due to the cancellation effect of the reflections of horizontally polarized waves from a perfect ground, the horizontal plane diagram, for a horizontal antenna, is merely a point. The way to view directivity is properly in its solid form, but the calculations and plotted representations are somewhat laborious. The designer is in real need of knowing the horizontal width of the major lobe of the directional characteristic as would be seen from a plan view. This angular width, as measured between null points, is not altered by ground effects; therefore a useful simplification of the calculations may be had by ignoring the cancellation effect of the ground reflection. It should be pointed out that the amplitudes are slightly erroneous when this is done, but the null point locations are accurate. If this is done, we obtain the following equation:

$$I_R = k'\left[\frac{1+\cos\beta}{\cos^2\phi-\sin^2\beta}\right]\left[1 \pm e^{-j2\pi l \sin(\phi+\beta)/\lambda}\right]\cdot\left[1 \pm e^{-j2\pi l \sin(\phi-\beta)/\lambda}\right]$$

where, as shown in Fig. 14,

β = wave angle in horizontal plane.
ϕ = tilt angle of elements.
l = element length in wavelengths.
k' = proportionality factor.
I_R = receiver current.

Note: In the second and third bracketed quantities use, in the $\pm$ sign, $-$ when l is an even integral multiple of $\lambda/2$ and $+$ when l is an odd integral multiple of $\lambda/2$.

Fig. 18 is a plot similar in character to that of Fig. 16, giving the location of nulls and maxima in the same manner. In our previous example, vertical plane considerations indicated that a tilt angle of 65 degrees was desirable. An examination of Fig. 18 gives a rapid estimate of the approximate plan view of the directive diagram and Fig. 19 is the more complete plan diagram for this tilt angle. It will be noted in Fig. 18 that the lines indicating factor maxima and minima frequently intersect. This property can be utilized for the suppression of particular minor lobes of the directive diagram by a proper selection of the tilt angle.

Frequency Range

Previously, it was stated that the V form of antenna counteracts the slight tendency for a change in optimum direction when the frequency is altered. The correctness of this statement is verified in Figs. 19, 20, and 21. The linear dimensions and tilt angle were unaltered as the wavelength was varied over a two-to-one range. The optimum direction is maintained although, as would be expected, the directivity becomes less sharp as the wavelength is increased in respect to the antenna dimensions.

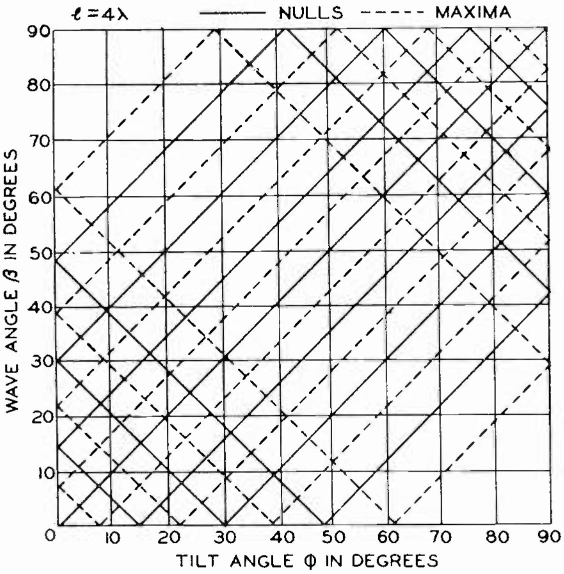
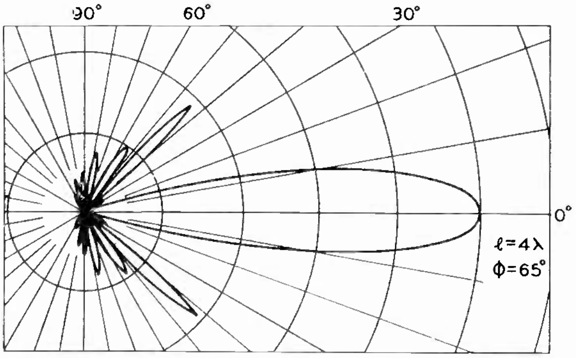


Fig. 18—Plan view design chart.



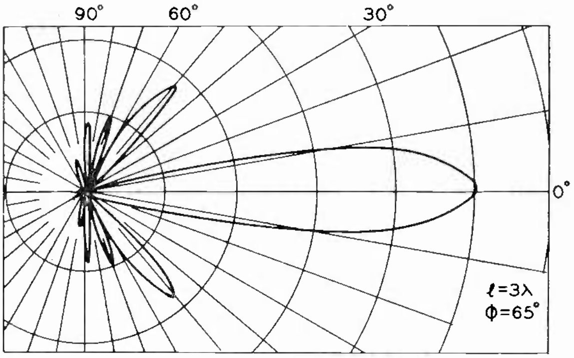
NOTE: GROUND CANCELLATION IS IGNORED

Fig. 19—Plan view directive diagram.

Due to the variability of the wave directions in the vertical plane, this desirable direction is not well defined. As the wavelength is increased, a broadening characteristic counteracts the possibility of losing signal due to the optimum direction of the characteristic moving slightly upward.

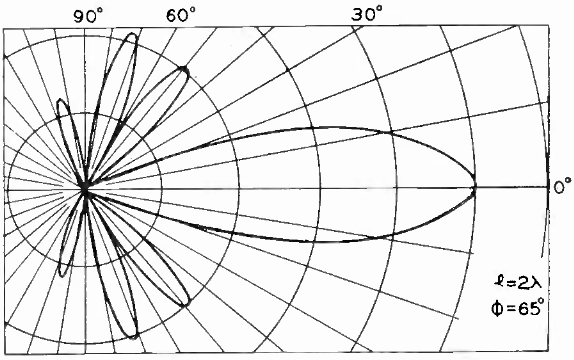
Antenna Coupling Circuit

A two-wire transmission line has been used as the connecting link between the antenna and the coupling circuits at the receiver. With this arrangement, the circuits must be carefully balanced against vertical waves to obtain local noise reduction and to avoid reradiation losses from the transmission line. This is not difficult for a single frequency but if the coupling circuits are to maintain this balance for a range of frequencies, very careful designing of the coupling circuits is required.



NOTE: GROUND CANCELLATION IS IGNORED

Fig. 20—Plan view directive diagram.



NOTE: GROUND CANCELLATION IS IGNORED

Fig. 21—Plan view directive diagram.

The present practice is to place these coupling circuits in an elevated position directly at the antenna terminals to reduce the necessity for finical balancing adjustments. These circuits are connected to the receiver through a concentric pipe transmission line with its accompanying low loss, freedom from "pick-up," and substantial weatherproof construction. Multipeaked coupling circuits have been devised so that no readjustment is required over quite a frequency range.

Measured Performance

From the inception of our short-wave experience, we have been accustomed to compare the performance of antennas with a half-wave vertical antenna. The lower end of this standard of comparison is near the ground and connected to a coupling circuit in such a manner that matched impedances are realized. Although the antenna under consideration is intended for the reception of horizontally polarized waves, the same vertical comparison standard has been maintained.

As previously mentioned, automatic signal recorders of the type shown in Fig. 4, are connected to each antenna. This recorder indicates an integrated average signal during each ten-second period, thus removing the wide amplitude excursions due to fading. It is an interesting fact that, although the instantaneous fading of two antennas may be different, the average signal over ten seconds usually has corresponding rises and falls in amplitude. This effect is so marked that any possible inaccuracies in the timing axis are readily detected, when comparing records. To promote accuracy in amplitude comparisons, only corresponding peaks or hollows of the curves are used. It is obvious that the employment of steep sides of curves would put a premium on very accurate timing. The relative timing of recorders is usually very good, as their synchronous motors are run by the same a-c power supply. The relative signal strength accuracy of the recorders is better than one db.

The antenna reported in the following data is an experimental antenna, at Holmdel, N. J., shown in the photograph of Fig. 22. This picture illustrates the extreme simplicity of this type of antenna. The antenna dimensions are the same as those in the previously discussed directive diagrams when used at 16 meters. As has been said so many times before, the gain of the antenna over the standard may be expected to vary with the varying wave directions. The following data are the results of several hundred hours of tests, made at Holmdel, N. J., during the fall and winter months. Three different wavelengths were used with no alteration whatever in the antenna, its termination, or its transmission line coupling circuits. The standard of comparison, however, was always a half wavelength for the signal under test. It has been thought desirable to plot the gain data as the percentage of total time the antenna gain was above the indicated value in order to show the gain distribution with time. This summary of gains is given in Fig. 23.

I am indebted to a member[7] of our laboratories for an interesting variation which has been used in the application of this type of antenna to the transmitting problem. A simple terminating resistance is often undesirable in the transmitting case since it may be called upon to dissipate several kilowatts, in fact, that portion of the energy which would be radiated backward if no terminating resistance were employed. A

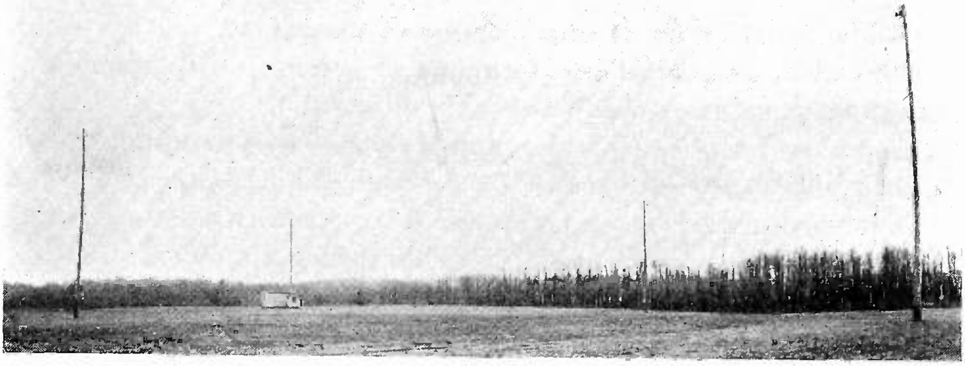


Fig. 22—An experimental horizontal diamond-shaped antenna.

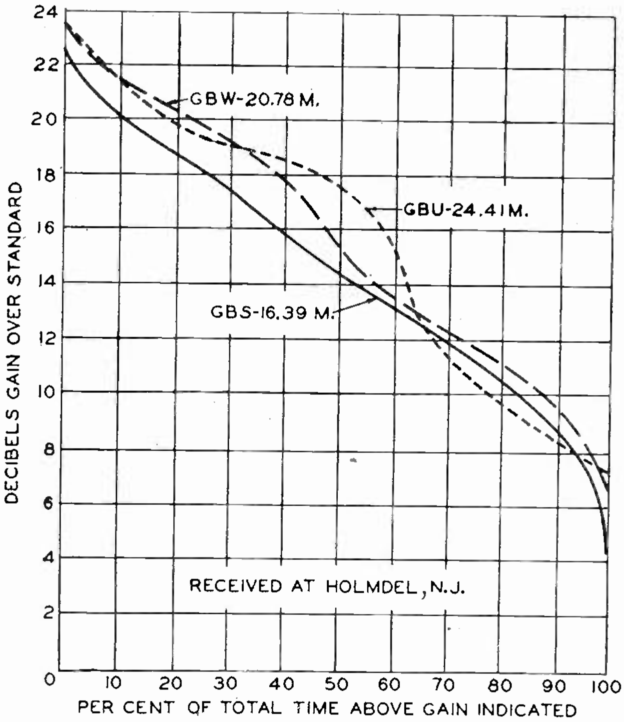


Fig. 23—Gain-time distribution curves.

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[7] E. J. Sterba, Bell Telephone Laboratories.

long, two-wire iron transmission line shorted at the far end has been found to be one useful terminating load of the required dissipating ability.

The terminated diamond-shaped antenna possesses a broad impedance-frequency characteristic. This property may be augmented by reducing the characteristic impedance of the antenna. One convenient scheme for reducing the impedance is to employ several conductors in parallel in each leg of the antenna. The characteristic impedance may in this manner be dropped to a value for which matching iron wire lines are readily constructed.

The terminating load which produces the most desirable impedance characteristic does not necessarily produce the best front-to-back ratio. In the transmitting case, however, the deep directed nulls required in reception, to eliminate interference of some particular station, are not necessary. It is sufficient to reduce by 10 or more decibels the field in the back directions. Thus the modified diamond-shaped antenna may be employed as a unidirectional transmitting array accepting power over a two-to-one frequency range.

In conclusion, I should like to point out that the work described in this paper was possible only through the assistance, coöperation, and advice of many people in the Bell System, to all of whom I render my sincere thanks. In particular, I wish to mention Messrs. A. C. Beck and L. R. Lowry who supervised the construction and did most of the testing of the experimental models. Mr. H. T. Friis, not only contributed many suggestions and constructive criticisms of the work, but took steps to have developed apparatus which was essential for the automatic measurement of received signal levels.

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RADIO TRANSMISSION STUDIES OF THE UPPER ATMOSPHERE*

BY

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Summary—*In this paper are given a number of measurements which show time variations in the virtual height of the ionized regions of the upper atmosphere. These measurements were usually made simultaneously on two frequencies, 1604 kc and 3088 kc. Single frequency data are also given. The following are the main points of interest presented.*

(1) The data indicate the existence of two distinct ionized regions or layers. The changes in virtual height are sometimes very abrupt. The existence of the lower layer even at night is indicated by an occasional return to low virtual heights during this period.

(2) Experimental evidence has been found of large retardations in group velocity near the critical conditions for which the waves just penetrate the layer to the point of maximum ionization. (Fig. 1.) Absorption is especially marked at such times.

(3) Except at these critical periods the records for the simultaneous transmissions show that the virtual heights of the upper layer are greater for the higher frequency than they are for the lower frequency. This statement would probably hold for the lower layer but no evidence on this point is presented.

(4) In the discussion several possible methods of two-layer formation are suggested, one of which involves the formation of negative ions in the region between the layers.

I. INTRODUCTION

DURING the last five or six years numerous investigators have been making theoretical and experimental studies of the electrical structure of the upper atmosphere, especially as to the effect on radio transmission. One of the commonest fields of investigation has been the attempt to determine the virtual height of the Kennelly-Heaviside layer.

This paper gives the results of certain experiments which were performed at the Deal Radio Laboratories during the early part of 1929 in an endeavor to add to the available fund of information on this subject. The experimental method which was employed is that originated by Breit and Tuve[1] in which the virtual height of the upper ionized regions is calculated from measurements of short time echoes obtained from pulse transmissions.

Our experiments differed from those of other investigators in that not only were frequencies chosen upon which data had not previously

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* Decimal classification: R113. Original manuscript received by the Institute, April 24, 1931. Presented before U. R. S. I., May 1, 1931, Washington, D. C.

[1] G. Breit and M. A. Tuve, *Phys. Rev.*, **28**, 554; September, 1926.

been published[2] but simultaneous transmission and reception on two frequencies were employed. In this manner the path differences for two frequencies at the same instant could be studied. The two frequencies used were 1604 and 3088 kc. (187 and 97 meters.)

II. APPARATUS

In order to perform these pulse transmission experiments two transmitters were necessary. One transmitter consisted of a push-pull oscillator employing two 250-watt tubes and the second was a high power oscillator-amplifier transmitter employing two 5-kw water-cooled tubes in the final stage. The small transmitter was operated at the lower frequency of 1604 kc and gave an output of 300 to 400 watts. The second transmitter was operated at 3088 kc and was adjusted to give an output of approximately 3 kw. The antennas used for both frequencies were of the simple "L" type construction. The pulse transmission was obtained by means of a contact wheel driven by a synchronous motor. This contactor controlled the grid bias of the transmitters in such a manner as to prevent transmission except for one short interval during each revolution of the wheel. This arrangement gave pulses of approximately 0.0005-second duration spaced at intervals of 1/30th of a second. Both transmitters were controlled in parallel from the same contactor to insure simultaneous transmission of the pulses on both frequencies. The transmitters were located at the Deal, N. J., laboratory.

For reception, field strength measuring sets operating on the double detection principle were used. These receivers were followed by amplifiers in order that currents of the order of 100 milliamperes could be obtained to operate a string oscillograph. These amplifiers used resistance coupling between stages and were so adjusted that the d-c output was proportional to the logarithm of the high-frequency input of the receiver. In this manner it was possible to measure differences in the amplitude of the radio signal of about 20 db (i.e., a range of ten to one) for a given adjustment of the receiver. If a linear amplifier had been used the range would only have been about 10 db. A square-law second detector is assumed in both cases. A large range in amplitude variation was desirable in order to insure observation of weak reflections without overloading the receiver on the ground wave. Long horizontal wires were used as antennas so as to obtain an amplitude discrimination in favor of the downcoming waves.

The output circuits of the two receivers were connected to separate

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[2] Since this work was done other investigators have performed experiments using frequencies in the same range but not simultaneously. 3009 kc—E. V. Appleton and A. L. Green, *Roy. Soc., Proc. A.*, **128**, 159, 1930. 1410 kc—P. A. de Mars, T. R. Gilliland, and G. W. Kenrick, PROC. I. R. E., January, 1931.

oscillograph vibrators and a timing wave voltage was applied to the third vibrator. This timing wave was obtained from a 1000-cycle tuning fork controlled oscillator. The latter was purposely operated in a manner which gave a distorted output wave shape so as to obtain very sharp peaks. This increased the accuracy and ease of measuring time intervals on the oscillograms.

The receiving apparatus for these experiments was located at the Cliffwood, N. J., laboratory. This gave a base line distance between the transmitter and receiver of 25 km (15 miles).

III. EXPERIMENTAL RESULTS

Numerous oscillographic records of reception of the pulse transmissions on the two frequencies were obtained for different hours of the day and night over a period of several months. The results are presented in terms of virtual height of the ionized regions, and in general verify the results of other investigators as to the variation in these heights with frequency and time of the day. The hypothesis of the existence of two distinct layers has been advanced at various times and has recently been supported by the experimental results of Appleton.[3] This hypothesis is also strengthened by the results of our tests. Other peculiarities which have been observed in this series of experiments will be explained as the individual data curves are discussed.

Test of March 27, 1929—Single Frequency of 1604 kc. Sunset 6:12 P.M.

It is interesting that the sharp rise and discontinuity in the virtual height curve of Fig. 1 is to be expected at the time when the signal reflection jumps from one layer to the other. It will be assumed that the curve shown in Fig. 2 represents the effective electron ionization at any given time, on the basis of two layers. The term "layer" is here applied to each of the two regions which are separated by a region of minimum ionization. By assuming different functions for the ionization, it is possible to calculate the different group times required to transmit a signal of any given frequency through the medium. For the case where the distribution is assumed to be parabolic,[4] it may be shown that the group time is infinite for the frequency which just penetrates the layer to the height of maximum ionization. The distribution shown in Fig. 2 would closely approximate a parabola near the maximum in the curve.

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[3] E. V. Appleton,—*Nature*, Sept. 3, 1927 and March 23, 1929; *Roy. Soc., Proc. A.*, **126**, 1930.

[4] A rectangular hyperbola having one asymptote vertical represents another distribution which gives an infinite group time. This distribution is therefore possible, but it is not probable because of the sharp drop in virtual height at sunrise. (See Fig. 8.) See also W. de Groot, *Phil. Mag.* S. 7, **10**, No. 65; October, 1930.

This critical condition may be found in two ways. In the first place it is possible to vary the frequency, assuming the ionization remains constant; in the second place it is possible to find the time at which the critical condition is obtained for any given frequency, providing the ionization varies with time in such a way that the shape of the ionization curve remains substantially the same with changes in ionization.

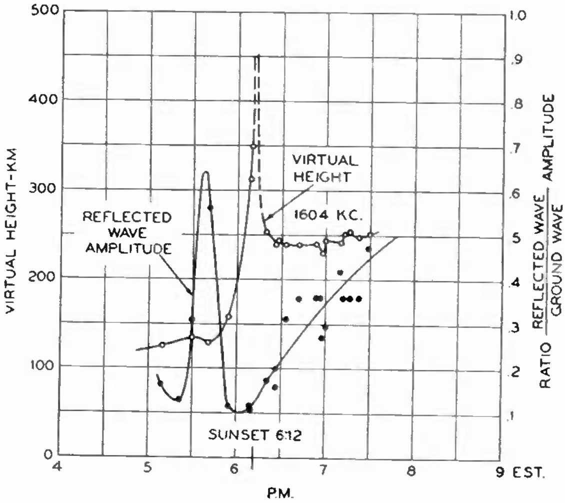

Fig. 1—Curves showing the variation in virtual height and reflected wave amplitude for a frequency of 1604 kc on March 27, 1929. The abnormally large values of virtual height at the critical point, where reflections change from the low to the high layer, are probably due to a decreased group velocity and not to an actual increase in layer height.

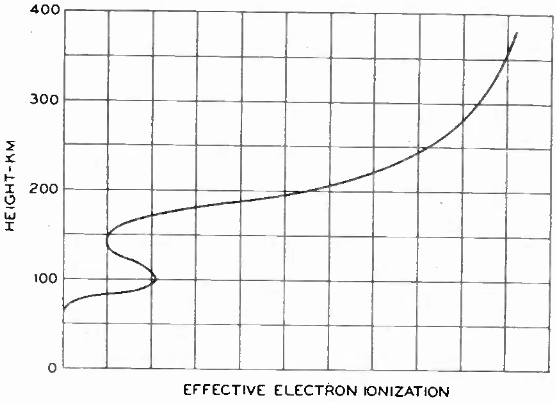

Fig. 2—Hypothetical curve showing variation in ionization as a function of height.

When the ionization in the reflecting regions decreases due to recombination and the decreased ionizing effect of the sun, the signal will have to penetrate deeper into the lower layer in order to be reflected. At the time the signal penetrates the layer just to the height of maximum ionization the virtual height should become infinite. As the ionization continues to decrease with time the signal would have to jump to the upper layer in order to be reflected. However the virtual heights will still be especially large due to the slow group velocities in the lower layer. Thus the virtual heights observed during this critical period are much greater than the actual height to which the signal penetrates the layers. The increasing virtual heights during this period (Fig. 1) correspond to reflection from the lower layer while decreasing virtual heights correspond to reflection from the upper layer.

The curves in Fig. 1 show that absorption was especially large at the time corresponding to the large values of virtual height. At 5:15 P.M. the virtual height was increasing slowly, and due to increased obliqueness of the sun's rays, ionization was decreasing in the lower absorbing regions. On two other occasions the absorption was so great that no reflections were received at all during this period. (See Figs. 3 and 7.)

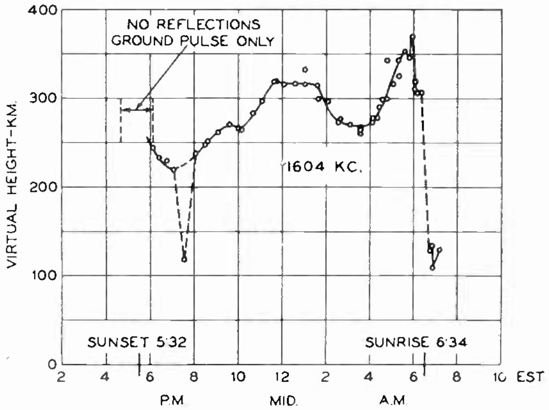

Fig. 3—Curve showing the variation in virtual height beginning before sunset on February 19 and continuing until after sunrise on February 20, 1929. Note absence of reflections during the first hour which is probably due to the high absorption at this time.

On this particular day, however, conditions happened to be favorable for receiving reflections before sunset and at 5:30 P.M. they were improving rapidly due to recombination at absorbing levels. But at 6:14 P.M. the signal had penetrated the layer to the critical height and the absorption had again increased due to the unusually slow group velocities. After the critical period had passed, the absorption again decreased and strong reflections were received from the upper layer.

Probably the reason why these large values of virtual height were not obtained more often is due to the high absorption present at such times. Although these observations might indicate that the slow group velocity explanation of the long time delay echoes is unlikely, this argument is not conclusive since it is probable that, if this explanation be true, the delay occurs in the upper layer where absorption should not be as strong as in the lower layer.

Test of February 19 and 20, 1929. Frequency 1604 kc. Sunset 5:32 P.M. February 19; Sunrise 6:34 A.M. February 20.

Fig. 3 again shows the variation of virtual height with time for a single frequency. The decrease in virtual height from 6 to 7 P.M. can be explained, (as in the curve of Fig. 1), as being due to the decreasing effect of the large group time during the critical period. The rise from 8:00 P.M. to 1:00 A.M. is probably due to decreasing ionization as recombination takes place. From 1:00 A.M. to 3:00 A.M., the virtual height fell off again showing a minimum at 3:30 A.M. It then rose to 370 km at about 45 minutes before sunrise. The dip at 3:30 A.M. has been observed on other days and at other frequencies. It also has been indicated in curves published by other experimenters.[5] The rise to a maximum at 6:00 A.M. is possibly an approach to a critical condition of ionization due to recombination. In this case the large virtual heights would be due to a slow group velocity in the upper layer. A similar effect is also found in the data for 3088 kc (Figs. 7 and 8). From the point of maximum virtual height a rather rapid falling off is observed for a period of 30 minutes, then a sharp drop to a low layer height of about 125 km. This sharp discontinuity is difficult to explain on a single layer basis. Probably what happens during this period is that the effect of radiation from the sun in the upper layer begins about 45 minutes before ground sunrise, and this radiation causes an increase in ionization in this upper layer. This would result in the first decrease in virtual height. Then about 15 minutes before ground sunrise the ionization of the lower layer is effected sufficiently so that the waves are reflected from the lower layer. This latter change would correspond to a condition of critical ionization in the lower layer and therefore cause the abrupt change shown on the curve.

One phenomenon observed was that in the early night period, about 7:30 P.M. reflection from the lower layer was found, indicating a critical condition of the ionization in this layer at this time which for a short time was great enough to cause a change in reflection from the upper to the lower layer. This low layer value of virtual height indicates that two layers are present at all times. Hollingsworth's data at 21 kc and Appleton's at 400 meters show low layer virtual heights for both day and night conditions.[6]

Test of March 28, 1929—1604 and 3088 kc observed simultaneously—Sunset 6:12 P.M.

In general the curves of Fig. 4 for two frequencies indicate that the

[5] G. W. Kenrick and C. K. Jen, PROC. I. R. E., April, 1929.

[6] J. Hollingsworth, *Jour. I. E. E.* (London), 64, 1926. E. V. Appleton and M. A. F. Barnett, *Roy. Soc., Proc.* **112–113**, 1926–1927.

virtual heights for 3088 kc are about 30 km greater than those for 1604 kc. An actual crossing of the curves at about 6:30 P.M. is shown which indicates a greater virtual height for the lower frequency than for the higher frequency. This again is due to the retarded group velocities for 1604 kc near the critical condition of ionization. During the late afternoon and through the sunset period the virtual heights measured for 3088 kc were practically constant at 250 km. This is unlike the condition obtained for the lower frequency where an abrupt change from the low to a high value was obtained at sunset. (See Fig. 1). Apparently the ionization in the lower layer had become insufficient to give reflections at 3088 kc at any considerable time before actual sunset, due to the obliqueness of the sun's rays through the atmosphere. The wave therefore penetrates the lower layer and is reflected from the

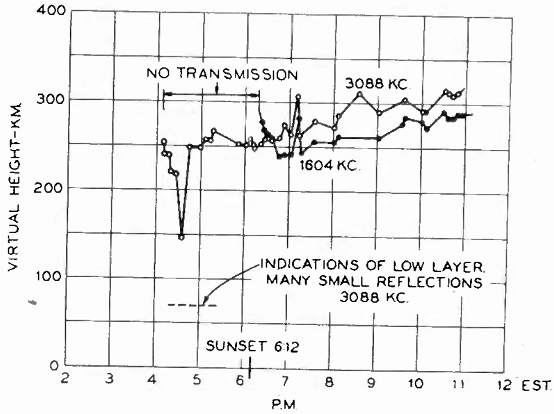

Fig. 4—Curves showing the variation in virtual height for two frequencies simultaneously (1604 and 3088 kc) on March 28, 1929. In general the curves show that the virtual height for 3088 kc is about 30 km greater than that for 1604 kc.

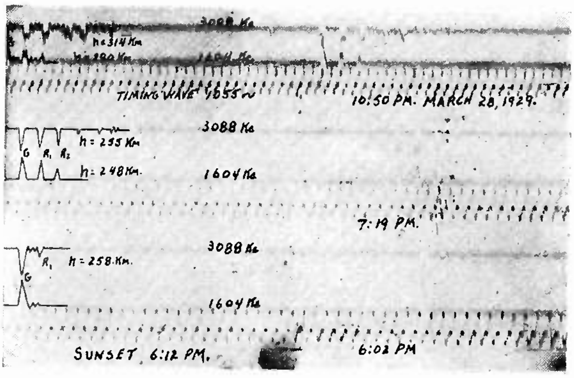

Fig. 5—Oscillograms showing reflections for two frequencies simultaneously on March 28, 1929.

upper. Appleton[2] has published curves showing this same condition at this frequency with a jump from the lower to the upper region at about 3 P.M. The presence of more than one layer is also indicated in Fig. 4 where a virtual height of about 145 km was obtained at 4:30 P.M. During this same time, 4:15 to 5:30 P.M., the oscillograms show the presence of many small multiple reflections between the main reflections, and these multiple reflections correspond to very low virtual heights of about 70 km. It is possible that this is a case of true reflection from a surface of discontinuity in the lower ionized region.

Fig. 5 shows a group of oscillograms taken between 6 and 11 P.M. and indicates the occasional complex nature of the received pulses.

Test of March 17, 1929—1604 and 3088 kc observed simultaneously—Sunset 6:03 P.M.

An unusual condition in which the virtual height for the lower frequency changed from 250 km to 125 km between 8 P.M. and midnight, was present on this day. At midnight the lower frequency gave values of virtual heights corresponding to the lower layer (125 km), while the higher frequency simultaneously gave the usual values of virtual heights corresponding to the upper layer (320 km). Fig. 6 is an oscillographic record of the received pulses on two frequencies at this time.

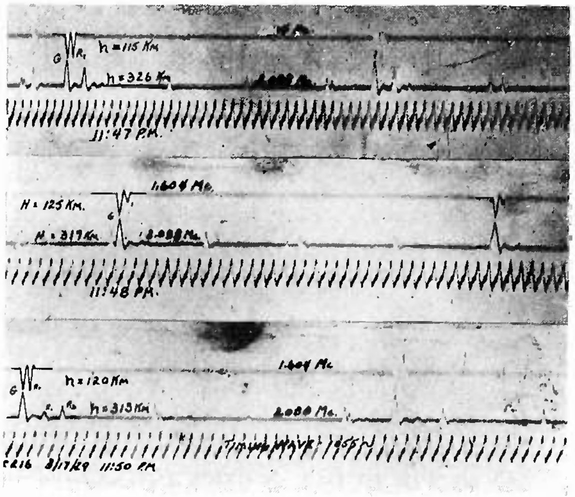

Fig. 6—Oscillograms showing reflections for two frequencies simultaneously on March 17, 1929. Note the unusual night condition of reflections taking place from the lower layer for 1604 kc and from the upper layer for 3088 kc.

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[2] *Loc. cit.*

Test of March 19 and 20, 1929—1604 and 3088 kc observed simultaneously —Sunset 6:03 P.M. March 19; Sunrise 6:04 A.M. March 20.

In Fig. 7 the part of the curve for 3088 kc from midnight through sunrise has not previously been shown. The virtual height varies in a manner similar to that for 1604 kc shown on Fig. 2 with the exception that the sharp discontinuous drop to the lower layer value is not obtained at sunrise. The virtual height falls off at sunrise but reflections continue to occur from the upper region until a considerably later time.

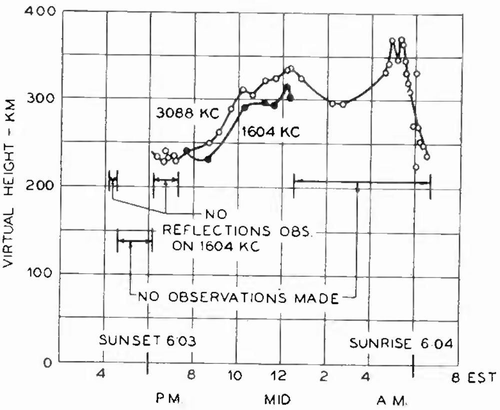

Fig. 7—Curves showing the variation in virtual height for two frequencies simultaneously (1604 and 3088 kc) on March 19–20, 1929. Note that although the virtual height decreases at sunrise for 3088 kc, the sharp drop to the lower layer is not obtained at this time.

The drop to the lower daytime value would probably occur sometime later when the ionization in the lower region becomes sufficient to give reflections for the 3088-kc frequency.

Test of April 6, 1929—1604 kc and 3088 kc observed simultaneously—Sunrise 5:26 A.M.

The parts of the curves in Fig. 8 drawn in solid lines represent actual data taken on this day. The dotted portions represent composite curves and are believed to represent a typical condition for the spring of this particular year. Since the simultaneous records show that the peak preceding sunrise for 3088 kc is much more pronounced than for 1604 kc, it appears that the rise in the virtual height curves for this period may be due to an approach to a critical condition of ionization in the upper layer.

The discontinuity in the values of virtual heights near sunrise for 1604 kc is more marked than that shown in Fig. 3, as only a few minutes elapsed between observations of virtual heights representing the upper layer and those representing the lower layer. It was possible to make the measurements of virtual heights until about 10 A.M. for this lower frequency.

For the higher frequency (3088 kc) the decrease in virtual height was rather gradual up until an hour or more after sunrise but the reflections were still taking place from the upper layer. The comparison of the behavior of these two frequencies at this time confirms in a striking manner the idea that a discontinuity should be found when the point of reflection jumps from the higher to the lower layer, while only a continuous change should be found in the case of a frequency which throughout is reflected from the upper layer. Observations were continued but the reflections were no longer observed, indicating high absorption and a probable drop to the lower layer. The curve has been continued in a dotted line and drawn to show this drop at a somewhat later time but the exact point of change was not determined. That this drop may occur at this frequency has been demonstrated by Appleton.[2]

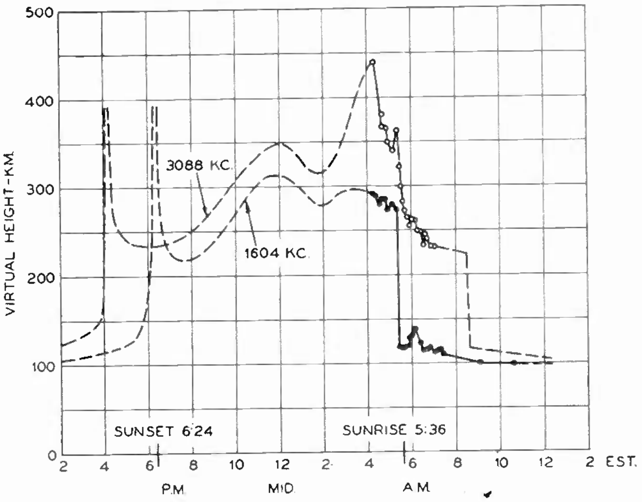

Fig. 8—Composite curves showing the variation in virtual height for two frequencies (1604 and 3088 kc) during the spring of 1929. Note particularly the solid line portion of the curves which represent actual data taken on April 6, 1929 and shows a discontinuity in the curve for 1604 kc at sunrise but not for the higher frequency of 3088 kc.

The dotted portions preceding the actual data curves are taken from average data of other days except that the discontinuity for 3088 kc where the reflections change from the lower to the upper layer (at 4 P.M.) has not been determined experimentally but has been found on certain days by Appleton.[2] This discontinuity is assumed to be present at this frequency due to the same cause which gives this phenomenon at 1604 kc.

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[2] *Loc. cit.*

DISCUSSION

One of the objects of this series of experiments was to obtain a relation between frequency and virtual height by simultaneous transmission on two frequencies. This relation was to be used in the manner indicated by Schelleng[7] to determine the ionization of the layers. Because of this discontinuity of the virtual height curve at the critical frequency this relation cannot be used to determine the ionization above the lower layer. Since no simultaneous records were obtained for the lower layer no new calculation of the ionization of this layer has been made.

Except during the critical period, the simultaneous records for the upper layer have shown that the higher frequency always corresponded to the greater virtual heights. No simultaneous records were obtained which showed upper layer reflections for 1604 kc and lower layer reflections for 3088 kc.

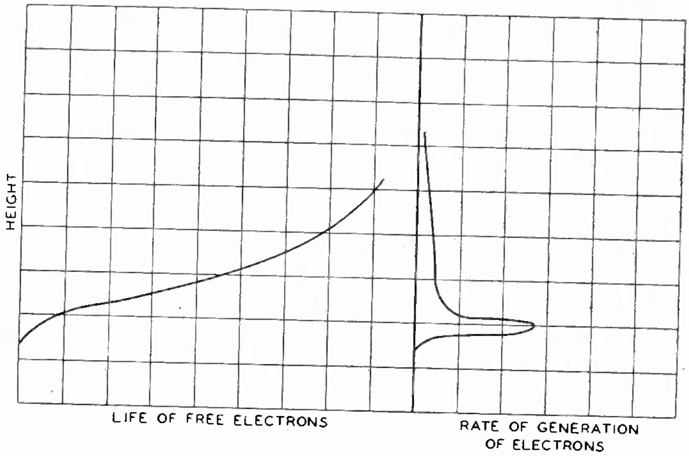

Fig. 9—Hypothetical ionization curves.

It has already been pointed out that by making the assumptions of two layers it is possible to explain the experimental data. How more than one layer could exist might be explained in several ways. In the first place if the rate of ionization and of recombination were represented as in Fig. 9, then the resulting curve for the ionization as a function of height would be represented by Fig. 2. In this case only one ionizing agency would be necessary although more than one could be effective. The second type of explanation might assume two distinct ionizing agencies with the corresponding optimum conditions for maximum ionization at different altitudes. According to this type of explanation the rate of generation of ions or electrons would actually be less in

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[7] J. C. Schelleng, PROC. I. R. E., **16**, November, 1928.

the intermediate region. Another type of explanation would assume no peculiar conditions of generation in this region, but would make the assumption that here conditions are peculiarly favorable for formation of heavy negative ions which are relatively inactive from the radio point of view. On this assumption there would be fewer electrons in this region than in the regions above and below.

Oxygen molecules and water vapor are both known to form negative ions. It may be that conditions of temperature and pressure in the region between the two layers are especially suitable for the formation of these negative ions. Consideration of this hypothesis may result in an increased knowledge of the composition of the atmosphere in this region.

The rather rapid minute-to-minute variations which we observed at intervals during the night indicate either that the atmosphere is disturbed at these heights, or that some ionizing agency is functioning which does not produce uniform ionization.[8] Since there is strong evidence that direct electromagnetic radiation from the sun is an effective ionizing agency, these rapid variations might be regarded as indicating the presence of more than one ionizing agency. Further evidence supporting this view is obtained from a study of the data presented in this paper which show that after sunset the signal reflections may jump from the upper to the lower layer. As before, however, this effect may be due to disturbed conditions in the atmosphere. It would appear that the ionization caused by other effects than electromagnetic radiation from the sun may not be uniform and such irregularities as exist may be due to this cause.

The experiments which have been described in this paper were begun at the suggestion of J. C. Schelleng and a continuation of this work is now in progress at the laboratory at Deal, N. J. It is hoped that a simpler method of obtaining virtual heights may be worked out so that accurate results may be obtained without the use of photographic equipment.

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[8] See also R. A. Heising, PROC. I. R. E., **16**, January, 1928.

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INVESTIGATION OF THE ATTENUATION OF ELECTROMAGNETIC WAVES AND THE DISTANCES REACHED BY RADIO STATIONS IN THE WAVE BAND FROM 200 TO 2000 METERS*

By

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Summary—*The paper gives the results of a large number of field strength measurements in which the relative radiation characteristics of the transmitting and receiving stations remain constant even on changing their distance.*

The constant of attenuation as a function of the wavelength was determined for the propagation of waves over land from these measurements. The method used eliminates the errors in the absolute determination of the received field strength.

The attenuation values that were obtained were used for the calculation of the received field strengths as a function of wavelength and distance in two examples (long-wave Zeppelin and long-wave airplane station).

We see that it is more important to use a long wave over land than over water. The best waves as regards propagation and radiation can be deduced from this work. The dependence of the internal efficiency of the antenna circuit and of disturbing reflections on the wavelength must be considered especially carefully.

The results of the work make possible the numerical calculation of the range of radio waves over land in the wave band between 200 and 2000 meters. They should be more reliable than the rule-of-thumb methods hitherto used.

1. Introduction

The present investigation, which was made in 1928 and 1929, takes up the attenuation of electromagnetic waves over land. Rule-of-thumb methods have hitherto been used in planning radio communication over land while over water the Austin-Cohen formula gives sufficiently reliable results. Sommerfeld[1] has indeed given the theoretical foundation according to which the received field strength can be calculated if the earth conductivity, dielectric constant, and permeability are assumed to be known. However, not only are average values for the earth constants not known, but also the effect of the space radiation on the received field strength is not considered by Sommerfeld, although it exerts an influence even in the region of "medium waves." Therefore in practice the results of experimental research must be used.

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[1] Notes refer to bibliography.

In our investigations a decision was first made as to the scope of the work. We knew from earlier investigations that the character of the surface of the earth affects the propagation of the waves, so that tests completely explaining the action of waves over land, would have to be made over all possible land formations. Such an investigation would exceed by far the available means of the Institute. Therefore we decided to select at first certain stretches for detailed investigation, and by means of differences, as compared with the action over other stretches, to see to what extent our results are applicable in general. From our later tests we considered the most used stretch, that from Berlin to Hannover, as suitable for a rather large number of measurements.

The change in the attenuation with distance over land, has previously been determined with automobile sets. This method involves very long investigations, and in addition is subject to several fundamental defects.

2. DESCRIPTION OF THE METHOD OF MEASUREMENT

We tried to avoid these difficulties by using an airplane station for our tests. This made it possible to determine a large number of measuring points in a comparatively short time, so that we even were rather independent of atmospheric changes during a series of measurements. Since the course of the airplane can be chosen without regard to the terrain, there is no great difficulty in making a measurement in any selected direction. If the flying altitude is high enough, there is also the assurance that the radiation characteristics of the test station have not been changed by ground influences during the trip.

The transmitter was in the airplane, and the receiving or measuring was done on the ground. The results thus obtained can be made directly applicable to the reverse case with the transmitter on the ground and the receiving set in the airplane, by means of the Sommerfeld reciprocity principle.[2] The following method was selected in order not to lose time by communicating from the airplane to the receiving station during the measurements that will be described later. The airplane—in this case a Junkers F 13—had a reliable clock that checked exactly with the observer's clock at the receiving station. Transmitting was done in the airplane according to a prearranged time schedule: after a call signal of one-half minute, a dash was sent several times, followed by a 20-second dash. The field intensity measuring instrument was read during this time. At the same time, the transmitting current was recorded in the airplane. This decreased slowly during a rather long series of measurements, because of the heating of the *FT* generator.

Since the deflections of the electrometer filament never were constant but fluctuated greatly and rather frequently with the shorter waves,[3,4] an average value was estimated and used as a basis in the later calculation. In many tests the fluctuation limits were determined and plotted in the diagrams (Fig. 14). In the airplane a second observer with a clock agreeing exactly with the other two clocks, drew the path-time diagram. In this manner all observed magnitudes depending on time were known, and the relation of the received field strength to distance could be derived later.

The received field strength was determined in the usual manner with the Anders field intensity measuring apparatus. Since the values

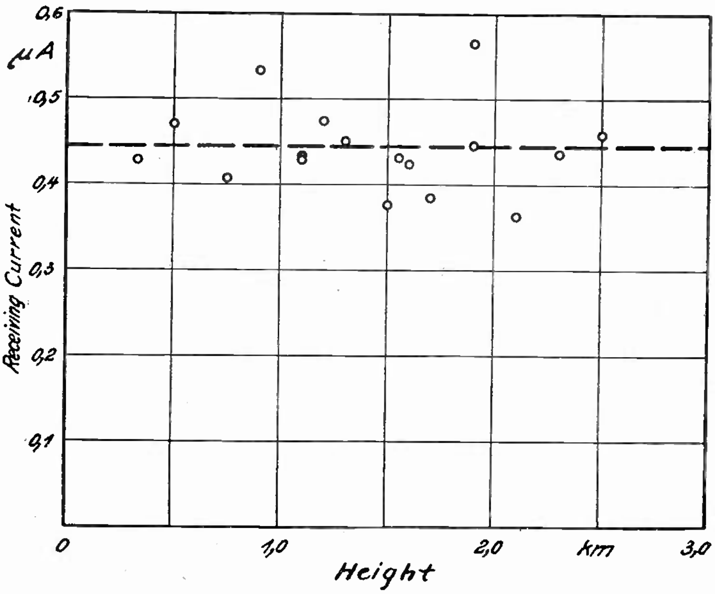

Fig. 1—Received current as a function of the altitude of the airplane, at constant distance. $\lambda = 2000$ meters.

measured for the received field strength dropped to the noise field intensity, an elevated antenna had to be used for reception. This was a 6-wire symmetrical umbrella antenna with an insulated iron mast 9 meters high. The effective height of this antenna was determined in the usual way by comparison with a loop measurement at the different wavelengths used in the investigation. The ohmic resistance R_a of the receiving antenna circuit was measured before and after each series of tests, using the resistance substitution method.

In an earlier work,[5] it was found for short distances that the field strength depended only to a slight extent on the flying altitude. In order to make certain that there would not be great errors due to changes in altitude (which averaged 300 meters in all flights) even with great distances, field intensity measurements were made in Adlershof while the airplane transmitted signals with 650- and 2000-meter wavelengths above Hannover airport at different altitudes between 100 and 2500 meters. Fig. 1 shows the result for the 2000-meter wavelength. We see that we cannot detect a systematic effect due to the flying altitude. The same is true for the 650-meter wavelength. Nevertheless the flying altitude was recorded in the measurement flights described here.†

The method of working out the results will be shown briefly in the following. The time and receiving current J_2 were observed at the field intensity measurement set, and the time and current J_s in amperes at the transmitter. In addition, we must know the distance d_{km} between transmitter and receiver, and the wavelength λ_{km}. For the absolute measurement of the received field strength, the effective height h_2 of the receiving antenna also must be known. The received field strength $\mathfrak{F}$ in v/m, which is referred to a transmitting current of one ampere, is then given by the following formula from the observations at the field intensity measurement set:

$$\mathfrak{F} = \frac{I_2 \cdot R_a}{h_2 \cdot I_S} \cdot 10^{-3} \tag{1}$$

where R_a is the resistance of the receiving antenna circuit.

In order to obtain the attenuation the further observations make use of the propagation formula in the following form:

$$\mathfrak{F} = \frac{0.12 \cdot \pi \cdot I_S \cdot h_1}{\lambda \cdot d} \cdot \sqrt{\frac{\partial}{\sin \partial}} \cdot e^{-\alpha \cdot d / \sqrt{\lambda}} . \tag{2}$$

In this formula the first factor is the electromagnetic field strength that is valid if we assume a no-loss propagation. The effective height of a transmitting antenna is defined in a general way by the formula:

$$h = \frac{1}{i_0} \int_0^l i \cdot dl \tag{3}$$

where i_0 is the current at the current loop of the antenna, and l the length of the antenna. For airplane antennas we make a change, defining h by the effective height of a ground station that is vertically

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† Recently R. L. Jones and F. M. Ryan found that the height of the airplane exerted a considerable effect on the received field strength for waves between 53.3 and 198 meters. According to their measurements, a 198-meter wave, for example, at a distance of 90 km increased the receiving field strength fourfold when the airplane climbed from 800 meters to 3000 meters. (R. L. Jones and F. M. Ryan, "Air Transport Communication," presented at the Great Lakes District Meeting of the A.I.E.E., Chicago, Ill., December 2–4, 1929.)

under the airplane and produces the same received field strength with equal transmitting current strength at the set for measuring the field strength. This definition generally gives a value differing from that in (3). The value for h in airplane sets, corresponding to this formula, cannot be measured directly. In the method given below for the determination of the constant for attenuation it is not necessary to know the value of h from either definition. The second factor in (2) takes account of the spherical shape of the earth. For the distances of not more than 600 km used by us, this factor can be made 1. The exponential function, according to Austin, takes account of the absorption of the waves during propagation between transmitter and receiver. The so-called attenuation constant is indicated by α. We have deliberately retained the original form of the exponent especially as this also can be derived theoretically.

If the product $d \cdot \mathfrak{F}$ is plotted logarithmically with d, the slope of the straight line that we expect, is a measure of α. We calculate from two pairs of values $\mathfrak{F}_1$, d_1 and $\mathfrak{F}_2$, d_2 according to the equation:

$$\alpha = \frac{\sqrt{\lambda}}{d_2 - d_1} \log \cdot \frac{\mathfrak{F}_1 \cdot d_1}{\mathfrak{F}_2 \cdot d_2} \cdot \qquad (4)$$

The representation of $\mathfrak{F} \cdot d$ as a function of d according to our observations, however, does not give a straight line for the outward or return flight, but a wavy line. In some cases there is even an increase in $d \cdot \mathfrak{F}$ at greater distances. The investigations of Barfield[7] have shown that dense woods exert an especially strong attenuating action. The observed difference in attenuation on the outward and return flights probably cannot be explained by the woods in the Letzlinger Heide only because of a lack of more exact knowledge of the underground conditions we have not made other speculations.

In planning radio installations we are not so greatly interested in the attenuation value in a given section of the range, but in its average that can be used generally for propagation over land. In order to obtain a reliable average for the attenuation along the stretch being measured, we divide the entire curve into a large number of equal parts. For each of these parts (for every 10 km in our case) we calculate α from the two values of the electromagnetic field strength at the end points according to (4). The arithmetical average of all the values obtained from this curve gives the most probable value for the desired attenuation.

Equation (3), which gives the magnitude of the received field strength taking attenuation into consideration, assumes that the attenuation by a more or less poorly conducting surface of the earth is, nevertheless, uniform. But all observations on the propagation of waves over land show that a uniform distribution of the cause of attenuation cannot be assumed. It is easy to prove, however, that our equations are correct for a small part of the wave path, if, as in our case, the reception is far from this part.

The method described here for determining the attenuation differs from other methods by the fact that it was possible for us to plot a large number of measured points quickly. In a flight between Berlin and Hannover with a flying time averaging two hours, 120 points could be obtained, in spite of interference. Later the speed of measuring was increased so that generally one point could be taken every half minute. The next question is the accuracy of the results. For the determination of the received current J_2 at the field strength measuring instrument, a large number of single readings and adjustments and several calibration curves are available. These are subject to errors in determining time and location, and particularly the inaccuracy due to the time-average for the received current, which greatly reduce the accuracy of the measurements. (See also the observations, on accuracy of measurement, given by Kiebitz during the testing of wave propagation with

TABLE I

Tabulation of All Flights

No.	Direction	Date	Time of day	Wavelength
1	Berlin-Königsberg	8/ 3/28	1047–1252	950 meters
2	Königsberg-Berlin	8/ 4/28	0840–1105	1350 meters
3	Berlin-Hannover	9/13/28	0930–1130	950 meters
4	Hannover-Berlin	9/13/28	1310–1530	950 meters
5	Berlin-Hannover	9/27/28	0930–1205	950 meters
6	Hannover-Berlin	9/27/28	1345–1550	950 meters
7	Berlin-Hannover	10/ 5/28	0915–1105	950 meters
8	Hannover-Berlin	10/ 5/28	1300–1500	950 meters
9	Berlin-Hannover	10/11/28	1417–1638	950 meters
10	Hannover-Berlin	10/12/28	0950–1210	950 meters
11	Berlin-Königsberg	10/16/28	0934–1125	950 meters
12	Königsberg-Berlin	10/16/28	1320–1625	950 meters
13	Berlin-Königsberg	10/25/28	0837–1021	950 meters
14	Königsberg-Berlin	10/25/28	1319–1545	950 meters
15	Berlin-Kassel	11/23/28	1302–1531	950 meters
16	Kassel-Berlin	11/24/28	0953–1136	950 meters
17	Berlin-Hannover	3/ 5/29	1038–1211	300 meters
18	Hannover-Berlin	3/ 5/29	1211–1308	300 meters
19	Berlin-Hannover	3/20/29	0954–1200	450 meters
20	Hannover-Berlin	3/21/29	1430–1522	450 meters
21	Berlin-Hannover	4/10/29	1420–1604	650 meters
22	Hannover-Berlin	4/12/29	1144–1610	650 meters
23	Kreisflug $\phi = 100$ km	6/18/29	0911–1209	950 meters
24	Kreisflug $\phi = 50$ km	7/10/29	0938–1054	950 meters
25	Kreisflug $\phi = 50$ km	9/ 6/29	0758–0920	950 meters
26	Berlin-Amsterdam	10/31/29	1020–1510	950 meters
27	Amsterdam-Berlin	11/ 1/29	1034–1612	950 meters
28	Berlin-Hannover	12/ 2/29	1130–1251	200 meters
29	Hannover-Berlin	12/ 2/29	1251–1345	200 meters
30	Berlin-Hannover	12/ 5/29	1143–1254	2000 meters
31	Hannover-Berlin	12/ 5/29	1254–1354	2000 meters
32	Berlin-Hannover	12/ 6/29	1048–1158	1350 meters
33	Hannover-Berlin	12/ 6/29	1158–1248	1350 meters
34	Berlin-Hannover	12/16/29	1310–1442	450 meters
35	Hannover-Berlin	12/16/29	1442–1525	450 meters

the Deutschland transmitter.8) But the relative accuracy is greater, which is shown by the fact that the points in any one series of measurements lie comparatively well along the curve. As the measurement points are plotted in our diagrams, it is possible to get an idea of the degree of relative accuracy to be expected. It can be assumed that atmospheric conditions are sufficiently constant during the comparatively short observation time. Consequently, while the value of $\mathfrak{F}$ in one series can be different from that in another series under the same conditions, it will yield correct results in any one series, that is, as regards attenuation in space. Since we are particularly interested in the determination of the value of α from the slope of the $d\mathfrak{F}$ line, all errors in measurement that are the same for each measurement point, for example, incorrect determination of the resistance or of the effective height of the receiving antenna, have no effect on the value of α.

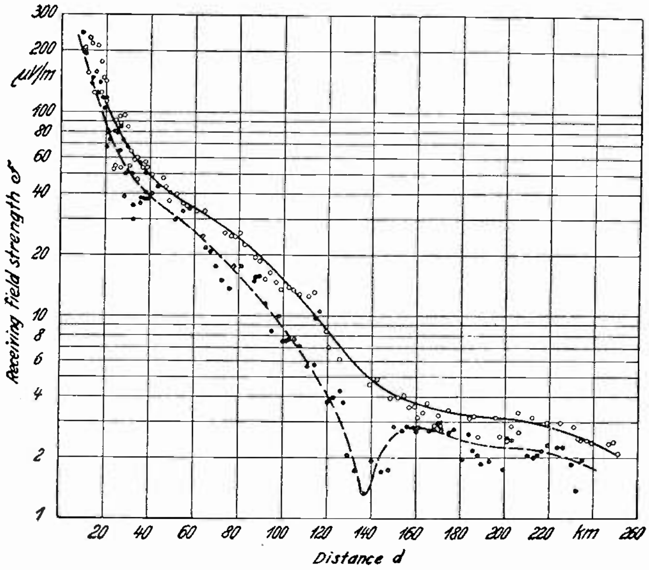

Fig. 2—Measurement points for the received field strength on the Berlin–Hannover flight (Numbers 5 and 6 in Table I) on September 27, 1928. Wavelength 950 meters.

All flights made during this investigation are summarized in Table I, which also shows the time of day.

Several series of measurements on the Berlin-Hannover flight with a wavelength of 950 meters showed that the results of the individual measurements agreed closely. Fig. 2, for instance, shows the measurement points for flights 5 and 6.

Flights 3 to 8 were made from Adlershof to Hannover in the morning, and in the opposite direction in the afternoon of the same day. All the values for the afternoon flights 4, 6, and 8 were lower than those for morning flights 3, 5, and 7. We drew average curves through the observed points, and Fig. 3 shows the three round trips. Figs. 4 gives the curves for 𝔉 determined from Fig. 3, and the $d \cdot \mathfrak{F}$ curves for flights in both directions calculated accordingly.

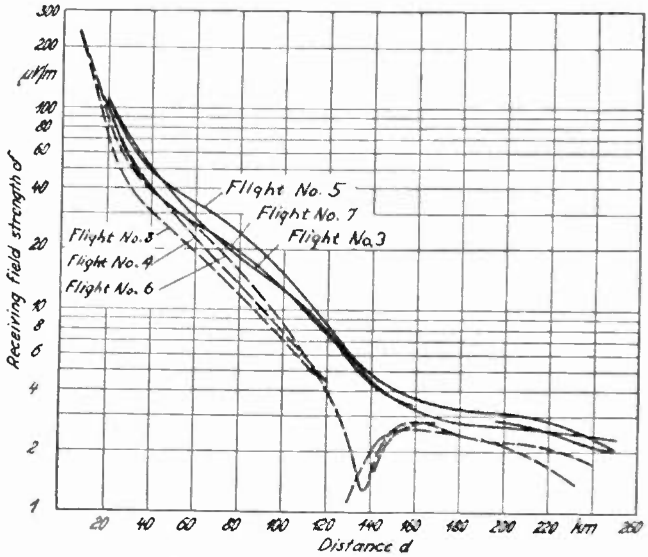

Fig. 3—Field strengths on the Berlin-Hannover route, as functions of the distance (wavelength 950 meters) for flights 3 to 8 in Table I.
———— Outward flight
– – – – – Return flight

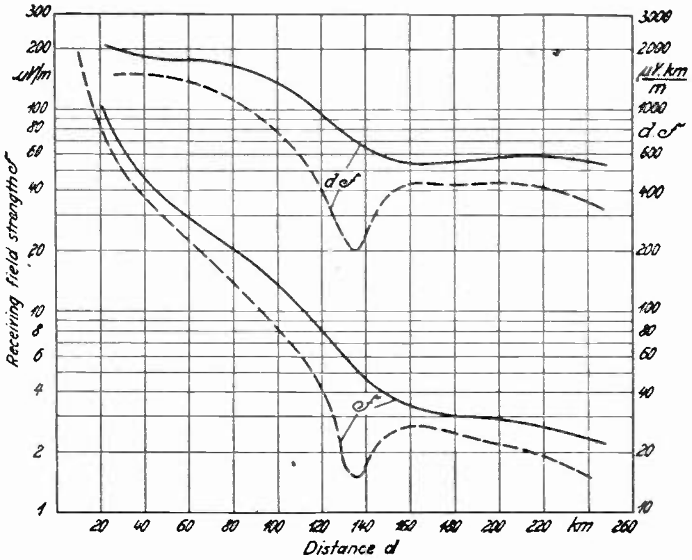

Fig. 4—Average curves for 𝔉 and $d \cdot \mathfrak{F}$ on the Berlin-Hannover route; wavelength 950 meters.

During the tests we also made flights 1 and 11 to 14 in the Berlin-Königsberg direction in order to determine the change in attenuation in other directions. They are shown in Fig. 5 for the round trips by average curves drawn through the individual points. Here also, we see that there is satisfactory agreement between the results of the individual flights. (The curves for flights 12 and 14 in Fig. 5 show a considerable displacement which, however, has no effect on the determination of the constant α by our method.)

In order to find why the points of the return flights were always below those of the outward flight, a flight (number 9) was made in the afternoon from Adlershof to Hannover, and the return flight (number 10) was made on the morning of the next day. Fig. 6 shows the observed measurements with the average curves. The points for the trip from Adlershof are shown by circles, and those taken on the trip to Adlershof are shown by points. Comparison with the curves in Fig. 4 shows that the relation of the out bound flights in Fig. 4 is of the same character as in Fig. 6. The same is true of the return flights, but $\mathfrak{F}$ values for the morning return flight are higher than those for the afternoon return flight, and the morning outward flights are also higher than the afternoon outward flight. The effect of the time of day is shown clearly by this observation.

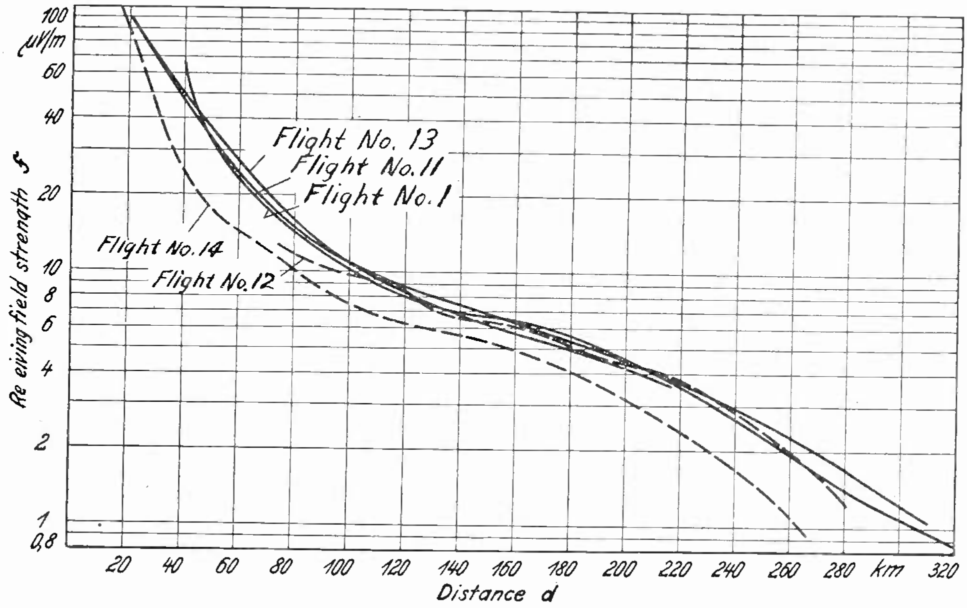

Fig. 5—Field strengths on the Berlin-Königsberg route, as functions of the distance (wavelength 950 meters) for flights on August 3, October 16 and 25, 1928 (Number 1, and 11 to 14 in Table I).

———— Outward flight

– – – – – – – Return flight

It is striking that the curves for all the return flights from Hannover to Berlin show a strong band about 130 km from Berlin, which is not found as pronounced on the outward flight. Since, as we saw above,

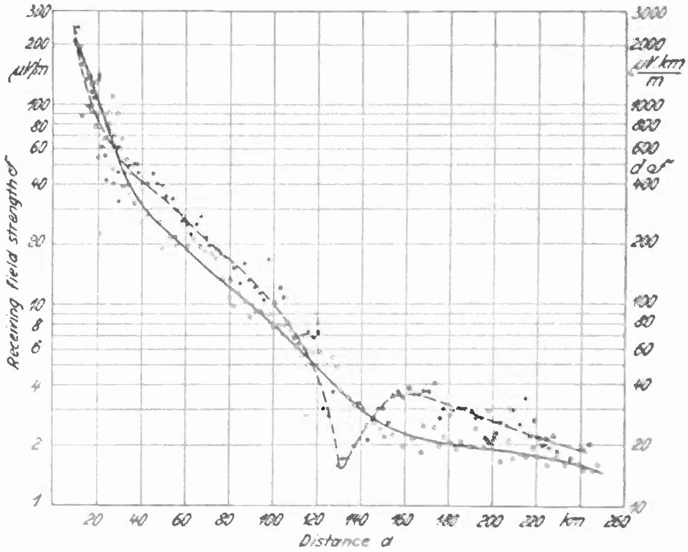

Fig. 6—Measurements of received field strengths on the Hannover-Berlin flight (Numbers 9 and 10 in Table I) on October 11 and 12, 1928; wavelength 950 meters.

———— Outward flight in afternoon
- - - - - - - Return flight in morning

Fig. 7—Measurements and average curves for $\mathfrak{F}$ and $d\cdot\mathfrak{F}$ on the Berlin-Hannover route; wavelength 200 meters. Flight on December 2, 1929 (Numbers 28 and 29, in Table I).

the characteristic shape of the curve does not depend on the time of day, and hence can be dependent only on the character of the terrain,

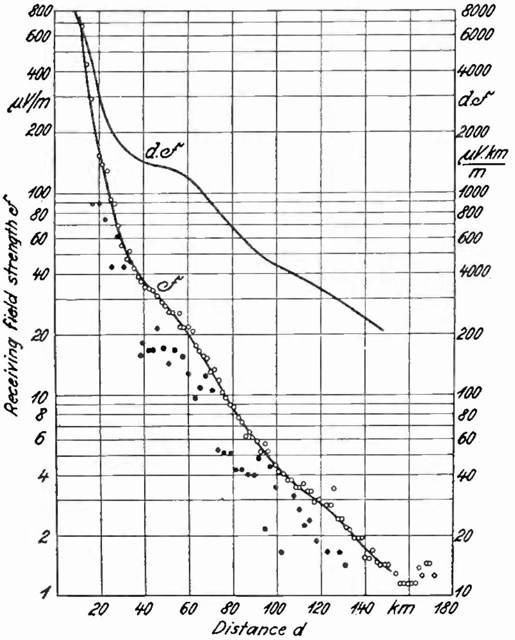

Fig. 8—Measurement points and average curves for $\mathfrak{F}$ and $d \cdot \mathfrak{F}$ on the Berlin-Hannover route; wavelength 300 meters. Flight of March 5, 1929 (Numbers 17 and 18 in Table I).

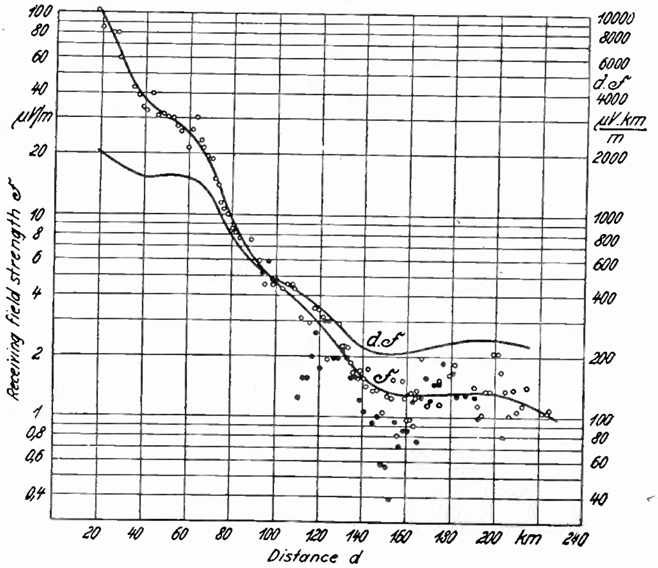

Fig. 9—Measurement points and average curves for $\mathfrak{F}$ and $d \cdot \mathfrak{F}$ on the Berlin-Hannover route; wavelength 450 meters. Flight of March 20 and 21, 1929 (Numbers 19 and 20 in Table I).

this difference can be explained only by the assumption that the oblique airplane antenna on the outward and return flights is coupled to the terrain in a different manner, thus causing the discrepancy.

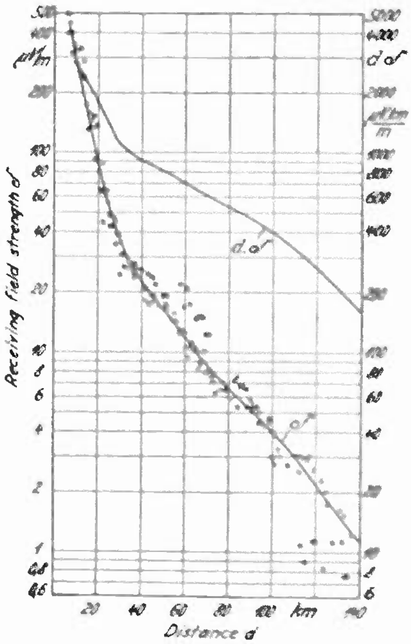

Fig. 10—Measurement points and average curves for $\mathfrak{F}$ and $d\cdot\mathfrak{F}$ on the Berlin-Hannover route; wavelength 450 meters; repeated flight. Flight on December 16, 1929 (Numbers 34 and 35 in Table I).

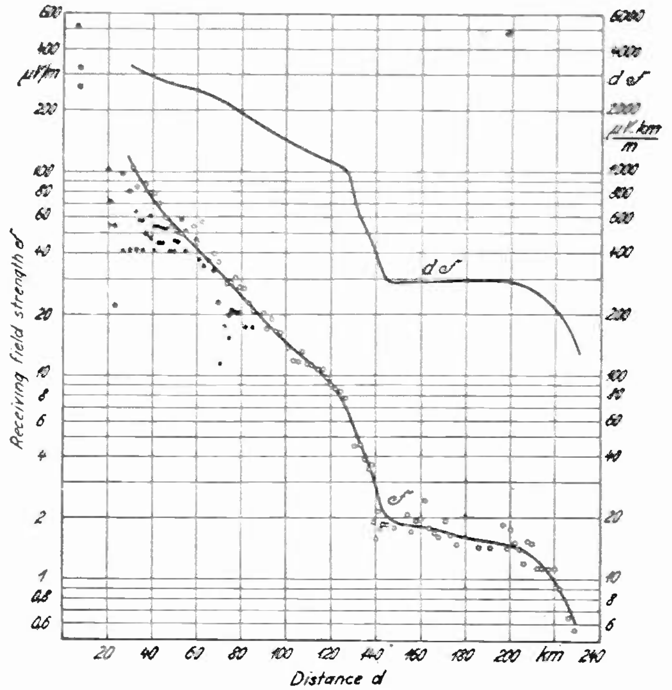

Fig. 11—Measurement points and average curves for $\mathfrak{F}$ and $d\cdot\mathfrak{F}$ on the Berlin-Hannover route; wavelength 650 meters. Flight of April 10 and 12, 1929 (Numbers 21 and 22 in Table I).

3. RESULTS OF THE TESTS

We calculated the values of α according to the above method, for waves of 200, 300, 450, 650, 950, 1350, and 2000 meters. Figs. 7 to 13, in which are shown the results of observations on flights with these waves, mark the observed values for the outward flight by circles, and the values for the return flights are marked by dots. Fig. 4 shows the corresponding curves for the 950-meter wave. The average curves shown are for the $\mathfrak{F}$ values for the outward flights. Curves $d \cdot \mathfrak{F} = f(d)$ are derived and drawn from the values $\mathfrak{F} = f(d)$. While the curves are irregular, it must be assumed that at these wavelengths a repetition of the flights would not give very divergent values for the values of attenuation. At any rate the same value of α was obtained in the two flights 19, 20 and 34, 35, with the 450-meter wave.

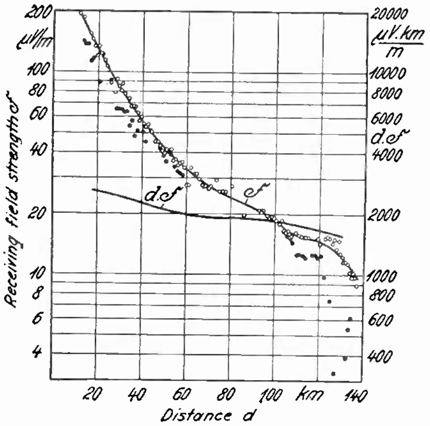

Fig. 12—Measurement points and average curves for $\mathfrak{F}$ and $d \cdot \mathfrak{F}$ on the Berlin-Hannover route; wavelength 1350 meters. Flight of December 6, 1929 (Numbers 32 and 33 in Table I).

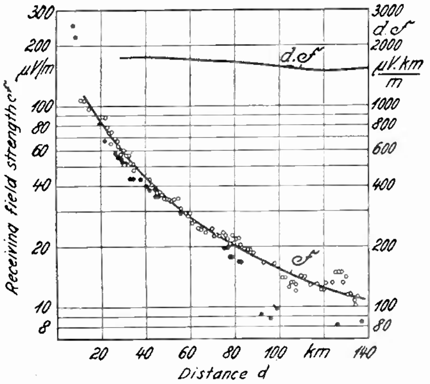

Fig. 13—Measurement points and average curves for $\mathfrak{F}$ and $d \cdot \mathfrak{F}$ on the Berlin-Hannover route; wavelength 2000 meters. Flight of December 5, 1929 (Numbers 30 and 31 in Table I).

The α values were obtained from all these curves for every 10 km, according to the method described above. The average was taken of all α values from one flight, and the averages thus obtained are given in Table II.

TABLE II

α VALUES MEASURED ON THE BERLIN-HANNOVER ROUTE (OUTWARD FLIGHT)

λ m	Flight Number in Table I	α For Total Route		For $d = 30\ldots120$ km
		Stretch	α	
200	28	30...120	0.0096	0.0096
300	17	10...150	0.0143	0.0100
450	19	20...210	0.0076	0.0118
	34	10...140	0.0138	
650	21	30...230	0.0134	0.0099
950	3, 5, and 7	20...250	0.0056	0.0076
1350	32	20...140	0.0058	0.0044
2000	30	20...130	0.0006	0.0027

In the Königsberg-Berlin direction a value of $d = 0.0061$ was obtained for the 1350-meter wave in the measured route 20—340 km.

In order to prove that approximately the same propagation constants would be found in other directions, we made flights with $\lambda = 950$ meters, first to Königsberg and then to Kassel, where we expected to find great differences as compared with other directions, because the flight was over the Harz mountains. The measurement points and curves for $\mathfrak{F}$ and $d \cdot \mathfrak{F}$ from the Berlin-Königsberg flight with a 950-meter wave are given in Fig. 5. The measurement points and curves for the Berlin-Kassel flight are shown in Fig. 14. The values of α for the entire flight are again calculated as the average of all individual values over

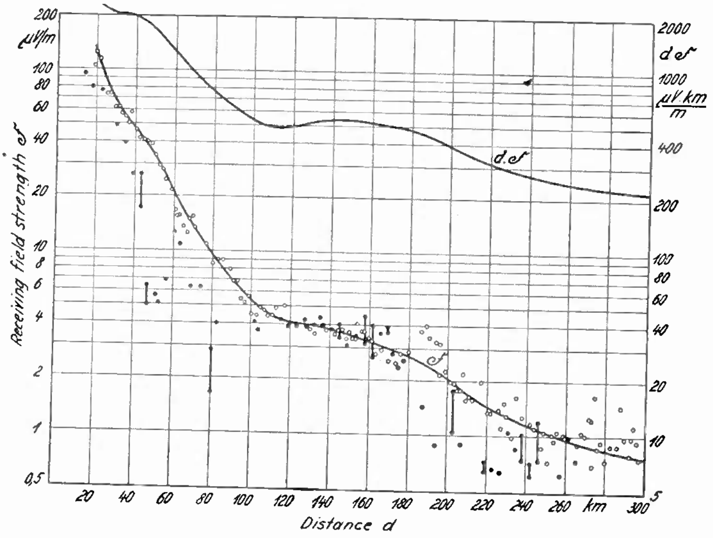

Fig. 14—Measurement points and average curves for $\mathfrak{F}$ and $d \cdot \mathfrak{F}$ on the Berlin-Kassel route; wavelength 950 meters. Flight of November 23 and 24, 1928. (Numbers 15 and 16 in Table I).

10-km stretches and are given in Table III. As a matter of fact, there is greatly increased attenuation for the Berlin-Kassel stretch. The average of all three directions is $\alpha = 0.0073$.

TABLE III

α FOR DIFFERENT DIRECTIONS (OUTWARD FLIGHT)
$\lambda = 950$ m

Direction	α	Average α	Measurement Route in km	Flight No. in Table III
Berlin-Amsterdam	0.0058		40–580 km	26
Berlin-Königsberg	0.0067	0.0073	20–320 km	1, 11, 13
Berlin-Kassel	0.0093		20–260 km	15

Fig. 15 shows the result of a Königsberg-Berlin flight (number 2) with a 1350-meter wave. Within the limits of anticipated accuracy it gives values similar to flight 32 from Berlin to Hannover. The calculated α value is 0.0061.

In order to trace the attenuation over a greater distance, a flight (numbers 26, 27) was made on October 31 and November 1, 1929 beyond Hannover to Amsterdam. Fig. 16 shows the points that were obtained, and the curves for $\mathfrak{F}$ and $d\mathfrak{F}$. This flight to Amsterdam was studied especially thoroughly, and the results are listed in Table IV. The α value for the outward and return flight was calculated for every 10 km of the total distance of more than 500 km. We see that these values fluctuate greatly. The α values were also determined from individual sections about 100 km long. The values in Table IV show the fluctuations of the α value over a rather great distance. This flight shows

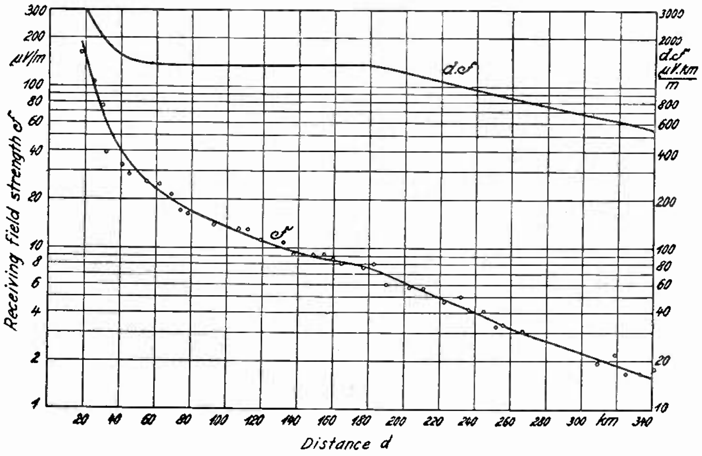

Fig. 15—Measurement points and average curves for $\mathfrak{F}$ and $d \cdot \mathfrak{F}$ on the Königsberg-Berlin route; wavelength 1350 meters. Flight of August 4, 1928 (Number 2 in Table I).

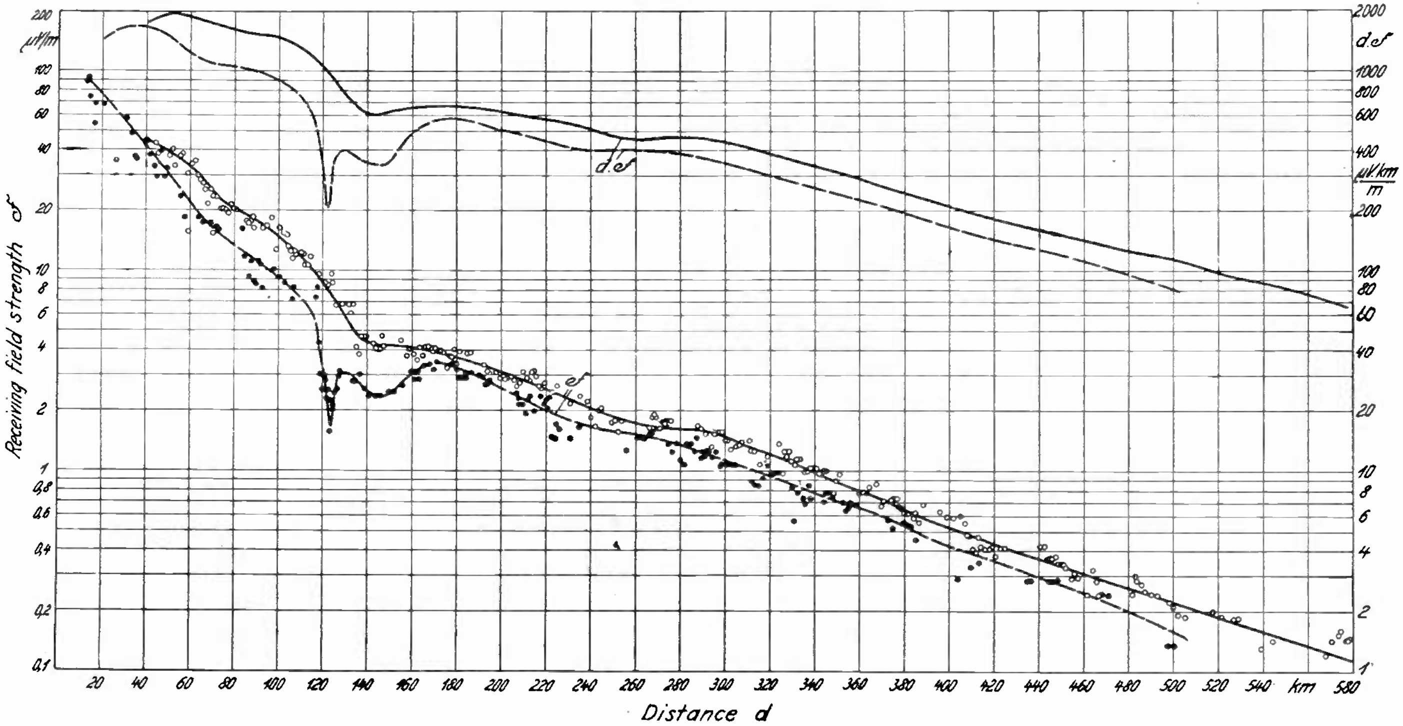

Fig. 16—Measurement points and average curves for 𝔉 and $d \cdot 𝔉$ on the Berlin-Amsterdam flight; wavelength 950 meters. Flight of October 31 and November 1, 1929 (Numbers 26 and 27 in Table I).

clearly how incorrect it would be to try to derive a generally valid α value from the values measured between two fixed points.

TABLE IV

α VALUES MEASURED ON THE BERLIN-AMSTERDAM ROUTE
$\lambda = 950$ m Flights 26 and 27

Section	Outward	Return	Average	Average			Average
				Section	Outward	Return	
40– 50 km	−0.0106	+0.0110	+0.0002	40–120 km	0.0058	0.0220	0.0139
50– 60	+0.0040	+0.0225	+0.0132				
60– 70	+0.0104	+0.0094	+0.0099				
70– 80	+0.0072	+0.0045	+0.0058				
80– 90	+0.0011	+0.0047	+0.0029				
90–100	+0.0052	+0.0117	+0.0084				
100–110	+0.0126	+0.0142	+0.0134				
110–120	+0.0234	+0.0975	+0.604				
120–130 km	+0.0311	−0.0372	−0.0030	120–220 km	0.0058	−0.0042	+0.0008
130–140	+0.0202	+0.0142	+0.0173				
140–150	−0.0032	±0	−0.0016				
150–160	−0.0047	−0.0326	−0.0186				
160–170	−0.0014	−0.0128	−0.0071				
170–180	±0	−0.0016	−0.0008				
180–190	+0.0014	+0.0052	+0.0033				
190–200	+0.0063	+0.0072	+0.0068				
200–210	+0.0047	+0.0062	+0.0054				
210–220	+0.0034	+0.0096	+0.0065				
220–230 km	+0.0054	+0.0052	+0.0053	220–320 km	0.0036	0.0035	0.0036
230–240	+0.0054	+0.0022	+0.0038				
240–250	+0.0081	+0.0014	+0.0048				
250–260	+0.0043	+0.0011	+0.0027				
260–270	±0	+0.0018	+0.0009				
270–280	−0.0022	+0.0032	+0.0005				
280–290	+0.0014	+0.0031	+0.0022				
290–300	+0.0034	+0.0062	+0.0048				
300–310	+0.0067	+0.0067	+0.0067				
310–320	+0.0047	+0.0061	+0.0054				
320–330 km	+0.0092	+0.0065	+0.0078	320–420 km	0.0075	0.0071	0.0073
330–340	+0.0061	+0.0076	+0.0068				
340–350	+0.0090	+0.0049	+0.0070				
350–360	+0.0040	+0.0070	+0.0055				
360–370	+0.0061	+0.0065	+0.0063				
370–380	+0.0114	+0.0094	+0.0104				
380–390	+0.0074	+0.0072	+0.0073				
390–400	+0.0085	+0.0072	+0.0078				
400–410	+0.0063	+0.0090	+0.0076				
410–420	+0.0065	+0.0058	+0.0062				
420–430 km	+0.0065	+0.0054	+0.0060	420–580 km	0.0062	0.0065	+0.0063
430–440	+0.0014	+0.0065	+0.0040				
440–450	+0.0108	+0.0072	+0.0090				
450–460	+0.0040	+0.0068	+0.0054				
460–470	+0.0067	+0.0072	+0.0070				
470–480	+0.0052	+0.0079	+0.0066				
480–490	+0.0063	+0.0085	+0.0074				
490–500	+0.0067	+0.0072	+0.0070				
500–510	+0.0072						
510–520	+0.0047						
520–530	+0.0072						
530–540	+0.0063						
540–550	+0.0083						
550–560	+0.0063						
560–570	+0.0081						
570–580	+0.0043						

The question arises, which average α value should be used in forecasting results on new stretches. We believe results of general validity are obtained if the calculation of the α values to be used for such purposes is limited for all wavelengths to those measured over similar stretches having similar initial and end points whose 𝔉 values show no noteworthy deviations from a regular curve. This is the case to some extent in a region between 30 and 120 km on the Berlin-Hannover route.

The averages for α and $\alpha/\sqrt{\lambda}$ for this stretch are listed in Table V and are shown graphically in Fig. 12. These values for α are also given in Table II for comparison with the values determined for the entire stretch.

TABLE V

VALUES FOR $\frac{\alpha}{\sqrt{\lambda}}$ (30. . . . 120 km)

f kHz	λ m	α	$\frac{\alpha}{\sqrt{\lambda}}$
1500	200	0.0096	0.0214
1000	300	0.0100	0.0183
667	450	0.0118	0.0176
462	650	0.0099	0.0123
316	950	0.0076	0.0078
222	1350	0.0044	0.0038
150	2000	0.0027	0.0019

The results of our tests require further explanation on two points. Table IV shows several negative values for the attenuation. This means that there is a greater field strength at a receiver than would be expected without the action of space attenuation, so that the field strength decreases less rapidly than according to the distance law $1/d$.

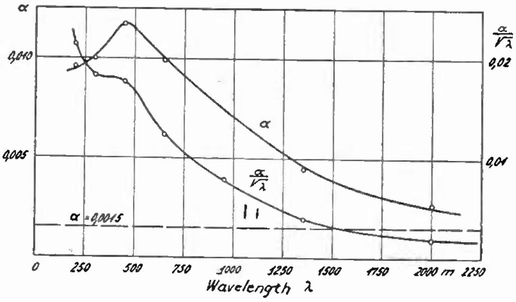

Fig. 17—Attenuation α and $\alpha/\sqrt{\lambda}$ as functions of the wavelength.

The negative attenuation at some points can be explained by the fact that electromagnetic energy from adjacent regions flows in behind a region of strong absorption according to Huyghen's principle. In certain cases the phenomenon can be explained more simply by directional conductivity influences.

In addition, the curve $\alpha = f(\lambda)$ in Fig. 17 shows a maximum at about $\lambda = 500$ meters. Since, in general, a maximum appears due to the opposing action of two functions increasing at different rates, it must be possible to ascribe the attenuation to two different causes. As a matter of fact, the so-called "limiting waves" and the "short" waves act far differently than the "medium" and "long" waves. In the region of limiting and short waves there enters a new phenomenon in addition to the simple propagation of the Hertz solution and the Austin attenuation. The wave reflected from the Heaviside layer causes a larger average field strength at a receiving station than without this reflection, and therefore it acts to reduce the attenuation. In the presence of space radiation there is generally fading, which makes is extraordinarily difficult to measure the received field strength. The values observed for the 450-, 300-, and 200-meter waves, shown in Figs. 7 to 10, cannot claim the same reliability as the measurements with the longer waves.

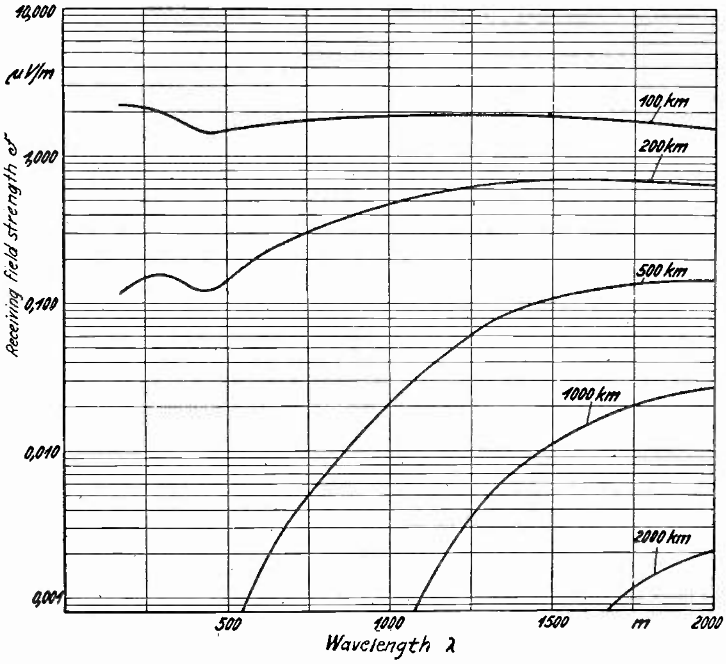

Fig. 18—Received field strength for 1 meter-ampere as a function of the wavelength, for different distances over land.

If the propagation is determined by field strength measurements near the transmitter, where the contribution from space radiation in the case of limiting and short waves is so small that it can be disregarded as compared with the amount transmitted direct by propagation along the ground, we get very high values for α. For example, Bäumler (verbal communication) finds a value of $\alpha/\sqrt{\lambda}=0.6$ and $\alpha=0.147$ for $\lambda=66$ meters at a distance of 4 wavelengths. If more such measurements were available, these values and those given for the long waves could be used to find a law for $\alpha=f(\lambda)$ representing the effect of earth damping alone.

4. CALCULATION OF THE RANGE OF RADIO STATIONS

The values in Table V are used in the following in order to calculate the field strengths as functions of the wavelengths for different distances.

TABLE VI

RECEIVED FIELD STRENGTHS IN $\mu v/m$ FOR 1 METER-AMPERE FOR DIFFERENT WAVELENGTHS AND DISTANCES OVER LAND

(Calculated From the Values in Table VII)

λ m	$d=100$ km	$d=200$ km	$d=500$ km	$d=1000$ km	$d=2000$ km
200	2.20	0.130	$0.086\cdot 10^{-3}$	—	—
300	2.00	0.158	$0.262\cdot 10^{-3}$	—	—
450	1.43	0.122	$0.243\cdot 10^{-3}$	—	—
650	1.69	0.246	$2.58\cdot 10^{-3}$	—	—
950	1.81	0.415	$16.0\cdot 10^{-3}$	$0.166\cdot 10^{-3}$	—
1350	1.90	0.660	$84.0\cdot 10^{-3}$	$6.25\cdot 10^{-3}$	$0.070\cdot 10^{-3}$
2000	1.55	0.640	$144.0\cdot 10^{-3}$	$27.8\cdot 10^{-3}$	$2.07\cdot 10^{-3}$

In Table VI the received field strengths for different wavelengths and distances are calculated for one meter-ampere. In Fig. 18 the values from Table VI are shown with λ as abscissa. It should be noted in the curves for 100 and 200 km, that the field strengths below $\lambda=450$ meters increase again with decreasing wavelengths, which is exactly the opposite of what one might expect. Obviously this is a result of space radiation, the effect of which we have already determined in Fig. 17. The form shown in Fig. 19, with d as abscissa, is more convenient for many practical purposes.

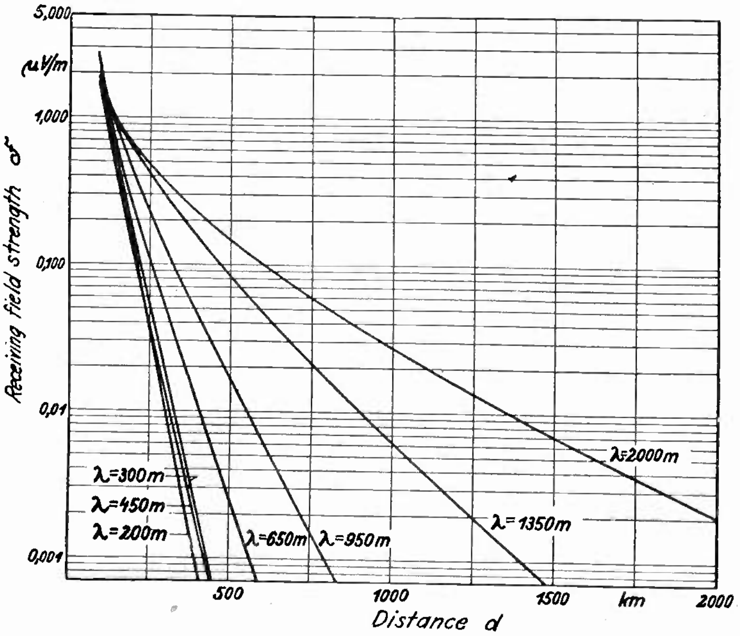

Fig. 19—Received field strength for 1 meter-ampere as a function of the distance over land, for different wavelengths.

In Table VII are given the calculated field strengths over sea water with the value for $\alpha = 0.0015$, as measured by Austin. The values of Table VII are plotted in Figs. 20 and 21. In Fig. 20, the best waves for the various distances are shown by a dotted line.

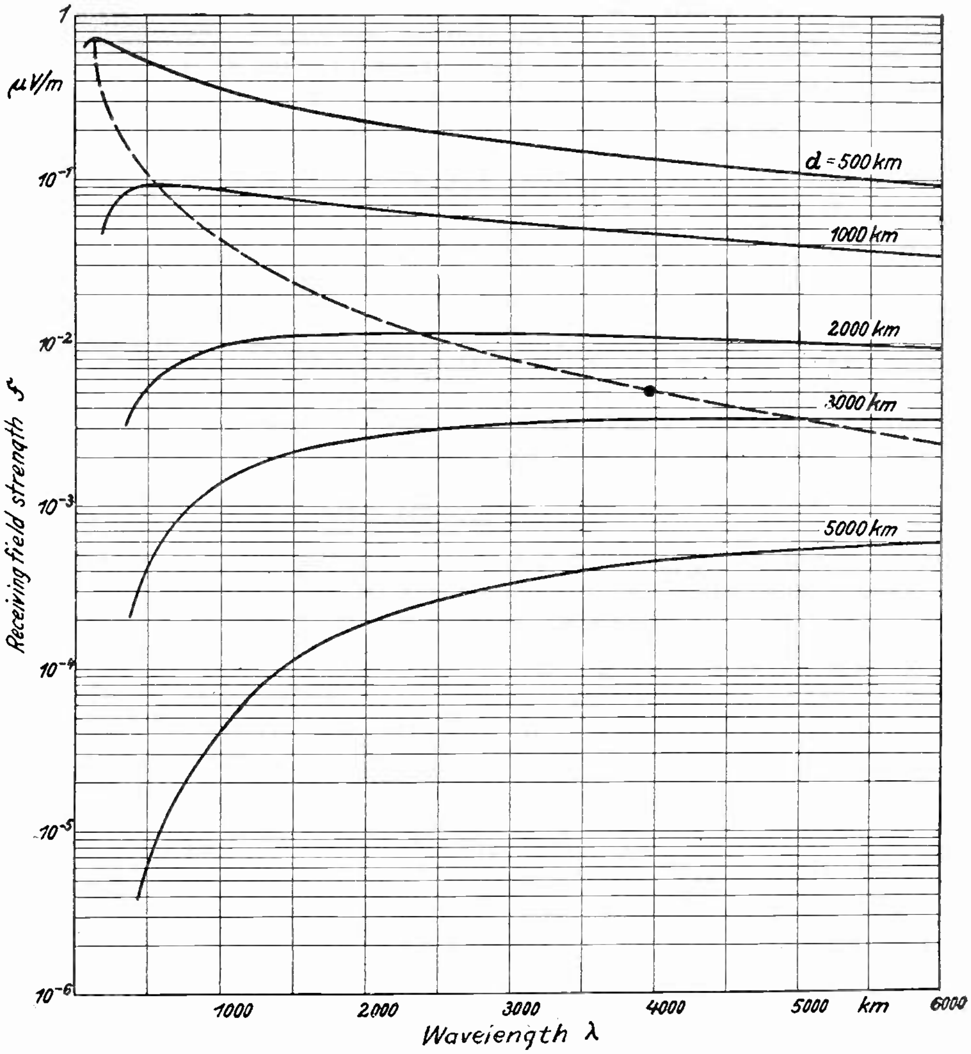

Fig. 20—Received field strength for 1 meter-ampere as a function of the wavelength, for different distances over water.

A more convenient representation of the range of a radio installation is obtained if, for a constant received field strength, one states the distances that can be covered with a given antenna power. From the values in Table VI and VII we can derive the amount of the product $J \cdot h$ in meter-amperes that is necessary in order to reach a certain distance with a field strength of $1 \mu v/m$. For example, we see from Table VI that with $\lambda = 450$ meters at a distance of 200 km, $0.122 \mu v/m$ can be produced by 1 meter-ampere and we calculate from this, that at the same distance with the same wave, $1\mu v/m$ will be produced by $1/0.122 = 8.2$ meter-amperes. Table VIII was derived in this manner, which gives

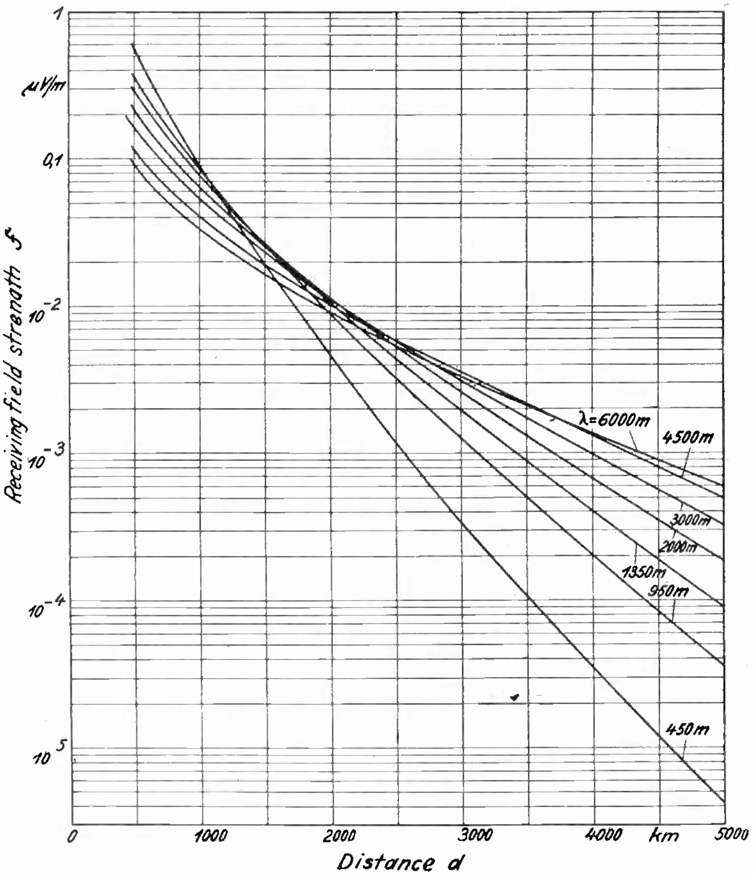

Fig. 21—Received field strength for 1 meter-ampere as a function of the distance over water, for different wavelengths.

TABLE VII

RECEIVED FIELD STRENGTHS IN $\mu v/m$ FOR 1 METER-AMPERE FOR DIFFERENT WAVELENGTHS AND DISTANCES OVER WATER

(Calculated with the Austin value $\alpha = 0.0015$)

λ m	$d = 500$ *km*	$d = 1000$ *km*	$d = 2000$ *km*	$d = 3000$ *km*	$d = 5000$ *km*
450	0.547	$89.0 \cdot 10^{-3}$	$4.72 \cdot 10^{-3}$	$0.335 \cdot 10^{-3}$	$4.19 \cdot 10^{-6}$
950	0.367	$85.3 \cdot 10^{-3}$	$9.13 \cdot 10^{-3}$	$1.27 \cdot 10^{-3}$	$35.8 \cdot 10^{-6}$
1350	0.292	$77.5 \cdot 10^{-3}$	$10.6 \cdot 10^{-3}$	$1.95 \cdot 10^{-3}$	$89.4 \cdot 10^{-6}$
2000	0.221	$65.4 \cdot 10^{-3}$	$11.2 \cdot 10^{-3}$	$2.60 \cdot 10^{-3}$	$187.0 \cdot 10^{-6}$
3000	0.163	$53.9 \cdot 10^{-3}$	$11.2 \cdot 10^{-3}$	$3.13 \cdot 10^{-3}$	$326.0 \cdot 10^{-6}$
4500	0.118	$41.4 \cdot 10^{-3}$	$10.2 \cdot 10^{-3}$	$3.36 \cdot 10^{-3}$	$491.0 \cdot 10^{-6}$
6000	0.092	$34.1 \cdot 10^{-3}$	$9.2 \cdot 10^{-3}$	$3.33 \cdot 10^{-3}$	$588.0 \cdot 10^{-6}$

the necessary value of $J \cdot h$ in communicating over land, for a desired receiving field strength of $1\mu v/m$. Table IX gives the corresponding figures for radio communication over water. The diagram in Fig. 22

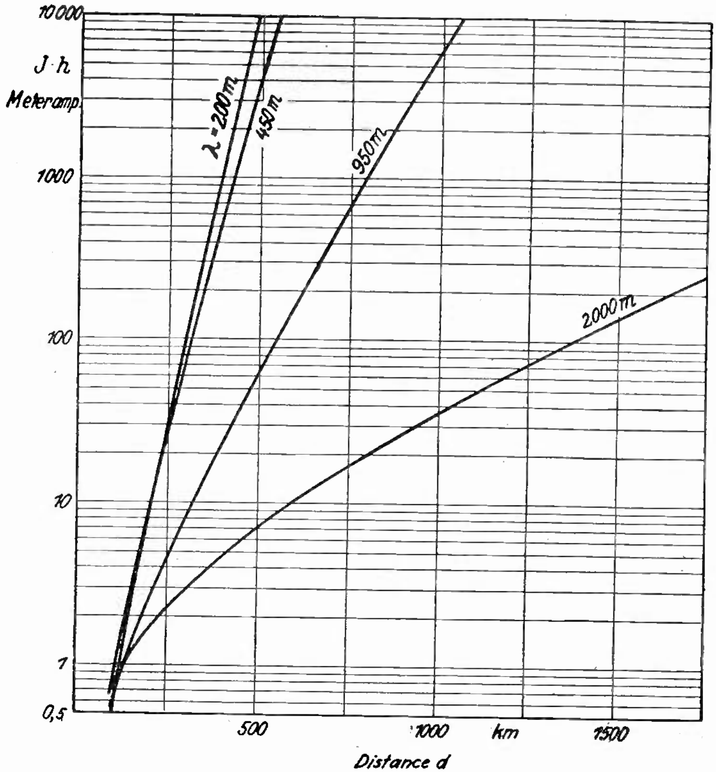

Fig. 22—$J \cdot h$ in meter-amperes for a received field strength of $1 \mu v/m$ at different wavelengths, as a function of distance d over land.

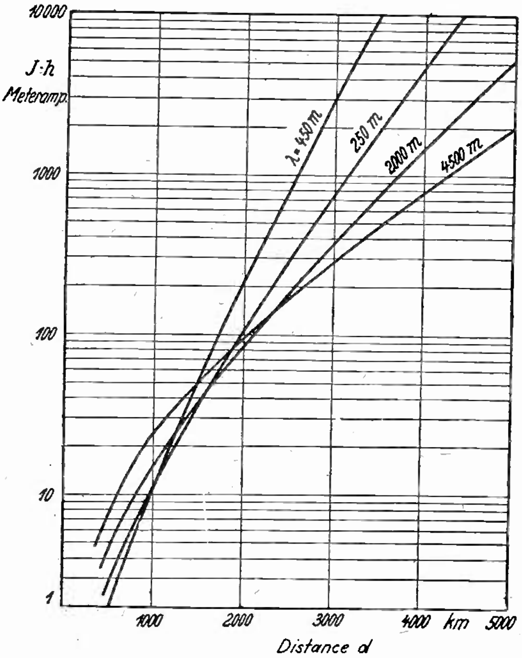

Fig. 23—$J \cdot h$ in meter-amperes for a received field strength of $1 \mu v/m$ at different wavelengths, as a function of the distance d over water.

shows the curves for Table VIII and the corresponding curves for Table IX are given in Fig. 23.

TABLE VIII

$J \cdot h$ IN METER-AMPERES FOR A RECEIVED FIELD STRENGTH OF $1\mu v/m$ WITH DIFFERENT WAVELENGTHS λ, IN RELATION TO DISTANCE d OVER LAND

λ = 200 m		λ = 450 m		λ = 950 m		λ = 2000 m	
d km	$J \cdot h$ Meter-amp.	d km	$J \cdot h$ Meter-amp.	d km	$J \cdot h$ Meter-amp.	d km	$J \cdot h$ Meter-amp.
100	0.455	100	0.70	100	0.552	100	0.645
200	7.70	200	8.20	200	2.41	200	1.56
500	11600.0	500	4120.0	500	62.5	500	6.95
				1000	6020.0	1000	36.0
						2000	483.0

TABLE IX

$J \cdot h$ IN METER-AMPERES FOR A RECEIVING FIELD STRENGTH OF $1\mu v/m$ WITH DIFFERENT WAVELENGTHS λ, IN RELATION TO DISTANCE d OVER WATER

λ = 450 m		λ = 950 m		λ = 2000 m		λ = 4500 m	
d km	$J \cdot h$ Meter-amp.	d km	$J \cdot h$ Meter-amp.	d km	$J \cdot h$ Meter-amp.	d km	$J \cdot h$ Meter-amp.
500	1.83	500	2.72	500	4.52	500	8.48
1000	11.1	1000	11.7	1000	15.3	1000	24.2
2000	212.0	2000	109.5	2000	89.2	2000	98.0
3000	2990.0	3000	788	3000	385.0	3000	298.0
				5000	5340.0	5000	2004.0

For a given value of $J \cdot h$ in meter-amperes these curves permit the determination of the distance that can be covered if a field strength $1\mu v/m$ is to appear at the receiver. These ranges for $1\mu v/m$ received field strength are shown in Fig. 24 and 25 as functions of the wave-

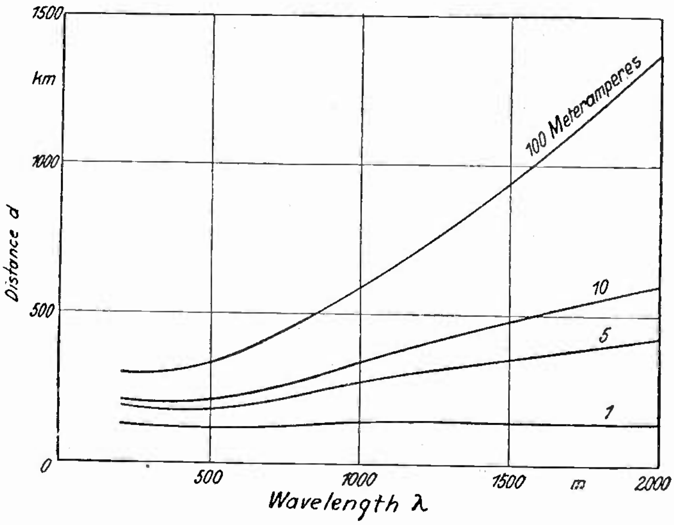

Fig. 24—Range d over land for a received field strength of $1\mu v/m$ as a function of wavelength, for different values of $J \cdot h$.

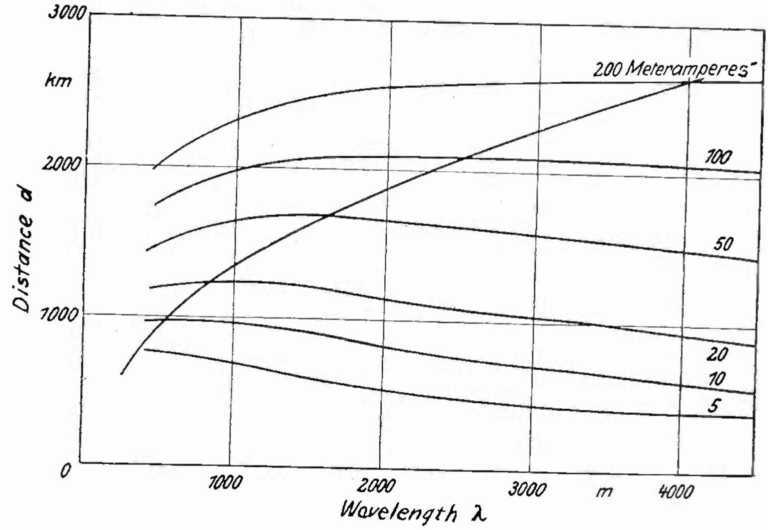

Fig. 25—Range d over water for a received field strength of $1\mu v/m$ as a function of wavelength, for different values of $J\cdot h$.

length. They give maximum ranges. With the propagation of waves over water it can be seen that these maxima lie on a parabola with the equation $d=2\sqrt{\lambda}/\alpha$, and with $\alpha=0.0015$, we get

$$d_{km} = 1330\cdot\sqrt{\lambda}_{km}.$$

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ON THE CALCULATION OF RADIATION RESISTANCE OF ANTENNAS AND ANTENNA COMBINATIONS*

BY

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Summary—*It is shown that there are two methods for calculating the radiation of antennas and antenna systems. One depends on the consideration of the electromagnetic field produced by the radiating system, that is, on the integration of the Poynting vector over a surface enclosing the system. The other is based on a consideration of the electromagnetic phenomena on the conductor itself. The identity of the two methods is demonstrated. The second method is much simpler to treat formally and gives clearer results. The calculation of the radiated power is especially simple, using a law that provides a connection between the radiated power and the Hertzian vector for the system under consideration. This hitherto unknown law is derived. The radiated power of any arbitrarily loaded antenna and of short-wave antenna systems with parallel elements is calculated by means of this law.*

THERE are two methods for calculating the radiation or radiation resistance of antennas and antenna combinations. One depends on the integration of the Poynting vector produced by the radiating system in question, over a surface enclosing this system. The other is based on direct integration along the conductor, and specifically on the calculation of the power which in turn is determined by the phase displacement between the electric and magnetic field around the conductors. This phase displacement is a result of the finite propagational velocity of the electromagnetic field and is generally specified by the retarded expressions for all fields. Both methods are equivalent and one can be converted into the other by means of Gauss' law. A. Pistolkors[1] has mentioned the formal and essentially simpler second method for the calculation of the radiation of antenna combinations whose elements are several half waves long.

The derivation used by A. Pistolkors can be considerably simplified by applying a law which relates the radiation to the Hertzian vector for the system. We shall show in the following that a simple relation exists between the radiating properties of a conductor and its Hertizan vector and that the radiation resistance may then be obtained without further integration. The Hertzian vector for any oscillating linear conductor can be readily determined. This has been done by the author[2]

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[1] A. Pistolkors, PROC. I.R.E., 17, 562, 1929.

[2] R. Bechmann, PROC. I.R.E., 19, 461, 1931; *Jahrbuch für drahtl. Telegr.*, 36, 182 and 201, 1930.

in a general manner. The electric fields from antennas and antenna systems may then be obtained in a straightforward manner from the Hertzian vector. The resulting expressions are of simple form as was previously shown.[2]

The two above-mentioned methods for determining the radiation from antennas and antenna systems are identical. This is a result of the Poynting energy law of electrodynamics. In the following we shall consider briefly the connection between the two methods and derive the above-mentioned relation between the radiation and the Hertzian vector.

The radiation from an oscillating electric system such as an antenna or an antenna array is given by the Poynting vector $\overline{P}$. This is defined in the usual manner by

$$\overline{P} = \frac{c}{4\pi} [\overline{E}, \overline{H}] \tag{1}$$

in which $\overline{E}$ and $\overline{H}$ are the electric and magnetic fields produced by the radiators considered. The total radiation S of the system is obtained by integrating the normal components of $\overline{P}$ over surface σ enclosing it. This is given by

$$S = \int_\sigma \overline{P}_n d\sigma \tag{2}$$

This expression (2) for the total radiation can be transformed easily by means of Green's theorem.

$$S = \int_\sigma \overline{P}_n d\sigma = \int_\tau \operatorname{div} \overline{P} d\tau. \tag{3}$$

where $d\tau$ is an element of volume in the space enclosed by σ. It is evident that only elements of the conductor under consideration contribute to the value of the integral. But according to (1):

$$\operatorname{div} \overline{P} = \frac{c}{4\pi} \operatorname{div} [\overline{E}, \overline{H}] = \frac{c}{4\pi} \{(\overline{H}, \operatorname{curl} \overline{E}) - (\overline{E}, \operatorname{curl} \overline{H})\} \tag{4}$$

In addition we have the two Maxwell equations for $\overline{E}$ and $\overline{H}$. If $\overline{I}$ is the current density, these read

$$\begin{aligned} c \operatorname{curl} \overline{H} &= \epsilon \frac{\partial \overline{E}}{\partial t} + 4\pi \overline{I} \\ c \operatorname{curl} \overline{E} &= -\mu \frac{\partial \overline{H}}{\partial t} \cdot \end{aligned} \tag{5}$$

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[2] *Loc. cit.*

For the space outside the conductor we take the dielectric constant $\epsilon = 1$ and the permeability $\mu = 1$. By using the above relations (5), we can change (4) to the form:

$$\text{div } \overline{P} = -(\overline{E}\overline{I}) - \frac{1}{8\pi}\frac{\partial}{\partial t}(\overline{E}^2 + \overline{H}^2). \tag{6}$$

The second member on the right side of (6) represents the rate of change in total energy. We now make the assumptions that all observed oscillations of the conductor are represented as the product of a space and a time, in which the time factor is harmonic, and that all conductors are excited with the same frequency. Under these conditions the second member in (6) disappears. Consequently, according to (3) and (6), we get for the total radiation:

$$S = -\int_{\tau}(\overline{E}\overline{I})d\tau. \tag{7}$$

Let us now consider a system of N conductors. For such a case the integration indicated by (3) is extended over all radiators of elements $d\tau$ in equation (7), which together produce the resulting field from the system. It follows from the above that if we designate the electric field strength produced by the $\bar{r}$-th radiator by $\overline{E}_r$, we get for (7):

$$S = -\sum_{s=1}^{N}\int_{\tau_s}\left(\sum_{r=1}^{N}\overline{E}_r,\ I_s\right)d\tau_s. \tag{8}$$

where I_s is the current distribution on the s-th conductor. In view of the above assumption we use, for the current distribution on the s-th conductor:

$$\overline{I}_s = \overline{J}_s e^{-i\nu t}. \tag{9}$$

Further, we may express the complex electric field $\overline{E}_r$ by

$$\overline{E}_r = E_r e^{-i\nu t}. \tag{10}$$

Taking (9) and (10) into consideration we obtain for the time average of the total radiation $\bar{S}$ the complex expression:[3]

$$\bar{S} = -\tfrac{1}{2}Re\left\{\sum_{s=1}^{N}\int_{\tau_s}(E_r\overline{J}_s^{*})d\tau_s\right\} \tag{11}$$

in which $\overline{J}_s^{*}$ is the conjugated complex value of $\overline{J}_s$ If we remove the complex current amplitudes and observe that the field E_r produced by the r-th conductor is proportional to the current amplitude A_r of this conductor, we set:

[3] *Re* means that the real part of this expression is to be taken.

$$J_s = A_s J_s' \text{ and } E_r = A_r E_r' \tag{12}$$

and if we make a further simplification by introducing the coefficients U_{rs} by means of the equation

$$U_{rs} = -\int_{\tau_s} (E_r' J_s') d\tau_s \tag{13}$$

we get, for (11), the general law

$$\tilde{S} = \overline{Re}\left\{ \sum_{r=1}^{N} \sum_{s=1}^{N} \frac{A_r A_s^*}{2} U_{rs} \right\}. \tag{14}$$

This is a generalization of the expression given by A. Pistolkors[4] for the radiation of an antenna system. In this expression the coefficient U_{rs} represents the coupling coefficients between the r-th and s-th conductors, and the coefficient U_{ss} is the radiation resistance of the s-th conductor if it is present alone. In particular if all current amplitudes are assumed to be equal in amplitude and phase, and designated by A, we get for the radiation,

$$\tilde{S} = \frac{|A|^2}{2} Re\left\{ \sum_{r=1}^{N} \sum_{s=1}^{N} U_{rs} \right\}. \tag{15}$$

Assuming equal current amplitudes, this gives us the resultant radiation resistance of the system:

$$R_{\text{res}} = \overline{Re}\left\{ \sum_{r=1}^{N} \sum_{s=1}^{N} U_{rs} \right\}. \tag{15a}$$

In order to be able to calculate the radiated power, we must form the coefficients U_{rs} in accordance with (13). For this purpose we may carry out the integration given by (13). The coefficient U_{rs} is obtained in a much simpler manner if we start from the well-known Hertizan vector for the radiation. This is taken from a previous paper.[2] Without restricting the generality we assume that all radiators are parallel to each other, and that the common direction of the radiators is along the z axis of a cylindrical coördinate system. The electric field strength $\overline{E}$ and the Hertzian vector $\overline{Z}$ are related according to the equation

$$\overline{E} = \text{grad div } \overline{Z} - \frac{1}{c^2} \frac{\partial^2 \overline{Z}}{\partial t^2}\,. \tag{16}$$

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[1] *Loc. cit.*

[2] *Loc. cit.*

If the Hertzian vector produced by the s-th conductor, (whose z component is other than zero according to the above assumption) is

$$\overline{Z}_s = (\overline{Z}_z)_s = \Pi_s e^{-i\nu t} \tag{17}$$

and we again remove the current amplitude A_s as before, by setting

$$\Pi_s = A_s \Pi_s'. \tag{18}$$

We get from (16) and (17) the z component of the electric field produced by the s-th radiator, thus:

$$(\overline{E}_z)_s = \frac{\partial^2 \Pi_s}{\partial Z^2} + k^2 \Pi_s \tag{19}$$

in which $k = 2\pi/\lambda = \nu/c$, where λ is the wavelength of the exciting oscillation, and c is the velocity of light. Taking (19) into consideration and observing that J' is a real magnitude and designating the element of the conductor $d\tau_s$ by dz_s we get in place of (13):

$$U_{rs} = -\int_{z_s} \left(\frac{\partial^2 \Pi_r'}{\partial z^2} + k^2 \Pi_r' \right) J_s' dz_s. \tag{20}$$

If we consider the case of sinusoidal current distribution for the radiator, we get for J_s the differential equation:

$$\frac{d^2 J_s}{dz^2} + k^2 J_s = 0. \tag{21}$$

By double partial integration of (20) and taking into consideration (21), we may readily obtain the following expression

$$U_{rs} = \left[\Pi_r' \frac{\partial J_s'}{\partial z} - \frac{\partial \Pi_r'}{\partial z} J_s' \right]_{z_{s1}}^{z_{s2}} \tag{22}$$

which must be taken between the upper and lower extremities of the s-th conductor designated respectively by z_{s1} and z_{s2}. The law given by (22) represents the above-mentioned simple combination of the radiation resistance with the Hertzian vector. In particular if the conductors considered here have a length that is a multiple of a half wavelength, that is, if there are current nodes at the ends of the conductors, so that $J_s(z_{s2}) = J_s(z_{s1}) = 0$, then (22) simplifies to:

$$U_{rs} = \left(\Pi_r' \frac{\partial J_s'}{\partial z} \right)_{z_{s2}} - \left(\Pi_r' \frac{\partial J_s'}{\partial z} \right)_{z_{s1}}. \tag{23}$$

As an example we now shall calculate the coefficient U_{rs} of two conductors each of whose lengths is a multiple of a half wavelength of the exciting oscillation, and which are parallel but at different heights. We thus get a general expression. All further expressions for coupling coefficients of special arrangements and radiation resistances of conductors whose lengths are a multiple of a half wavelength, are obtained from this merely by specialization.

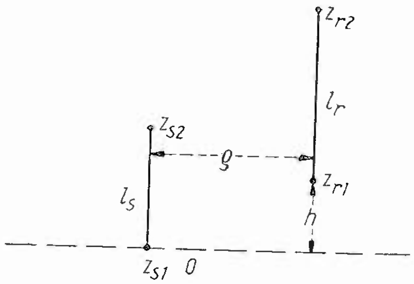

Fig. 1

Let us consider two conductors r and s (Fig. 1). The lengths of the r-th and s-th conductors are l_r and l_s respectively, where $l_r = m\lambda/2$ and $l_s = n\lambda/2$ with m and $n = 1, 2, 3 \cdots$. The coördinates of the endpoints of the conductors are x_r, y_r, $z_{r2} = l_r + h$, $z_{r1} = h$ and x_s, y_s, $z_{s2} = l_s$, $z_{s1} = 0$. Let the current function be given by $\bar{J}'_r = \sin k(z_r - h)$ and by $\bar{J}'_s = \sin kz_s$. The U_{rs} coefficient is given by (23). We take the Hertzian vector Π for the arrangement under consideration from the earlier paper.[2] If A_r is the current amplitude of the r-th conductor, the above[2] becomes:

$$\Pi_r = \frac{A_r}{2kc} \left[e^{ik(z-h)} \{ Ei(iku_2) - Ei(iku_1) \} + e^{-ik(z-h)} \{ Ei(iku_2') - Ei(iku_1') \} \right] \tag{24}$$

With the values

$$\begin{aligned} u_1 &= \sqrt{\rho^2 + (z - z_{r1})^2} - (z - z_{r1}); \\ u_1' &= \sqrt{\rho^2 + (z - z_{r1})^2} + (z - z_{r1}) \\ u_2 &= \sqrt{\rho^2 + (z - z_{r2})^2} - (z - z_{r2}); \\ u_2' &= \sqrt{\rho^2 + (z - z_{r2})^2} + (z - z_{r2}) \end{aligned} \tag{24a}$$

in which ρ is the distance to the base of the two conductors. The function $Ei(x)$ in (24) is the integral exponential function. Accordingly, for the coefficients U_{rs} defined by (23), we get

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[2] *Loc. cit.*

$$U_{rs} = \frac{1}{2c}\left[e^{ikh}\left\{Ei(iku_{22}) + Ei(iku_{11}) - Ei(iku_{21}) - Ei(iku_{12})\right\}\right.$$
$$\left. + e^{-ikh}\left\{Ei(iku_{22}') + Ei(iku_{11}') - Ei(iku_{21}') - Ei(iku_{12}')\right\}\right] \quad (25)$$

and for the sake of brevity:

$$\begin{aligned} u_{22};\ u_{22}' &= \sqrt{\rho^2 + (h + l_r - l_s)^2} \mp (h + l_r - l_s) \\ u_{12};\ u_{12}' &= \sqrt{\rho^2 + (h + l_r)^2} \mp (h + l_r) \\ u_{21};\ u_{21}' &= \sqrt{\rho^2 + (h - l_s)^2} \mp (h - l_s) \\ u_{11};\ u_{11}' &= \sqrt{\rho^2 + h^2} \mp h. \end{aligned} \quad (25a)$$

The coefficients U_{rs} have the dimension of an ohmic resistance. We used the Gauss system in the above. In order to change to technical units we must remember that for the electrostatically measured unit of resistance, R' (electrostatic units) $= 30$ c.R.Ω. Thus we have found the most general expression for the radiation coupling of two parallel conductors whose lengths are a multiple of the half wavelength. All expressions given by A. Pistolkors[1] are special cases of (24). It is unnecessary to repeat the expressions obtained by specialization.

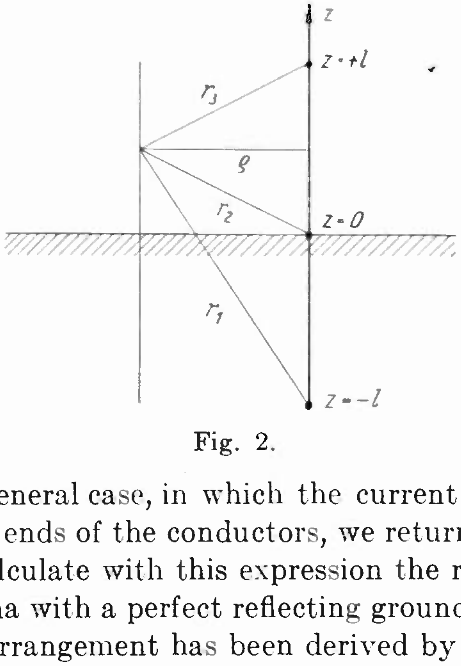

Fig. 2.

In the more general case, in which the current distribution has a finite value at the ends of the conductors, we return to (22). As an example we shall calculate with this expression the radiation resistance of a loaded antenna with a perfect reflecting ground. The radiation resistance for this arrangement has been derived by van der Pol[4] using the method of Poynting vectors.

Let us consider a loaded conductor of length l, placed along the z axis of a cylindrical coördinate system (ρ, z). The coördinates of the

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[1] *Loc. cit.*

[4] van der Pol., *Jahrb. d. drahtl. Telegr.*, 13, 217, 1918.

end points of the conductors are $z_{01}=0$, $z_{02}=l$ (Fig. 2). The current distribution along the conductor is given by the expression $J'_1=\cos(kz-\beta)$. For brevity, we introduce for the upper end of the conductor $kl=\Theta$ and $kl-\beta=\alpha$. A radiating conductor above an infinitely conducting earth gives rise to an image whose energy contribution must be added to that of the conductor. For this mirrored conductor the current distribution is $J'_2=\cos(kz+\beta)$ with the limits $z_{01}=-l$ and $z_{02}=0$. The amplitudes of the currents in the antenna and its mirror counterpart are equal. The radiation resistance R of the latter system is calculated from the relation

$$R = -\int_{-l}^{+l} E_z' J' dz. \tag{26}$$

The total current function here is $J=J_1+J_2$. According to the current distribution on the conductor and its mirror image, E_z consists of two parts. Let the part associated with J_1 be E_1, and that with J_2 be E_2. Therefore, $E_z=E_1+E_2$. Consequently (26) may be resolved into four parts as follows:

$$\begin{aligned} R = -\Big\{ &\int_0^l E_1'J_1'dz + \int_0^l E_2'J_1'dz \\ &+ \int_{-l}^0 E_1'J_2'dz + \int_{-l}^0 E_2'J_2'dz \Big\}. \end{aligned} \tag{27}$$

For reasons of symmetry, the first and fourth and also the second and third members are respectively equal. Consequently taking (22) into consideration we obtain for the radiation resistance of the loaded conductor above an infinitely conducting earth, the expression:

$$R = \lim_{\rho=0} 2\left\{ (\Pi_1' + \Pi_2')\frac{dJ_1'}{dz}\Big]_0^l - \frac{\partial}{\partial z}(\Pi_1' + \Pi_2')\cdot J_1'\Big]_0^l \right\} \tag{28}$$

in which Π_1 is the Hertzian vector of the loaded antenna, Π_2 is the Hertzian vector of its mirror image. The expressions Π_1 and Π_2 are taken from an earlier paper.[2] If A is the current amplitude for the conductor and its mirror image, we have:

$$\begin{aligned} \Pi_1 &= \frac{iA}{2kc}\Big[e^{i(kz-\beta)}\{Ei(iku_{21}) - Ei(iku_{11})\} \\ &\quad - e^{-i(kz-\beta)}\{Ei(iku_{21}') - Ei(iku_{11}')\}\Big] \\ \Pi_2 &= \frac{iA}{2kc}\Big[e^{i(kz-\beta)}\{Ei(iku_{22}) - Ei(iku_{12})\} \\ &\quad - e^{-i(kz-\beta)}\{Ei(iku_{22}') - Ei(iku_{12}')\}\Big] \end{aligned} \tag{29}$$

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[2] R. Bechmann, *loc. cit.*

with the values

$$\begin{aligned} u_{21};\ u_{21}' &= r_3 \mp (z - l) \\ u_{11};\ u_{11}' &= r_2 \mp z \\ u_{22};\ u_{22}' &= r_2 \mp z \\ u_{12};\ u_{12}' &= r_1 \mp (z + l) \end{aligned} \tag{29a}$$

in which, for brevity, we make

$$r_3 = \sqrt{\rho^2 + (z - l)^2}$$

$$r_2 = \sqrt{\rho^2 + z^2}$$

$$r_1 = \sqrt{\rho^2 + (z + l)^2}\,.$$

We next obtain (28) for finite values of ρ making use of (29) and (29a). For this we get

$$\begin{aligned} U = \frac{1}{2c}\Big[&4E\mathrm{i}(ik\rho) - 2E\mathrm{i}(ik(\sqrt{\rho^2 + l^2 - l})) - 2E\mathrm{i}(ik(\sqrt{\rho^2 + l^2 + l})) \\ &+ e^{2i\beta}\{2E\mathrm{i}(ik(\sqrt{\rho^2 + l^2 - l})) - E\mathrm{i}(ik(\sqrt{\rho^2 + (2l)^2} - 2l)) \\ &- E\mathrm{i}(ik\rho)\} + e^{-2i\beta}\{2E\mathrm{i}(ik(\sqrt{\rho^2 + l^2} + l)) \\ &- E\mathrm{i}(ik(\sqrt{\rho^2 + 2l)^2} + 2l)) - E\mathrm{i}(ik\rho)\}\Big] \\ &+ \frac{1}{c}\left(\frac{e^{ik\sqrt{\rho^2+(2l)^2}}}{ik\sqrt{\rho^2 + (2l)^2}} - \frac{e^{ik\rho}}{ik\rho}\right)\cos^2\alpha. \end{aligned} \tag{30}$$

The real part of (30) reads:

$$\begin{aligned} Re(U) = \frac{1}{2c}\Big[&4C\mathrm{i}(k\rho) - 2C\mathrm{i}(k(\sqrt{\rho^2+l^2} - l)) - 2C\mathrm{i}(k(\sqrt{\rho^2+l^2} + l)) \\ &+ \cos 2\beta\{2C\mathrm{i}(k(\sqrt{\rho^2+l^2} - l)) - C\mathrm{i}(k(\sqrt{\rho^2 + (2l)^2} - 2l)) \\ &- 2C\mathrm{i}(k\rho) + 2C\mathrm{i}(k(\sqrt{\rho^2+l^2} + l)) - C\mathrm{i}(k(\sqrt{\rho^2+(2l)^2} + 2l))\} \\ &+ \sin 2\beta\{2S\mathrm{i}(k(\sqrt{\rho^2+l^2} + l)) - S\mathrm{i}(k(\sqrt{\rho^2+(2l)^2} + 2l)) \\ &- 2S\mathrm{i}(k(\sqrt{\rho^2+l^2} - l)) + S\mathrm{i}(k(\sqrt{\rho^2 + (2l)^2} - 2l))\}\Big] \\ &+ \frac{1}{c}\left\{\frac{\sin k\sqrt{\rho^2 + (2l)^2}}{k\sqrt{\rho^2 + (2l)^2}} - \frac{\sin k\rho}{k\rho}\right\}\cos^2\alpha. \end{aligned} \tag{31}$$

We still must make the limit transition to $\rho = 0$ in (31) in accordance with (28). For this purpose we develop the members which approach infinity. The integral cosine $Ci(x)$ for small values of x has the development

$$Ci(x) = C + \log x - \frac{1}{2}\frac{x^2}{2!} + \cdots \tag{32}$$

in which $C=0.577$, the Euler constant. Let us develop the infinite expressions for $\rho=0$ in (31) in the vicinity of zero. We get

$$\begin{aligned} Ci(k(\sqrt{\rho^2+l^2}-l)) &= Ci\left(\frac{k\rho^2}{2l}\right) = C + \log\frac{k}{2l} + 2\log\rho \\ Ci(k(\sqrt{\rho^2+(2l)^2}-2l)) &= Ci\left(\frac{k\rho^2}{4l}\right) = C + \log\frac{k}{4l} + 2\log\rho. \end{aligned} \tag{32a}$$

The limiting value in (31) remains finite, as the infinite members in $\log\rho$ drop out for $\rho=0$. Changing from the Gauss to the technical system of units we get for the expression, for the radiation resistance R of the loaded conductor.

$$\begin{aligned} R = {} & 15\sin 2\beta\{2Si(2\Theta) - Si(4\Theta)\} \\ & + 15\cos 2\beta\{2Ci(2\Theta) - Ci(4\Theta) - \log\Theta - C\} \\ & + 30\left\{\cos^2\alpha\left(\frac{\sin 2\Theta}{2\Theta} - 1\right) - Ci(2\Theta) + 2\log 2\Theta + C\right\} \end{aligned} \tag{33}$$

This expression is identical with the law for the radiation resistance of a loaded conductor over an infinitely conducting earth previously calculated by van der Pol.[9]

It is found that the calculation of the radiation resistances by the direct method, shown in more detail here, is much simpler and leads to clearer results in the case of rather complicated antenna systems.

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[9] van der Pol, *loc. cit.*

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Proceedings of the Institute of Radio Engineers
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A SIMPLE METHOD OF HARMONIC ANALYSIS FOR USE IN RADIO ENGINEERING PRACTICE*

By

HANS RODER

(General Electric Company, Schenectady, New York)

Summary—*A simple method of harmonic analysis is applied for a-c waves with certain properties. Curves having such properties often occur in audio- and radio-frequency applications in the form of so-called "characteristics." This paper presents a graphical method of finding the amplitudes of the harmonics by working directly from the "characteristic." For obtaining the results, a polar planimeter is used. The design of a new mechanical harmonic analyzer is based on this method.*

In general, this method is a special case of a method of harmonic analysis formerly found by English and German authors.

INTRODUCTION

THE problem of analyzing the wave produced by an audio or radio-frequency amplifier is often met in engineering practice. In such cases a so-called "characteristic," i.e., a function of the exciting versus the output voltage, is given. For an audio amplifier, for example, the dynamic characteristic is known (Fig. 1), and we may wish to ascertain how far the grid may swing without introducing more than a specified amount of higher harmonics in the output. In broadcast transmitters the relation between audio input voltage and the amplitude of the radiated radio frequency should be investigated very accurately. This relation should be a straight line as nearly as possible. Each deviation from the straight line will introduce audio distortion and, what is worse, result in an increase of the width of the frequency band.

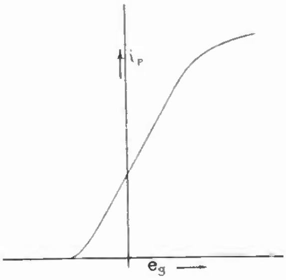

Fig. 1

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* Decimal classification: 537.7. Original manuscript received by the Institute, January 17, 1931.

THEORY

In Fig. 2 the curve $Z(x)$ represents the given characteristic. x is the value of the exciting voltage, Z the output voltage or current. That branch of $Z(x)$ being above the operating point may be called $z_1(x)$; that branch below is called $z_2(x)$. If the exciting term x varies with the time according to a sine function, these relations will exist:

$$x = A \sin y$$

and,

$$Z(y) = f(x) = f(A \sin y)$$

wherein y is proportional to the time.

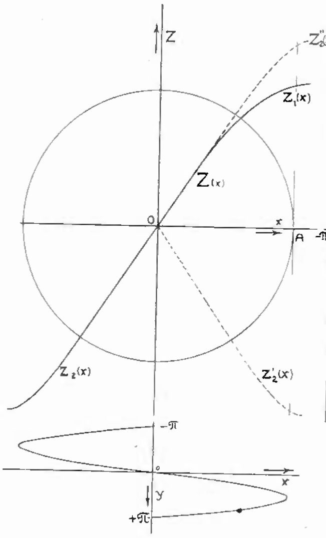

Fig. 2

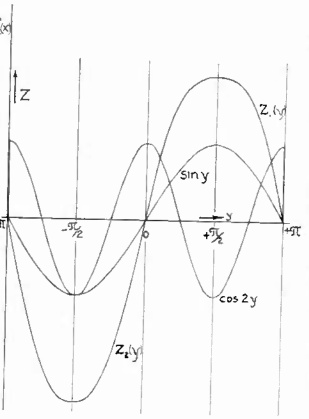

Fig. 3

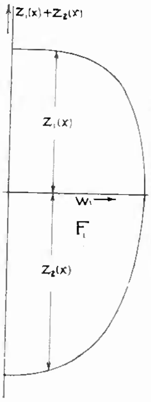

Fig. 4

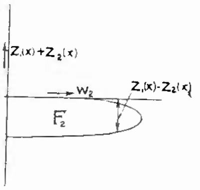

Fig. 5

From Fig. 3 the following may be readily seen. The branch $z_1(y) = f(A \sin y)$ is symmetrical to $+\pi/2$ and the branch $z_2(y) = f(A \sin y)$ is symmetrical to $-\pi/2$, no matter what the shape of $z_1(x)$ and $z_2(x)$ may be. Hence it follows: If $Z(y) = f(A \sin y)$ is developed in a Fourier series, all those sine and cosine members which are asymmetrical to $\pm\pi/2$ will become equal to zero. Thus the Fourier series of $Z(y)$ will take on the following form:

$$Z(y) = f(A \sin y) = a_0 + a_1 \sin y + b_2 \cos 2y + a_3 \sin 3y + b_4 \cos 4y + a_5 \sin 5y + \cdots.$$

According to Fourier, the *coefficient of the fundamental* is given by:

$$a_1 = \frac{1}{\pi} \int_{-\pi}^{+\pi} Z(y) \sin y \, dy.$$

As Fig. 3 shows this may be written:

$$a_1 = \frac{1}{\pi}\left(\int_{-\pi}^{0} z_2(y) \sin y\, dy + \int_{0}^{+\pi} z_1(y) \sin y\, dy\right)$$

$$a_1 = \frac{1}{\pi}\int_{0}^{\pi} (z_1(y) + z_2(y)) \sin y\, dy$$

$$a_1 = \frac{2}{\pi}\int_{0}^{\pi/2} (z_1(y) + z_2(y)) \sin y\, dy \tag{1}$$

This integral can be represented very simply. If we draw a curve with $(z_1(y)+z_2(y))$ as ordinate and $w_1=\cos y$ as abscissa we find for the area of this curve (Fig. 4):

$$F_1 = \int_{w_1=0}^{w_1=1} (z_1(y) + z_2(y))dw_1$$

Because,

$$w_1 = \cos y$$

and,

$$dw_1 = \sin y\, dy$$

we may write this equation:

$$F_1 = \int_{0}^{\pi/2} [z_1(y) + z_2(y)] \sin y\, dy. \tag{2}$$

Hence the coefficient of the fundamental is represented by

$$a_1 = \frac{2}{\pi}\cdot F_1. \tag{3}$$

y is the parameter interconnecting x and w_1:

$$w_1 = \cos y$$
$$x = A \sin y$$

wherein A is the peak value of the exciting sine function.

The Fourier *coefficient of the second harmonic* is:

$$b_2 = \frac{1}{\pi}\int_{-\pi}^{+\pi} Z(y) \cos 2y\, dy.$$

Fig. 3 shows that this is equal to

$$b_2 = \frac{2}{\pi} \int_0^{\pi/2} (z_1(y) - z_2(y)) \cos 2y \cdot dy. \tag{4}$$

If we plot a curve with

$z_1(y) - z_2(y)$ as ordinate and $w_2 = \sin 2y$ as abscissa,

(Fig. 5), the resulting area will become:

$$F_2 = \int_0^0 (z_1(y) - z_2(y)) dw_2$$

or,

$$F_2 = 2 \int_0^{\pi/2} [z_1(y) - z_2(y)] \cos 2y \, dy. \tag{5}$$

Comparing (4) with (5) we find:

$$b_2 = \frac{2}{\pi} \cdot \frac{F_2}{2}. \tag{6}$$

The value of w_2 is determined by:

$$x = A \sin y$$

$$w_2 = \sin 2y.$$

In the same manner the computation of the harmonics can be continued. For each harmonic a curve can be found the area of which is proportional to the amplitude of this harmonic. The results can be tabulated as follows:

TABLE I

	1st harm.	2nd harm.	3rd harm.	4th harm.
Ordinate	$z_1(y) + z_2(y)$	$z_1(y) - z_2(y)$	$z_1(y) + z_2(y)$	$z_1(y) - z_2(y)$
Abscissa	$w_1 = \cos y$	$w_2 = \sin 2y$	$w_3 = \cos 3y$	$w_4 = \sin 4y$
Amplitude	$a_1 = \frac{2}{\pi} F_1$	$b_2 = \frac{2}{\pi} \frac{F_2}{2}$	$a_3 = \frac{2}{\pi} \frac{F_3}{3}$	$b_4 = \frac{2}{\pi} \frac{F_4}{4}$

GRAPHICAL METHOD

The graphical method has already been outlined in the previous chapter. It consists in drawing the curve $z_1(y) \pm z_2(y) = f(w)$ and measuring the area of this curve by means of a planimeter. There remains only to give a few hints so as to reduce the time of performance of this method to a minimum.

When drawing the curves $z_2'(x)$ and $z_2''(x)$ (Fig. 2) ($z_2'(x)$ being symmetric to $z_2(x)$ with respect to the Z-axis and $z_2''(x)$ is symmetric to $z_2'(x)$ with respect to the x-axis), then the vertical distance between $z_1(x)$ and $z_2'(x)$ equals $z_1(y)+z_2(y)$, while the vertical distance between $z_1(x)$ and $z_2''(x)$ represents $z_1(y)-z_2(y)$. The abscissas w must be calculated from the equations shown in Table I and from the relation:

$$\frac{x}{A} = \sin y$$

For greater convenience, w has been calculated for a certain number of values x/A in steps from 10 to 10 per cent. The results are shown in the following table, which also contains a few more values of x/A so as to secure a greater accuracy when drawing the function $z_1(y) \pm z_2(y) = f(w)$.

TABLE II

	y Parameter: $\frac{x}{A} = \sin y$		$w_1 = \cos y$ $w_2 = \sin 2y$ $w_3 = \cos 3y$ $w_4 = \sin 4y$	
$\frac{x}{A}$	w_1	w_2	w_3	w_4
0,000	100,0	0,0	100,0	0,0
0,100	99,5	19,9	95,5	29,6
0,200	98,0	39,2	82,3	72,1
0,300	95,4	57,2	61,1	93,9
0,383	—	—	—	100,0
0,400	91,6	73,3	33,0	99,7
0,500	86,6	86,6	0,0	86,6
0,600	80,0	96,0	− 35,2	35,2
0,700	71,4	99,9	− 68,6	4,0
0,707	70,7	100,0	− 70,7	0,0
0,800	60,0	96,0	− 93,6	− 53,8
0,866	50,0	86,6	−100,0	− 86,6
0,900	43,6	78,4	− 97,6	− 97,3
0,920	39,2	72,1	− 93,5	− 99,9
0,924	—	—	—	−100,0
0,940	34,1	64,1	− 86,5	− 98,4
0,960	28,0	53,8	− 75,2	− 90,6
0,980	19,9	39,0	− 56,5	− 71,8
1,000	0,0	0,0	0,0	0,0

Let us summarize the method:

(1) Draw $z_2'(x)$ and $z_2''(x)$

(2) Divide the distance OA into 10 equal parts

(3) Plot a curve of corresponding values of $z_1(x) \pm z_2(x)$ and of w. w may be taken from Table II.

(4) Measure the area of the curve so found. This area is proportional to the desired amplitude.

The usual method[1] of graphical harmonic analysis requires much more work for obtaining a harmonic coefficient. The function $Z(x)$ must be redrawn in the function $Z(y) = f(A \sin y)$. The period of $Z(y)$ is divided into n equal parts and in each dividing point the amplitude is

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[1] C. Runge, *Elek. Zeit.* **26**, 247, 1905.

measured. The values thus found are arranged according to certain laws, multiplied by certain coefficients and finally added. The sum thus obtained is the desired coefficient.

MECHANICAL HARMONIC ANALYZER

The foregoing method can be extended by the design of a mechanical harmonic analyser. This is a small and comparatively simple device which, when combined with a normal polar planimeter, will read the harmonic amplitudes directly. Thus, all drawing and calculating work is minimized and the time for finding 3 or 4 harmonic components may be reduced to a few minutes. For further details, reference is made to a later article.

APPLICATION OF THIS METHOD FOR THE ANALYSIS OF GENERAL PERIODICAL CURVES

We started above with the characteristic $Z(x)$ of Fig. 2. We transformed this characteristic into the periodic function $Z(y)=f(A \sin y)$.

Let us consider a periodic function $W(y)$ (Fig. 6). We derive from this function the curves A and B. Both A and B have W as ordinate. The abscissa of A is $\cos y$, the abscissa of B is $\sin y$. We see that B corresponds to our former $Z(x)$ and A corresponds to our former function $z_1(x)+z_2(x)=f(w_1)$. For the areas we find:

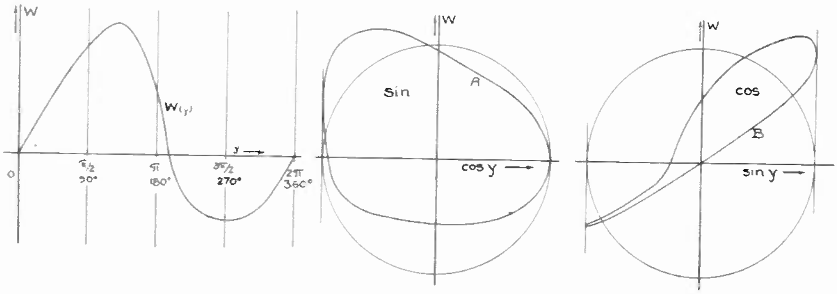

Fig. 6

$$F_A = \text{Area of } A = \int_0^{2\pi} W \cdot d(\cos y) = -\int_0^{2\pi} W \cdot \sin y \, dy$$

$$F_B = \text{Area of } B = \int_0^{2\pi} W \cdot d(\cos y) = \int_0^{2\pi} W \cdot \cos y \, dy.$$

Thus we get:

$$a_1 = \frac{1}{\pi} F_A;$$

$$b_1 = \frac{1}{\pi} F_B;$$

where a_1 is the sine—and b_1 the cosine-coefficient of the first *a-c* term in the Fourier series of $W(y)$. To find the coefficients of the second member, we plot curves *A* and *B* with the abscissas cos $2y$ and sin $2y$, respectively. In similar manner the higher harmonics can be obtained.

This method is not new. It has been outlined first by Clifford[2] and Finsterwalder.[3] These authors assume, that the curve with the period *T* is wound up on a cylinder of the circumference T/n, whereby *n* denotes the order of the harmonic coefficients to be found. The space curves so obtained are projected on planes through the cylinder axis and *x*-axis and cylinder axis and *y*-axis respectively. The area of these projections is proportional to the coefficients of the n^{th} harmonic. It can readily be seen that these projections are indentical with the curves called above *A* and *B*.

It may be mentioned further that the investigations of Clifford and Finsterwalder were the basis for the design of several mechanical harmonic analyzers.[4] In their principle, these analyzers are drawing apparatus which draw the curves *A* and *B* when the tracing point is carried along the function $W(y)$. A planimeter is combined with these apparatus by means of which the areas are measured simultaneously. The apparatus is designed such that the reading of the planimeter immediately gives the value of the coefficient to be found.

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[2] Clifford, *Proc. Lond. Math. Soc.*, 5.

[3] S. Finsterwalder, *Zeit. f. Math. u. Physik*, 43, 85, 1898.

[4] (a) Yule, *Proc. Phys. Soc.* (London), 13, 403, 1894–1895.
(b) Le Conte, *Phys. Rev.*, 7, 27, 1898.
(c) O. Mader, *Elek. Zeit.* 30, 847, 1909.
(d) L. W. Chubb, *Elec. Jour.*, 11, 91, 1914. (In this harmonic analyzer, the polar diagram of the periodical function is used.)

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GRAPHICAL REPRESENTATION OF THE THREE CONSTANTS OF A TRIODE*

By

Itomi Miura

(Electrotechnical Laboratory, Ministry of Communications, Tokyo, Japan)

Summary—*It is shown that the three constants of a triode, i.e., amplification factor, internal resistance, and transconductance, can be represented by one point in an equilateral triangle logarithmically scaled, and the author gives, as an example, a graph in which are plotted the constants of twenty-four kinds of typical tubes now used in practice.*

THERE have recently appeared a variety of vacuum tubes in the field of radio communications, including small receiving tubes of a few milliwatts and high power transmitting ones of some hundred kilowatts. It goes without saying that these various sizes of vacuum tubes have different characteristics according to their usages. The characteristics of a triode may simply be represented by the three constants, i.e., amplification factor, internal resistance, and transconductance. It is of great importance for us to have a knowledge of the constants of triodes before they are put in operation, in order that their performance may be predetermined, the accompanying circuits may suitably be designed, and thus, the triodes may be operated under their best conditons.

Now, if the constants of the different kinds of tubes be simply put on one graph so as to make them clear at a glance, it will be very convenient for the users of such tubes.

It has been suggested by Decaux[1] and Meyer[2] that the three constants of a triode can be represented by one graph. A method proposed here is of a similar nature but it is believed to be much simpler, more accurate and of a greater practical value than those previously given.

As is well known, the following relation exists among the three constants of a triode:

$$\mu = s_m \cdot r_p$$

or,

$$s_m \cdot r_p \cdot \frac{1}{\mu} = 1$$

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* Decimal classification: R131. Original manuscript received by the Institute, February 20, 1931. Abbreviated translation of the original paper in Japanese, *J.I.T.T.E.* (Japan), No. 91, October, 1930.

[1] Decaux, B., "Un abaque de classification pour les triodes de réception," *L'Onde Elec.*, 8e année, p. 37, 1929.

[2] Meyer, E., "Das Röhrendreieck," *Telefunken Zeitung*, Nr. 54, S. 54, April, 1930.

where,

μ = amplification constant,
s_m = transconductance (mho or amp./volt),
r_p = internal resistance (ohm).

In the logarithmic expression, the above relation becomes

$$\log s_m + \log r_p - \log \mu = 0.$$

In actual cases, s_m takes a value less than unity and $\log s_m$ is a negative quantity, so that

$$\log r_p - \log \mu - |\log s_m| = 0.$$

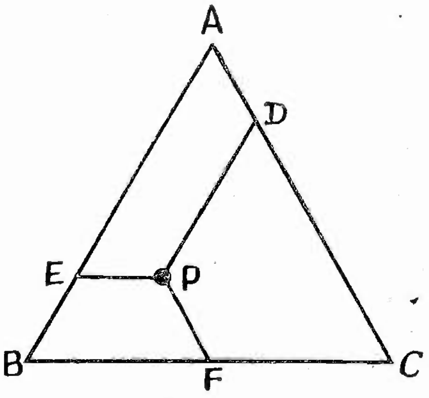

Fig. 1

On the other hand, in Fig. 1, let P be a point in an equilateral triangle $A\ B\ C$, and D, E, and F be the points where the straight lines drawn from P parallel to the three sides intersect the adjacent sides. Then the following relation holds as given in the elementary geometry:

$$\overline{AE} - \overline{AD} - \overline{CF} = 0.$$

This relation may be utilized for the presentation of the above formula, if these lengths represent the following values.

$$\overline{AE} = \log r_p \quad \overline{AD} = \log \mu \quad \overline{CF} = |\log s_m|.$$

Fig. 2 is constructed on this principle and there are plotted the constants of twenty-four kinds of typical receiving and transmitting tubes now regularly used. The numbers given close to the plotted points on the graph correspond to those given in the accompanying tables, from which the names of vacuum tubes will be known.

As seen in Fig. 2, for the tubes of similar usages their points on the graph will gather together on account of similarity of their constants. For instance, No. 1 to No. 11 are all receiving tubes, of which No. 1, No. 2, No. 4, and No. 6 are the detecting and amplifying tubes (No. 4 is the so-called a-c tube and not used for detection, but its characteristics and constants are about the same as those of No. 2 or No. 6, differing in just one point in that its filament may be supplied from a-c

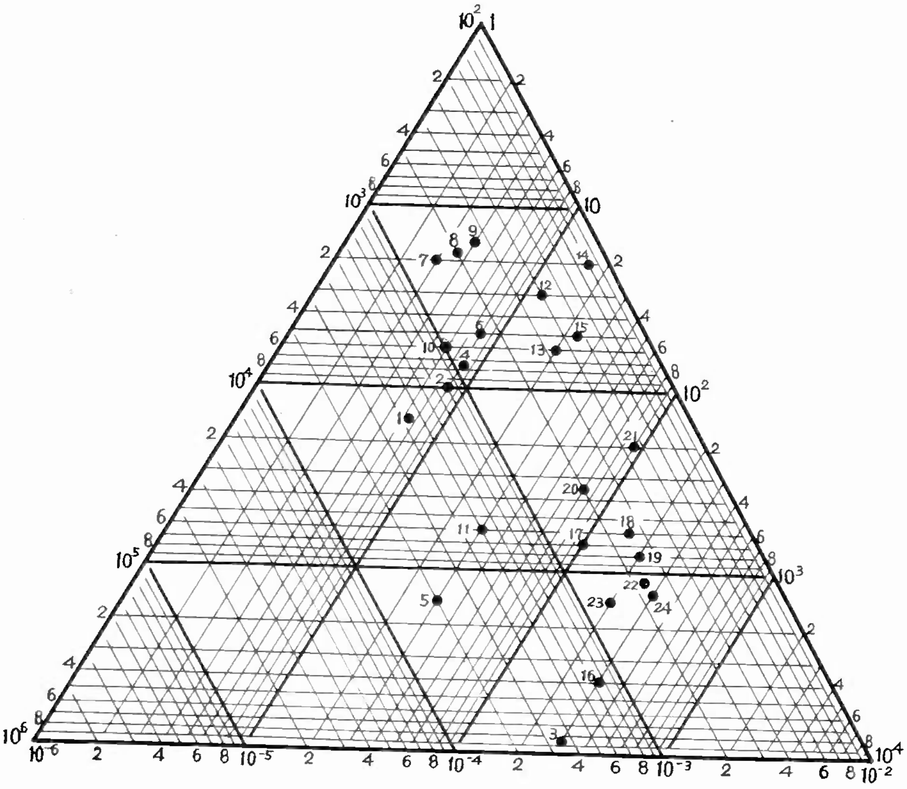

Fig. 2

TABLE I

○ Receiving tubes

Number	Name of Tubes	Filament		Anode Voltage V	μ	r_p Ω	s_m A/V
		Voltage V	Current A				
1	UX 199	3.0	0.06	90	6.6	15,500	0.000425
2	UX 201A	5.0	0.25	135	8	10,000	0.0008
3	*UX 222	3.3	0.132	135	300	850,000	0.00035
4	UX 226	1.5	1.05	135	8.2	7,400	0.0011
5	UX 240	5.0	0.25	135	30	150,000	0.0002
6	UX 112A	5.0	0.25	135	8	5,000	0.0016
7	UX 171A	5.0	0.25	180	3	2,000	0.0015
8	UX 245	2.5	1.50	180	3.6	1,800	0.002
9	UX 250	7.5	1.25	400	4	1,600	0.0025
10	101-D	4.4	0.97	130	5.9	6,000	0.001
11	102-D	2.0	0.97	130	30	60,000	0.0005

* Tetrode (for high-frequency amplification).

source); No. 7, No. 8, and No. 9, the last stage amplifying tubes; No. 10 and No. 11, the telephone repeater tubes, the former being for the power amplifier, the latter for the voltage amplifier. Nos. 12 to 22 are the transmitting tubes, of which No. 12, No. 13, No. 14, and No. 15 are the low voltage and low power transmitting tubes, while Nos. 16 to 22 are relatively high in voltage and power.

TABLE II

○ Transmitting tubes

Number	Name of Tubes	Filament		Anode		μ	r_p Ω	s_m A/V
		Voltage V	Current A	Voltage V	Loss W			
12	211D	10	3	1,000	65	12	3,000	0.004
13	UV 203A	10	3.25	1,000	75	20	6,000	0.0033
14	R 212D	14	6	2,000	200	16	2,000	0.008
15	UN 204A	11	3.85	2,000	200	23	5,000	0.0046
16	UN 154	10	6	8,000	200	300	400,000	0.00075
17	UN 155	12	6	10,000	400	100	70,000	0.00143
18	UN 156	15	10	10,000	600	150	60,000	0.0025
19	UN 157	17	15	10,000	1,500	200	80,000	0.0025
20	UN 158	12	24	10,000	1,000	70	35,000	0.002
21	UN 159	17	24	10,000	1,500	90	20,000	0.0045
22	UV 206	11	14.75	10,000	360	250	110,000	0.00227
23	*UX 860	10	3.25	2,000	100	200	150,000	0.00133
24	*UX 861	11	10	3,000	400	300	133,000	0.00225

* Tetrode (for short-wave transmission).

The constants of power tubes shown in the tables are those measured at a particular point on the static characteristics.

This graph has an advantage of representing the three constants of a triode in equal weight, and it may also be considered as another merit of the graph that, when any two of the three constants are known, the remaining one can readily be found on the graph.

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Proceedings of the Institute of Radio Engineers
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DISCUSSION ON THE RESISTANCE OF SPARK AND ITS EFFECT ON THE OSCILLATIONS OF ELECTRICAL OSCILLATORS*

JOHN STONE STONE

E. Amelotti:[1] The current i in a circuit containing resistance, self-induction and capacity in series, is given by the expression:

$$i = A_1 e^{-\alpha t} \sin \omega t \tag{1}$$

$$\alpha = \frac{R}{2L}; \quad \omega = \sqrt{\frac{1}{LC} - \left(\frac{R}{2L}\right)^2}.$$

where A_1 is a constant, $e^{-\alpha t}$ the damping factor, and $\sin\omega t$ the oscillatory characteristic.

If one expands $e^{-\alpha t}$ in a power series one obtains:

$$e^{-\alpha t} = 1 - (\alpha t) + \frac{(\alpha t)^2}{2!} - \frac{(\alpha t)^3}{3!} + \cdots. \tag{2}$$

The curve representing the solution (1), which is due originally to Lord Kelvin is:

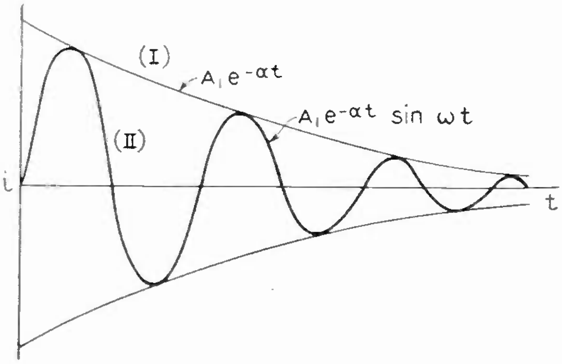

FIG. 1

(I) can be considered as the envelope of the maximums of (II).

In the article published by Stone the author calls the above theory the "Logarithmic Decrement Theory," and proposes at the same time a new method of attack of the same problem which he calls the "Straight Line Decrement Theory." He assumes that Lord Kelvin's theory does not give results which tally with experimental data when the frequency is very high and thus spark resistance is the predominant factor. By this remark the author seems to overlook the fact that high-frequency currents through conductors are accompanied by a significant increase in resistance due to what is known as "skin effect."

The author begins by introducing the tentative solution:

$$i = (A_1 - B_1 t) \sin \omega t \tag{3}$$

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* PROC. I.R.E., 2, 307–327, 1914.

[1] Student, University of Illinois, Urbana, Illinois.

instead of the one given in (1), where A_1 has the same interpretation as above, and $(-B_1)$ is the decrement factor. He calls $A_1 - B_1t$ the straight line decrement, and $\sin \omega t$ has the same interpretation as in (1). The curves representing the phenomena at different times is as shown below:

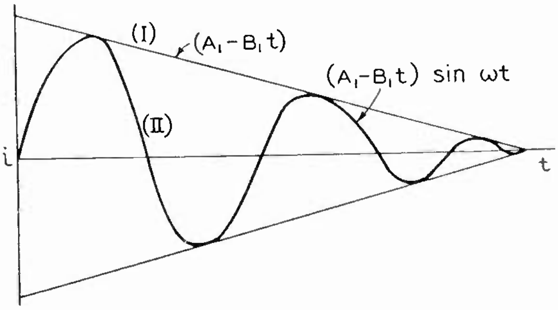

FIG. 2

Let us now consider (1) and (3) simultaneously and note if much difference exists between them.

If (2) converges fairly rapidly, depending upon the value of (αt), then within experimental error one can neglect all powers of (αt) beyond the first. So that (1) may be written as:

$$\begin{aligned} i &= A_1(t - \alpha t) \sin \omega t \\ &= (A_1 - \alpha A_1 t) \sin \alpha t. \end{aligned} \tag{4}$$

If one calls $\alpha A_1 = B_1$, one has essentially the author's proposed equation. So it seems to me that the new expression for the current introduced by the author is nothing more than a special, restricted case of Lord Kelvin's theory developed in 1853.

Let us now take the standard differential equation for a circuit discharging through the constants R, L, C, which has the form:

$$L \frac{di}{dt} + Ri + \frac{1}{c} \int i dt = 0. \tag{5}$$

Differentiating (5) with respect to the time t we have:

$$L \frac{d^2 i}{dt} + R \frac{di}{dt} + \frac{i}{c} = 0. \tag{6}$$

Taking the first and second derivatives with respect to the time of the expression for i, equation (3), we have:

$$\begin{aligned} \frac{di}{dt} &= (A_1 - B_1 t)\omega \cos \omega t - B_1 \sin \omega t \\ \frac{d^2 i}{dt^2} &= -(A_1 - B_1 t)\omega^2 \sin \omega t - B_1\omega \cos \omega t - B_1 \omega \cos \omega t \\ &= -(A_1 - B_1 t)\omega^2 \sin \omega t - 2B_1\omega \cos \omega t. \end{aligned}$$

Substituting these in (6) in the usual manner we have:

$$L[-(A_1 - B_1 t)\omega^2 \sin \omega t - 2B_1\omega \cos \omega t] + R[(A_1 - B_1 t)\omega \cos \omega t - B_1 \sin \omega t] + \frac{1}{c}(A_1 - B_1 t) \sin \omega t = 0. \tag{7}$$

Rearranging the terms and collecting them as coefficients of $\cos \omega t$ and $\sin \omega t$ respectively we have:

$$[-2LB_1\omega + R(A_1 - B_1t)\omega]\cos\omega t$$
$$+\left[-L(A_1 - B_1t)\omega^2 - RB_1 + \frac{1}{c}(A_1 - B_1t)\right]\sin\omega t = 0. \qquad (7')$$

Take all values of t for which $\cos\omega t = 0$. Those values are:

$$t = \frac{(2n \pm 1)}{2\omega}\cdot\pi \qquad n = 0, 1, 2, \cdots, \quad -1, -2, \cdots.$$

Then we have:

$$-L(A_1 - B_1t)\omega^2 - RB_1 + \frac{1}{c}(A_1 - B_1t) = 0$$

$$(A_1 - B_1t)\left(\frac{1}{c} - L\omega^2\right) - RB_1 = 0$$

$$\left[A_1 - \frac{B_1(2n \pm 1)\pi}{2\omega}\right]\left(\frac{1}{c} - L\omega^2\right) - RB_1 = 0$$

$$\frac{1}{c} - L\omega^2 = \frac{RB_1}{A_1 - B_1\dfrac{(2n \pm 1)}{2\omega}\cdot\pi}. \qquad (8)$$

The author's result is:

$$\frac{1}{c} - L\omega^2 = 0$$

$$\omega^2 = \frac{1}{LC}. \qquad (9)$$

In order to obtain the same result in (8), R must equal zero since B_1 cannot. But $R = 0$ is a very restricted case, and furthermore there would be no damping.

Further if we take all values of t for which $\sin\omega t = 0$ which would be:

$$t = \frac{n\pi}{\omega} \qquad n = 0, 1, 2, \cdots, \quad -1, -2, \cdots,$$

then we have:

$$\left[-2LB_1\omega + R\left(A_1 - B_1\frac{n\pi}{\omega}\right)\omega\right] = 0 \qquad (10)$$

$$R = \frac{2LB_1}{A_1 - B_1\dfrac{n\pi}{\omega}}. \qquad (11)$$

which depends on frequency instead of being independent of it as stated by the author.

CONCLUSIONS

The following is a summary of the above criticisms:

(1) The new proposed expression for the current is merely a special case of Lord Kelvin's theory, obtained by using only the first two terms of the damping factor $e^{-\alpha t}$.

(2) The substitution of an approximate solution into the differential equation is mathematically incorrect as it may lead to serious errors depending upon the function.

(3) The substitution of a solution into a differential equation leads to an identity, true for all values of the variable involved, but in the case discussed by the author it seems to be true only for the integral values of t. Furthermore the coefficients of the variables should be purely constants.

(4) Even though one goes through the development using the author's solution, the final results obtained are altogether different as is easily seen by comparison of the results of (8) and (11) of the above with those of the author.

John Stone Stone:[1] Since few of those who read this discussion of my 1914 paper on the resistance of the spark and its effect on the oscillations of electrical oscillators will have read that paper, and since, owing to the early date of its publication, it will be difficult of access to very many who see this discussion, I deem it most advisable to review briefly its historical background and to give a much condensed abstract of its substance.

HISTORICAL NOTE

In 1914, the date of my paper in question, and prior thereto, most radio transmitters depended, for the production of their high-frequency currents, on the oscillatory discharge of a condenser across a spark gap. For want of a more appropriate theory by which to interpret the performance of these oscillators, radio engineers and inventors very generally made use of the well-known Thomson[2] logarithmic decrement theory of the oscillatory discharge of a condenser through a circuit of *constant resistance* and inductance. They were well aware that the theory in question did not contemplate a discharge circuit comprising the variable spark resistance so characteristic of their oscillator circuits, but there was no theory at hand which applied to such circuits.

There were no empirical data establishing directly the relation between the resistance of a radio-frequency spark and the current flowing across the gap. Some realized that a spark in which the charge of the condenser surged across the gap a million or more times a second must partake more of the nature of an arc than of a succession of ordinary isolated sparks, but it was equally apparent to us that we were not justified in applying to these high-frequency sparks the existing data as to the resistance of the low-frequency arc or the isolated spark.

It appeared to me that, under the circumstances then existing, the most valuable data we possessed in the premises was that supplied by F. Richarz and W. Ziegler[3] and J. Zenneck.[4] This showed that when the resistance of the spark was the dominant resistance in a radio-frequency oscillator, then the subsidence of the oscillations no longer followed the logarithmic decrement curve but followed a well defined linear decrement law illustrated in Fig. 6 of my paper, and expressed analytically by equation (2) of that paper.

ARGUMENT OF THE PAPER

The paper begins by giving the origin and a statement of the Thomson logarithmic decrement theory of the oscillatory discharge of a condenser through

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[1] Associate, Department of Development and Research, American Telephone and Telegraph Company, San Diego, California

[2] The distinguished author of that theory was not Lord Kelvin in 1853 when he published it. He was not even Sir William Thomson at that time, but was plain Prof. William Thomson.

[3] *Ann. Phys.*, 1, 468, 1901.

[4] *Ann. Phys.*, 13, 822, 1908.

a circuit of *constant resistance* and inductance. It indicates why the theory should be more nearly applicable to the case of a low-frequency oscillator than to that of a radio-frequency oscillator. It points out that the theory does not contemplate the resistance of the variable nature of that of the spark, and in this connection it compares numerically an oscillator of natural frequency 1000 with one having the same energy, but, a natural frequency of 1,000,000. This comparison shows that the spark in the low-frequency oscillator is negligible in length compared to that in the high-frequency oscillator.

The paper then describes the manner in which the subsidence of oscillations in a radio-frequency oscillator has been measured and made evident through the use of the Braun cathode ray tube, and illustrates in Figs. 4 and 5 cases of observed linear and nonlinear subsidences of the oscillations in such oscillators.[5] It is shown by these and other observations that when the conductor resistance of the oscillator is negligible compared to the spark resistance the subsidence is a linear function of the time instead of a logarithmic function of the time.

The paper then shows that, in the case of an oscillator in which the conductor resistance is negligible compared to the spark resistance, the *observed* current is given by the simple expression.

$$i = (A_1 - P_1 t) \sin \omega' t \tag{1}$$

in which no assumption is made as to A_1, B_1, or ω' except that they are not functions of the current i nor the time t.

Since this is the expression of an *actually observed* oscillatory discharge in a real radio-frequency oscillator, it *must* be a special solution of the well-known circuital equation for such a circuit. This being the case, if the value of the current given by (1) is substituted in the circuital equation of the oscillator, it must satisfy that equation, and with the aid of the boundary conditions supply not only the values of A_1, B_1, and ω', but also give us a definite expression for the resistance of the oscillator.

Accordingly the substitution in question is made, and the boundary conditions are imposed. An explicit expression for the current in terms of the initial charge of the condenser, the natural periodicity of the circuit and the time interval occupied by the oscillation train is secured. This is given in (9) of the paper.

The all important result attained from the standpoint of this paper, however, is the determination of the fact that the resistance of the radio-frequency spark is inversely proportional to the amplitude of the oscillations. That is to say, it is at any moment inversely proportional to the envelop of the maxima of the successive oscillations.

The paper next digresses to discuss the linear decrement theory of radio-frequency oscillators, to compare it with the logarithmic decrement theory of low-frequency oscillators, and to show how the former lends itself to the analytical expression of a periodic succession of oscillation trains such as occur in all spark radio transmitters.

The paper then returns to the main question of the resistance of the high-frequency spark and illustrates by a nonmathematical discussion, making use of the property of arc resistance hysteresis, why the resistance of radio-frequency sparks should be inversely proportional at any time to the amplitude of the high-frequency current.

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[5] These diagrams are in part redrawn from Dr. Zenneck's well-known treatise "Electromagnetische Swingungen und Drahtlose Telegraphie," to which credit was given. Curves 2 and 3 of these diagrams are computed from my expression for the resistance.

Next, the paper points out that both the logarithmic and the linear decrement theories of the oscillatory discharge of a condenser are the extreme limiting cases of a more general theory. I there express the hope soon to present this more general theory to the Institute. In this general theory, neither the conductor nor the spark resistance is assumed to be negligible compared to the other. I presented this more general theory before the Institute on February 3, 1915.[6]

The paper then closes with a digresssion on the subject of the effect of the dielectric hysteresis of the condenser on the impedance of the oscillation circuit, and it shows that the effect of the hysteresis may be expressed either as a conductance in parallel with the condenser or as a resistance in series with the condenser.

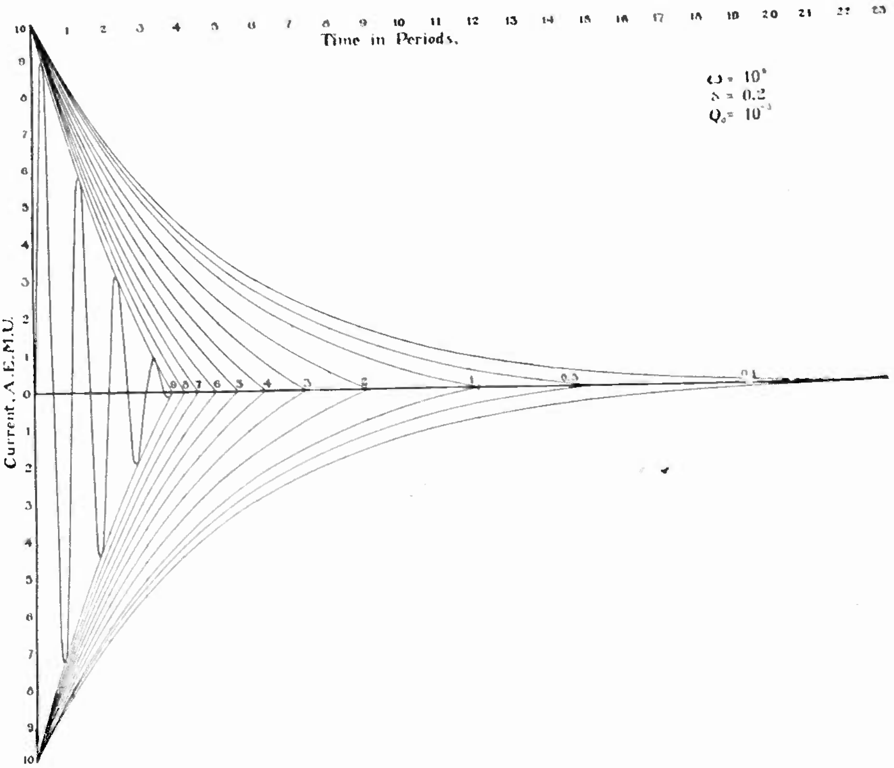

FIG. 1

E. AMELOTTI'S CRITICISM

The above historical note and statement of the argument of my paper disposes of all the implied criticism in E. Amelotti's communication, and most of its explicit criticism as well. Nevertheless these latter are answered below in the order in which he gives them.

(1) The linear decrement expression for the current is not a special case of the logarithmic decrement theory of the oscillatory discharge of a condenser. They are each limiting cases of a more general theory in which neither the conductor nor the spark resistance is assumed to be negligible compared to the other. This fact is graphically illustrated in the appended diagram taken from my 1915 paper to which I have referred. In the extreme left of this diagram is illustrated the oscilla-

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[6] PROC. I.R.E., 4, 463–482, 1915.

tions in the case of a circuit in which the initial resistance of the spark is nine times the conductor resistance, the other curves give the envelopes of the amplitudes when the ratio of these resistances is successively smaller till it reaches the value 0.1 in the extreme right-hand curve.

(2) The linear decrement expression for the oscillatory discharge current is not an "approximate" solution of the circuital equation of the oscillator nor is its substitution therein "tentative" in any sense. It is the correct expression of the current, as observed by means of oscillographs, in the limiting case when the conductor resistance is negligible compared to the spark resistance.

(3) and (4) Both of these criticisms rest upon the critic's alleged mathematical analysis of the problem, but this analysis has no physical or mathematical significance, since its first step consists in the differentiation of the well-known circuital equation of the oscillator under the assumption that the resistance of that circuit is constant, while its next step is to substitute in the resulting equation a value of the current which is incompatible with the assumption that this resistance is constant. Unless you assume that the resistance of the oscillator is in no way a function either of the time t or of the current i, you can not derive his equation (6) from his equation (5). But by making this unwarranted assumption, he in effect begs the question and practically assumes the logarithmic decrement solution of the circuital equation of the oscillator, and he forfeits the right to substitute therein any other solution, especially one that requires the resistance to vary explicitly with the time.

CONCLUDING REMARKS

Viewed with the gained perspective of 17 years of elapsed time, I find my paper to be far from impeccable as to the manner in which it presents its mathematical analysis of the problem, even though I find this analysis to be correct and its conclusions to be sound. This is because the boundary conditions used are largely left to implication instead of being explicitly stated.

The condition that the discharge of the condenser is complete is clearly implied by the fact that the integral of the energy dissipated in the circuit during the time of an oscillation train is equated to the energy of the initial charge in the condenser, but this boundary condition need not have been left to implication.

The condition that the spark quenches at a time when the current is normally zero is indeed explicitly stated, and is a well-known property of oscillating circuits, but it is not pointed out with sufficient clarity that this condition, when combined with the condition that the condenser is then completely discharged, results in a special and limiting case of the linear decrement discharge.

The paper does indeed point out in effect that when the discharge is complete there are an integral number of oscillations in the oscillation train, and it points out in effect that when this condition is departed from to the maximum possible extent, a change, which is quantitatively small, is necessitated in the expression for the oscillatory current. Nevertheless, it could just as well have been explicitly pointed out that this latter condition brings about another special and limiting case of the linear decrement discharge.

However, the all important result reached in the paper, namely, that the resistance of the spark is inversely proportional to the amplitude of the oscillatory current, is not dependant upon the particular limiting case of the type of oscillatory discharge analytically discussed in this paper. It is equally true when either limiting case is used in this analysis, or when the general case of the linear decrement oscillation train is employed.

Of course, as pointed out in the paper, the conclusions reached as to the resistance of the spark do not apply to oscillation circuits equipped with special kinds of gaps or with special appliances for minimizing the effects of arc or spark hysteresis such as air blasts, blow-out magnets, hydrogen or hydrocarbon atmospheres, or nonarcing electrodes for the gap. In this connection, it should be noted that when such conditions prevail the oscillations, if any occur, are not of the linear-decrement type.

To those who are not mathematically inclined, the nonmathematical discussion on pp. 319–322 in connection with Figs. 7 and 8 will amount to a demonstration of the fact that when the frequency of an arc or train of sparks is sufficienty high, its resistance will become inversely proportional to the amplitude of the high-frequency current passing across the gap, provided means are not taken to minimize the hysteresis of the arc or spark train.

This 1914 paper must be credited with supplying a new method of studying the resistance of a spark or arc, and with furnishing the material by which the more general theory of the oscillatory discharge of condenser at radio frequencies was determined in my 1915 paper whereby it became evident that the experimentally observed linear decrement discharge and the theoretically deduced logarithmic decrement discharge are opposite limiting cases in which the ratio of the conductor resistance to the spark resistance is respectively zero and infinite.

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Proceedings of the Institute of Radio Engineers
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DISCUSSION ON KENNELLY-HEAVISIDE LAYER HEIGHT OBSERVATIONS FOR 4045 KC AND 8650 KC*

T. R. GILLILAND

Frederick K. Vreeland:[1] The author of this very interesting paper is perhaps wise in not offering any explanation of his observations. To attempt such an explanation in the discussion may be rash, yet the observations show a phenomenon of such extraordinary interest as to merit very earnest attention.

Considering the lower graph marked (1) in Fig. 1 and the graph marked (3) in Fig. 5, representing the "virtual height of the Heaviside layer," there is a very marked periodic fluctuation in the height, with minima separated by intervals of approximately four weeks. And the dates of these minima coincide very nearly with the times of new moon. The lowest minimum occurred on the date of the solar eclipse and the other minima both before and after this are less marked.

Remembering that the date of new moon is the time when the moon comes closest to being in a direct line between the sun and the earth it is noteworthy that on the eclipse date, when the moon came precisely between the sun and the earth, the minimum was very low and on the other dates, where the moon was more or less out of line, the minima were not so low, and they are less marked in proportion to the amount by which the moon was out of line.

The coincidence is striking, and suggests very strongly, even though the data extend over a period of only six months, that the position of the moon has an important effect on radio transmission.

The way this works out can be seen more clearly from the graphs presented herewith and marked Fig. 9 and Fig. 10, showing how the angular position of the moon with respect to the sun, as viewed from the earth, and its angular position with respect to the earth, as viewed from the sun, change throughout the year. In Fig. 9 the full line curve represents the trajectory or path of the sun through the heavens. The broken curves is the trajectory of the moon. The moon goes through its cycle a little more than twelve times for each cycle of the sun, and in each cycle follows this curve approximately enough for present purposes. The point in each cycle where the moon overtakes the sun is at the time of new moon. These points are marked on the curve. When new moon occurs at the intersection of the curves at *C* or *D* we have an eclipse of the sun. This is what happened on April 29th, 1930. At other times the sun and the moon can never be closer together than the angular distance between the two curves. This angular distance becomes a minimum at each time of new moon. The relative positions of the sun and moon on successive days before and after new moon are shown by the light and dark circles on the two curves, and the dotted lines joining these circles represent the angular separation of the sun and moon on these successive days.

Transferring our point of view from the earth to the sun—a more rational though less usual procedure—the moon would appear to approach and recede from the earth. Plotting their apparent separation against time their relative positions would appear as shown in Fig. 10.

This graph is plotted on the same time scale as graph 1 of Gilliland's Fig. 1, which is here reproduced for direct comparison.

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* Proc. I.R.E., 19, 114; January, 1931.
[1] Research Engineer, Vreeland Corporation, New York City.

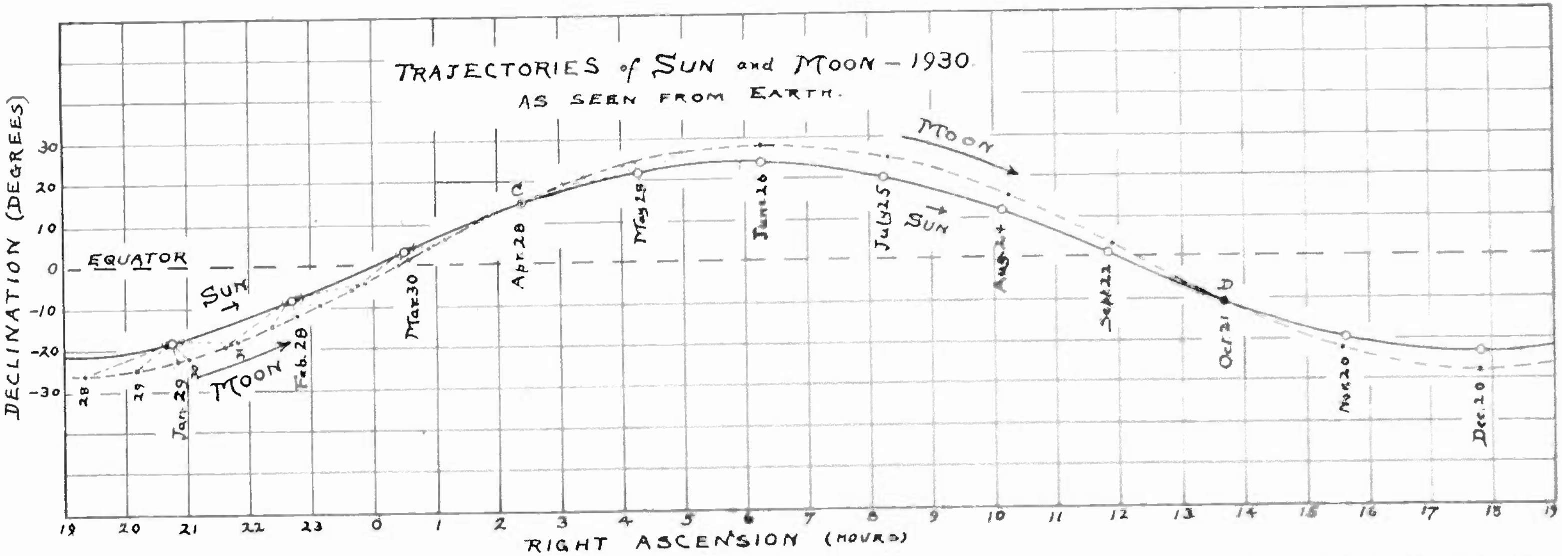

Fig. 9—The moon goes through its cycle following the broken curve more than twelve times for each cycle of the sun, overtaking the sun on the dates noted. The moon overlaps the sun's disk on the eclipse dates of April 28th and October 21st. At other times of new moon the moon's nearest approach to the sun is five degrees or less.

The very close correspondence of the minima of this angle with the minima of Gilliland's graphs is striking. This correspondence is so conspicuous that it cannot be dismissed as mere coincidence. It seems to indicate quite definitely that

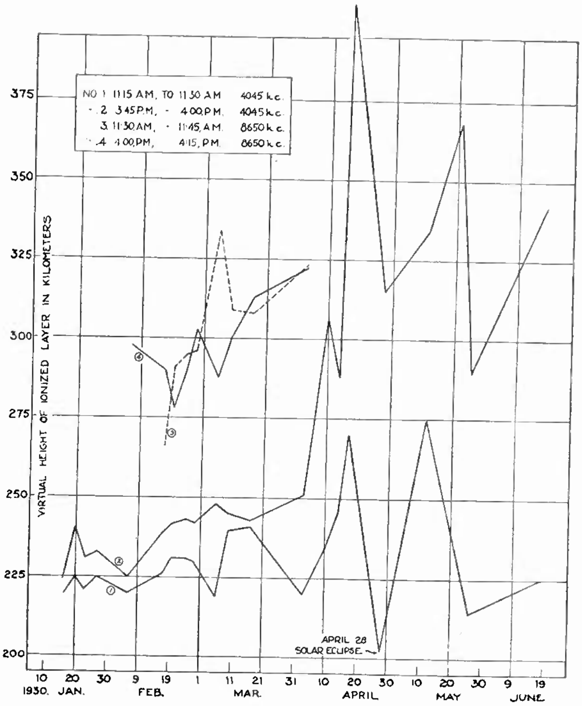

Fig. 1

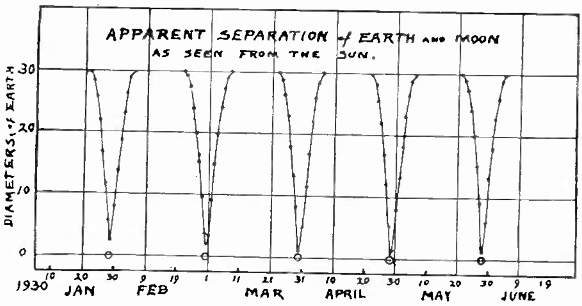

Fig. 10—Apparent separation of the earth and the moon plotted against time on the same time scale as that of Gilliland's Fig. 1, reproduced above. The unit of the vertical scale is the diameter of the earth. The circles and dots representing the earth and the moon are drawn approximately to scale. The half cycle during which the moon is behind the earth is not plotted since the effect under consideration is presumably small during this period. It will be noted that the dates of the minima coincide with the minima of Gilliland's Fig. 1 within the limit of accuracy of his biweekly observations.

as the moon passes between the sun and the earth, what Gilliland calls the "virtual height of the Heaviside layer" is lowered.

This raises two very important questions.

First. What are the factors that determine the height of the Heaviside layer?

Second. How does the moon influence these factors?

While one cannot attempt now a complete answer to these questions, which must be very intricate, we can at least formulate a working hypothesis which may point the way to this answer.

In 1904 the writer ventured to suggest what is believed to be the first published explanation of the fading of radio signals. Fading had been previously observed by Marconi but its cause was a mystery. The writer's suggestion was as follows:[2]

> "This (the breaking down of a spark gap when exposed to ultra-violet light) may be explained on the hypothesis**** that ultra-violet light has the power of "ionizing" a gas or splitting up some of its molecules into smaller bodies or ions***

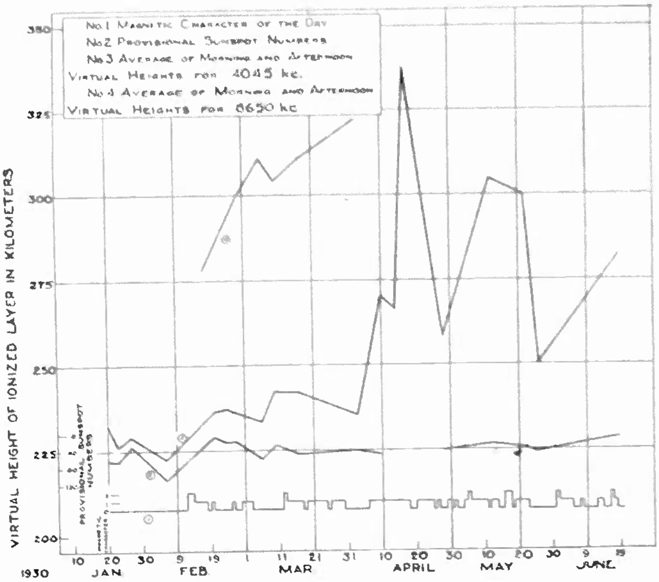

FIG. 5

> "If now we postulate a similar splitting up of the atoms of air into oppositely charged ions, under the influence of sunlight, we may explain the anomalous attenuation of electromagnetic waves. Under the influence of the electrostatic stresses which accompany the waves, the oppositely charged ions will move in opposite directions, and by their motion will produce what is practically a conduction current, or more strictly a convection current, in the air, just as the ions of electrolysis are the seat of the current in an electrolyte. Such currents will fritter away the energy of the wave, in the same way the the convection of heat from the warm compressed portions of a sound wave to the cooler rarified portions results in a dissipation of energy and an attenuation of the sound."

This explanation appears still to be valid as far as it goes. The powerful ultra-violet radiation from the sun unquestionably ionizes the upper air which absorbs it. Also the Heaviside layer unquestionably is a zone of atomspheric ionization. But we can hardly conceive of ultra-violet radiation being influenced by the position of the moon. Clearly there is another factor.

It is now known that the terrific upheavals on the sun which we call sun

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[2] Maxwell's Theory & Wireless Telegraphy, Poincare-Vreeland, McGraw Pub. Co., 1904.

spots belch forth an electron stream of great power. This electron emission shows itself in producing the aurora borealis, and in the "magnetic storms" which play havoc with our telegraph lines. Such an electronic bombardment has a powerful ionizing effect on the rarified upper air. Gilliland's observations indicate a certain coördination between sun spot activity and the height of the Heaviside layer, shown in Fig. 5, but this effect is completely masked by some other effect except for a short period.

We must therefore look for some emission from the sun besides ultra-violet radiation and besides sun spot bombardment whose effect is modified by the position of the moon.

Is it not reasonable to suppose that there is a steady electron emission from the intensely heated mass of the sun which goes on constantly whether sun spot outbursts are present or not, and that this emission is an important factor in producing the Heaviside layer? And may we not assume, until proof is forthcoming, that this emission has a high enough initial velocity to overcome the retardation due to its own space-charge effect, escape the influence of the sun's field, and finally reach the earth?

The mind of the radio engineer will perhaps picture the solar system, as far as the earth is concerned, as a gigantic three-element vacuum tube of which the thermionic cathode is the sun, the anode is the earth, and the control electrode is the moon. These elements are however all free and without external connections. The extent to which the electron stream reaches the earth will be determined by the velocity of emission, by space-charge considerations and by the electrostatic charges of the three bodies.

If we postulate that the moon has an electrical charge, its field should have an important effect on the distribution and the penetrating power of the electron stream that finally reaches the earth's atmosphere, particularly on that portion of the electron stream that has relatively low velocity. This effect would be greatest when the moon is directly in line with the stream, as it was on April 28th, and the effect should be less and less on the dates of March 30th, Feb. 28th, and Jan. 29th, when the moon was farther out of line. This is just what Gilliland's curves indicate.

When we go further and try to explain just what happens when the modified electron stream reaches the upper air we get into difficulties, because of the incompleteness of the observations and our imperfect knowledge of what we really mean by the "virtual height of the Heaviside layer" in these reflection experiments. Clearly this is not a simple case of reflection from a definitely located reflecting body, for the "virtual height" is quite different for the 8650-kc transmission and for the 4045-kc transmission. The graphs for 4045-kc (Gilliland's Fig. 1), which are the most complete, show quite definitely a tendency of the virtual height to increase as the season progresses from winter to summer, that is, as the sun moves northward and comes more nearly over the point of observation. The graphs indicate also that this effect of the total solar emission on any given day is progressive, being uniformly larger in the afternoon observations than in the forenoon observations. The periodic lowering of the reflecting layer, which coincides with the periods of new moon, indicates that the effect of the moon, when it comes between the sun and the earth, is to *diminish* the effect of the general radiation in raising the reflecting layer. This diminution is marked in the morning but is partly, though not entirely, overcome by the cumulative effect of the general radiation, as shown by the afternoon observations.

It is also worthy of notice that the effect of sun spot activity shown in Fig. 5, in so far as a correspondence between the curves can be observed, appears to be a lowering of the reflecting layer. This effect is completely masked by the larger general effect as the season advances.

May we not suppose that the sun spot emission, having relatively high velocity, is able to penetrate the atmosphere obliquely when the angle of incidence is low, while the less penetrating emission and radiation become effective as the sun rises overhead in the summer?

Other observations show that these sun spot effects are strong only when the earth is nearly in the direct line of the projected stream, giving another reason why their general effect is not greater, notwithstanding their tremendous activity.

The general emission from the whole surface of the sun may be assumed to have less velocity but greater total volume. Much of the initial velocity is probably lost through space-charge retardation before it reaches the earth, and because of the reduced velocity the effect of the moon's electrostatic field will be correspondingly large.

These conclusions also seem to be in harmony with the observed results, as far as the observations go.

Of course it is unwise to go too far in attempting to explain such phenomena from a single incomplete set of observations. But the agreements of these observations with what we might expect on theoretical grounds are so striking that we cannot dismiss them lightly.

The possibilities of this hypothesis are so interesting and the uncertainties arising from the meager data so tantalizing that it seems highly desirable that those who have great accumulations of data should study their records for further light on this important question. It will be of interest to study attenuation data as well as reflection data, to see whether the horizontal transmission is affected by the moon's position in the same way that the vertical transmission seems to be.

May we hope that those who have such data at their disposal will give us their contribution to the subject?

It would seem that the reason why the supposed lunar effect stands out so prominently in Gilliland's observations is that the measurements were made in the forenoon, when the immediate effect of the modified solar emission can be observed. In the afternoon we observe chiefly an integrated result of the various factors, while at night we have the residual effect that remains after the restoring factors have been at work—including the erratic effects of the weather from day to day. Hence the far greater complexity of nighttime studies.

Incidentally it would seem that our forefathers were not so foolish as we thought when they said that the new moon was a time of change of the weather; for certainly such powerful causes as these must have an effect on weather conditions. Meteorologists please note and give us your contribution to the discussion.

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CORRECTION

F. Guarnaschelli and F. Vecchiacchi, authors of the paper "Direct-Reading Frequency Meter" published in the April, 1931, issue of the PROCEEDINGS have requested the following material be published in elaboration of the sections entitled "Constants of the Device" and "Limits of Operation" (page 660), some details of which were not made sufficiently clear in the original translation.

CONSTANTS OF THE APPARATUS

A schematic of the complete circuit is given in Fig. 2. It will be noted that the elementary circuit of Fig. 1 is preceded by two triodes which act as amplifiers for low voltages and as limiting devices for higher voltages. This results in a greater sensitivity and in addition makes the device more independent of the form and value of the applied voltage.

The triodes, 2, 3, and 4 have intermal plate resistances of 2000 ohms and amplification factors of 5. The plate voltage, E_0, is 120 and the bias voltages, P_1 and P_2, are 20 volts each.

In practice, two sets of transformers are used; one for the frequency range between 20 and 300 cycles per second; the other for frequencies between 200 and 10,000 cycles per second. The former has a primary inductance of about 10 h while the latter has a primary inductance of about 0.6 h. These transformers must be carefully shielded and have low interwinding capacity.

Suitable resistors are shunted across the primary windings of the of the transformers, T and T_1, to prevent oscillations at a frequency determined by the inductance and capacity of the windings. It is obvious that other values than those given in Fig. 2 will be needed in cases where the transformer characteristics are different.

LIMITS OF OPERATION

(*a*) *Frequency Limits of the Control Transformer*

As already pointed out, the frequency range from 20 to 10,000 cycles per second is covered by means of two control transformers, each of which has a lower frequency limit of operation corresponding to the frequency at which the inductive reactance of the primary winding becomes too low in relation to the internal resistance of the tube to permit a satisfactory transfer of power.

There is, moreover, an upper frequency limit which is due to the two secondary voltages, V_1 and V_2, getting out of phase as a consequence of flux dispersion and the mutual capacity of the transformer windings.

These factors result in an imperfect control of the system by the two triodes, 3 and 4, which permits a flow of plate current even when the capacity, C, is zero. We cannot consider these factors as if they were merely parasitical capacities because even with $C=0$ the plate current varies in relation to the input voltage. If, however, the capacity, C, is sufficiently large, this effect, which causes higher readings than the theoretical formula indicates, can be neglected.

(b) *Maximum Values of Capacity*

Each frequency range is limited to a certain maximum value of C. This is due to the time required to charge C through the plate resistance of the two triodes and there is an upper limit of frequency above which C cannot be fully charged during a half cycle. For frequencies above this limit, the plate current, I, becomes perceptibly smaller than that indicated by the formula and this limit must not be exceeded or the value of the input voltage will again affect the plate current. The maximum value of plate current permissible in the equipment used was between 7 and 10 ma and was independent of the value of the capacity or the input frequency.

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CORRECTIONS

Bibliography on Piezo-Electricity

W. G. Cady has brought to the attention of the editors the following corrections to his paper "Bibliography on Piezo-Electricity" which appeared in the April, 1928, issue of the PROCEEDINGS, on pages 521 to 535.

Page 522, line 17. *For* Corning, New York, *read* Minneapolis, Minn.
Page 524, No. 71. *For* Gieger *read* Geiger.
Page 526, No. 112. *For* November, 1927, *read* November and December, 1927.
Page 527, No. 142. *For* Morecroft, J. M., *read* Morecroft, J. H.
Page 528, No. 167. *For* Widemann's *read* Wiedemann's.
Page 529, No. 211. *For* (C) *read* (B).
No. 212. *For* Verbeek, D. C., *read* Verbeek, C. C.
Page 531, Patent 1,495,429. *For* Rochelle *read* Rochelle salt.

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BOOK REVIEWS

Radio Frequency Measurements, by E. B. Moullin. Second Edition. Charles G. Griffin & Co., Ltd., London, and J. B. Lippincott Co., Philadelphia. Printed in Great Britain, 1931. 487 pp., 289 illustrations. Price $12.50.

This "handbook for the laboratory and textbook for advanced students" has increased in size some seventy per cent over the original edition. While it has been largely rewritten, the general plan and purpose of the book have been kept as before. Preliminary to the subject of measurements and of apparatus, new chapters are inserted on the electromagnetic field and on circuit formulas. In the former chapter the electromagnetic equations are developed and the calculation of the field intensity near circuits and aerials is treated. In the very brief treatment of circuits, vector methods and circle diagrams are neglected entirely. Filter circuits receive two pages, while the cable claims nine. The remaining chapters of the book have retained their former titles, and for the most part the material covered is the same with generous amplification here and there. The chapter dealing with the vacuum tube oscillator has been trebled in content.

It is to be regretted that a book of as general possible interest as the present one should be as one-sided as it appears to be from a glance at the name index. The names of many of those most familiar to American readers for their contributions to this field do not appear at all or receive attention which is entirely inadequate. The references to the work of a few are quite numerous. It is questionable whether a general text of this kind should devote so much attention to the author's own instruments as is done here.

There is no mention of Pierce's piezo oscillator or of push-pull circuits. The only use of the tetrode referred to is as a relaxation oscillator.

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Standards Year Book. Compiled by the National Bureau of Standards, George K. Burgess, Director. Bureau of Standards Miscellaneous Publication No. 119. Washington, D. C., 1931. For sale by the Superintendent of Documents, Washington. Price (cloth) $1.00. 339 pp.

This volume follows the pattern set by the four previous issues of the Standards Year Book in picturing the standardization movement and the national and international agencies involved. The present volume features in its first section a symposium on standardization in transport in which aeronautics, the marine field, railway, automobile, and elevator transportation call for brief articles contributed by experts, as well as the fields of power and speech transmission, and oil and gas pipe-line systems. Among the fifteen or twenty topics that are considered in a section of international interest the two following topics selected at random may be mentioned. One is the proposed new primary standard of light, the Widner-Burgess standard. This is a one-square centimeter opening in a black body at the freezing point of platinum and emits light equal to 58.84 International candles. The other is a statement of the progress being made internationally in the proposed simplification of the calendar.

In the third section of the book the work of national standardizing laboratories outside of the United States is treated. Interesting, brief summaries of the research work of the National Physical Laboratory, at Teddington, the corresponding German institution, and the two French laboratories are to be found here. After a long section devoted to the various standardizing agencies of the U. S. government comes that devoted to the National Bureau of Standards. A brief statement of some of the activities and accomplishments of the Bureau is here given. Items of radio interest considered include the blind landing of aircraft, the installation and maintenance of a primary frequency standard at the Bureau, field intensity, and Kennelly-Heaviside layer height measurements, and various items of piezo-electric research.

In a later section devoted to the standardizing activities of technical societies and trade associations, the Institute of Radio Engineers, is, of course, listed and a statement is made of the work which its Committee on Standardization has under way and its coöperative program with other agencies.

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Report of the Delegation of the U. S. A. to the First Meeting of the C. C. I. R., The Hague, September 18 to October 2, 1929. 532 pp. Superintendent of Documents, U. S. Government Printing Office, Washington, D. C. 90 cents.

The report of the American Delegation to the First Meeting of the International Technical Consulting Committee on Radio Communications held at The Hague September 18 to October 2, 1929, formed an especially valuable reference book because of the fact that it contains much information hitherto unpublished, which has been collected in the field of practical communication engineering.

The report, divided as it is into the first section which is a general discussion of the problems and accomplishments of the conference, and the second section which contains as appendixes all the papers submitted during the Conference by the various Government and company delegates attending, is presented in such a way as to make available the latest thought in communication engineering concerning such important subjects as frequency tolerances for various types of services, stabilization of frequency, definition of power, classification of service, methods of comparing frequency standards, degree of precision of frequency meters, selectivity of receiving apparatus, elimination of nonessential transmissions, coördination between land-line radiotelephony and various other topics of this nature which have never received specific treatment in general reference works on radio engineering.

This volume also contains the various proposals made prior to the Conference by the representatives participating, and the conclusions as to the state of the art then existing. This book together with the report of the Second Meeting of the C. C. I. R. to be held this year in Copenhagen, Denmark, beginning May 27, should form the most up-to-date text book on the specific problems of radio communication considered which is available to the radio profession.

*Gerald C. Gross

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* Federal Radio Commission, Washington, D.C.

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BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed on this page may be obtained gratis by addressing a request to the manufacturer or publisher.

The Ferris model 5B standard signal generator is described in a 24-page booklet prepared by Malcolm Ferris, Bonton, N. J. The model 5B generator is normally supplied with coils covering the broadcast spectrum, although through the use of interchangeable coils the range from 150 kc to 7500 kc may be covered. The output is continuously variable from 1 microvolt to 500,000 microvolts. A self-contained audio oscillator of 400 cycles provides modulation although an external modulator may be used if this is desirable.

The Weston Electrical Instrument Corporation of Newark, N. J., has recently issued several publications adopted to the requirements of the service man. "Uses of Electrical Instruments for Radio Testing" are given in a 24-page booklet of this title. A multi-range a-c and d-c voltmeter and milliammeter which may be made by the service man is described in mimeographed data sheets entitled "Model 301 Universal Meter." "Model 576 Mutual Conductance Meter" is the title of mimeographed data sheets describing an instrument intended primarily for production testing of thermionic tubes. The instrument measures transconductance (mutual conductance) in micromhos directly on a calibrated scale.

An 8-page folder issued by Herbert H. Frost, Inc., Elkhart, Ind., describes a number of fixed and variable resistors and attenuators, telephone jacks, microphones, and telephone headsets.

A 68-page supplement to catalog F-129 entitled "New and Improved Laboratory Apparatus and Instruments" manufactured by the Central Scientific Co. of Chicago describes equipment which is frequently used in various physical and electrical laboratories, especially those of colleges and universities. Among the items listed which should be of interest to radio men are the Cenco-Hypervac pump which will reduce air pressure to 0.1 micron within a few minutes, and the high-frequency vacuum tube oscillator for demonstrating the presence of standing waves.

A series of new relays manufacturered by Struthers Dunn, Inc., 139 N. Juniper St., Philadelphia, Pa., are described in two catalogue sheets recently issued as a supplement to their catalog.

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REFERENCES TO CURRENT RADIO LITERATURE

THIS is a monthly list of references prepared 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, which appeared in full on pp. 1433–56 of the August, 1930 issue of the Proceedings of the Institute of Radio Engineers. The classification numbers are in some instances different from those used in the earlier version of this system used in the issues of the Proceedings of the Institute of Radio Engineers before the October, 1930, issue.

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

R055 ×R113 Bureau of Standards. Bibliography on radio wave phenomena and measurement of radio field intensity. Proc. I.R.E., **19**, 1034–1089; June, 1931.

An extensive bibliography compiled by the Bureau of Standards.

R100. Radio Principles

R113.6 Edes, N. H. The multiple refraction and reflection of short waves. Proc. I.R.E., **19**, 1024–1032; June, 1931.

This paper discusses the theory that normal long-distance short-wave communication is brought about by a series of refractions and reflections. Single-hop characteristics given by the author in a previous paper are used in this way to derive the characteristic for longer ranges in daylight. The result is in close accord with a curve given by Lloyd Espenschied which shows the result of actual experiment over long ranges.

R116 Roosenstein, H. O. High-frequency feeders. *Exp. Wireless and the Wireless Eng.* (London), **8**, 294–297; June, 1931.

Methods are given for measuring the characteristic impedance and damping of high-frequency transmission lines, together with a procedure for proportioning such lines in a way to avoid loss by reflection.

R116 ×621.385 Saglio, M. Téléphonie par courants porteurs sur lignes a haute tension (Carrier telephony on high-tension lines). *L'Onde Electrique*, **10**, 189–220; May, 1931.

After a brief review of carrier telephony as applied to high tension lines, the author discusses present solutions of various practical problems involved and points out possibilities of the system.

R125.3 Bashenoff, V. I. and Mjasoedoff, N. A. The effective height of closed aerials. Proc. I.R.E., **19**, 984–1018; June, 1931.

Formulas are given for the calculation of the effective height of coil aerials with non-quasi stationary distribution of current. Aerials of several forms, suitable for use in radio beacons, are treated.

R125.31 Horton, C. E. The practical correction of a wireless direction-finder for deviations due to the metal work of a ship. *Jour. I.E.E.* London), **69**, pp. 623–636; May, 1931.

A general treatment is given of the field of a wireless wave in the vicinity of a ship. It is shown that both semicircular and quadrantal correctors will generally be required to obtain readings of high precision. The best position for a direction finder in a ship is discussed and the results of actual measurements are given. The principles of a new form of sensefinder are also explained.

R132 Campbell, A. G. The variable-mu tube and distortion in radio receivers. *Radio Eng.*, **11**, 29–30; June, 1931.

A brief study of the principles underlying the operation of the new variable-mu vacuum tubes is given, with an explanation of how crosstalk is eliminated and how supercontrol is accomplished.

R132 Couillard, L. Amplificatéurs a bande de fréquences (Band-pass amplifiers). *L'Onde Electrique*, **10**, 171–188; April, 1931.

Some simple amplifier arrangements are described which have band-pass frequency characteristics. These amplifiers are shown to be particularly useful as intermediate-frequency amplifiers in broadcast, superheterodyne receiving sets.

R133 Müller, F. and Zimbalin, W. Untersuchungen an einem Kurzwellen-Gegentaktsender (Experiments with a high-frequency push-pull oscillator). *Elek. Nach.-Tech.*, **8**, 207–213; May, 1931.

A push-pull circuit arrangement for generating ultra-high frequencies is described in which the filament leads are set up as a pair of Lecher wires. Tuning the filament circuit is shown to have no effect on the frequency of the oscillator but does increase the oscillation amplitude.

R134 Colebrook, F. M. Avoiding detection distortion. *Wireless World and Radio Review*, **28**, 529–532; May, 1931.

A grid rectification circuit is described which practically eliminates the distortion of higher modulation frequencies, which is usually encountered in this type of circuit.

R134 Colebrook, F. M. A new development in power-grid detection. *Wireless World and Radio Rev.*, **28**, 625–628; June, 1931.

The difficulties of distortion normally encountered in power-grid detectors may be overcome by dividing the functions of rectification and amplification normally performed by the single rectifier stage, and allocating each to a separate stage.

R139 Nottingham, W. B. A note on the time required to set up conduction in an FG-17 thyratron as determined by a linear time axis circuit of an oscillograph. *Jour. Frank. Inst.*, **211**, 751–755; June, 1931.

Recently published results seemed to indicate that 1000 microseconds were required to set up conduction but an investigation using a cathode-ray oscillograph definitely shows that good conduction can be set up in 10 to 20 microseconds.

R146 Labus, J. W. and Roder, H. The suppression of radio-frequency harmonics in transmitters. PROC. I.R.E., **19**, 949–962; June, 1931.

The effects of several types of circuit arrangement on the suppression of harmonics are tabulated and the advantages of the push-pull amplifier and another circuit arrangement which inherently compensates harmonics are discussed.

R161 Case, N. P. Receiver design for minimum fluctuation noise. PROC. I.R.E., **19**, 963–970; June, 1931.

The effects of various changes in both tube and circuit conditions have been investigated with regard to their influence on the limitation which fluctuation noise sets on the sensitivity of a receiver.

R200. RADIO MEASUREMENTS AND STANDARDIZATION

R210 Polkinghorn, F. A. and Roentken, A. A. A device for the precise measurement of high frequencies. PROC. I.R.E., **19**, 937–948; June, 1931.

A description is given of equipment that was used for the measurement of radio frequencies between 5000 and 30,000 kc with a precision of better than three parts in a million.

R211.1 Mittelmann, E. and Wald, M. Zeigerfrequenzmesser (Indicating frequency meter). *Zeit. für Hochfrequenz.*, **37**, 187–191; May, 1931.

The principles underlying the theory of a new type of frequency meter are discussed. The instrument is direct reading and very sensitive to slight variations of frequency.

R211.1 Mittelmann, E. and Mittelmann, R. Messung geringer Frequenzabweichungen mit direkter Anzeige (Measuring small frequency changes with direct indication). *Zeit. für Hochfrequenz.*, **37**, 191–199; May, 1931.

Several of the new type direct-reading frequency meters covering different frequency ranges were constructed and tests made to check the theory of a preceding paper. Measurements indicated a higher sensitivity than that predicted by theory.

R213 Vormer, J. J. and van Geel, C. Frequency measurements of high accuracy. *Exp. Wireless and the Wireless Eng.* (London), **8**, 298–303; May, 1931.

A harmonic method of measuring frequency is described. This method is used in the radio laboratory of the Dutch State Telegraphs and is accurate to the order of 1:100,000.

R243 Rohde, L. Eine Spannungsmessmethode für Frequenzen bis zu 1.5 $\times 10^8$ Hertz (A method for measuring voltages at frequencies up to 150 megacycles per second). *Zeit. für tech. Physik*, **12**, 263–265, No. 5, 1931.

A compensation method for measuring voltages at ultra-high frequencies is described. A special diode is used and an accuracy of 1 per cent is claimed.

R261 Smith-Rose, R. L. Testing wireless receivers. *Wireless World and and Radio Rev.* **28**, 636–638; June, 1931.

A brief discussion of equipment for making standard over-all performance measurements of radio receivers is given.

R265.2 Clarke, H. M. The moving coil loud speaker. *Experimental Wireless and the Wireless Eng.* (London), **8**, 304–306; June, 1931.

Supplementing a previous paper, the author describes a series of experiments for determining the frequency characteristics of a moving coil speaker and suggests the possibility of extending his method to determine the output and losses for a nonrigid diaphragm.

R265.2 Turner, P. K. Some measurements on a loud speaker in vacuo. *Jour. I.E.E.*, (London), **69**, 591–622; May, 1931.

A useful method for determining the characteristics of a moving coil loud speaker involves the measurement of the electrical impedance of the moving coil under three different conditions, viz: with the coil held fast; with the coil free in a vacuum; and under normal conditions.

R281 Vogler, H. Die Untersuchung dielektrischer Verluste flüssiger Isolierstoffe bei kurzen Wellen mit dem Kalorimeter (The investigation of high-frequency dielectric losses in liquid non-conductors by means of the Calorimeter). *Elek. Nach.-Technik*, **8**, 197–207; May, 1931.

Methods of procedure and results for several materials including turpentine, paraffin oil, and transformer oil are given.

R300. Radio Apparatus and Equipment

R350 Schäffer, W. and Lubszynski, G. Messung der Frequenzcharakteristik mit Hilfe des Lichttongenertators. (Measuring the frequency characteristic with the help of the photo-electric tone generator). *Elek. Nach.-Technik*, **8**, 213–217; May, 1931.

It is shown that the photo-electric generator offers several advantages as an audio-frequency source in frequency characteristic measurements.

R355.9 Franks, C. J. A laboratory oscillator for receiver testing. *Electronics*, **2**, 668–670; June, 1931.

An improved audio-frequency oscillator for use in testing and rating radio receivers is described. The oscillator covers a large range with constant output and is direct reading in frequency.

R355.9 Franks, C. J. and Ferris, M. The design and construction of standard signal generators. *Radio Eng.*, **11**, 37–44; June, 1931.

A detailed discussion of the general requirements of a standard frequency generator, suitable for testing modern, highly sensitive and selective radio receivers, is followed by a description of three particular generator models, each of which was designed to fulfill the general and some special requirements.

R361
×R134 Petrasco, E. Sur une méthode de reception des ondes courtes entre tenues (On a method of continuous short-wave reception). *L'Onde Electrique*, **10**, 141–170; April, 1931.

A method of continuous short-wave reception is described, which involves ordinary super-regeneration plus an additional local oscillator, for which several advantages are claimed.

R381 Rhodes, H. E. Intermediate-frequency tuning condenser requirements. *Electronics*, **2**, 690–691; June, 1931.

Precautions to be observed in the design and production of intermediate-frequency amplifiers are discussed.

R400. RADIO COMMUNICATION SYSTEMS

R423.4 New telephony system. *Wireless World and Radio Rev.* **28**, 590–593; June, 1931.

The apparatus used in a recent demonstration of a short-wave, single side-band, duplex telephony system is described.

R423.5 Karplus, E. Communication on the quasi-optical frequencies. *Electronics*, **2**, 666–667; June, 1931.

The potential possibilities of ultra-high-frequency communication are briefly discussed and some typical apparatus is described.

R430 Conrad, F. and Schöne, A. Die Aufsuchung von Störern des Funkempfanges (Searching for sources of radio interference). *Elek. Zeit.*, **52**, 697–700; May, 1931.

A description is given of several methods that have been successfully used in locating sources of man-made static.

R500. APPLICATIONS OF RADIO

R550 Wenstrom, W. H. Low-frequency high-power broadcasting as applied to national coverage in the United States. PROC. I.R.E., **19**, 971–983; June, 1931.

With P. P. Eckersley's general theory derived from north European practice as a starting point, the possibilities of broadcasting in the United States on frequencies around 200 kc are examined from the viewpoint of national coverage.

R550
×R355.5 Schwandt, E. Ultra-short-wave broadcasting. *Wireless World and Radio Rev.*, **28**, 526–628; May, 1931.

A brief description is given of the methods and apparatus used in Germany for local broadcasting tests at ultra-high frequencies.

R566 Graham, V. M. A radio receiver for police service. *Radio Eng.*, **11**, 31–32; June, 1931.

A modern police radio receiver is described. This set is a superheterodyne with automatic gain control and was specially designed for automobile service.

R590 Williams, H. L. Typical public address installations. *Radio Eng.*, 11, 47–48; June, 1931.

The need for careful planning of a public address installation is pointed out and a typical example illustrates various fundamental requirements common to all such installations.

R600. RADIO STATIONS

R612.1 Lubszynski, G. and Hoffmann, K. Die rundfunktechnischen Einrichtungen im neuen "Haus des Rundfunks" in Berlin. (The arrangement and equipment of the "House of broadcasting" in Berlin). *Elek. Zeit.*, **52**, 561–566; April, 1931.

A detailed description of the newly built broadcast center in Berlin.

R800. NONRADIO SUBJECTS

347.7 Rogan, J. J. Patent review on receiver circuits and tubes. *Electronics*, **2**, 672–673; June, 1931.

Important patents covering radio and sound amplification are listed and their status discussed.

530 Nordheim, L. Zur Elektronentheorie der Metalle (The electron theory of metals). *Ann. der Physik*, **9**, 607–640, No. 5, 1931; 641–678, No. 6, 1931.

A comprehensive mathematical treatment of the electron theory of metals, in which a complete theoretical structure is set up.

537.65
×R111 Errera, J. Dispersion von Hertzschen Wellen in fester Körpern (The dispersion of Hertzian waves in solid bodies). *Phys. Zeit.*, **32**, 369–373; May, 1931.

An experimental study of the dispersion of Hertzian waves in certain piezo-electric crystals is given.

537.65 Grossman, E. and Wein, M. Über den Einfluss der Umgebung auf die Frequenz eines Schwingquarzes. (The influence of ambient conditions on the frequency of a quartz crystal). *Phys. Zeit.*, **32**, 377–378; May, 1931.

Reference is made to the reflection of sound energy to the oscillating crystal as a serious cause of frequency variation.

537.65 Koga, I. Note on the piezo-electric quartz oscillating crystal regarded from the principle of similitude. PROC. I.R.E., **19**, 1022–1023; June, 1931.

It is pointed out that the principle of similitude in the vibrating periods of an aeolotropic elastic body has several useful applications. As an example the case of X-waves in X-cut quartz plates is discussed.

537.87 Ancelme, P. Applications médicales des ondes ultracourtes. (Medical applications of ultra-high frequencies). *L'Onde Electrique*, **10**, 221–232; May, 1931.

A résumé of experiments performed by the author to determine the biological effect of ultra-high frequencies (40 to 80 megacycles) on living tissue.

621.313.7 Schottky, W., Störmer, R. and Waibel, F. Über die Gleichrichterwirkungen an der Grenze von Kupferoxydul gegen aufgebrachte Metallelektroden. (On the rectifying action of cuprous oxid in contact with other metals). *Zeit. für Hochfrequenz.*, **37**, 162–167; April, 1931; 175–187; May, 1931.

The results are given of a series of experiments carried out to determine the rectifying properties of the oxides of copper in contact with other metals.

621.375.1 Verman, L. C. and Richards, L. A. A vacuum-tube voltage regulator. *Elec. Eng.*, 50, 436–8; June, 1931.

A vacuum-tube device is described which not only provides voltage regulation for alternators, but short-circuit protection as well. Saturation current from filament to plate forms the control element while stabilization is accomplished by a feed-back system.

621.383.21 Huff, C. Automatic time-delay relay. PROC. I.R.E., 19, 1019–21; June, 1931.

The design of an automatic time-delay relay is described. For certain types of mercury-rectifier tubes, no plate voltage should be applied for at least one-half minute after the filament current is turned on if the peak inverse potential exceeds 2100 volts. During a six-months' test of this relay on a special radio transmitter, it has not failed to function at any time.

621.385.91 MacCoun, T. D. Radio program distribution over lighting circuits. *Electronics*, 2, 682–683; June, 1931.

A method of distributing radio programs by using the regular lighting wires without the expense of rewiring the building for audio circuits is described.

629.13 Sonic altimeter for fog flying. *Sci. Amer.*, 144, 264–266; April, 1931.

Describes an altimeter designed for aircraft use. The air supply for the whistle is obtained by "bleeding" one of the engine cylinders through a check valve into a small storage tank. Each time the whistle valve sends out a blast, a pointer starts moving uniformly around its scale, from which the height is read directly.

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Hansell, Clarence Weston: Born January 20, 1898 at Medaryville, Indiana. Student, training course of Commonwealth Edison Company, Chicago, summer, 1918. Received E. E. degree, Purdue University, 1919. General Electric Test Course, 1919–1920; radio engineering department, 1920; engineering department, Radio Corporation of America, 1920 to date; at present in charge of transmitter development laboratory of R. C. A. Communications at Rocky Point, New York. Member, A.I.E.E., American Association for the Advancement of Science, and Franklin Institute. Associate member, Institute of Radio Engineers, 1926; Member, 1929.

Hastings, Harris F.: Born April 14, 1896. Graduated, Emerson Institute, 1928. Amateur radio operator and experimenter, 1910–1925. Commercial radio operator, 1920–1921. Elliot Woods Laboratory, Washington, spare time, 1912–1920; Naval Research Laboratory, Bureau of Standards, 1918–1919; Marcus Hopkins Laboratory, Washington, 1923–1925; Naval Research Laboratory, Bellevue, D.C., in conjunction with schooling; general engineer, Universal Wireless Communication Company, 1928–1930; assistant radio engineer, Naval Research Laboratory, 1930 to date. Associate member, Institute of Radio Engineers, 1918.

Kurlbaum, Georg: Born May 2, 1902 at Berlin-Charlottenburg. Studied in Tübingen and Berlin; received degree of diploma-engineer, May, 1926. Scientific assistant, Federal Central Post Office, radio division, Berlin-Tempelhof, 1926–1927; assistant, German Experimental Institute for Air Navigation, E. V., Berlin-Adlershof, electrical engineering and radio division, September 1927 to April, 1929; Colonial Airways Corporation, New York, and development department, Automatic Electric, Inc., Chicago, July, 1929 to July, 1931. Associate member, Institute of Radio Engineers, 1930.

McNamara, Francis T.: Born July 6, 1896 at Clinton, Massachusetts. Received Ph. B. degree, Yale University, 1921; E.E. degree, 1924. General Electric Company Test Course, 1922; Transmission and Distribution, Philadelphia Electric Company, 1923. Instructor, Yale University, 1923–1928; assistant professor, 1928 to date. Associate member, Institute of Radio Engineers, 1931.

Miura, Itomi: Born May 7, 1904 at Sado, Japan. Graduated, electrical department, Tokyo Higher Technical School, 1925. Radio section, Electrotechnical Laboratory, Ministry of Communications, Japan; engaged in work on radio transmitters and vacuum tubes, 1925 to date. Nonmember, Institute of Radio Engineers.

Patterson, Edward B.: Born 1902. Received B.S. degree Haverford College, 1924; graduate work, University of Pennsylvania. Amateur radio and commercial operating, 1913 to date. Radio editorial work, 1924–1926; engineering department, Victor Talking Machine Co. and RCA Victor Co., 1926 to date. Member, American Institute of Electrical Engineers and Franklin Institute. Associate member, Institute of Radio Engineers, 1924; Member, 1930.

Peterson, Harold O.: See PROCEEDINGS for April, 1931.

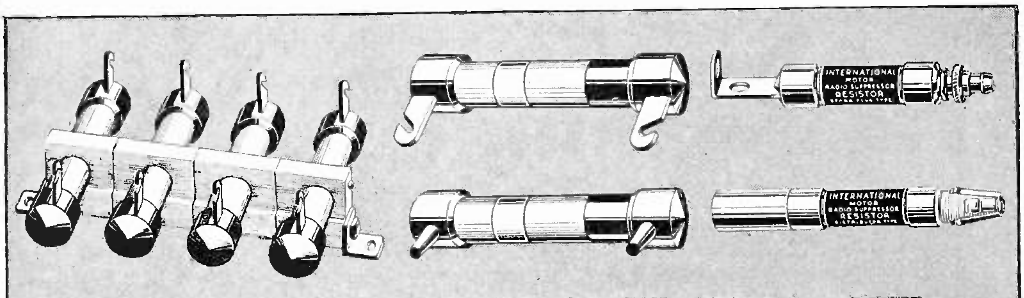
Roder, Hans: Born September 27, 1899 at Mengersreuth, Bavaria, Germany. Institute of Technology, Munich, 1919–1923; received M. S. in E.E. degree, 1923. Development and research of transmitting apparatus, Telefunken

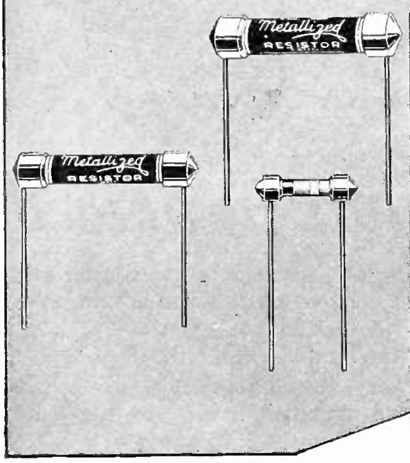
Gesellschaft für drahtlose Telegraphie, 1923–1929; radio engineering department, General Electric Company, 1930 to date. Member, Institute of Radio Engineers, 1929.

Schafer, J. Peter: Born October 29, 1897 at Brooklyn, New York. Received B.S. in E.E. degree, Cooper Union, 1921; E.E. degree, 1925. Member, technical staff, research department, Western Electric Company and Bell Telephone Laboratories, 1915 to date. New York Laboratories, 1915–1922; Rocky Point Transatlantic Radiotelephone Station, 1922–1928; Deal Radio Laboratories, 1928 to date. Member, A.I.E.E. Associate member, Institute of Radio Engineers, 1924; Member, 1930.

Taylor, A. Hoyt: see PROCEEDINGS for February, 1931.

Vreeland, Frederick King: Born March 4, 1874 at Bergen, New Jersey. Received M.E. degree, Stevens Institute, 1895; Sc.D. degree, 1921; M.A. degree, Columbia University, 1909. Assistant engineer, Crocker-Wheeler Electric Company, 1898–1900; research work, 1900 to date; president, Vreeland Apparatus Company, 1905 to date. Nonresident lecturer, Stevens Institute, 1907–1909; Columbia University, 1909. Fellow, Physical Society, Electrical Engineers, Mathematical Society, Franklin Institute, New York Academy. Member, Institute of Radio Engineers, 1916; Fellow, 1926.





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Special types reduce assembly costs. Advanced designs and complete ranges anticipate the needs of new industries. As pioneers in the business we have long co-operated with the world's leading engineers. We welcome a discussion of your own problems.

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- TELEPHONE APPARATUS
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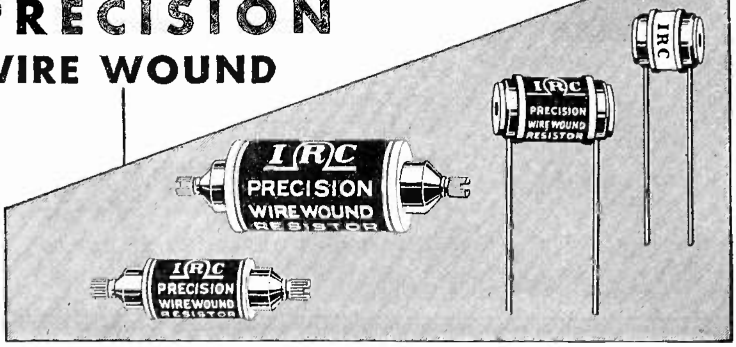
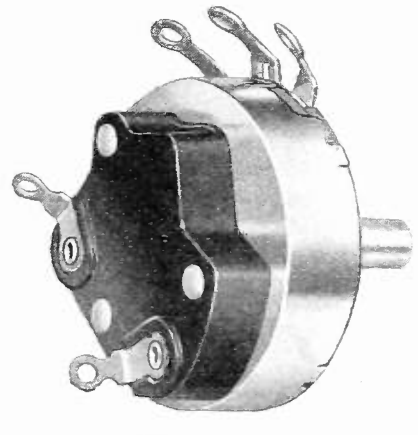
- *And other industrial uses*

PRECISION
WIRE WOUND



METALLIZED — made in 1/3 Watt, 1/2 Watt, 1 Watt, 2 Watt and 3 Watt Sizes.

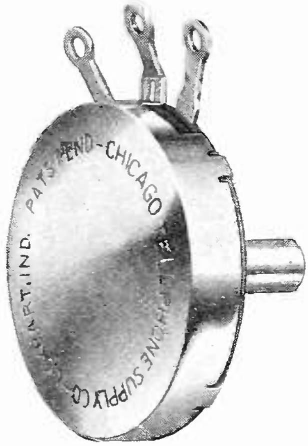
PRECISION WIRE WOUND — made in ranges of 1 ohm to 2 1/2 megohms. Tolerance 1% or better.



At left: No. 70 Series, with "Underwriters' Laboratories Inspected" Switch, Rated 1.5 Amps. 250 Volts, 3 Amps. 125 Volts.

At right: No. 70 Series, without switch.

Both illustrations actual size

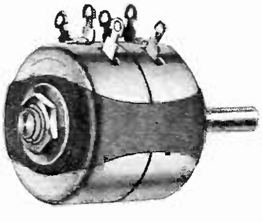


Smaller...better...*lower priced*

ANNOUNCING the New "Seventy" Series Composition Element Volume and Tone Controls, furnished either with or without built-in switch and in whatever resistance gradient and resistance value that may be desired.

These new units have been developed in response to a demand for high class, small controls, which can be purchased for less money. The surprisingly low price is made possible by design economies and improved manufacturing methods and not through the employment of lower grade materials or less painstaking workmanship.

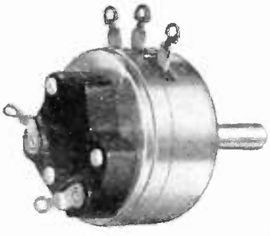
No. 20 Series Single Control.



No. 20 Series Tandem Unit.

The outstanding feature of the No. 70 Series Control, is the full wiping contact between the movable contact member and the hard smooth composition resistance element. A new formula developed in our Engineering Laboratory has made possible this ideal contacting method and despite the fact that we employ a relatively high unit pressure between the contacting surfaces, the frictional component is so low that life tests of 100,000 complete operations leave these members in even better condition than before fatiguing was begun.

We shall be glad to send literature to interested engineers in which we have more fully described the seasoned principles involved in the construction of these controls and upon receipt of specifications, will send without charge, samples possessing electrical and physical characteristics that will exactly suit your circuit conditions.



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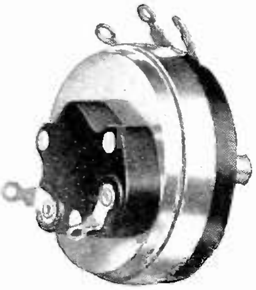

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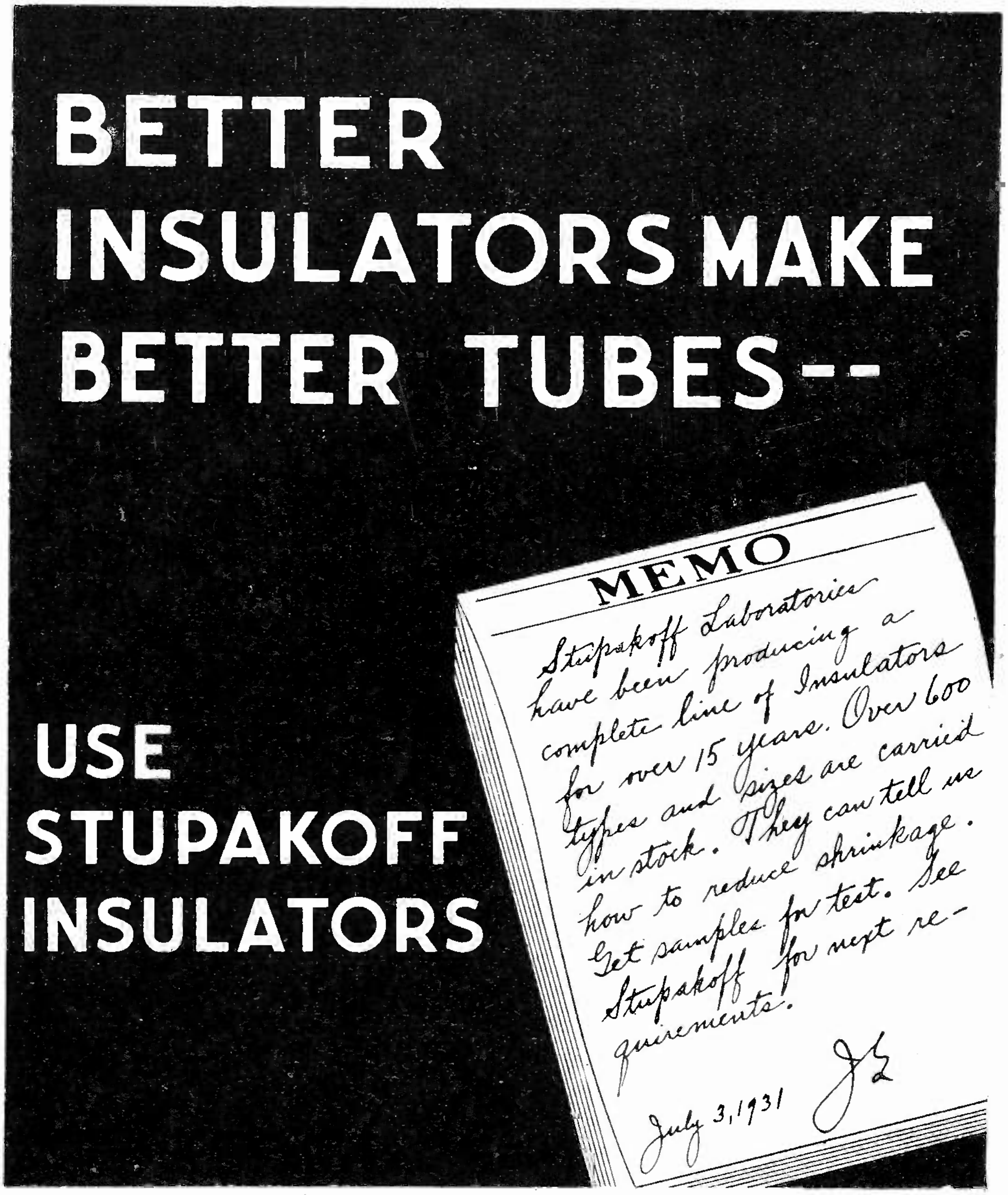

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July 3, 1931

STUPAKOFF LABORATORIES, Inc.

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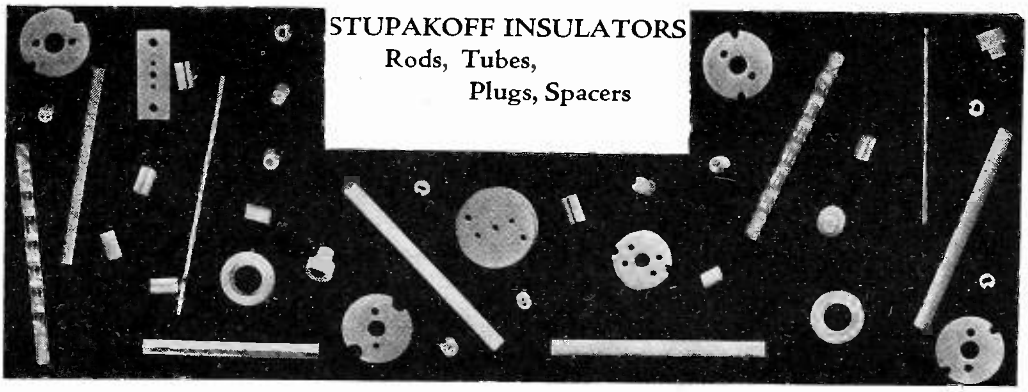
Cable address, Stupakoff, Pittsburgh



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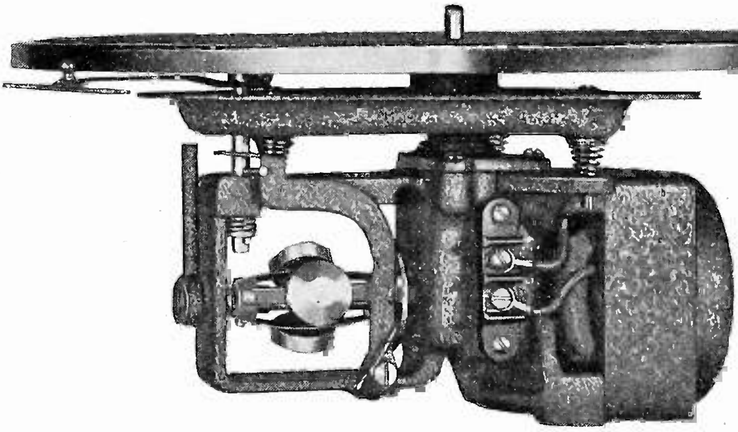

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Specially designed for Radio-Phonographs. Self-starting, amply powerful, supplying uniform speed to perfection, and silent, uninterrupted service. Made for all voltages and frequencies. (For direct current, either 110 or 220 volts, ask for the d.c. Green Flyer.) Open construction with complete ventilation. Furnished complete, with mounting plate, turntable and speed regulator. Automatic stop is optional.

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Right from the start

The right motors for Radio-Phonographs

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FROM the first sample submitted to leading engineers for examination, well over two years ago, Flyer Electric Radio-Phonograph Motors have been known as the motors that started out right. They have *always* been right. Everywhere, doing their work under conditions of every kind, they have steadily built up a loyal following of representative large users.

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In ordering samples give voltage and frequency.

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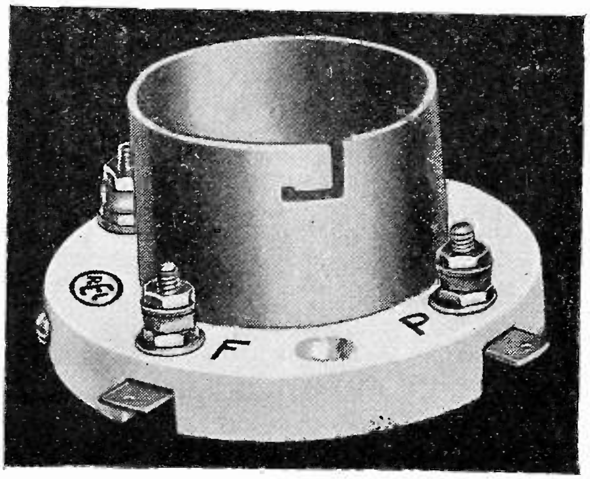
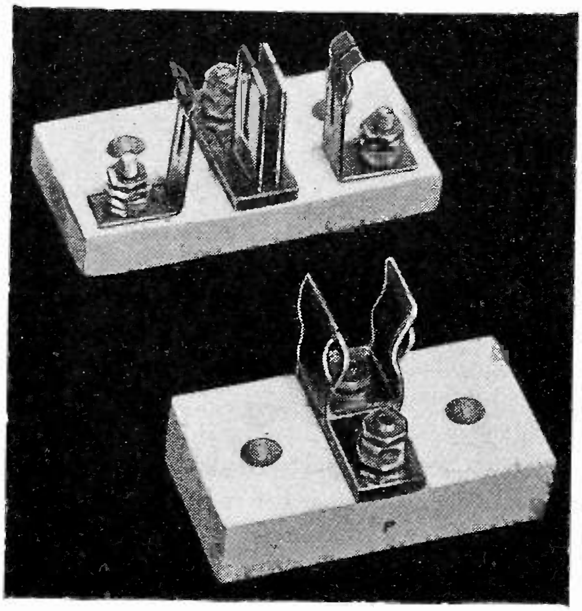
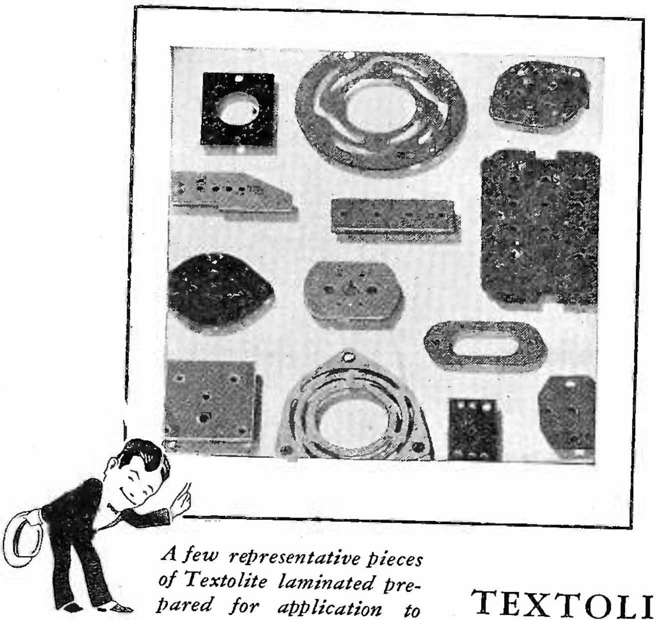
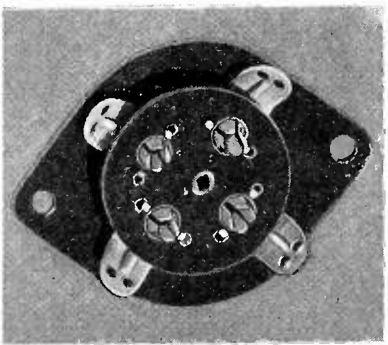


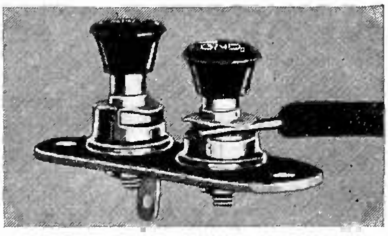




YOUR PROBLEMS ARE *OUR* PROBLEMS!



CINCH SOCKETS



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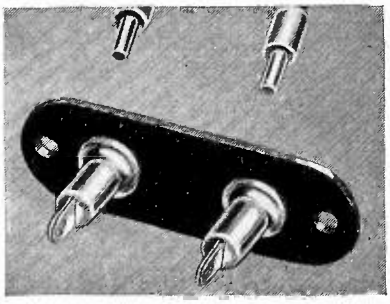

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IT COST BILLIONS OF DOLLARS TO BUILD

YET YOU CAN USE IT FOR A FEW CENTS A DAY

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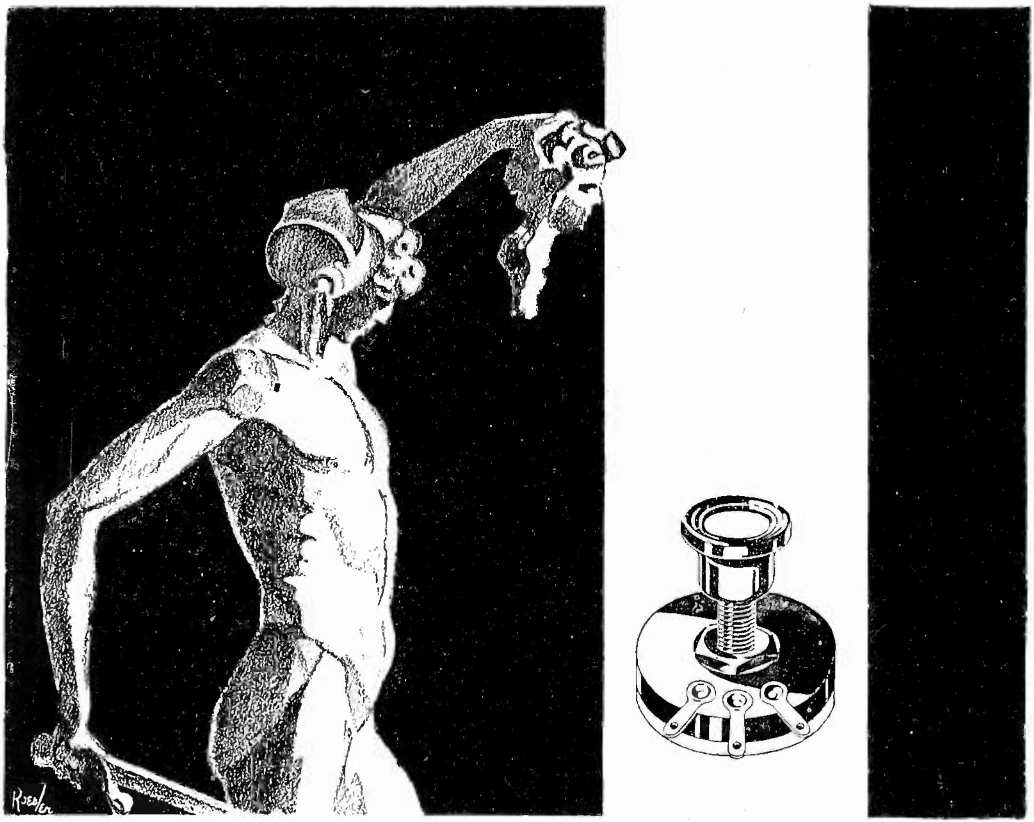
At any hour of the day or night, the telephone stands ready and waiting to carry your voice to any one of twenty million other telephone users in this country. It knows no rest or sleep, or class or creed. All people —everywhere—may use it equally. Its very presence gives a feeling of security and confidence and of nearness to everything.

Many times during the day or week or month, in the ordinary affairs of life and in emergencies, you see the value of the telephone and realize the indispensable part it plays in every business and social activity.

The growth of the Bell System through the past fifty-five years and the constant improvement in service may well be called one of the great achievements of this country. Greater even than that are the policies, improvements and economies that make this service possible at such low cost.

Of all the things you buy, probably none gives so much for so little as the telephone.

★ *AMERICAN TELEPHONE AND TELEGRAPH COMPANY* ★



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These 19 characteristics will convince any engineer of Acracon's superiority in electrical efficiency and mechanical construction:

1. Metal cover for protection and appearance.
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3. Nipple spun into aluminum shell. Absolutely leak-proof.
4. Anode spiral, cold welded into anode, giving rigid construction.
5. One piece extruded aluminum container.
6. Retaining flange for rubber gasket.
7. Tapered anode stem for snug fit.
8. Large cadmium plated steel mounting nut, concave to insure tight connection.
9. ¾"-16 thread neck for mounting.
10. Metal washer.
11. Anode nut.
12. Anode soldering tab.
13. Large size insulating washer.
14. Tapered hole to take tapered anode.
15. Special live, rubber insulating gasket free from impurities.
16. Heavy, rigid, anode stem of high purity aluminum.
17. High purity anode, spiral, so wound as to eliminate the necessity of insulating liner between anode and container.
18. Special, high, critical voltage electrolyte, well over anode to insure long life.
19. Leak-proof rolled seam as used in canning industry.

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Condenser Corporation of America

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Factory Representatives In:

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8-31

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.

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. I furthermore 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.

Yours respectfully,

....
(Sign with pen)

....
(Address for mail)

....
(Date) (City and State)

References:

(Signature of references not required here)

Mr. .... Mr. ....
Address .... Address ....
Mr. .... Mr. ....
Address .... Address ....
Mr. ....
Address ....

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: * * * (d) Associates, who shall be entitled to all the rights and privileges of the Institute except the right to hold the office of President, Vice-president and Editor. * * *

Sec. 5: An Associate shall be not less than twenty-one years of age and shall be: (a) A radio engineer by profession; (b) A teacher of radio subjects; (c) A person who is interested in and connected with the study or application of radio science or the radio arts.

ARTICLE III—ADMISSION

Sec. 2: * * * Applicants shall give references to members of the Institute as follows: * * * for the grade of Associate, to five Fellows, Members, or Associates; * * * Each application for admission * * * shall embody a concise statement, with dates, of the candidate's training and experience.

The requirements of the foregoing paragraph may be waived in whole or in part where the application is for Associate grade. An applicant who is so situated as not to be personally known to the required number of members may supply the names of non-members who are personally familiar with his radio interest.

(Typewriting preferred in filling in this form) No..................

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

1 Name ....
(Give full name, last name first)

2 Present Occupation ....
(Title and name of concern)

3 Permanent Home Address ....

4 Business Address ....

5 Place of Birth........................ Date of Birth.............Age......

6 Education ....

7 Degree ....
(college) (Date received)

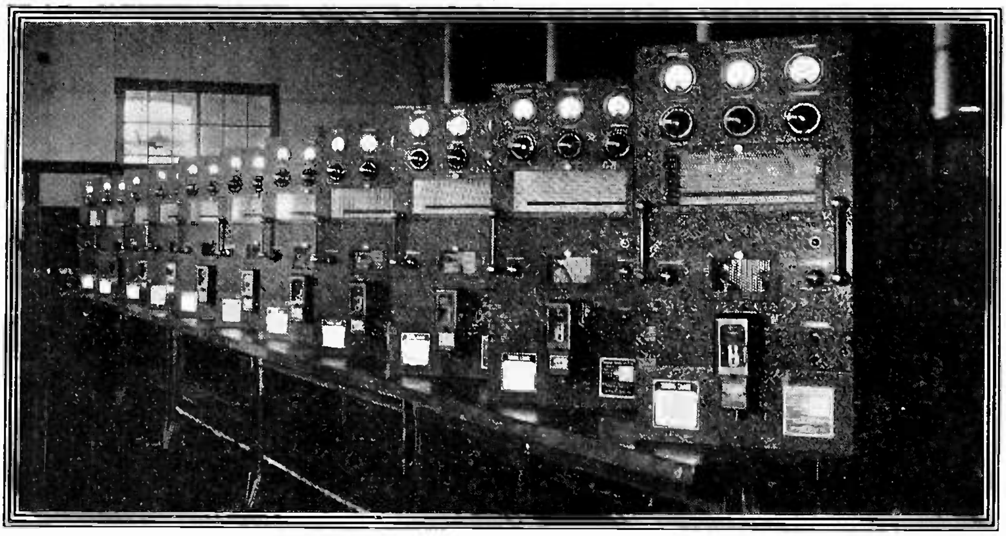
8 Training and Professional experience to date...............................

NOTE: 1. Give location and dates. 2. In applying for admission to the grade of Associate, give briefly record of radio experience and present employment.

DATES HERE	

9 Specialty, if any..

Receipt Acknowledged............ Elected............ Deferred............
Grade........... Advised of Election.......... This Record Filed..........



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Ten Type LSR-303 Radio Beacon Transmitters have been built by DeForest for the Bureau of Lighthouses, Department of Commerce. Each transmitter covers the frequency range of 255-335 K.C., with a power output rating of 10-30 watts. 100% modulation is obtained by an unique method. Provision is made for a motor-driven key, whereby the transmitter may automatically repeat its identifying call letters. Operation is reduced to the mere pressing of buttons by lay hands. A 110-volt 60-cycle A.C. supply is employed.

To complete the installation, four DeForest Transmitting Audions are employed—two 510s, one 511 and one 545—together with two DeForest mercury vapor rectifiers.

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After All, There's No Substitute for 25 years' Experience

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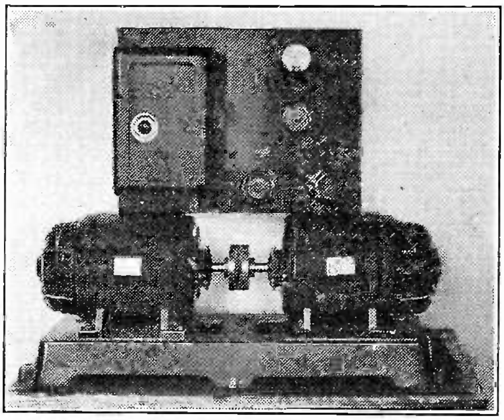
ESCO *Motor-Generators*



"Fit" the application **"Fit" the power supply**

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Motors—Generators—Dynamotors—Rotary Converters

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Designed · Developed · Manufactured
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"ESCO" is a company where "Special" does not mean "soak 'em" or "Let 'em wait." Because for twenty years "ESCO" has specialized in the "Special." "Special" voltages, frequency, speed. and mechanical design is the objective of our equipment and organization.
OUR EXPERIENCE IS BROAD. Below is a list of some of the special applications of our motors and generators, manufactured during this last December.

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Pianos
Electric indicators
Speed reducers
Ventilators
Telechron resetters
Automatic chucks
Recording instruments
Elevator door control
Elevators
Flexible shafts
Riveters
Gasoline pumps
Bottle washers
Phase shifters
Frequency changers
Hair dryers
Arc lamps
Anti-corrosion
Oil burners
Automatic whistle
Ore-vibrators
Organ generators
Hoists
Freight trucks
Temperature control
Polishers
Stock quotations
Sun lamps
Bookkeeping unit
Forced draft
Knitting mills
Printing press
Electro plating
Remote control
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COMPANY

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Manufacturers of motors, generators, dynamotors and rotary converters.

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and ITS SOLUTION!

WITH home recording promising to be an outstanding feature of next year's radio receivers, the problem of HUM is more serious than ever.

Even a slight hum not noticeable in radio reproduction will, when recorded and augmented through reproduction, make itself heard with sufficient volume to prove disagreeable. In this modern day of radio, hum is not to be tolerated in radio or phonograph reproduction.

For more than twenty years, Pacent engineers have concerned themselves exclusively with problems relating to sound amplification. That's why Pacent amplifiers are the best in the field today.

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A Pacent Amplifier, the 170 Recordovox and the 107 Hi-Output Phonovox make a remarkable combination for recording and reproducing. With this apparatus, it is possible to assure professional results.

The Recordovox and Phonovox are available in special manufacturers' types. Write for additional information.

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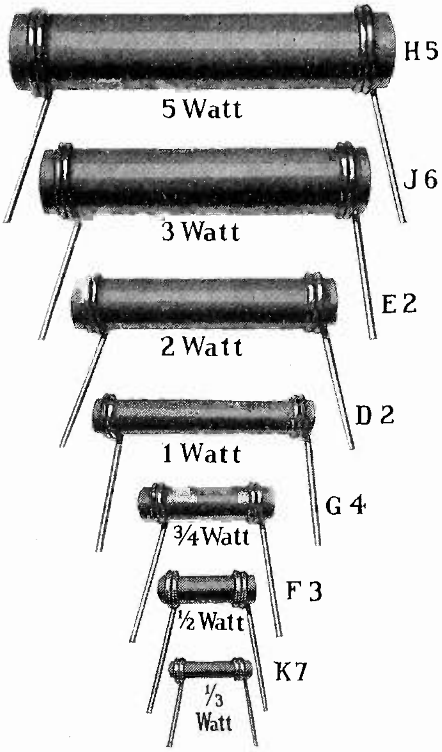
PACENT

Continental Resistors



RESISTOR CHARACTERISTICS:

NOISELESS
RESISTANCE CONSTANT
SMALL TEMPERATURE COEFFICIENT
RUGGED
FIRMLY SOLDERED TERMINALS
DEPENDABLE
COLOR CODED

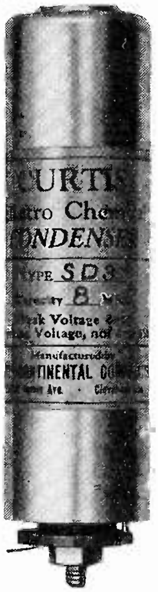
UNITS SHOWN HALF SIZE

CONTINENTAL SUPPRESSORS

For Radio-Equipped Cars

Stop ignition noise in the automobile radio set when used with suitable by-pass condensers. The distributed capacity is extremely small. The resistors are enclosed in a tough ceramic tube of high crushing and tensile strength, of low coefficient of expansion and of high dielectric strength. They are hermetically sealed in the tube and this renders the suppressors moisture proof.

Suppressors shown full size



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Essential Characteristics

- Full capacity at all voltages
- Uniform capacity at all frequencies
- Low freezing point
- Low internal resistance
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- No nipples, new breather principle
- One hole mounting terminal at bottom

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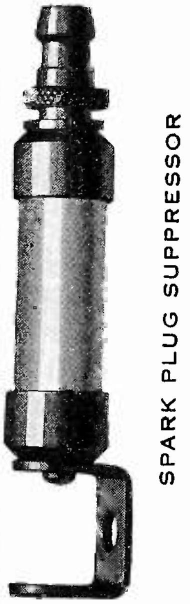
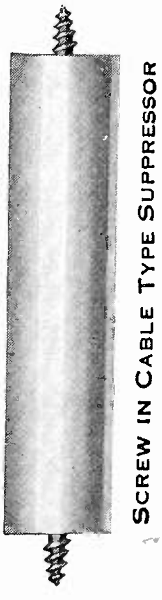

SPARK PLUG SUPPRESSOR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DISTRIBUTOR SUPPRESSOR

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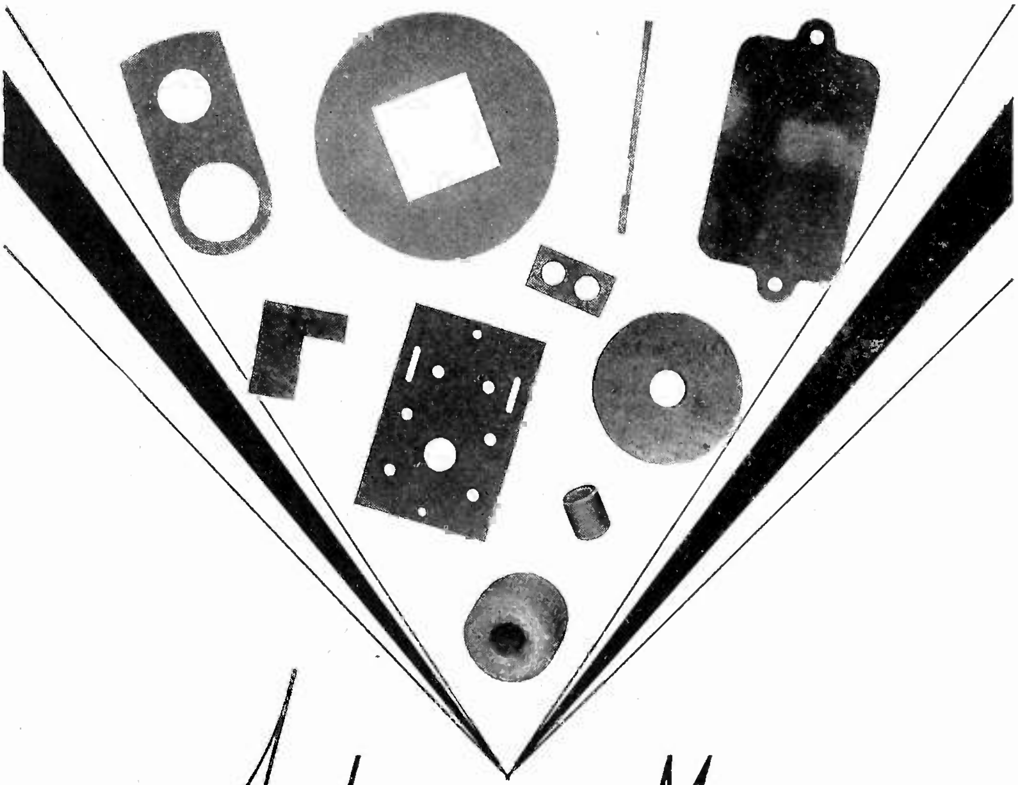

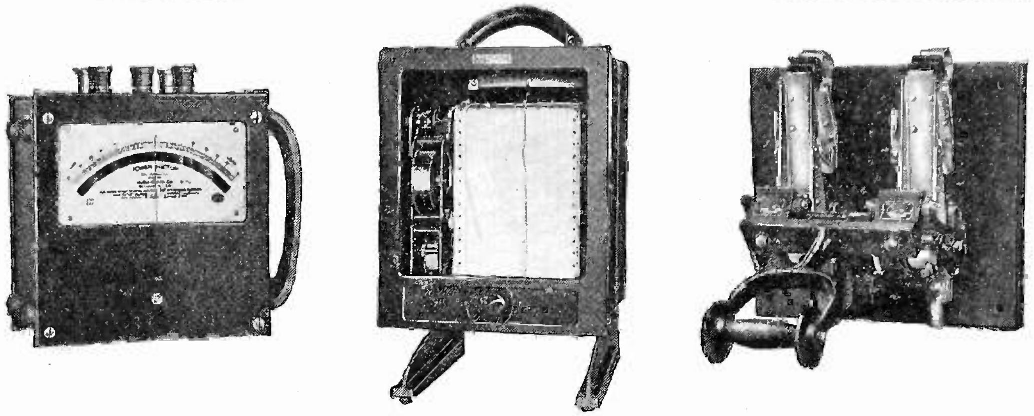

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Special Prices Will Be Quoted in Quantities of Ten or More

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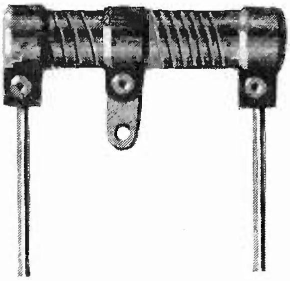

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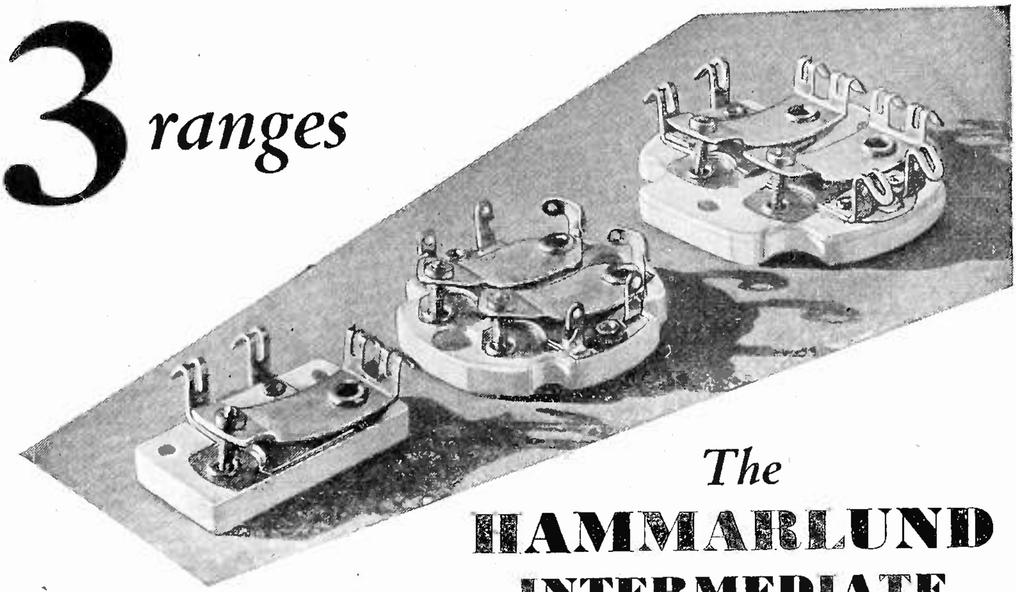

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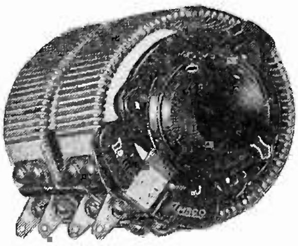

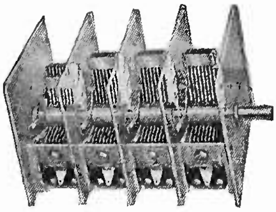

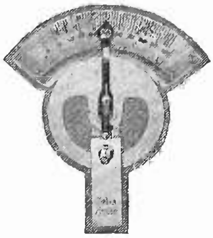

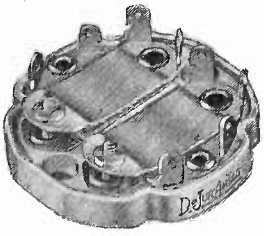

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Frequency Range: 75 KC to 10,000 KC (4,000 meters to 30 meters) uses six shielded plug-in coils removable from front of panel.

RF Voltage Output: ¼ microvolt to 2 volts; continuously variable, push-pull RF oscillator, output substantially constant over frequency range.

Modulation: Internal *400 cycle* sine wave oscillator, capable of modulating output up to *80% in* 10% steps. Harmonic content less than 5%. Provision for external modulation. Self calibrating modulation meter employed.

Output System: Special designed, low impedence, resistance network attenuator accurate within ± 3% even at the high frequencies. Output available through internal standard IRE dummy antenna or directly from attenuator. Long scale precision type meter with mirror and knife edge pointer used to measure voltage input to attenuator; calibrated directly in microvolts.

Frequency Calibration: Accurate within ± 0.5 of 1%.

Frequency Modulation: Less than .02% at 30% modulation.

Leakage: Potential difference between any external grounded point, meters or controls, is less than 0.1 microvolt. The stray field is not sufficient to effect measurements within the range of the instrument.

Mechanical Features: Enclosed in heavy aluminum shielded case finished with black crackle lacquer. All internal parts carrying RF are enclosed in individual shields. Dimensions 20″ x 11″ x 8″. Weight, 30 lbs.

Tubes Required: Three RCA-231, One RCA-230.

Power Supply: External batteries; filament supply 6 volts, 0.425 amps. Plate Supply 90 volts, .018 amps.

Price: $875.00 less tubes and batteries.

(*Write for Bulletin "A" covering RCA Victor "Laboratory and Test Instruments"*)

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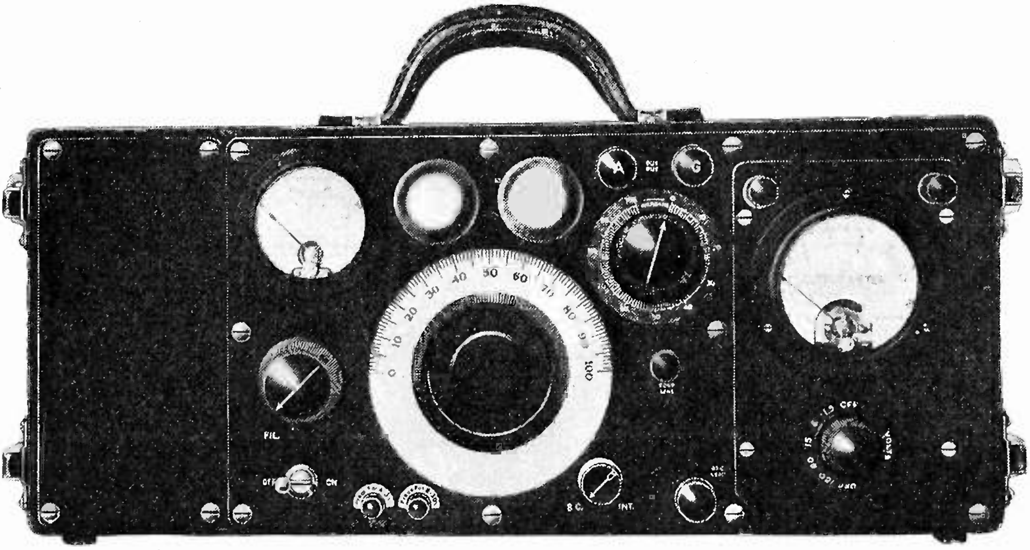
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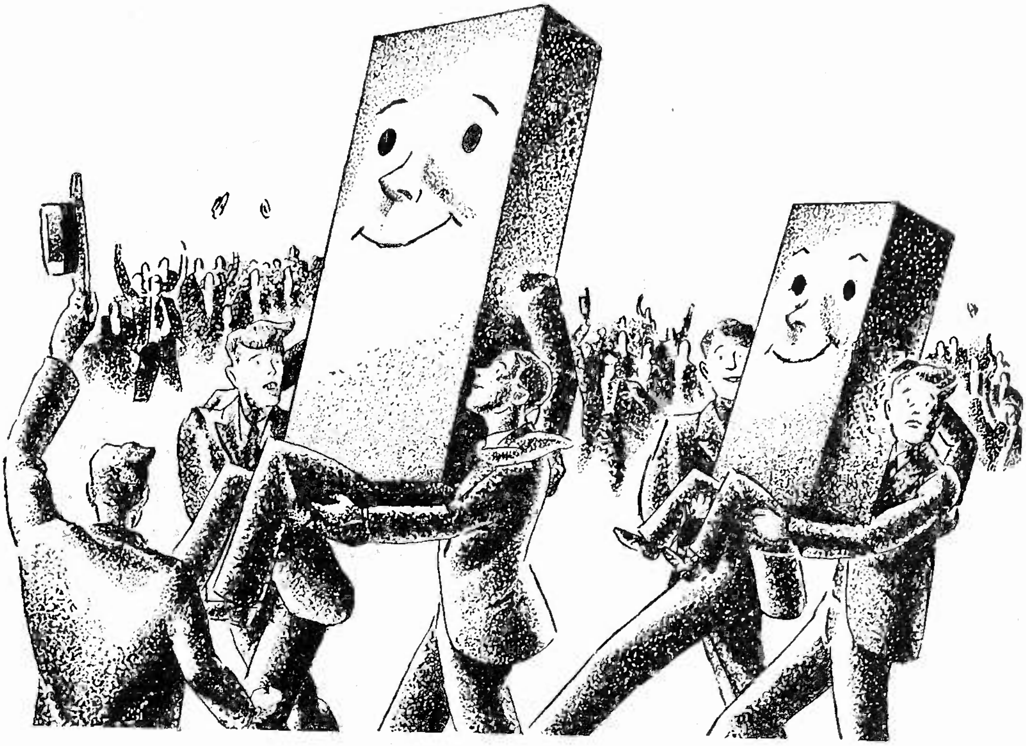

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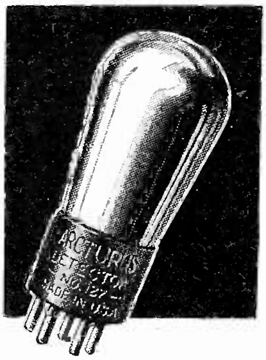
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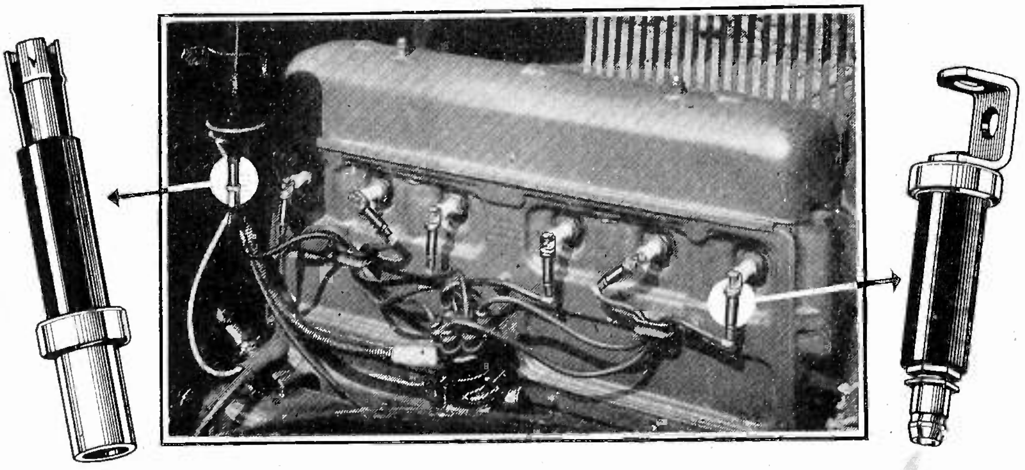

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No Shielded Ignition Cable Needed

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They increase the resistance of the high tension ignition system and minimize the disturbing oscillations in the ignition circuit which interfere with the radio receiver in the car. When used with suitable by-pass condensers in other parts of the circuit, shielded ignition cables are unnecessary.

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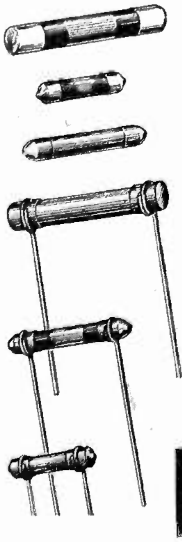

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Bradleyunits are solid molded resistors unaffected by temperature, moisture or age.

Their accurate calibration, great mechanical strength and performance make them ideal for providing correct C-bias, plate voltage, screen grid voltage and for use as grid-leaks and as fixed resistors in resistance-coupled circuits.

All units are color-coded to meet any radio set manufacturer's specifications.

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The Bradleyometer is a potentiometer with approximately fifty solid resistance discs interleaved between metal discs.

The total number of discs can be arranged in accordance with any resistance-rotation curve.

One or more Bradleyometers can be arranged to operate with one knob. Mixer controls, T-pad and H-pad attenuators and other complex controls can be provided.

Type A. Single Bradleyometer

Type AA. Double Bradleyometer

Type AAA, Triple Bradleyometer

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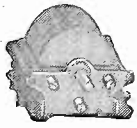
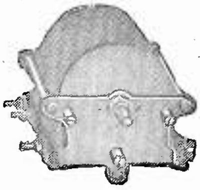
Produced by the makers of Allen-Bradley Control Apparatus

A WORTHY ADDITION TO A FAMOUS LINE—THE CARDWELL 16-B

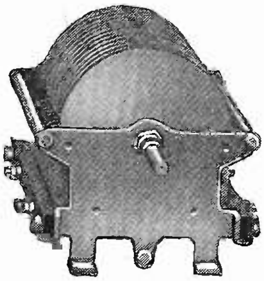
For Transmitting



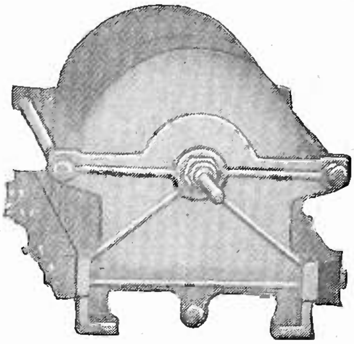
Midway
(featherweight)



"Standard"



16-B



166-B

This transmitting model is primarily intended to meet requirements where condensers for moderately high voltages are indicated of a size between our Construction Design 166-B and our smaller condenser designs such as T-183, T-199, etc.

It is possible to furnish in the Construction Design 16-B, greater airgaps, and higher capacities in relation to airgaps, than could well be done using our smaller transmitting condenser design as mentioned above, and still retain adequate structural strength and a proper balance between the various elements. Likewise condensers of low capacity with airgap equivalent to that in our standard 166-B may be furnished in this construction.

The 16-B can be supplied with promptness and within reasonable limits as to capacity and breakdown voltage. Standard airgaps (actual airgap between adjacent rotor and stator plates) are .168 inches and .294 inches, but condensers with airgaps of .231, .122 and .090 inches can be furnished on special order. We solicit inquiries for special sizes.

The figures given below indicate a few possible sizes. The number of plates determining increase or decrease in capacity can be accommodated to suit special requirements, provided that an overall depth behind panel of 11 inches, as indicated for Type 3276, is not exceeded.

Construction Design 16-B, Type Nos.	Max. Cap.	*Air Gap	No. of Plates	Depth Behind Panel *(Overall)*	List Price
3279	315 mmf.	.168 in.	31	9 9/32″	$34.00
3280	147 "	.168 "	15	5 13/16″	31.00
3281	84 "	.168 "	9	4 1/2″	28.00
3276	160 "	.294 "	25	11″	32.00
3277	80 "	.294 "	13	6 7/8″	30.00
3278	47 "	.294 "	7	4 13/16″	28.00

* Actual airgap between adjacent rotor and stator plates. All plates have well rounded edges and are highly polished overall.

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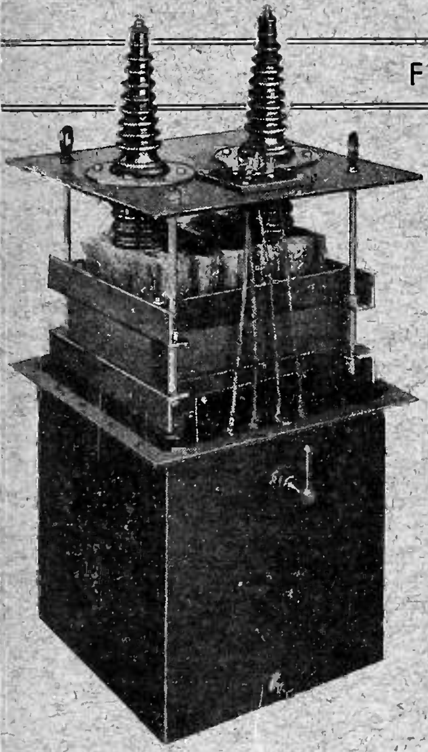
FROM TRANSFORMER HEADQUARTERS



TRANSFORMERS

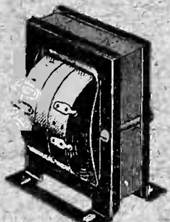
BUILT to ANY SPECIFICATIONS

Fifty thousand volt secondary potential with 220 and 440 volt input, to test insulation breakdown—this was the only specification given C. T. C. engineers by a large manufacturer of insulated wire.

C. T. C. quickly produced this oil-immersed, mica-insulated high potential transformer. A control panel was also designed with an input voltage regulator and a circuit breaker that operates automatically to disconnect the transformer when the insulation under test breaks down.

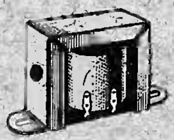
The production of this special transformer is only one of the many instances where the C. T. C. organization has proven its ability to meet unusual demands. If you are facing transformer problems of design, cost, or quick delivery in large quantities, C. T. C. service can solve them for you. Let us quote you today.

Immediate Production With C. T. C. Semi-Standard Designs



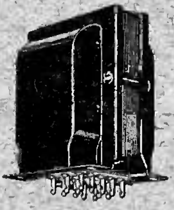
Inter-Stage Transformers

The flexible design of C. T. C. Semi-Standard units can be adapted to fill your requirements for coupling transformer and impedance audio stages. Don't delay your production by developing special designs.



Speaker Coupling Transformers

Transformers to couple the output of single '47, push-pull '47, and '50 stages to dynamic speaker voice coils may be selected without delay for your production from the C. T. C. Semi-Standard line.



Power Supply Transformers

Sets employing '80 type rectifiers to supply assemblies using from 4 to 11 tubes can go into production weeks earlier with C. T. C. Semi-Standard power transformers adapted to their needs.

CTC

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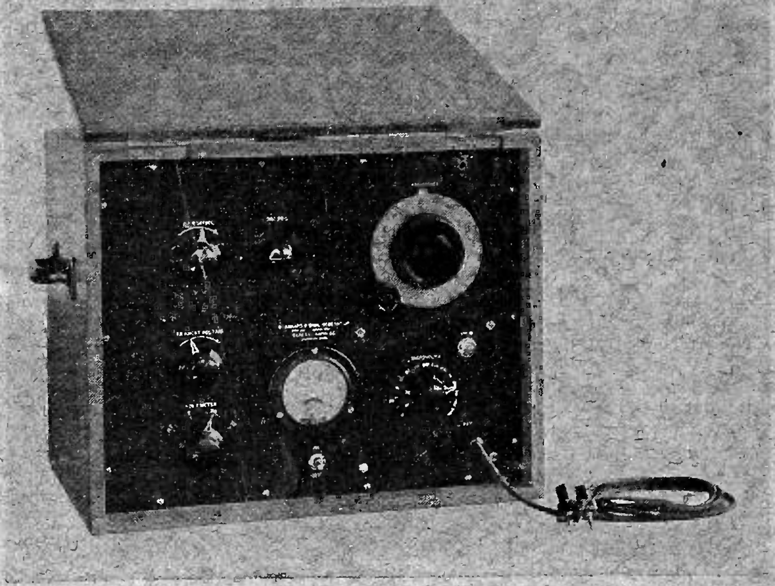
CHICAGO TRANSFORMER CORP., 2622 WASHINGTON BLVD., CHICAGO, U. S. A.

A High-Frequency Standard-Signal Generator

90 kc. to 6000 kc.

For aircraft, police, and broadcast receivers



TYPE 601—A Standard-Signal Generator
Price $210.00, Inductors extra

FEATURES

1. Self-contained batteries for convenience and portability.
2. Wide frequency range.
3. Ruggedness with accuracy.

CONDENSED SPECIFICATIONS

OUTPUT VOLTAGE: 0-150,000 microvolts from a calibrated attenuator in steps of 0-1-2-5-10, etc., microvolts, with a multiplying factor of 1x and 1.5x, indicated by means of a self-contained vacuum-tube voltmeter.

FREQUENCY RANGE: 90-6000 kc. covered by the use of six plug-in inductors, any two of which may be plugged in at one time and a selection between them made by means of a switch. A special inductor is available for the range 150-200 kc. of the superheterodyne broadcast receivers.

MODULATION: An internal modulating oscillator gives 400 cps. at 30%, although 50% modulation will be supplied on special order at no extra charge.

TUBES: Three 230-type tubes are used. The total plate-battery drain is only 2 mils, thus assuring long battery life.

Catalog Supplement F-306 contains a complete description of this and the two other General Radio standard-signal generators. Write for it.

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