VOLUME 18 DECEMBER, 1930 NUMBER 12

PROCEEDINGS

of

The Institute of Radio Engineers

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Institute of Radio Engineers
Forthcoming Meetings

CINCINNATI SECTION

December 16, 1930

January 13, 1931

NEW YORK MEETING

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January 7, 1931

SAN FRANCISCO SECTION

December 17, 1930

SEATTLE SECTION

December 19, 1930

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The Institute of Radio Engineers

Volume 18 — December, 1930 — Number 12

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The Institute of Radio Engineers

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Harold P. Westman, *Secretary*

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

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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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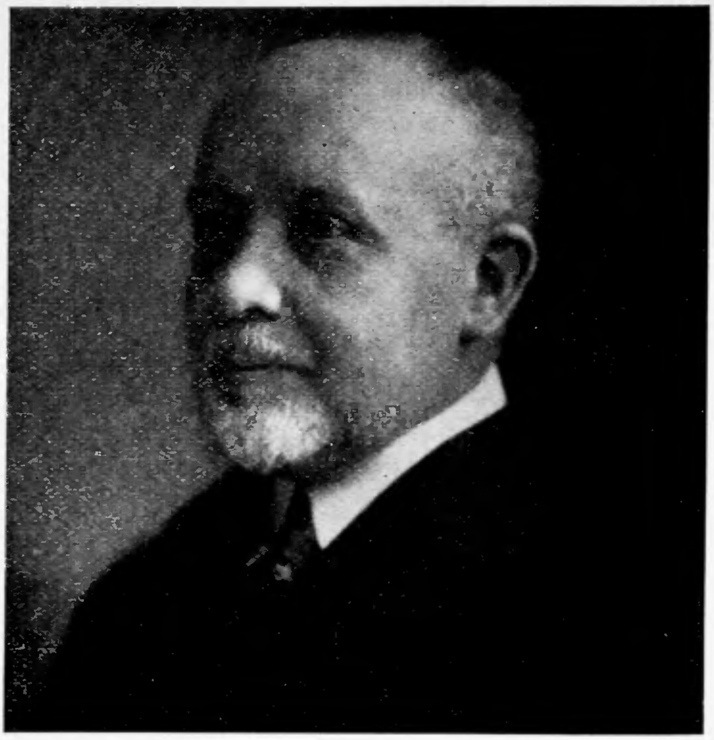


PEDER OLUF PEDERSEN
Medal of Honor Recipient, 1930

Peder Oluf Pedersen was born in Sig, Denmark, June 19, 1874, the only son of a farmer.

He attended the village school and in 1889 sent to the King of Denmark proposals for a perpetual motion device and a calculating machine. These so interested the King in his further education that he was transferred to Copenhagen where, after the necessary preparatory education, he entered the Royal Technical College, graduating with honor in civil engineering in 1897. The expense of his studies was contributed to by the King.

He soon became interested in electrical research work and in 1899 became associated with Valdemar Poulsen in his development work on the telegraphone. He later aided in the same inventor's development of the system for continuous wave wireless telegraphy and telephony which has become known as the Poulsen Arc.

In 1909 he was appointed Assistant Professor in Telegraphy, Telephony and Radio at the Royal Technical College at Copenhagen, becoming Professor in 1912, which chair he is still holding. In 1922 he was appointed principal of this college.

Professor Pedersen has contributed a great number of papers on scientific matters in electrophysics and electrotechnics, mainly on experimental researches carried out by himself. He received his Ph.D. at the University of Copenhagen in 1929.

He is a member of the Royal Danish Society of Sciences, the Royal Academy of Natural Sciences (Lund, Sweden), the Academy of Engineering Sciences (Stockholm), and the Royal Society of Sciences (Trondhjem, Norway).

In 1911–1912 Professor Pedersen was Chairman of the Danish Physical Society. In 1916–1920 he was President of the Danish Society of Electricians, and in 1920 he became President of the State Control board for Licensed Telephone Companies of Denmark. From 1920 to 1923 he was President of the Society of Danish Civil Engineers.

He is keenly interested in standardization work and in 1926 was elected President of the Danish Section of the International Electrotechnical Commission, which office he still holds.

In 1907 he was awarded the Gold Medal of the Royal Danish Society of Sciences of Copenhagen for a paper on an experimental investigation of the oscillations in liquid jets. He was awarded the H. C. Oersted Medal in 1927.

Professor Pedersen is a Fellow of the American Institute of Electrical Engineers and a Member of the Institution of Electrical Engineers (Great Britain). He became a Fellow of the Institute of Radio Engineers in 1915.

INSTITUTE NEWS AND RADIO NOTES

November Meeting of the Board of Direction

The November meeting of the Board of Direction was held at 4:15 P.M., Wednesday, November 5, 1930 at the office of the Institute, 33 West 39th Street, New York City. The following members were present: R. H. Marriott, (acting chairman); Melville Eastham, treasurer; R. A. Heising, J. V. L. Hogan, L. M. Hull, R. H. Manson, A. F. Van Dyck, and H. P. Westman, secretary.

A petition nominating the following officers for 1931 was presented to the Board and accepted.

For President—Captain S. C. Hooper, Director of Naval Communications, U. S. Navy, Washington, D. C. Fellow of the Institute.

For Manager – H. E. Hallborg, Engineer, Radio Corporation of America, New York City. Member of the Committee on Constitution and Laws. Fellow of the Institute.

For Manager—H. W. Houck, Chief Engineer, Dubilier Corporation, New York City. Member of the Institute.

Proceedings Binders

During the last year two sizes of binders for the Proceedings have been made available. The smaller size was suitable for containing annual sets of the Proceedings issued before 1928. The yearly copies for 1928 and later years were too bulky for inclusion in the smaller size and the larger size binder was made available to accommodate them.

At the present time there is such small demand for the smaller size binder that it is now being discontinued and all future orders will be assumed to call for the larger size.

Incorrect Addresses

On pages XLIX, L, and LI of the advertising section of this issue will be found the names of two hundred and two members of the Institute whose correct addresses are not known. It will be appreciated if anyone having information concerning the present addresses of any of the persons listed will communicate with the Secretary of the Institute.

Associate Application Form

For the benefit of members who desire to have available each month an application form for Associate membership, there is printed in the PROCEEDINGS a condensed Associate form. In this issue this application will be found on page XXXIII of the advertising section.

Application forms for the Member or Fellow grades may be obtained upon application to the Institute office.

The Committee on Membership asks that members of the Institute bring the aims and activities of the Institute to the attention of desirable and eligible nonmembers. The condensed form in the advertising section of the PROCEEDINGS each month may be helpful.

Radio Signal Transmissions of Standard Frequency

The following is a schedule of radio signals of standard frequencies for use by the public in calibrating frequency standards and transmitting and receiving apparatus as transmitted from station WWV of the Bureau of Standards, Washington, D. C.

Further information regarding these schedules and how to utilize the transmissions can be found on pages 10 and 11 of the January, 1930, issue of the PROCEEDINGS, and in the Bureau of Standards Letter Circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C.

Eastern Standard Time	Dec. 22
10:00 P.M.	550
10:12	600
10:24	700
10:36	800
10:48	1000
11:00	1200
11:12	1400
11:24	1500

Committee Meetings

Committee on Admissions

A meeting of the Committee on Admissions was held at 9:30 A.M. on November 5th at the office of the Institute, the following members being present: R. A. Heising, chairman; A. V. Loughren, R. H. Marriott, E. R. Shute, A. F. Van Dyck, and H. P. Westman, secretary. One of three applications for admission to the grade of Member was approved and five applications of eight for transfer to the grade of Member were approved. An application for transfer to the Fellow grade was also approved.

Committee on Broadcasting

The Committee on Broadcasting held a meeting at 1:30 P.M. November 5, 1930, in the office of the Institute, the following being in attendance: L. M. Hull, chairman; B. Ray Cummings (represented), J. V. L. Hogan, C. W. Horn, P. A. Greene, R. Guy, and R. H. Marriott. The Committee drafted an answer to a letter from the Federal Radio Commission asking for comments upon a Proposed General Order on the Limitation of Harmonics Emitted by Radio Stations. The Board of Direction later in the day granted approval of this proposed answer.

Committee on Membership

At 5:30 P.M. in the office of the Institute on Wednesday, November 5, a meeting of the Committe on Membership was held. The meeting was attended by I. S. Coggeshall, chairman; H. C. Gawler, and S. R. Montcalm.

Committee on Standardization

A meeting of the Institute's Committee on Standardization to consider the reports of the various technical committees for adoption as I.R.E. standards was held October 31 and November 1, 1930. During this two-day meeting a large number of reports of technical committees were discussed and approved for adoption as I.R.E. standards and will be published in the forthcoming 1931 Year Book of the Institute. The following were present on October 31st: J. H. Dellinger, chairman; Stuart Ballantine (Chairman of the Technical Committee on Vacuum Tubes), L. G. Bostwick (alternate of H. A. Frederick), R. D. Brown, (representing W. E. Holland), W. A. R. Brown (representing Robert Morris), C. M. Burrill, E. T. Dickey (Chairman, Technical Committee on Radio Receivers), P. H. Evans, H. A. Frederick (Chairman of the Technical Committee on Electro-Acoustic Devices), V. M. Graham, H. E. Hallborg, C. M. Jansky, Jr., O. T. Laube (representing A. F. Rose), W. H. Murphy, E. L. Nelson, N. C. Olmstead, (representing L. E. Whittemore), H. F. Olson, Haraden Pratt, H. M. Turner, J. C. Warner (representing W. C. White) L. E. Whittemore, Adney Wyeth (representing L. G. Pacent), H. P. Westman, secretary, and B. Dudley, assistant secretary.

The following were present November 1st: J. H. Dellinger, chairman; J. B. Blanchard (representing W. Wilson), L. G. Bostwick (representing H. A. Frederick), R. D. Brown (representing W. E. Holland), W. A. R. Brown (representing Robert Morris), E. T. Dickey, V. M. Graham, C. M. Jansky, Jr., O. T. Laube (representing A. F. Rose), W. H. Murphy, E. L. Nelson, Haraden Pratt, H. M. Turner, L. E. Whittemore, W. Wilson, Donald Whiting, Adney Wyeth (representing L. G. Pacent), H. P. Westman, secretary, and B. Dudley, assistant secretary.

As a result of these two sessions, the Committee on Standardization is now in a position to forward its report to the Board of Direction for final approval.

Technical Committee on Radio Transmitters and Antennas—I.R.E.

A meeting of the Technical Committee on Radio Transmitters and Antennas was held at 9:30 A.M. on Tuesday, October 21, at the office of the Institute. Haraden Pratt, chairman; J. B. Blanchard (nonmember), W. E. Downey, H. E. Hallborg, D. G. Little, W. F. Diehl, William Wilson, and B. Dudley, secretary, were present.

Technical Committee on Radio Receivers—I.R.E.

A meeting of the above Technical Committee was held in the office of the Institute on Thursday, November 6, at 10 A.M., the following being in attendance: E. T. Dickey, chairman; C. M. Burrill, Malcolm Ferris, E. J. T. Moore (representing V. M. Graham), and B. Dudley.

Subcommittee on Miscellaneous Tests of the Technical Committee on Radio Receivers—I.R.E.

A meeting of the above subcommittee was held at 10 A.M. on Tuesday, October 14, 1930, at the office of the Institute, attended by F. X. Rettenmeyer, chairman; W. Wilson, and B. Dudley, secretary.

Subcommittee on High-Frequency Receivers of the Technical Committee on Radio Receivers—I.R.E.

This subcommittee met at 10 A.M. October 16, 1930, in the office of the Institute, the following being present: C. M. Burrill, chairman; T. McL. Davis, H. M. Lewis, H. O. Peterson (representing H. H. Beverage), and B. Dudley, secretary.

Subcommittee on Aircraft Receivers of the Technical Committee on Radio Receivers—I.R.E.

The following members were present at a meeting of the above subcommittee which was held at 10 A.M. on Thursday, October 23, 1930, at the office of the Institute: E. J. T. Moore, acting chairman (representing V. M. Graham); H. Diamond, B. Trevor (representing H. O. Peterson), L. A. Hyland (representing T. McL. Davis), H. B. Fischer (representing S. E. Anderson), W. H. Murphy, R. M. Wilmotte, and B. Dudley.

Subcommittee on Nomenclature of the Technical Committee on Radio Transmitters and Parts—A.S.A.

At 10 A.M. on Wednesday, October 8, a meeting of the above subcommittee was held in the office of the Institute and was attended by R. M. Wilmotte, chairman; W. Wilson, and B. Dudley.

Institute Meetings

New York Meeting

Two papers were presented at the New York meeting of the Institute held in the Engineering Societies Building at 7:30 P.M., November 5, 1930.

The first paper, which was presented by Professor W. G. Cady was entitled "Rochelle Salt Crystals in High-Frequency Circuits." It is summarized as follows:

"Although the remarkable electrical properties of Rochelle salt have been known since 1894, it is only in the last thirteen years that any serious attempts have been made to turn them to practical account. In the present paper the chief physical characteristics of Rochelle salt are discussed, including its mechanical, elastic, and electrical properties. The technique of preparing Rochelle salt plates is described, and the limitations of the substance from the point of view of practical applications are pointed out. Numerical data on the elastic and electrical constants are presented, based largely on the work of Mandell, and on the writer's own experiments on high-frequency vibrations of Rochelle salt plates cut in various orientations. A consideration of the theory of the effect of the piezo-electric constants upon the dielectric constant leads to an explanation of the enormously large values of the dielectric constant which some investigators have found. The effects of temperature upon the electric constants and upon the frequency of vibration are discussed. Experiments were carried out on the various modes of vibration of Rochelle salt plates, and also on their behavior as high-frequency resonators, stabilizers, and oscillators."

The second paper of the evening by Professor K. S. Van Dyke was on "The Measurement of the Decrement of Piezo-Electric Resonators." A summary of this paper is also given.

"The dying away of the vibrations of a piezo-electric resonator is watched with a cathode-ray oscillograph. The crystal is either driven by a separately tuned oscillator or forms part of a crystal-controlled oscillator circuit, and is switched from the oscillator to the oscillograph, where the e.m.f. generated by the crystal is watched or photographed through all or part of its period of decay. This interval is less than one-tenth of a second in duration for a 1000-kc quartz resonator. Both the Western Electric low voltage oscillograph and the General Electric cathode-ray oscillograph are used, the former necessitating a low impedance load on the crystal during its decay, and thus yielding decrements greater than characteristic of the crystal on open circuits, while the latter requires for most crystals an amplifier system to step up the crystal e.m.f.'s before application to the deflection plates of the oscillograph.

"Values of decrement for quartz resonators obtained to date indicate that the values of resistance quoted four years ago, when presenting the network equivalent of the crystal resonator, should be decreased by a factor between 5 and 10. This brings them into agreement with observations by Terry on the conditions for oscillations of crystal-controlled vacuum tube circuits and with recent measurements of decrement by Vigoreaux."

The meeting was attended by three hundred members and guests.

Atlanta Section

The October 3rd meeting of the Atlanta Section was held at the Cecil Hotel in Atlanta, Chairman Harry F. Dobbs, presiding.

A paper by J. T. Gardberg on "Volume Control used in Modern Radio Receiving Sets" brought forth a lively discussion which was participated in by the majority of the members present.

At the conclusion of this paper, the members proceeded to the office of the U. S. Supervisor of Radio in Atlanta and were shown the practical operation of the latest type of frequency measuring equipment used by the supervisors. After this demonstration a number of questions were asked of George Llewellyn of the supervisor's office.

The meeting was attended by twenty members and guests.

Buffalo-Niagara Section

A meeting of the Buffalo-Niagara Section was held on October 14th at the University of Buffalo, A. B. Chamberlain, chairman, presiding.

A paper on "Sound Reproduction" was presented by George Stringfellow, Division Superintendent of Service of Electrical Research Products, Inc. Mr. Stringfellow described the method used in the motion picture studios to record on both film and wax. Some of the difficulties encountered in the commercial application of these devices by the motion picture industry since 1926 were explained. The difficulties due to interference caused by arc lights and other equipment and the developments to obviate these were discussed. The methods and equipment used in both recording and reproduction were considered in detail.

Messrs. Chamberlain, Hector, Johnson, Waud and others participated in the discussion which followed the presentation of the paper.

Forty-nine members and guests were in attendance.

Cincinnati Section

The October meeting of the Cincinnati Section was held on the 14th at the Cincinnati Chamber of Commerce, R. H. Langley, chairman, presiding.

A paper on "Electrolytic Capacitors; Theory and Application" was presented by F. E. Johnston, Product Engineer of the Crosley Radio Corporation.

The speaker briefly outlined the history of the electrolytic condenser. The manufacturing process was then described, special emphasis being placed on the necessity for pure materials and the care required in the forming process. The theory of operation of electrolytic condensers was touched on but it was stated that there is a wide variance in opinion on the subject among investigators in this field and much remains to be learned. The actual operating characteristics under a wide variety of service conditions were discussed in some detail. The useful life was stated to be indefinite provided the maximum peak voltage and operating temperature ratings are not exceeded. The methods of measuring the capacity of this type of condenser were given.

The paper was discussed by Messrs. Austin, Barton, Hoffman, Israel, Kilgour, Osterbrock, and Wilson of the forty-three members and guests in attendance.

A Nominating Committee composed of Messrs. Desh, Dixon, and

Kilgour was appointed to bring in nominations for election of officers for 1931. This election will take place at the annual meeting to be held December 16th.

Cleveland Section

A meeting of the Cleveland Section was held on September 19th in the Winton Hotel, D. Schregardus, chairman, presiding.

John F. Royal, Managing Director of WTAM, gave an outline of the relations of broadcasting and the vaudeville stage.

The attendance at the meeting was forty-five.

The October meeting of the Cleveland Section was held in the Physics Laboratory of the Case School of Applied Science on the 24th, S. E. Leonard, presiding in place of Dr. Schregardus who was unable to be present.

T. B. Owens, Assistant Professor of Electrical Engineering of the Case School of Applied Science presented a paper on "Filters in Electrical Circuits."

The paper covered the four types of wave filters, giving their circuit arrangements, transmission characteristics, and other pertinent data.

Following the presentation of the paper, Bruce David, Chairman of the Committee on the New Constitution, read the new Constitution approved by the Board of Direction. This Constitution was adopted by the Section.

The attendance at the meeting totaled thirty-five.

Detroit Section

L. N. Holland presided at the October 17th meeting of the Detroit Section held in the Detroit News Auditorium.

W. E. Jackson, Radio Engineer, Airways Division of the Bureau of Lighthouses, presented a paper on "Development of a Visual Type of Radio Range Transmitter having Universal Application to the Airways."

This paper is published in this issue of the Proceedings and it will, therefore, be unnecessary to describe it further.

The discussion which followed the paper was entered into by Messrs. Case, Firestone, Holland, Worel and others.

A motion picture showing the production of radio receiving sets was also shown.

Los Angeles Section

A meeting of the Los Angeles Section was held at the Engineers Club on September 22nd, T. C. Bowles, chairman, presiding.

T. E. Nikirk, Chairman of the Meetings and Papers Committee of the Los Angeles Section, presented a paper on "Three- and Six-Phase Rectifier Systems."

At its conclusion the paper was discussed by Dr. deForest, who was present, and Messrs. Breeding, McDonough and several others of the twenty-five members and guests in attendance.

The October meeting of the Los Angeles Section was held at the Engineers Club on October 21, T. C. Bowles, chairman, presiding.

A paper on "Recent Developments in Chain Broadcast Control" was presented by Paul Johnson, a transmission engineer for the Southern California Telephone Company.

Following the delivery of the paper and the discussion which was particpated in by Messrs. Anderson, Fox, Nikirk and others, members and their guests visited the broadcast control section of the long-distance control office of the Telephone Company.

Dr. Lee deForest introduced John Stone Stone, President of the Institute for 1915, to the one hundred and twenty-six members and guests in attendance.

Philadelphia Section

The September meeting of the Philadelphia Section was held on the 24th of the month at the Franklin Institute, W. R. G. Baker, chairman, presiding.

Prior to introducing the speaker of the evening the Chairman introduced to the Section members the new officers and also gave an outline of the various committees appointed together with their duties and personnel.

The speaker of the evening, C. W. Horn, General Engineer, National Broadcasting Company was then introduced. Mr. Horn presented a paper on "The Problems of Chain Broadcasting." Although only partly technical in its treatment of the subject, the paper covered the system of chain broadcasting now employed and outlined many of the minute details that enter into the successful production of a nation-wide broadcast.

A general discussion participated in by many of the sixty-six members and guests in attendance followed.

Pittsburgh Section

The October meeting of the Pittsburgh Section was held on the 21st of the month at Utility Hall, A. J. Buzzard, chairman, presiding.

A paper on "Shaking Down Electrons" was presented in two parts by R. C. Hitchcock and Lee Sutherlin of the Westinghouse Electric and Manufacturing Company.

The first portion of the paper presented by Mr. Hitchcock displayed and explained the operation of a number of types of photo-electric cells. The wide and varied field of usefulness which these cells serve was discussed and a practical demonstration of a cell as an "off" and "on" device was given.

The second part of the paper was presented by Mr. Sutherlin and covered systems for amplifying the output of these devices. Diagrams of suitable circuits and a special amplifying tube developed for use in this work were exhibited and discussed in detail.

After the presentation of both portions of the paper, a discussion participated in by Messrs. Armstrong, Best, Carman, Johnstone, Mag, McKinley, and Strainger took place.

The attendance at the meeting totaled forty-one members and guests.

Rochester Section

The October 2nd meeting of the Rochester Section was held at the Sagamore Hotel, H. J. Klumb, presiding.

A paper on "Recent Developments in Oscillographs" was presented by Clare Anderson of the Westinghouse Eectric and Manufacturing Company of Newark, N. J.

In his illustrated talk, Mr. Anderson discussed the constructional features of the new multielement oscillographs and the Osiso, a portable oscillograph. The older types of oscillographs were first discussed, particular emphasis being placed on the optical system consisting of a light source, the rays from which passed through adjustable slits and thence to the vibrating galvanometer and photographic plate or optical viewing window, the spot of light striking the photographic strip being the properly focused image of this adjustable slit. The speaker then described methods recently developed to overcome some of the objections to the older system. This included the use of a strip filament lamp for a light source, eliminating the use of the adjustable slits. In addition, self-energizing attachments which could be started to take a picture of line transients in less than two cycles and new types of vibrating elements were covered. The paper terminated with a description of some of the recent applications of oscillographs to radio and nonradio problems.

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

San Francisco Section

The October 15th meeting of the San Francisco Section was held at the Engineers' Club, C. H. Suydam, presiding.

A paper on "Portable Field Equipment" was presented by Ralph Heintz of Heintz & Kaufman, Ltd.

Various types of portable radio field equipment developed for the Army and Navy were described and several pieces of apparatus were on display. Power supply devices utilizing both foot power and a small single cylinder gasoline engine were also shown.

The attendance at the meeting was forty-nine.

Seattle Section

The October 10th meeting of the Seattle Section was held at Philosophy Hall in the University of Washington. Austin V. Eastman, chairman, introduced the speaker of the evening, J. R. Tolmie, who delivered a paper on the "Use of Quarter Wavelength Transmission Lines as Antenna Coupling Elements."

The paper covered the action of parallel transmission lines and their characteristics. The relative impedences of the antenna, the coupling elements, and the generator were considered in detail.

Messrs. Eastman, Foster, Williams, and Willson of the thirty-seven members and guests in attendance discussed the paper.

Washington Section

The October meeting of the Washington Section was held on the 9th in the Continental Hotel, L. P. Wheeler, chairman, presiding.

The paper on "Recent Developments in Radio-Acoustic Position Finding" was presented by Dr. Herbert G. Dorsey, Senior Electrical Engineer, Coast and Geodetic Survey.

The use of hydrophones in surveying had already been well established on the Pacific Coast prior to 1926. During that year experiments were started on the Atlantic Coast but were failures and continued as such until the fall of 1929. At that date a change in method was tried which consisted in using floating hydrophone stations instead of hydrophones near the shore. It was found that when the hydrophone was used in water of 20 or more fathoms depth, bombs could be heard at distances up to 70 miles. During the early spring of 1930 a complete trial was made in waters on the east coast of Florida and during the past summer was put into successful operation on Georges Bank where over 22,000 soundings were made with the author's fathometer in an area of over 6000 square miles and the surveying ship's position located by a radio-acoustic distance and radio compass bearing from an anchored ship. The position of the latter was located astronomically and a system of buoys used as references in a triangulation system.

A full description was given of the author's system of transmitting a series of radio dashes automatically by a thyratron from a crystal-controlled transmitter of 4135 or 8270 kc when a bomb signal was received on an electromagnetic receiver and amplifier of new design, and lantern slides showed diagrams and photographs of the circuits and apparatus.

During the summer a new submarine valley was discovered which will be of considerable importance to transatlantic navigation.

Part II
TECHNICAL PAPERS

Proceedings of the Institute of Radio Engineers
Volume 18, Number 12 *December, 1930*

SOLAR AND MAGNETIC ACTIVITY AND RADIO TRANSMISSION*

By

L. W. Austin, E. B. Judson, I. J. Wymore-Shiel

(Laboratory for Special Radio Transmission Research, Bureau of Standards, Washington, D. C.)

Summary—*In this paper curves are shown indicating a connection between the annual averages of sun spot numbers and daylight radio signal strength of Nauen, Germany, as received in Washington from 1915 to 1929. The curves of the monthly averages of sun spots and daylight transatlantic signals for the years 1924–1929 show little evidence of correlation. On the other hand, the correlation between these monthly average signals and terrestrial magnetic activity seems definite. Reception from Monte Grande (Argentina) shows less definite and on the whole inverse correlation with magnetic activity. This indicates that long waves, like the ultra short, are more influenced by magnetic activity when traveling across the earth's magnetic field than when traveling parallel to it.*

IN SOME earlier papers[1] dealing with solar and radio relations, it was shown that the yearly averages of sun spots and long-wave daylight transatlantic signal strength were apparently definitely connected, and at the same time the possibilities of shorter period correlation were also suggested. The signal observations which have been taken since the publication of these papers make it possible to extend the curves published at that time and to draw somewhat more definite conclusions as to the reality of these relations.

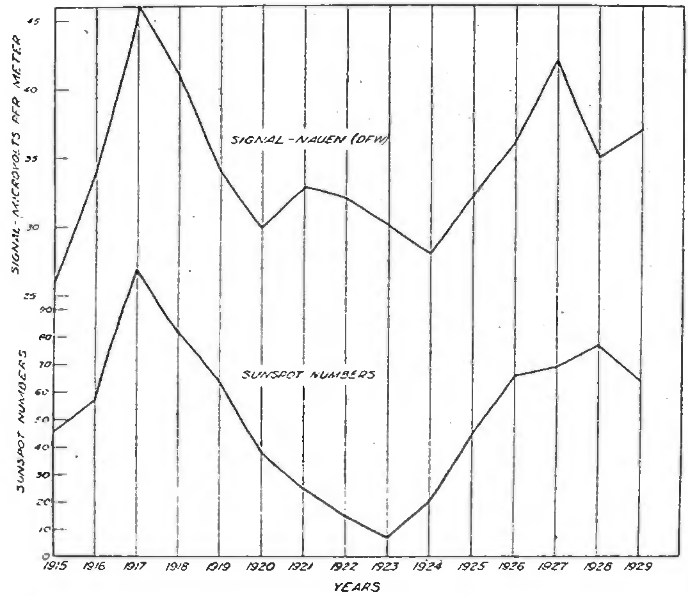
Fig. 1 shows the curves of annual averages of sun spots and the Nauen signals already mentioned, continued to include 1929. This covers approximately one and one-half sun spot cycles. Although the irregularities of the peaks of the two curves of the present maximum do not exactly correspond, and the signal measurements before 1922 are not of high accuracy, the evidences of correlation will be very strong if no marked discrepancies occur before the next sun spot minimum.

At the time of writing the earlier papers, it was thought from the observations that a fair degree of correlation might also exist between the monthly averages of sun spots and signals. The rapid rise in signal strength which followed closely on the sudden increase of sun spots during 1925 was very striking.

Fig. 2 shows the monthly average curves from 1924–1929 of sun

* Decimal classification: R113.5. Original manuscript received by the Institute, July 28, 1930. Publication approved by the Director of the Bureau of Standards of the U. S. Department of Commerce.

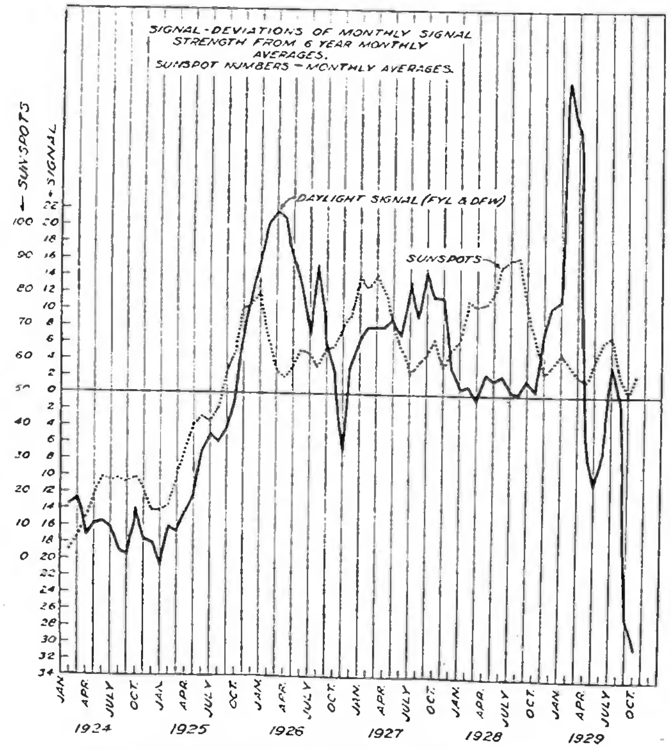
[1] Proc. I. R. E., 15, 825; October, 1927; 16, 166; February, 1928.



Annual average daylight signal intensity of Nauen (DFW) and sunspot numbers.

Fig. 1

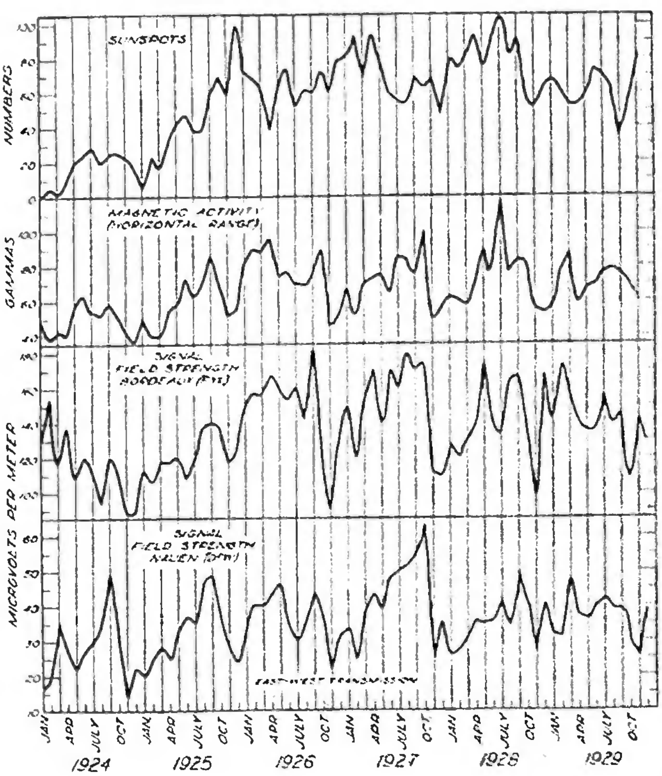
spots and of long-wave daylight transatlantic signals from Bordeaux, France (FYL) ($f=15.9$ kc, $\lambda=18{,}900$ m) and Nauen, Germany (DFW) ($f=23.4$ kc, $\lambda=12{,}800$ m).



Signals and sunspots. Three months running averages.

Fig. 2

In order to eliminate seasonal effects from the signal curve, the deviations of the individual monthly values from the average values of the corresponding months of the whole six years are plotted, while in the solar curve the sun spot numbers are used directly. Both curves are smoothed by three-month running averages. It is seen at once that there is no simple relationship between the two curves as a whole. While the number of peaks is the same in the two, the periods of high signals in the later years have lagged about six months behind the sun spot peaks, so that the peaks of one curve correspond roughly to



Curves of sunspot numbers, magnetic activity, and radio signal strength.

Fig. 3

the troughs of the other. At present it is impossible to say whether these relationships are significant or not.

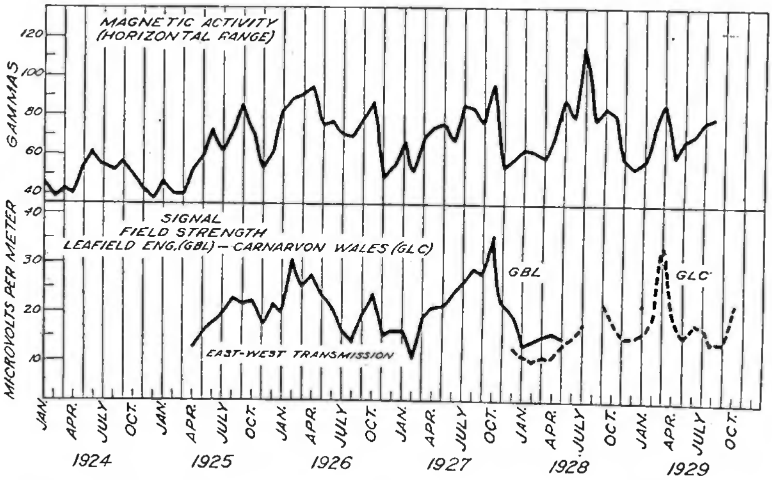
Fig. 3 which was published recently as a note in the Journal of the Washington Academy[2] shows the curves of the monthly averages of sun spots, magnetic activity as expressed in horizontal range[3] measured at Cheltenham, Md., and signals from Bordeaux (FYL) and Nauen (DFW). In these signal curves the actual monthly averages are

[2] *Jour. Washington Academy,* 20, 73, 1930.

[3] Horizontal range is defined as the difference between the maximum and minimum daily values of horizontal intensity and expressed in gammas. One gamma is equal to 1.10^{-5} gauss.

shown, not the deviations of monthly averages from the average monthly means for the six years. The use of the deviations in our earlier work prevented our noticing the close resemblance between the signal and magnetic activity curves.

The resemblance of the sun spot curve to the other three curves is not close, but the similarity in the changes in magnetic activity and signal strength seems to be unmistakable. The deep drop of both the magnetic and signal curves in November (more rarely in December) is especially striking. This early winter drop in signals has often been noticed, and in the case of transmission between Europe and America has been ascribed to the proximity of the signal path to the area of Arctic darkness at this season, or to a European sunset effect. These



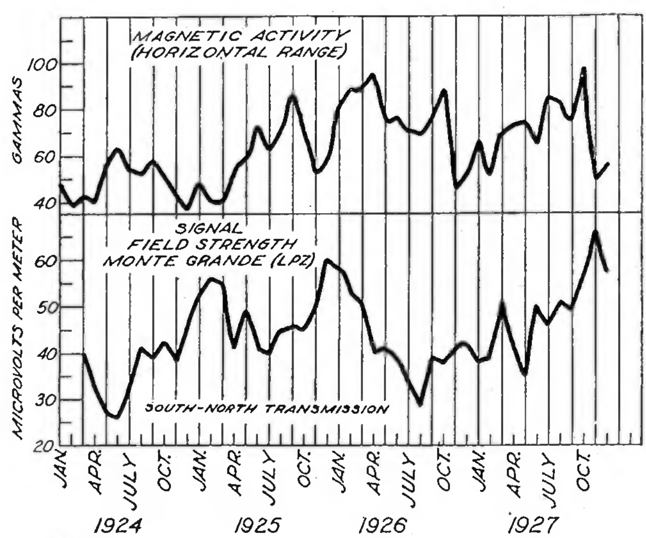
Curves of magnetic activity and radio signal strength.

Fig. 4

curves suggest that the connection between terrestrial magnetism and transmission may be more marked at the time of greater sun spot activity; at any rate, the correlation is poorer in 1924 than in the later years, and it appears to be decreasing in 1929 as the sun spots again begin to decrease.

Fig. 4 shows the magnetic-radio relationship in the case of reception from two English stations, Leafield (GBL) ($f = 24.4$ kc, $\lambda = 12{,}300$ m) and Carnarvon (GLC) ($f = 31.6$ kc, $\lambda = 9500$ m). Leafield in 1928 ceased to transmit regularly at an hour ensuring an all-daylight signal path across the Atlantic, and was therefore replaced for our purposes by Carnarvon, a station of approximately the same signal strength. In Fig. 4, the curves of magnetic activity and signal strength are not as strikingly alike to the eye as the curves of Fig. 3, but the percentage of months in which the two curves rise and fall together is slightly higher.

Thus far the radio signal paths which have been considered lie roughly in an east-west direction. In Fig. 5 the corresponding relations are shown for reception from Monte Grande in the Argentine (LPZ) ($f = 23.6$ kc, $\lambda = 12{,}700$ m), which lies nearly south of Washington. Here there seems to be an inverse rather than a direct correlation between the magnetic and signal curves. It is well known that in short-wave transmission there is a difference in the effect of magnetic activity on the north-south and east-west signal paths. According to T. L. Eckersley,[4] the number of occasions in the year October, 1927, to October, 1928, in which magnetic storms rendered the short-wave communication between England and Montreal impossible, were 49; between England and New York, 32; while between England and



Curves of magnetic activity and radio signal strength.

Fig. 5

points to the south (in Australia, South Africa, India, and South America), these varied from 4 to 7. These and other observations indicate that the connection between magnetic activity and radio transmission is most marked across the lines of the earth's magnetic field, and least marked parallel to them; and also that the effect is most noticeable in the east-west transmission nearer the poles, and less noticeable nearer the equator.

Attempts have also been made to show a relation between magnetic activity and signal strength for averages of shorter periods than a month. For this purpose the magnetic and signal observations used in Fig. 3 have been averaged in five- and ten-day periods. The resulting curves are not shown in this paper. The five-day averages

[4] T. L. Eckersly, *Jour. I.E.E.* (London), 992, 1929.

showed no definite magnetic-radio relationship. This poor correlation was perhaps to be expected, as Mrs. Wymore-Shiel[5] has shown that in the case of severe magnetic storms, at least, the transatlantic signals rise on an average on the second day after the most disturbed magnetic day. The ten-day averages show considerably better correlation than the five-day averages, the agreement sometimes being excellent for many weeks together and then failing completely over long periods of time.

[5] I. J. Wymore, "The relation of radio propagation to disturbances in terrestrial magnetism," Proc. I. R. E., 17, 1206; July, 1929.

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TEMPERATURE CONTROL FOR FREQUENCY STANDARDS*

By
J. K. Clapp
(General Radio Co., Cambridge, Mass.)

Summary—*A brief summary of the factors influencing the stability of temperature-control assemblies, in which control is obtained by adding heat and without the use of circulating mechanisms, is given. These factors include the degree of insulation; rate of application and method of distribution of heat; sensitivity, regularity of operation, and position of thermostat; degree of "ripple" attenuation; and the operating temperature. Examples of three types of control units regulating to within approximately ±0.5 deg., ±0.1 deg., and ±0.01 deg. C, respectively, at 50 deg. C are given, with heating rates and details of construction. Diagrams and photographs are included.*

In maintaining the frequency of a piezo-electric oscillator at as nearly a constant value as possible, it is necessary to control its temperature very carefully, in addition to controlling other important factors. This paper summarizes the problems of temperature control with regard to the final variations in temperature obtained in a given assembly. Some practical aspects of construction are considered with the object of obtaining the desired result with a simple assembly providing for reasonable accessibility.

The main factors which substantially influence the constancy of the temperature within the desired space are:

(1) Degree of insulation of entire assembly to external temperature changes
(2) Rate of application of heat
(3) Distribution of heat
(4) Sensitivity and regularity of operation of thermostats
(5) Position of thermostat
(6) Degree of "ripple" attenuation
(7) Operating temperature

These factors will each be given a brief discussion. The influence of these factors on the design of units for specific problems will then be illustrated through brief descriptions of certain types of temperature-control assemblies.

(1) The better the heat insulation of the entire assembly from the atmosphere (and, in general, the higher the operating temperature) the easier it is to regulate the inner temperature to a desired degree of

* Decimal classification: R214. Original manuscript received by the Institute, August 6, 1930.

constancy. The better the insulation, the smaller is the heat-flow through the walls. A given temperature within the unit is then maintained through the application of a smaller amount of heat than is the case when the heat loss is large. This condition is favorable for the regular operation of most forms of thermostats. It must be borne in mind that a *thermal balance exists and that under stable conditions, the average amount of heat applied is just equal to the average amount of heat lost through the walls of the assembly.* Balsa-wood was found to be most effective as the insulation material.

(2) If heat is applied too rapidly there is a definite tendency to produce "hot spots" near the heaters. If the thermostat is placed too far from the heaters, the rapid application of heat results in parts of the assembly being carried to temperatures which are far too high. When, finally, the heat "surge" reaches the thermostat, that device is also raised far above the temperature required for its operation. The result is that the temperature of the unit "overshoots" the thermostat operating temperature, resulting in poor regulation (even with sensitive and reliable thermostats), since the average temperature is controlled, not by the thermostat, but by the rate and amount of "overshooting" and by the rate of heat loss from the unit.

(3) The distribution of heat as applied to the unit is very important. Localization of heat application can only be successful in assemblies in which a rapid circulation of a fluid (air or liquid) is maintained. (Because the use of liquid baths is very inconvenient for general electrical laboratory work, and as the use of any rotating machinery is undesirable in units which are to operate continuously over long periods with minimum attention, emphasis has here been placed on units involving neither of these elements.)

If circulation of the air within the unit is to be avoided, it is necessary to distribute the heat-flow symmetrically around the space to be controlled. The ideal form would be a spherical space covered by a uniformly distributed heating surface. A close approach to this ideal is a cylinder, covered with a uniformly distributed heater around its circumference, with properly graded heating on the ends to simulate the conditions of an "infinite cylinder."[1]

For reasons of mechanical simplicity in manufacturing and for ease of access, a rectilinear form is preferable. In the units here described, heaters are mounted on all six faces of the distribution box

[1] W. A. Marrison, "A high precision standard of frequency," Proc. I.R.E., 17, 1103, July, 1929; *Bell Sys. Tech. Jour.*, 8, 493; July, 1929.
V. E. Heaton and W. H. Brattain, "Design of a portable temperature-controlled piezo oscillator," *Bureau of Standards Journal of Research*, 4, 345; March 1930; Proc. I.R.E., 18, 1239; July, 1930.

and are in the form of open grids,[2] covering as large a portion of the wall surfaces as possible, consistent with standardized construction of walls and heater elements.

(4) The greater the sensitivity of the thermostat, (i.e., the smaller the temperature difference required to open and close the operating contacts) the better the possible degree of control, other factors being equal. *The constancy of the operating temperature of the thermostat is just as important as high sensitivity.*

Examples of both bimetallic and mercury types which were rendered practically useless by extraneous variations in average operating temperature were encountered in the development work as follows: In the bimetallic type the friction against which the bimetallic element operated varied both with the adjusted operating temperature and with the degree of mechanical vibration to which the unit was subjected. In a mercury-type thermostat of experimental design having a thin-walled bulb of greatly flattened form, (for the rapid absorbtion of heat) the operating temperature varied markedly with air pressure.

(5) The position of the thermostat, in a given assembly, greatly influences the degree of control obtainable. It is at once apparent that *the thermostat should not be placed in the space to be controlled as to do so requires the temperature of the space to vary enough to cause the thermostat to operate.* The space to be controlled should be enclosed by the necessary distributing and attenuating walls, and the thermostat should be placed outside of these. In fact, the best position for the thermostat is *in intimate contact with the outer face* of the heat distribution wall, the distributed heaters being supported *outside* of the thermostat. Heat is transferred from the heater grids to the wall surface mainly by conduction; there is little convection if heat is not applied too rapidly.

Under these conditions the outer wall of the space is maintained very nearly at the average operating thermostat temperature (average of the "opening" and "closing" temperatures). The "ripples" caused by rise and fall of temperature necessary to operate the thermostat may be reduced to practically any desired extent[3] by suitable construction of the walls.

(6) If the heaters, thermostat, and heat distribution walls are properly arranged and operated, but little "ripple" attenuation is required to smooth out the short-period temperature fluctuations caused by the thermostat operation. In simple equipment where

[2] Manufactured by The States Company, Hartford, Conn., under the trade name of "Ohm-Spun" resistors.

[3] W. A. Marrison, "Thermostat design for frequency standards," Proc. I.R.E., 16, 976; July, 1928.

extreme constancy is not required, a distribution wall of aluminum (1/16 or 1/8 inch thick) will very largely eliminate the "ripples." To meet more rigorous requirements, several layers of distributing material (aluminum) and attenuating material (asbestos or felt) may be necessary. The distributing material should have a relatively low heat capacity and high conductivity; the attenuating material should have high heat capacity and low conductivity. The effectiveness of the layers is also dependent on the character of the surfaces; for example, polished aluminum is superior to coated.

(7) The ease with which a given degree of constancy of temperature may be maintained in a given type of assembly varies with the operating temperature. It is apparent that *if the temperature is to be maintained entirely by the addition of heat, the operating temperature must be substantially higher than the highest room temperature* under which control is to be maintained. As the heat lost from the unit depends on the difference between the operating and room temperatures, given variations in room temperature will produce less disturbance of the conditions internal to the unit when the temperature difference is large, i.e., when the operating temperature is high.

Unfortunately, high temperatures reduce the piezo-electric effects in quartz crystals, so that for control of these elements it is desirable to operate them at as low a temperature as can be maintained by the control assembly. General experience shows that an operating temperature of 50 deg. C (122 deg. F) is a fair compromise between these conflicting conditions, and this value is now quite widely accepted as standard for this work. In order that quartz crystals may be used interchangeably in various systems, it is necessary to choose a standard temperature by common agreement. For this reason, fine regulation of the crystal frequency should *not* be obtained by adjustment of the operating temperature. As there are other means for adjustment of the operating frequency, fixing the temperature entails no particular operating difficulties.

Examples of temperature-control units built upon the foregoing principles will now be given. In the diagrams, a cross section of one wall of the units is given, indicating the location of the thermostat. It is understood that the character of the other walls surrounding the controlled space is similar to that indicated in the sketches.

Type I—Simple Control Unit

The controlled space is surrounded by one-eighth inch aluminum walls over which, at a distance of one-quarter inch the heater cards are mounted. One card is supported somewhat farther away to permit the mercury thermostat to be mounted between the heater and the aluminum wall. The thermostat is in intimate contact with the aluminum and is held in place by an aluminum pocket secured to the wall. (See Fig. 1.) In units having controlled spaces of the order of 250 cubic inches, the instantaneous rate of heating required is 60 watts or somewhat less. The thermostat keeps the heat "on" about one-sixth to one-tenth of the time, so that the average rate of heating is from 6 to 10 watts. Such a unit will regulate to better than ±0.5 deg. C for room temperatures of 20 deg. C ± 11 deg. C.

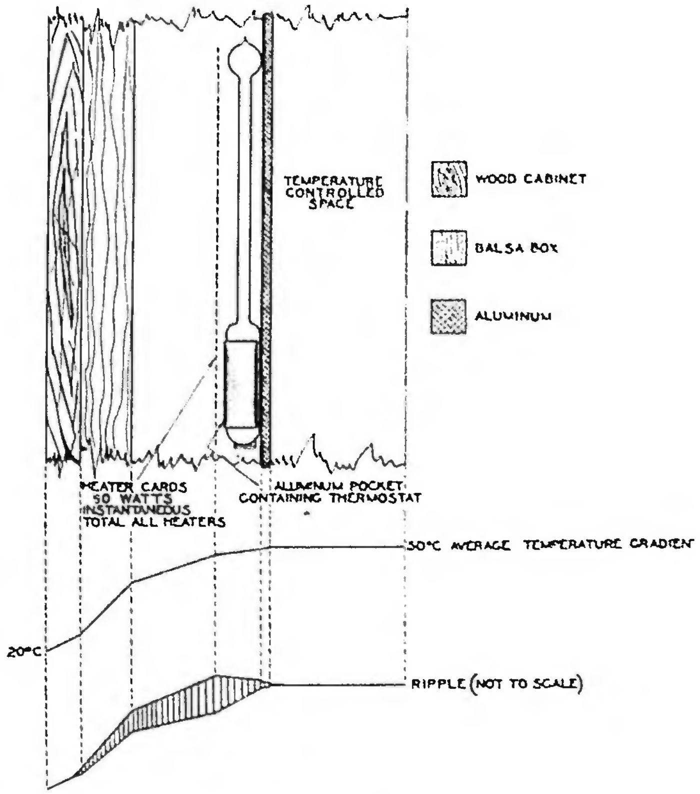

Fig. 1

Type II—Single-Stage Control Unit

This unit is very similar to the Type I described above, but includes a better attenuating and distributing system. (See Fig. 2.) Such units will regulate to about 0.1 deg. C under heating and temperature conditions given above.

Type III—Two-Stage Control Unit

Where great constancy of temperature is desired, it is in general simplest to obtain the desired result through the use of one temperature-control unit placed within another. The inner unit then has to operate

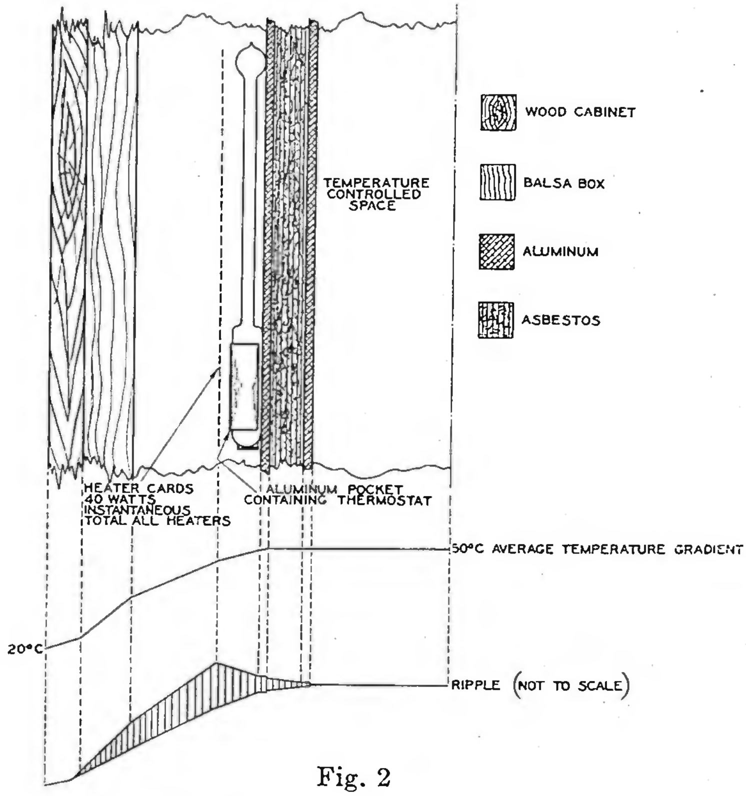
WOOD CABINET
BALSA BOX
ALUMINUM
ASBESTOS
TEMPERATURE CONTROLLED SPACE
HEATER CARDS 40 WATTS INSTANTANEOUS TOTAL ALL HEATERS
ALUMINUM POCKET CONTAINING THERMOSTAT
50°C AVERAGE TEMPERATURE GRADIENT
20°C
RIPPLE (NOT TO SCALE)

Fig. 2

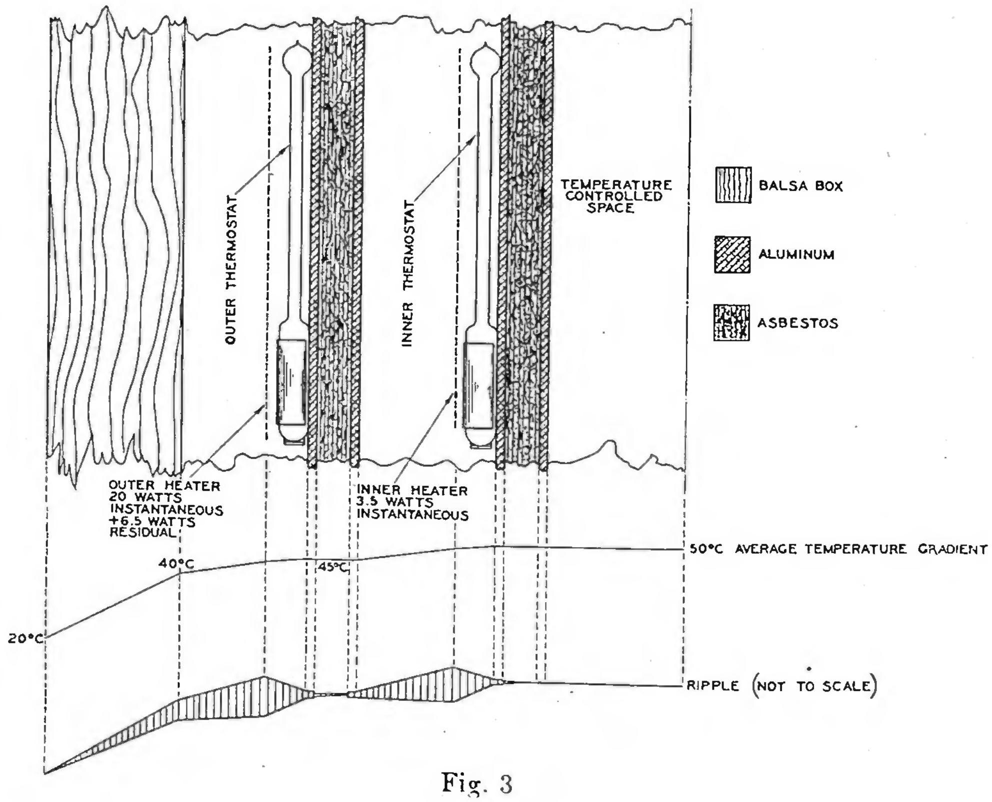
OUTER THERMOSTAT
INNER THERMOSTAT
TEMPERATURE CONTROLLED SPACE
BALSA BOX
ALUMINUM
ASBESTOS
OUTER HEATER 20 WATTS INSTANTANEOUS +6.5 WATTS RESIDUAL
INNER HEATER 3.5 WATTS INSTANTANEOUS
50°C AVERAGE TEMPERATURE GRADIENT
40°C
45°C
20°C
RIPPLE (NOT TO SCALE)

Fig. 3

only against the temperature fluctuations remaining from the operation of the outer unit. If the room-temperature fluctuations are reduced by one unit by a factor of 20, for example, so that a single-stage control would hold the temperature to within one degree for 20 degrees change in the temperature of the room, then a similar single-stage unit placed within the first would reduce the variations from one degree to something of the order of 1/20th degree. In frequency standards the quartz element may be placed in the inner unit, and the balance of the driving circuit in the outer to save space.

In the two-stage assembly indicated in Figs. 3 and 4 the inner and

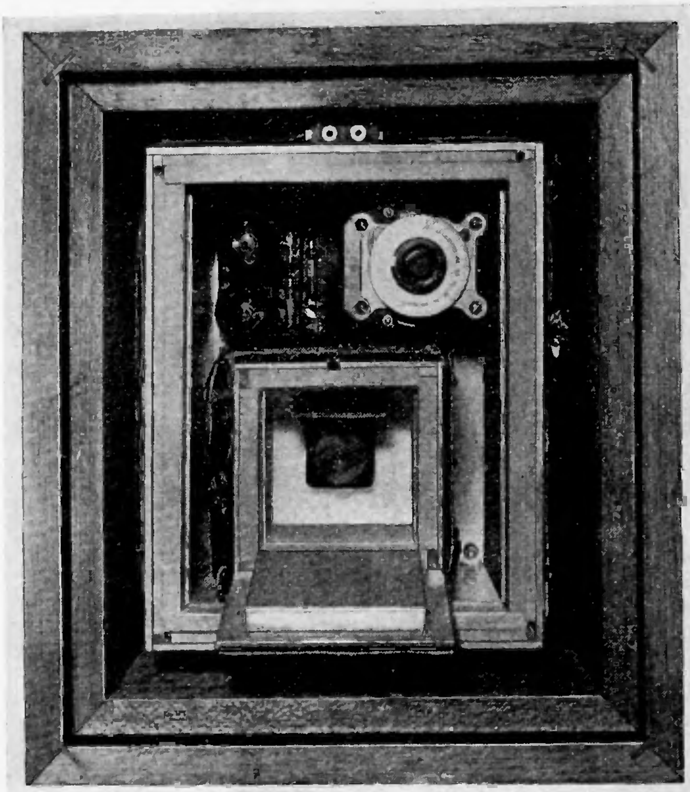

Fig. 4. Top view of complete crystal oscillator with double temperature control. Shows location of crystal holder in inner unit, with thermostat on right wall; oscillator tube, circuits, and isolating amplifier are shown in position in outer unit. Outer thermostat is mounted on left wall. The entire assembly is mounted in a balsa-wood box, with two-inch walls

outer units have instantaneous rates of heating of about 3.5 and 20 watts, the average rates being 0.3 and 3.0 respectively. A constant heating of 6 watts due to the filament heating and 0.5 watt due to the plate dissipation of the two vacuum tubes used is supplied to the outer unit. This residual heating alone is sufficient to hold the temperature of the entire assembly about 17 deg. C above a room temperature of 20 deg. C.

The arrangement of the unit is shown in the photograph, Fig. 4, the piezo-electric crystal being placed in the inner unit; the driving tube and circuits, as well as a coupling tube, are placed in the outer unit. The temperature of the crystal is maintained at 50 deg. ±0.01 deg. C while that of the circuit is maintained at 45 deg. ±0.1 deg. C.

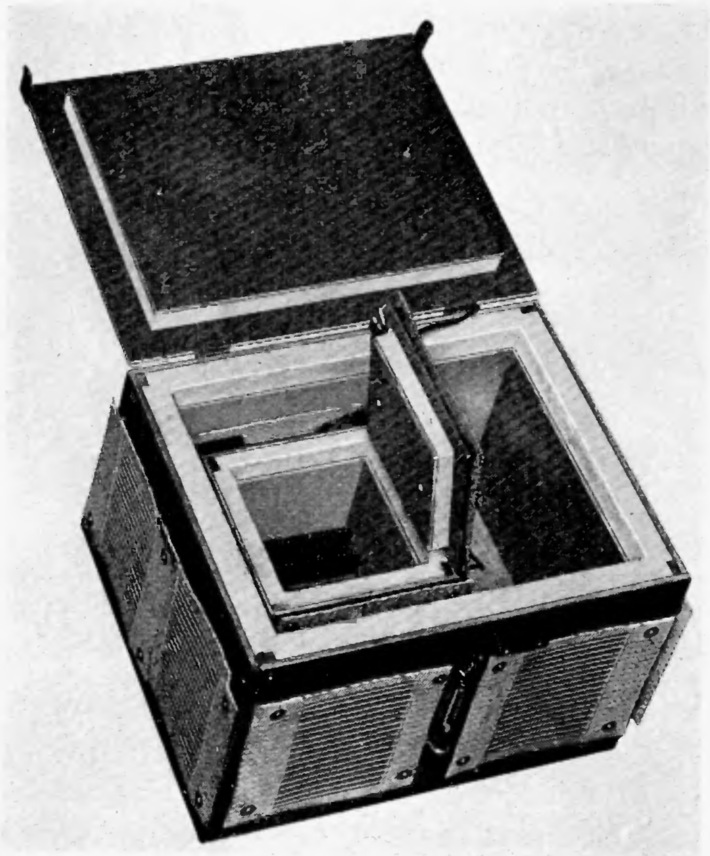

Fig. 5. The interior assembly of the double temperature-control box is here shown removed from the balsa-wood insulating container. The unit inside the larger box has heaters on all six faces, controlled by a thermostat mounted on the rear face. The outer unit also has heaters distributed over all six faces, the thermostat being mounted at the center of the front face.

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SOME EXPERIENCES WITH SHORT-WAVE WIRELESS TELEGRAPHY*

By

N. H. Edes

(Captain, Royal Signals, Catterick Camp, Yorkshire, England)

Summary—*An analysis is given of the factors that may be expected to determine the transmission phenomena peculiar to any given short-wave channel. It is considered that the wavelength that will give the greatest probability of satisfactory communication is determined by (a) the great circle distance between sender and receiver, (b) the geographical positions of the stations, (c) the time of year, and (d) the time of day. It is pointed out why the latitudes of the stations will be a pertinent factor.*

It is shown how in China in 1927–28 it became desirable to be able to predict the effects of the various factors, and in particular to predict the "best wavelength."

The development of the British military short-wave system in that country and the inauguration of a series of tests are described. The method employed in the tests is critically examined.

A summary is given of the working experiences and the results of the tests over the various "links." The diurnal and seasonal variations of best wavelength are recorded, and it is shown that the seasonal variations coincide approximately with the equinoxes.

The results of traffic-handling experience and tests are combined into a "range—best wavelength diagram" for the belt of latitude lying between 30 deg. and 40 deg. North, and the question as to how closely the diagram might apply to other parts of the world is discussed. It is anticipated that the curves for daylight would apply closely to any part of the world, but that those for darkness would need considerable modification for different latitudes.

I. Scope of the Article

THE writer's aim is to present some data concerning the performance of short-wave wireless telegraphy which may be of direct use to wireless engineers, and which may indirectly help to coördinate and confirm the results of previous physical research.

The data were obtained in China between August, 1927, and December, 1928, as a result of the working of traffic and organized tests between the British military wireless stations in that country. The stations had been improvised for the maintenance of communication between the British garrisons guarding "foreign"[1] lives and property during the civil war then raging.

The writer justifies himself for bringing the results to notice on the grounds that, in China, there became available an extensive "labora-

* Decimal classification: R113. Original manuscript received by the Institute, August 4, 1930.

[1] "Foreign," i.e., non-Chinese.

tory" stretching over 1200 miles and 17 deg. of latitude.[2] "Research workers" were not lacking, and willing ones at that.

As far as military exigencies would permit, advantage was taken of such an opportunity.

The writer has merely tried to apply an integrating process to the many hours of hard work put in by the noncommissioned officers and men who made up the operating and maintenance staffs of the stations concerned. This work was always carried out well and cheerfully, much of it under conditions of traffic congestion and emergency, and in a bad climate.

II. General Considerations

In the last few years the increasing use of the shorter waves for wireless telegraphy has brought into prominence certain peculiarities in their behaviour. The earlier instances of successful short-wave communication over long distances with the use of low power were at first ascribed to transient and unusual combinations of favourable circumstances. It was thought that they were "freaks."

Further experiment showed that such results were often obtainable. But the ratio of the number of failures to the number of successes was considerable.

Much of the earlier work was done by amateurs who were, however, closely followed up by professional investigators and by the commercial companies.

In the last two or three years research on the subject has proceeded apace. The existence of the phenomenon known as "skip effect" has come to be generally accepted, the occurrence of "fading" is established beyond doubt, and to provide at least a partial explanation of the observed effects the theory of the mechanism of propagation has had to be both trimmed and embellished.

The problem is complex. There are many variables that may be expected to affect the motion of a wave between transmitter and receiver. Any attempt to establish the influence that each factor contributes towards the observed effect amounts to a problem in statistics.

In considering the prospects of communication between two stations, the factors that we may expect to be operative are:

(a) The power and aerial system of the transmitting station.

(b) The aerial system and circuit arrangements of the receiving station.

[2] The writer has here taken the liberty of including Hong Kong in the "laboratory." It was a separate Command, and had built the first military short wave sets to be used in China, just before the Shanghai Defense Force landed.

(c) The great circle distance between the stations (hereinafter called the "range").
(d) The distribution of electrical conductivity over the earth's surface, particularly in the area near the great circle joining the stations.
(e) The distribution in space of the ionization of the earth's atmosphere, particularly in the region near the diametral plane through the two stations.
(f) The wavelength in use.

Now, although (a) and (b) will always affect the results, we shall, for the present, assume that they are given.

(c) we shall take to be an independent variable.

(d) is beyond our control and will vary, but it will lie between the lower limit for the earth's crust and the upper limit for sea water. As a first approximation it will be assumed that its effect is small compared with that of each of the other factors.

(f) may, for the present, be regarded as an independent variable.

(e) is beyond our control, and we have no direct method of measuring it. But it may be expected to depend in turn on the following more elemental factors:

(g) The intensity of solar and other ionizing activities at the time of the experiment.
(h) The geographical positions of the stations.
(i) The time of year.
(j) The time of day (taken at either station).

There is no simple way of measuring (g), which we shall, perforce, assume to be of sensibly constant value. But (h), (i), and (j) are known.

With the reservation, then, that there are other factors that may affect it, we can now restate our problem as being determined by the following:

(a) Range
(b) Geographical positions of stations
(c) Time of year
(d) Time of day
(e) Wavelength

If the stations are not to be more than a few hundred miles apart, we may, as a first approximation, for (b) substitute the belt of latitude in which both stations are situated.[3]

[3] This substitution may, however, introduce discrepancies at the periods round about dusk and dawn. For at those times the direction of the line joining the stations relative to the contour-direction of the shift-region of the Kennelly-Heaviside layer (or layers) may be expected to become significant.

The wireless engineer has no concern with what befalls those waves that fail to "get through." For him the problem is to be able to predict, for a given set of conditions, the band of wavelengths that will give the greatest probability of successful communication.

We shall call the central wavelength of this band the "best wavelength."

From this point of view the problem may be regarded thus:

Given conditions (independent variables)	Range Belt of latitude of stations Time of year Time of day
Required condition..... (dependent variable)	Best wavelength

Now the effects of varying range and time of day were apparent even in the early experiments with short waves. They have been taken into consideration by all investigators.

The seasonal change is, perhaps, less apparent, but it has been referred to in the work of some investigators.

The writer is not, however, aware that the effect of latitude has ever been noted.

Yet is it not to be expected that latitude will be a pertinent factor? For example:

(a) The arctic (or antarctic) regions experience for half the year the ionizing conditions associated with summer and daylight, and for the rest of the year those associated with winter and darkness. The variation is almost entirely seasonal.

(b) The tropics experience an almost regular diurnal change. Seasonal variation is almost nonexistent.

(c) The temperate zones experience both diurnal and seasonal variations, the relative effects of the two depending on the latitude.

The multiplicity of variables seems to indicate that the establishment of an empirical law for deriving "best wavelength" in terms of given conditions can be evolved only from the collection of a large number of statistics obtained under various sets of such conditions.

The data in this article are submitted in the hope that they may serve to fill a small part of the gap in the statistics hitherto available.

III. The Problem in China

In the winter of 1926–1927 the Chinese Nationalist movement and the resultant civil war assumed large proportions. The advance of the Cantonese forces threatened Shanghai. For the defense of British interests, the Shanghai Defense Force was dispatched. It reached Shanghai in February and March, 1927.

The military communication problems immediately encountered were mainly those connected with the local defense of the International Settlement.

In South China, however, Hong Kong had improvised two short-wave sets for use at Hong Kong and Shameen (an island near Canton containing a European settlement) and two more which were sent to Tientsin and Peking, in North China. The sets were capable of a wavelength variation from about 20 to 60 meters.

The Shanghai Defense Force had facilities for communication with North and South China by means of the ordinary cable routes, the Royal Navy and the Royal Air Force, the last-named working with the military stations at Tientsin and Hong Kong mentioned above.

The Nationalist drive having passed Shanghai, and the situation in and around that city having stabilized, endeavours were made to complete the military chain of communication.

The construction of a set at Shanghai was completed by August, 1927. Soon afterwards two sets were received from England, and one of these was manned and sent to Weihaiwei.

The positions of the stations and the ranges between them are shown in Fig. 1.

The earlier experiences with these stations revealed inconsistencies and discrepancies.

Hong Kong seemed to have little difficulty in working Shameen at all times and seasons on 30 meters. Yet communication during darkness between Tientsin and Peking, although the range was not very different, was often impossible on any wavelength. During the winter nights, communication always failed.

At night, Hong Kong and Tientsin could usually communicate on 30 meters.

The links Shanghai-Tientsin and Shanghai-Weihaiwei worked well by day on 27 to 29 meters, and at night on 43 meters until winter set in. It was found that the wavelength for night working then had to be increased to the upper limit of the sets.

The link Tientsin-Weihaiwei had characteristics similar to, but not so extreme as, those of Tientsin-Peking.

Broadly, the situation at the time was as follows: A short-wave chain had been established, lying roughly north and south. The power (input to the valves) of the stations varied between about 100 and 300 watts. It had been expected that the degree of successful communication would vary with the range and time of day. It had been expected that the wavelength used would influence results, but that it would not prove to be a critical factor. Some slight seasonal variation had been anticipated. Latitude had not been regarded as a pertinent factor.

Expectations were partly fulfilled, partly falsified.

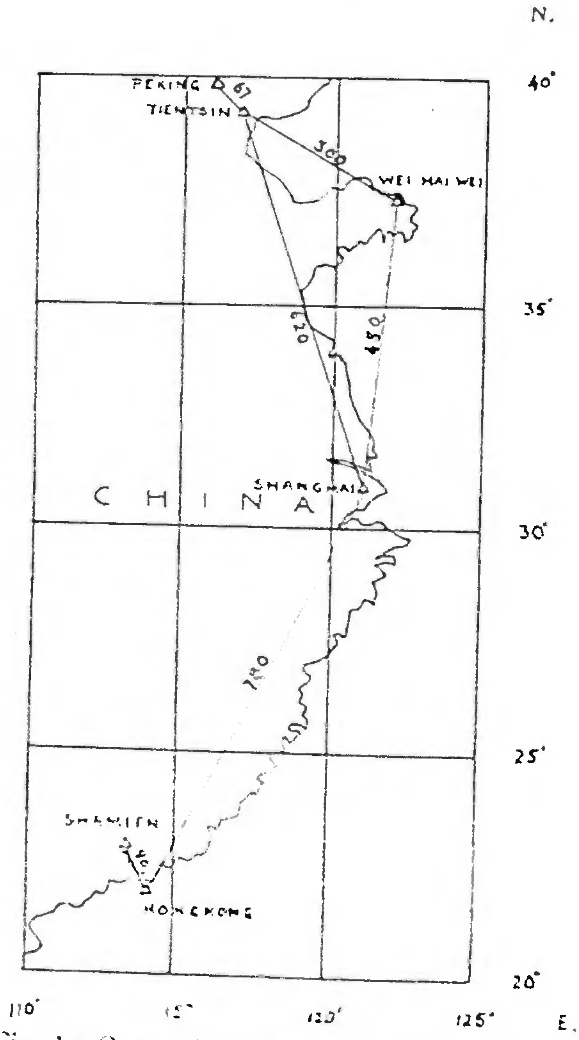

Fig. 1—Great circle ranges shown in miles.

It was thought at first that the discrepancies were attributable to the different aerial systems in use, or to the individualities of operators and sets.

The first supposition was tested by the trial of various types of aerials. But it was found that the supposition did not fit the facts. The general conclusions regarding aerials were:

(a) For short-wave work the directional effect was small compared with the effect of other factors.

(b) A horizontal half-wave Hertzian aerial gave good results for the wavelength for which it was designed. It was, however, inflexible to changes of wavelength.

(c) A vertical aerial (including Γ and T types) working on harmonics of fairly high order was flexible to changes of wavelength, though for a given wavelength its performance was not quite so good as that of a half-wave aerial designed for that particular wavelength.

(c) The influence of the design of the aerial was swamped by that of the other factors (to be discussed).

The second supposition (individualities of operators and sets) was to a large extent negatived by the results of the tests described later.

The pressing problem was how to ensure the speedy disposal of traffic which was liable to grow suddenly at short notice (and the peak might occur on any one of the links). How could the best wavelength for each link for the continually changing conditions be predicted? The relative effects of the various factors were not fully known.

The different stations favored different wavelengths, since each was confronted with a set of conditions peculiar to itself. Each link had its traffic peak-hours. These peak-periods often clashed at any station serving more than one link.

Factors other than purely physical conditions were involved. Each new phase in the political or military situation was liable to necessitate the complete recasting of the time and wavelength schedule. These phases could not always be predicted, and the difficulties of wireless control became considerable.

IV. Inauguration of Tests

It was decided that it was essential to be able to anticipate at least the changes in the purely physical conditions, and to be able to predict the wave-band that would give the greatest probability of successful communication over each link for any particular set of conditions.

In order to obtain some definite data, a series of tests was started in January, 1928, but completeness was prevented by the very heavy traffic from April onwards.

In May the already hard-pressed Wireless Section took over the Royal Air Force station in Shanghai which had been handling the traffic for home and much of the traffic internal to China. In June the Nationalist advance was approaching the Tongshan area, in North China, where there were European interests. A military wireless station was sent from Shanghai to accompany the force being dispatched to guard the Tongshan area. The station quickly established communication and immediately came in for heavy traffic with Shanghai, Tientsin, and Wei hai wei. The land-lines on either side of Tongshan were cut in succession by the rival Chinese armies as the Nationalist advance approached and passed Tongshan. The wireless station was

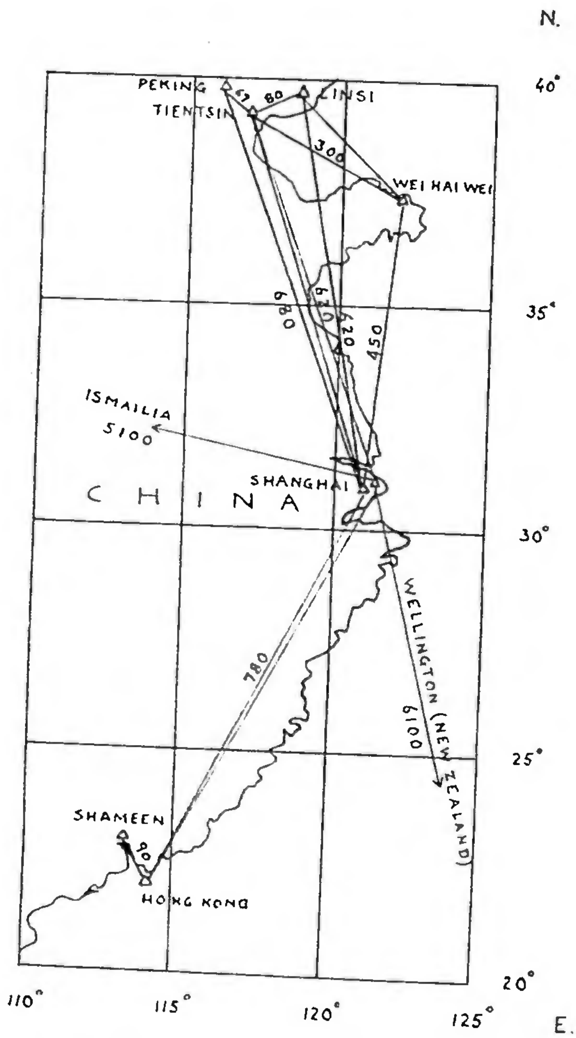

Fig. 2—Great circle ranges shown in miles.

therefore for some time the only means of signal communication between the garrison and the outside world.

The culminating point of the period of heavy traffic was reached in September when in one week the two Shanghai stations alone handled more than 23,000 words.

The links in operation from June to October 1928 are shown in Fig. 2.

V. Method Employed in the Tests

The sending station sent daily a series of transmissions on different wavelengths in accordance with a prearranged program, each wavelength being distinguished by a pair of letters allotted beforehand.

The receiving station tabulated these wavelengths with the signal strengths at which they were received. Signal strengths were estimated in the ordinary "R" code (0 to 9).

The records were then collected and divided into batches, each batch representing half a calendar month.

By taking the arithmetic mean of the signal strengths for each wavelength throughout each batch, figures were obtained which enabled the plotting of a wavelength signal-strength curve for each half month.

The results obtained between January and June 1928 are shown in Figs. 3 to 6.

VI. Weaknesses of the Method

The method was essentially qualitative rather than quantitative, and suffered from the following defects:

(a) The "R" code contains considerable ambiguity owing to the physiological differences between operators.

(b) Even if we assume all operators to have the same scale of estimation of signal strength, the figures of the "R" code bear no simple relationship to the actual field strength at the receiving station. They do, however, give an *indication* of its magnitude.

(c) Each set and aerial undoubtedly has its own characteristics, which will influence the results and prevent them from being necessarily and generally valid.

(d) Changes of operators and of receiving sets were from time to time unavoidable.

The only way to overcome these weaknesses would have been to conduct elaborate experiments involving the use of special sets and recording apparatus. This, of course, was impracticable.

VII. Probable Degree of Error

Referring to the above defects in turn:

(a) Reliefs of operators were allowed to follow their normal course, because the experiments were subordinate to the working of traffic, and because it was considered that the effect of preferences for different wavelengths by different operators would thus be merged in the general results.

(b) This does not invalidate the figures for the *best* wavelength.

(c) When the first curves were plotted it was thought that this source of error might be influencing results unduly. But the diurnal and seasonal variations of the curves go to prove the contrary.

(d) These took place without any consideration of the experimental work. Although possibly causing temporary error, over long periods they should tend to eliminate "personal prejudice."

The records of normal working experiences and of experimental work were filed and summarized separately, and compared afterwards.

It is thought that the agreement between the conclusions arrived at by the two methods lends strong support to a belief in the approximate validity of the results.

VIII. Summary of Working Experiences and Results of Tests over the Various Links, for the Period January to June, 1928, Inclusive

N. B. The links are dealt with in order of range.

(*a*) *Tientsin-Peking. Range 67 miles.*

Perhaps the most interesting of all the links because of the complete "skip effect" during winter darkness.

1. Daylight.

In January and February all wavelengths between about 36 and 58 meters gave reliable results.

Signals on 30 or 32 meters were very erratic.

In April, 54 meters gave best results, but all wavelengths, particularly the higher ones, were subject to occasional fading.

From the end of April till the end of June, 44 to 46 meters appeared to be best. Communication was reliable on about 5 days out of 7, but on the other 2 days signals were weak or inaudible over periods of 3 or 4 hours. On one occasion they were unreadable from 0900 to 1600 hours.

The acute fading of daytime signals on the 40- to 60-meter band, which was first noticed in April, tends to show that summer days are particularly subject to this effect. Similar results were observed in the summer of 1927. Day signals on this band during the winter were fairly steady.

2. Dusk.

At all seasons signals were good on all wavelengths from about 45 to 58 meters, the best wavelength being about 50 meters.

3. Darkness.

In January and February no signals were heard until February

27th. This was the first occasion on which night signals had been heard since early October, 1927.

In March, 58 meters became more and more reliable, wavelengths below 46 meters remaining unreliable.

In April, May, and June, communication remained fairly consistent on 58 meters. 36 and 32 meters were only occasionally heard.

Occasional interference was caused by dust storms, which sometimes delayed traffic for several hours. (The impact of grains of sand, etc., on the receiving aerial causes noise somewhat similar to that caused by atmospherics).

4. Experimental.

Curves for daylight and dusk periods in January and February are given in Fig. 3. They show a very definite "best wavelength" of the order of 50 meters.

It has already been stated that night signals were never heard in the winter months. The daylight curve for the second half of April shows a best wavelength of 54 meters. The dusk curve for the same period shows one of 50 meters, and the darkness curve one of 58 meters.

(*b*) *Tientsin-Linsi.*[4] *Range 80 miles.* (*June to October only*).

1. Daylight.

Linsi station came into operation on June 25th.

No tests were held, but while the 44- to 46-meter band did not seem nearly so good for day working as it was between Tientsin and Peking, 32 meters seemed rather better, though the latter was by no means reliable.

2. Dusk and darkness.

Results were practically identical with those between Tientsin and Peking.

(*c*) *Peking-Linsi. Range 112 miles.* (*June to October only*).

While Linsi's strength was often different at Peking from what it was at Tientsin, the mean strength seemed to be about the same. The only marked difference noticed was that Peking heard Linsi R8 to R6 on 32 meters at 2200 hours, Linsi being inaudible at Tientsin.

(*d*) *Tientsin-Weihaiwei. Range 300 miles.*

1. Daylight.

From January till March communication was reliable on all wavelengths between 28 and 62 meters. In the first half of March 46 to 58 meters was best. In April 28 meters was best. From early May

[4] Linsi was the actual site of the station serving the Tongshan area.

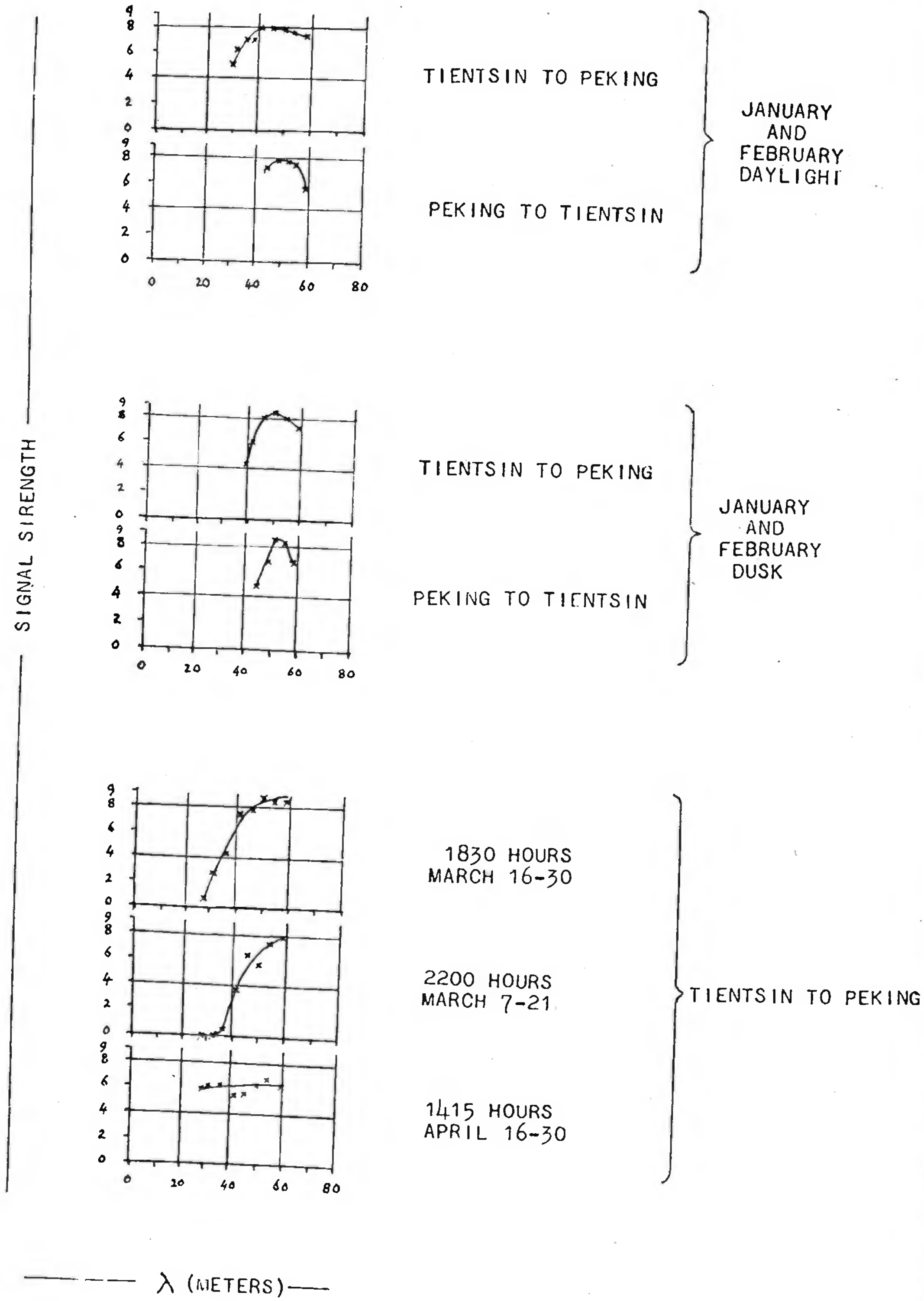

Fig. 3—Tientsin and Peking. Range 67 miles.

reliability decreased, the band 28 to 50 meters being best. Sometimes communication failed altogether.

2. Dusk.

From January till March 18th signal averaged R8 to R9 on 58 meters. From then onwards signals were consistently R9 on all wavelengths.

3. Darkness.

From January until March communication was reliable on the longer wavelengths, but wavelengths below 42 meters were practically useless.

From May onwards conditions became very bad, fading being common. Signals averaged R6 on 32 to 58 meters and R5 on 28 meters.

4. Experimental.

Results of tests are given in Fig. 4.

(e) *Shanghai-Weihaiwei. Range 450 miles.*

1. Daylight.

A wavelength of from 27 to 29 meters was used throughout the six months and proved consistently satisfactory, signals often being R9 and seldom less than R7.

2. Dusk.

Forty-three to 45 meters gave good results from January to March, signals usually being R9. Twenty-nine meters was tried in January, but although signals were sometimes R9, they often faded out completely.

After the March equinox, 38 and 46 meters seemed equally effective.

During May and June a wavelength of from 29 to 32 meters was best, but failed sometimes, and fading often caused difficulty.

3. Darkness.

In January and February it was only on two or three occasions that communication could be established. On these occasions signals were weak, and were obtained by the use of the highest wavelengths of which the sets were capable—about 60 meters.

As the spring equinox approached, the ratio of successful to unsuccessful attempts increased, the higher wavelengths still being best.

By the end of April conditions had changed to such an extent that communication was usually possible, and a wavelength of the order of 30 meters seemed to give best results.

During May communication on 29, 32, and 46 meters was very reliable, and much traffic was handled.

4. Experimental.

The only tests that it was possible to carry out with any degree of continuity were a series of daylight tests from April to June. For the second half of April they indicate a best wavelength of less than 25

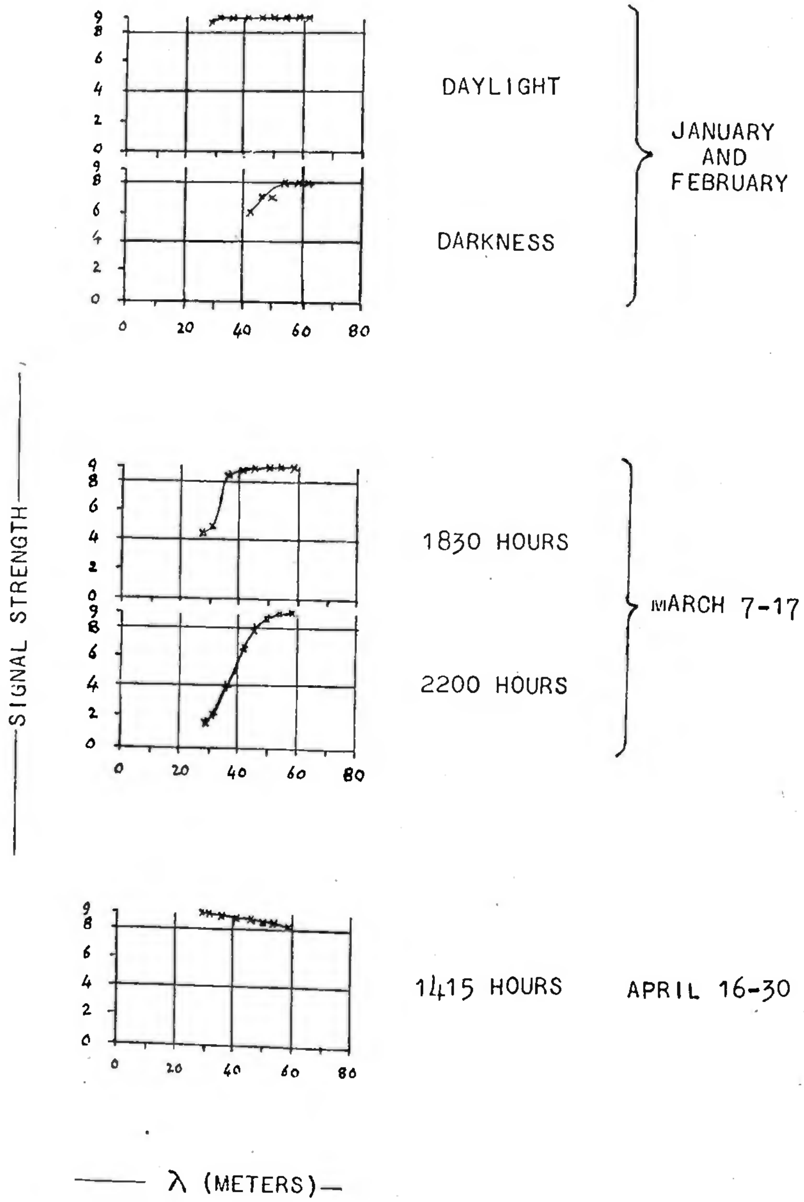

Fig. 4—Tientsin to Weihaiwei. Range 300 miles.

meters, for the first half of May one of 30 to 35 meters, and for June one of about 30 meters. See Fig. 5.

(*f*) *Shanghai-Tientsin. Range 620 miles.*

1. Daylight.

Throughout the six months daylight communication was altogether satisfactory, signal strength each way usually being from R6 to R8.

One of the Shanghai stations used 29 meters, the other 23 meters, and Tientsin 29 meters.

The 23-meter wave proved particularly reliable.

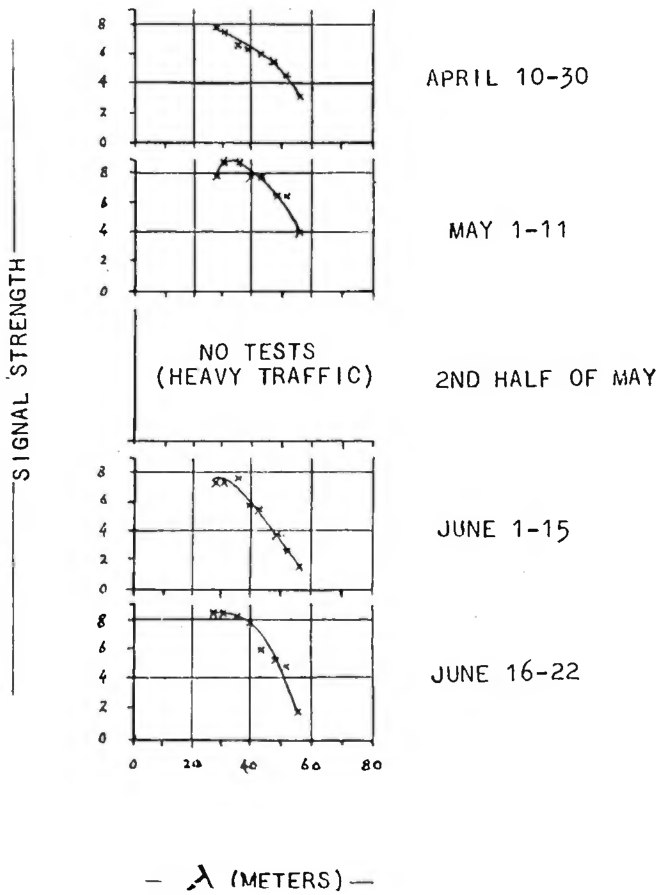

Fig. 5—Shanghai to Weihaiwei. Range 450 miles. 1100 hours.

2. Dusk.

During the greater part of January communication was uncertain on all wavelengths, but throughout February and until the middle of March it was generally good on 58 to 62 meters. During the second half of March communication failed in one direction or the other as often as it succeeded.

In April 32 meters usually gave good results. In May 32 meters gave generally good results, but atmospherics began to make their presence felt, and, even when atmospherics were slight, signals on several occasions were not heard at all.

From June 1st to 11th, 28–32 meters proved more reliable than it had been in May. On June 11th the time-table was recast to meet traffic requirements, and the period at dusk was abandoned.

3. Darkness.

During January all wavelengths obtainable by the sets were tried, but failed more often than not, particularly in the direction Shanghai to Tientsin. The higher wavelengths (58 to 62 meters) gave the best results.

In February there was a marked increase in reliability, a wavelength of 58 meters failing on one occasion only.

During March 58 meters was again satisfactory, but towards the end of the month it was found that all wavelengths down to 28 meters were as good as the higher ones.

Experiences in April, May, and June were similar to those at the end of March, i.e., all wavelengths from about 30 to 60 meters were equally satisfactory. From the beginning of May atmospherics increased but, owing to the considerable strength of signals, seldom prevented the passing of traffic.

4. Experimental.

Very little experimental work was possible in daylight owing to traffic requirements, but tests carried out in April and May indicate a best wavelength of about 30 meters for April and the first half of May, and something less than 25 meters for the second half of May.

Experiments round about dusk (actually at 1830 hours, China time) were carried out in March and April. For the first half of March the best wavelengths seem to be from 40 meters upwards. For the second half the best band is from 35 to 60 meters.

In the first half of April results were somewhat uncertain, 45 meters probably being about the best wavelength, while in late April the results were so discordant as to make it impossible to state a best wavelength. (See Fig. 6.)

The tests carried out over this link during darkness are of considerable interest, first, because they are the most complete of the series, and second, because they show to a marked degree the influence of the seasons. (See Fig. 6.)

In the latter half of January practically nothing below 40 meters was effective, but mean signal strength rose rapidly on wavelengths between 45 and 55 meters. From then until the end of March the shorter wavelengths came more and more "into the picture" until after the spring equinox (March 23rd) all wavelengths between about 30 and 60 meters (the upper limit of the sets in use) gave a mean strength of approximately R9.

The change from winter to summer conditions was most definite, and coincided closely with the actual astronomical date.

(*g*) *Shanghai-Linsi. Range 620 miles. (June to October only).*

Working experiences were almost identical with those for the link Shanghai-Tientsin.

The heavy traffic prevented experimental work.

(*h*) *Shanghai-Peking. Range 680 miles.*

Working experiences were similar to those for the link Shanghai-Tientsin, except that communication was not quite so reliable. Peking was much handicapped by local jamming.

Very little traffic was handled, though arrangements were made in May, when the fall of the native quarter of Peking to the Nationalists was imminent, for passing traffic between Peking and Tientsin via Shanghai in case of need.

The periods allotted to the link often had to be cancelled owing to pressure of work between Peking and Tientsin.

No experimental work was done.

(*i*) *Shanghai-Hongkong. Range 780 miles.*

Hongkong worked on a fixed wavelength of 30 meters.

1. Daylight.

Communication was maintained throughout January, but Hongkong seldom reported Shanghai's strength as more than R5. Shanghai used 27 to 29 meters.

In February signals in both directions weakened and there were several complete failures.

There was a slight improvement in early March, but towards the end of the month communication became very unreliable, usually failing completely in the direction Shanghai to Hongkong, while Hongkong's signals were so weak as to be unreadable.

In April and the first half of May communication failed in both directions almost without exception. The daylight period was cancelled towards the end of May.

In the summer and autumn it was found possible to send with a fair degree of reliability from Shanghai to Hongkong on 23 meters. But signals in the reverse direction (on 30 meters) were nearly always unreadable through the jamming prevalent in Shanghai.

2. Darkness.

Communication was satisfactory using the same wavelengths as for daylight, except that after summer conditions had set in 32 and

39 meters gave rather better results as sending wavelengths for Shanghai.

In May there were some failures due to atmospherics.

3. Experimental.

Practically no experimental work was possible, owing, chiefly, to traffic requirements between Hongkong and Shameen.

(j) Tientsin-Hongkong. Range 1130 miles.

Until March 14th no regular tests were held.

Communication had been maintained since January 30th at 1000 hours daily, and traffic passed, both stations using approximately 30 meters. Tientsin received Hongkong at a mean strength of R7, and Hongkong received Tientsin at a mean strength of R8 to R9.

Hongkong was received occasionally at strength R4 to R5 between 0800 and 0900 hours, and at strength R5 to R6 between 1600 and 1800 hours, but these periods were inconsistent.

Between March 14th and May 8th pressure of traffic at Tientsin prevented any daylight tests.

Hongkong on 30 to 32 meters was occasionally heard R3 to R4 between 0800 and 0900 hours, but then faded out and came in again between 1600 and 1800. Hongkong averaged about R6 at 1900 hours and heard Tientsin at strength R6 to R9 at this hour. Hongkong usually heard Tientsin's sending tests at 2200 hours at strengths varying between R5 and R8. The longer wavelengths were heard much more consistently than were the shorter ones.

From May 8th until the end of June daylight tests in both directions were carried out, but nothing was heard. Tientsin sent on 20 to 26 meters and Hongkong on about 30 meters. At 1900 hours signals in both directions were about R4 to R6, with a slight decrease from May to June. Hongkong heard Tientsin's tests at 2200 hours at about the same strength as in March and April.

(k) Shanghai-Ismailia (Royal Air Force). Range 5100 miles. Shanghai-Wellington (New Zealand). Range 6100 miles.

These links had been worked by the Royal Air Force in Shanghai until May, 1928, when the Royal Air Force Station in Shanghai was taken over by the Army.

The station at Wellington was erected by the Public Works Department, Wellington, with the assistance of a Royal Signals Warrant Officer, and manned by the Central Depot Corps of Signals (New Zealand Territorial Force).

The work in Shanghai was taken over from the Royal Air Force at the time of year when atmospherics and fading became troublesome. It happened, too, that the events in North China which led to a very large increase of traffic occurred soon afterwards.

Periods with the two distant stations were worked only while the terminal stations were in darkness, on a wavelength of 39 meters.

Throughout the summer and well into the autumn conditions between Shanghai and Ismailia were unfavorable, signal strength varying from R7 to R3 or less. Atmospherics were heavy, and fading usually set in at about 0600 hours (China time). 23 meters was tried from 0600 hours onwards, but with no success.

Conditions for the link Shanghai-Wellington were much more favorable. Wellington usually received Shanghai at strength R9, and signals in the opposite direction were almost as good.

In November and December conditions on the link Shanghai-Ismailia became favorable again, and traffic could nearly always be cleared.

IX. The Period July-December, 1928

In section VIII has been given a summary of the experiences over a period embracing both winter and summer conditions.

As winter once more approached, the changes of wavelength appropriate to each link had to be carried out in accordance with the experience derived in the previous season.

The transition from summer to winter conditions seemed to lag by about half a month behind the autumnal equinox. The curves derived from the results of tests (not reproduced in this article) showed a similar transition (in the reverse order) to those for spring, but the transition was not so abrupt, and was accompanied by discontinuities. The seasonal change was not perfectly symmetrical about the summer solstice. An interesting feature which also showed asymmetry was that atmospherics, although they did not begin to be troublesome until May, were heavy throughout the summer and early autumn, and did not become inconsiderable until the end of October.

Once winter conditions had fully set in, the values for best wavelengths obtained during the previous winter were completely confirmed.

X. An Attempt to Generalize the Results

The work in China showed the advantages possessed by short-wave wireless telegraphy as a means of handling comparatively heavy traffic over medium and long ranges with the use of low power and inexpensive apparatus.

But the limitations to its use were also revealed.

The limitations are chiefly due to the necessity for changing wavelengths to meet different sets of physical conditions. Unless the conditions under which a group of stations will be required to operate can be anticipated, and their influence predicted, difficulty will be found in searching for suitable day and night wavelengths for the various links, and in readjusting those wavelengths to meet the seasonal changes.

The question is of commercial as well as of military importance.

Given a large enough number of statistics it should be possible to construct tables or curves, or even an empirical law, to show the best wavelength in terms of the conditions. The breadth of the wave-band within which successful communication may be expected is also of practical importance.

In Fig. 7 is shown an attempt to incorporate the results of traffic experience and tests in a "Range-Best Wavelength" diagram, for the belt of latitude lying between 30 deg. and 40 deg.

The following notes supplement the diagram:

(a) The full line shows what is considered to be the best wavelength for given range.

(b) The dotted lines show the upper and lower limits of wavelength giving workable signals.

The effect of increasing the power of the sending station is to broaden the strip (shown shaded) bounded by the dotted lines.

For darkness (summer and winter) no upper limit was found. If there is an upper limit it lies beyond the highest wavelength that could be reached by the sets in use at the time.

(c) The curves for daylight are approximately the same for summer and winter. It is thought that the same curve would apply closely for any part of the world, whatever the latitude, provided *both* stations were in daylight.

An approximate law for best wavelength is given by:

$$\lambda = 0.046\ (1120\text{-}d)$$

where λ is the best wavelength in meters and d is the range in miles.

It appears that, with a power input of the order of that used at the stations in China, a daylight range of more than about 1200 miles cannot be expected.

(d) The curves for darkness are much more uncertain than those for daylight. It is thought that they would vary considerably with the latitudes of the stations.

(e) Best wavelengths for cases where the conditions of spring, autumn, dusk, or dawn prevail are intermediate between those shown for the extreme conditions. Within the transition periods near the

equinoxes and near dusk and dawn, the value of best wavelength cannot be accurately computed, since it varies throughout the transition period.

(f) To try and solve the difficulties of winter night communication between Tientsin and Peking, it was decided to raise the upper wavelength limit of the sets in use at those places. Since his return to this country the writer has learned that this step was carried out, with complete success. A wavelength of about 80 meters seems to have provided the solution.

The contrast between the results, as described, for Northern China and the experiences of Hongkong on 30 meters lends strong support to the belief that latitude is an important factor in the problem of nocturnal communication.

On the other hand the similarity of the range-best wavelength curves for summer and winter daylight suggests that the influence of latitude on daytime communication is almost negligible.

It would be interesting to obtain and compare equivalent statistics for regions of higher and lower latitudes than those to which the data in this article are directly applicable.

Proceedings of the Institute of Radio Engineers
Volume 18, Number 12 *December, 1930*

BASIS ESTABLISHED BY THE FEDERAL RADIO COMMISSION FOR THE DIVISION OF RADIO BROADCAST FACILITIES WITHIN THE UNITED STATES*

AT a session of the Federal Radio Commission held at its office in Washington, D. C., on June 17, 1930, the Commission adopted the following General Order:

WHEREAS, the Act of Congress approved March 28, 1928, entitled "An Act Continuing for One Year the Powers and Authority of the Federal Radio Commission under the Radio Act of 1927, and for other Purposes," provides and declares that "The people of all the zones established by section 2 of this Act (Radio Act of 1927) are entitled to equality of radio broadcast service, both of transmission and of reception," and

WHEREAS, said Act approved March 28, 1928, above referred to, also provides and requires that "in order to provide said equality the licensing authority shall as nearly as possible make and maintain an equal allocation of broadcast licenses, of bands of frequency or wavelengths, of periods of time for operation, and of station power," and

WHEREAS, it was the intent and purpose of Congress to secure an equal distribution of radio broadcast service, both of transmission and of reception between the five zones aforesaid, and

WHEREAS, it is necessary, in order to make and maintain such equal allocation between said five zones and fairly and equitably between the states within each of the zones, that the Commission determine the value of stations of various classes or of various powers in effecting such allocation, and

WHEREAS, the Commission has sought and obtained the best advice and information available and has given much time to an intensive study to such values of stations of various powers, and

WHEREAS, the Commission, through its engineers and from studies made by the Commission has considered all the elements required by Congress to be considered, and has allowed the paramount intent and purpose of the Act of March 28, 1928, above referred to, to control, i.e., "that the people of all the zones* * * * are entitled to equality of radio broadcast service, both of transmission and of reception," and

WHEREAS, it has been found that, according to the broadcast service rendered to the people of each zone and of the states within each zone by stations of various classes, both of transmission and of reception, each class of station is of the following value in units, to-wit:

* Decimal classification: R007. Original manuscript received by the Institute, July 7, 1930.

Classes of Stations	Value in Units
A. For Full Time Station	
(1) Stations of a power of 5 kw or more, one station only operating on the channel at night	5 units
(2) Stations of a power of 5 kw or more, two stations operating simultaneously on a common frequency and separated by 2000 miles or more	4 units
(3) Stations of a power of 5 kw or more, two or more stations operating on a common frequency and stations separated by less than 2000 miles	2 units
(4) Stations of a power of 1 kw, two or more stations operating simultaneously on a common frequency	1 unit
(5) Stations with 500-watts power with more than 2 stations operating simultaneously on a common frequency	0.6 unit
(6) Stations with 250-watt power with more than two stations operating simultaneously on a common frequency	0.4 unit
(7) Stations with 100-watts power or less, with two or more stations per zone operating simultaneously on a common frequency	0.2 unit
B. Day Stations	
(1) Stations of a power of 5 kw operating during daylight hours only simultaneously with stations of Class A(1), above	1.5 units
(2) Stations of a power of 2.5 kw operating during daylight hours only	0.75 unit
(3) Stations of a power of 1 kw operating during daylight hours only	0.5 unit
(4) 500-watt, 250-watt, or 100-watt stations operating during daylight hours only, one-half values given for corresponding full time stations above	
C. Full Time Stations Having Excess Day Power	
All stations shall have their values in units based on one-half the units for full time stations of same power as the stations have at night plus the value in units for a day station of the same power as the station has in daytime as follows:	
1 kw night $2\frac{1}{2}$ kw day equal	1.25 units
500 watts night 1 kw day equal	0.8 unit
250 watts night 500 watts day equal	0.5 unit
100 watts night 250 watts day equal	0.3 unit

D. Limited Time Stations

For stations of more than 5 kw the value of units will be the same for all powers. The units will be based on 5 units. The units for each station will therefore be 2.5 for day operation plus 2.5 times hours used between 6:00 P.M., and 12:00 P.M., Local Time, divided by 12.

Stations over 5 kw operating	1 night hour	2.7 units
	2 night hours	2.9 units
	3 night hours	3.1 units

For stations of 5 kw the basis shall be 1.5 units for day operation the same as a 5 kw day station given above, plus 2.5 units times hours used between 6:00 P.M. and 12:00 P.M., Local Time, divided by 12.

Stations of 5 kw operating	1 night hour	1.7 units
	2 night hours	1.9 units
	3 night hours	2.1 units

For stations operating with power of 1 kw, 500 watts, and 250 watts, the value in units shall be the same as for a day station plus the value in units of day station, times number of night hours used between 6:00 P.M. and 12:00 P.M., Local Time, divided by 12.

1000-watt stations operating	1 night hour	0.54 unit
	2 night hours	0.58 unit
	3 night hours	0.62 unit
500-watt stations operating	1 night hour	0.32 unit
	2 night hours	0.35 unit
	3 night hours	0.38 unit
250-watt stations operating	1 night hour	0.22 unit
	2 night hours	0.23 unit
	3 night hours	0.25 unit

For stations dividing time on the same frequency the value assigned will be in proportion to the time assigned.

It is therefore

ORDERED that the values of radio broadcast stations of the various classes, powers, and time of operation be and they are hereby fixed in units as above set forth, and

IT IS FURTHER ORDERED that each of the five zones created by Section 2 of the Radio Act of 1927 shall each have broadcast stations the total value in units of which shall be equal, and shall be fairly and equitably distributed among and allocated to the states within each of said zones in proportion to the population each of said states bears to the population of the zone, and that the quota of broadcast facilities to which each state is entitled shall be determined and fixed as herein provided and in accordance with values in units for various classes of stations above set out.

Explanatory Statement and Revised Quota Tables*

The "Davis Amendment" to the Radio Act, approved March 28, 1928, requires that the radio supervising authority of the United States " . . . . shall as nearly as possible make and maintain an equal allocation of broadcast licenses, of bands of frequency or wavelengths, of periods of time for operation, and of station power, to each of said (five) zones . . . . and shall make a fair and equitable allocation of licenses, wavelengths, time of operation, and station power to each of the States . . . . within each zone, according to population (of each state)."

The proportion of the maximum possible national broadcast facilities due each State is, therefore, fixed by law. The percentages or number of units due each State are based upon official estimates of 1930 populations prepared by the U. S. Census Bureau.

It is evident from a consideration of the estimated and variable factors and the different economic and geographic conditions in various parts of the United States that the quota allocation can never be exact. The ratios will vary from time to time as conditions are further improved by continued development of the radio art and decisions of the Radio Commission.

General Order No. 40, adopted by the Commission, August 30, 1928, is an outline basis for an equitable distribution of broadcast facilities in accordance with the "Davis Amendment," considering public interest, convenience, and necessity. As amended, it provides for a certain number of high power stations on interference free channels to serve rural and sparsely settled areas over long distances under favorable conditions. It also provides for a comparatively large number of smaller stations to serve state, regional, and city local areas. The stations assigned to these three classes of service, by virtue of power, frequency, and service area, have become known popularly as Clear, Regional, and Local Channel Stations.

Based on the frequency allocation specified in General Order No. 40, it was necessary for the Commission to determine the maximum number of stations of various powers which could operate simultane-

* Prepared by the Engineering Division of the Federal Radio Commission.

ously at night in the United States without objectionable interference, so that quota tables could be prepared showing the facilities assigned to each Zone, and each State within a Zone, for comparison with facilities due.

The following table was established in 1928:

40 night stations on clear channels, each 5 kw or more
130 night stations on regional channels, each 500 to 1000 watts
150 night stations on local channels, each 100 watts or less

In accordance with this table, two or more stations dividing time on one assignment were considered as one station. ("Limited Time Stations" operating on clear channels and "Day Stations" were not charged to "quota")

The "quota" system adopted in 1928 showed the number of full time station assignments of each of the three classes due each state as compared to the number of full time assignments licensed. These figures nearly all came out in fractions showing further the impossibility of an exact allocation among states based on population.

Under the 1928 system, if a State was "under quota" on one class of service and "over-quota" on another class, it was not practicable to determine the total value of the three classes of assignments so that one could be balanced against another to determine if a State was actually "under or over quota" on total radio facilities.

Therefore, the development of a "unit system" was undertaken to evaluate stations, based on type of channel, power, hours of operation, and all other considerations required by law.

The result of this research is General Order No. 92, adopted June 17, 1930, specifying the "unit value" of stations of various types, powers, etc., including "Limited Time" and "Day" stations as chargeable to "quota."

In order to calculate the number of units due each zone and each state in accordance with population, it was necessary to determine the number of channels of different classes and number of stations of various powers which could be used for simultaneous operation without objectionable interference, and calculate the unit values. The following table, based on General Order No. 40 and Amendments, was adopted by the Commission June 25, 1930.

CLEAR CHANNELS

40 Channels—5 kw up—full time	200 units
Day stations—5 watts to 5 kilowatts	10 "
Limited time stations—100 watts to 50 kilowatts	25 "
	235 units

Regional Channels

4 Channels—5 kw—two stations per channel	16 units
31 Channels—1 kw—three stations per channel	93 "
9 Channels—500 watts—four stations per channel	18 "
½—250 watts	127 units

(Including day stations and additional day power)

Local Channels

6 Channels—100 watts—thirty stations per channel	36 units
(Including day stations and additional day power)	36 units
Total	398 "
Round Total	400 "

To be Divided in Each Zone

Clear	Regional	Local	Total
47 Units	26 Units	7 Units	80 Units

A complete tabulation of the revised quota figures by zones and states follows, giving units due, based on preliminary population figures for 1930 issued by the U. S. Census Bureau, and units assigned as of July 31, 1930.

The number of units assigned will vary from time to time as changes are ordered or approved by the Commission.

TABLE I

First Zone

State	Population of Zones and States 1930	Percentage of Total Zone Facilities Due Each State 1930	Total Units Due Each State 1930	Total Units Assigned Each State as of July 31, 1930	Total Units Under or Over Quota as of July 31, 1930
Connecticut	1,604,711	5.58	4.46	3.62	−0.84
Delaware	238,380	.83	.67	0.7	+0.03
Dist. of Columbia	486,869	1.67	1.33	1.3	−0.03
Maine	800,056	2.78	2.22	2.0	−0.22
Maryland	1,629,321	5.70	4.56	3.8	−0.76
Massachusetts	4,253,646	14.80	11.85	10.08	−1.77
New Hampshire	455,293	1.64	1.31	0.2	−1.11
New Jersey	4,028,027	14.01	11.21	11.43	+0.22
New York	12,619,503	43.87	35.10	38.65	+3.55
Rhode Island	687,232	2.39	1.91	1.4	−0.51
Vermont	359,092	1.25	1.00	0.3	−0.70
Porto Rico	1,543,913	5.40	4.32	0.6	−3.72
Virgin Islands	22,012	.08	.06	.0	−0.05
	28,738,055	100.00	80.00	74.08	−5.92

TABLE II

Second Zone

State	Population Zones and States 1930	Percentage of Total Zone Facilities Due Each State 1930	Total Units Due Each State 1930	Total Units Assigned Each State as of July 31, 1930	Total Units Under or Over Quota as of July 31, 1930
Kentucky	2,623,668	9.42	7.54	7.62	+0.08
Michigan	4,842,280	17.36	13.88	10.90	−2.98
Ohio	6,639,837	23.80	19.05	18.55	−0.50
Pennsylvania	9,640,802	34.56	27.64	19.17	−8.47
Virginia	2,419,471	8.67	6.94	9.50	+2.56
West Virginia	1,728,510	6.19	4.95	3.60	−1.35
Total	27,894,568	100.00	80.00	69.34	−10.66

TABLE III

Third Zone

State	Population Zones and States 1930	Percentage of Total Zone Facilities Due Each State 1930	Total Units Due Each State 1930	Total Units Assigned Each State as of July 31, 1930	Total Units Under or Over Quota as of July 31, 1930
Alabama	2,645,297	9.23	7.39	4.50	−2.89
Arkansas	1,853,981	6.46	5.17	4.40	− .77
Florida	1,466,625	5.11	4.09	8.35	+4.26
Georgia	2,902,443	10.11	8.09	7.60	− .49
Louisiana	2,094,496	7.29	5.83	8.50	+2.67
Mississippi	2,007,979	7.00	5.60	2.60	−3.00
North Carolina	3,170,287	11.05	8.83	7.82	−1.01
Oklahoma	2,391,777	8.34	6.67	7.75	+1.08
South Carolina	1,732,567	6.03	4.82	1.70	−3.12
Tennessee	2,608,759	9.10	7.29	13.0	+5.71
Texas	5,821,272	20.28	16.22	22.77	+6.55
Total	28,695,483	100.00	80.00	88.99	+8.99

TABLE IV

Fourth Zone

State	Population Zones and States 1930	Percentage of Total Zone Facilities Due Each State 1930	Total Units Due Each State 1930	Total Units Assigned Each State as of July 31, 1930	Total Units Under or Over Quota as of July 31, 1930
Illinois	7,607,684	28.16	22.50	32.80	+10.30
Indiana	3,225,600	11.94	9.53	6.07	−3.46
Iowa	2,467,900	9.12	7.30	12.62	+5.32
Kansas	1,879,946	6.94	5.56	4.81	−0.75
Minnesota	2,566,445	9.48	7.59	9.01	+1.42
Missouri	3,620,961	13.40	10.72	12.0	+1.28
Nebraska	1,378,900	5.10	4.08	5.83	+1.75
North Dakota	682,448	2.52	2.02	2.30	+0.28
South Dakota	690,755	2.55	2.04	3.41	+1.37
Wisconsin	2,930,282	10.83	8.66	7.33	−1.33
Total	27,050,921	100.00	80.00	96.18	+16.18

TABLE V

FIFTH ZONE

State	Population Zones and States 1930	Percentage of Total Zone Facilities Due Each State 1930	Total Units Due Each State 1930	Total Units Assigned Each State as of July 31, 1930	Total Units Under or Over Quota as of July 31, 1930 1930
Arizona	435,833	3.53	2.83	2.60	−0.23
California	5,672,009	46.06	36.85	39.53	+2.68
Colorado	1,035,043	8.42	6.74	9.62	+2.88
Idaho	445,837	3.62	2.89	2.25	−0.64
Montana	536,332	4.35	3.48	2.90	−0.58
Nevada	90,981	.74	.59	0.80	+0.21
New Mexico	427,216	3.47	2.77	2.37	−0.40
Oregon	952,691	7.74	6.19	8.41	+2.22
Utah	502,582	4.08	3.27	6.60	+3.33
Washington	1,561,967	12.70	10.16	15.92	+5.76
Wyoming	224,597	1.82	1.46	.20	−1.26
Alaska	58,758	.48	.38	1.00	+0.62
Hawaii	368,336	2.99	2.39	1.60	−0.79
Total	12,312,182	100.00	80.00	93.80	+13.80

Table VI gives the number of units and the percentage of units by which the states are over or under quota.

TABLE VI

NUMBER OF UNITS OVER AND UNDER QUOTA		PERCENTAGE OF UNITS OVER AND UNDER QUOTA	
	Over		Over
Illinois	10.30	Alaska	163
Texas	6.55	Florida	104
Washington	5.76	Utah	102
Tennessee	5.71	Tennessee	78
Iowa	5.32	Iowa	73
Florida	4.26	South Dakota	67
New York	3.55	Washington	57
Utah	3.33	Louisiana	46
Colorado	2.88	Illinois	46
California	2.68	Nebraska	43
Louisiana	2.67	Colorado	43
Virginia	2.56	Texas	40
Oregon	2.22	Virginia	37
Nebraska	1.75	Oregon	36
Minnesota	1.42	Nevada	35
South Dakota	1.37	Minnesota	19
Missouri	1.28	Oklahoma	16
Oklahoma	1.08	North Dakota	14
Alaska	0.62	Missouri	12
North Dakota	0.28	New York	10
New Jersey	0.22	California	7
Nevada	0.21	Delaware	5
Kentucky	0.08	New Jersey	2
Delaware	0.03	Kentucky	1
	Under		Under
Dist. of Columbia	0.03	Dist. of Columbia	2
Virgin Islands	0.06	Ohio	3
Maine	0.22	Georgia	6
Arizona	0.23	Arizona	8
New Mexico	0.40	Maine	10
Georgia	0.49	North Carolina	11
Ohio	0.50	Kansas	13
Rhode Island	0.51	New Mexico	14
Montana	0.58	Massachusetts	15
Idaho	0.64	Arkansas	15
Vermont	0.70	Wisconsin	15
Kansas	0.75	Montana	17
Maryland	0.76	Maryland	17
Arkansas	0.77	Connecticut	19
Hawaii	0.79	Michigan	21

TABLE VI—(*Continued*)

Number of Units Over and Under Quota		Percentage of Units Over and Under Quota	
	Under		Under
Connecticut	0.84	Idaho	22
North Carolina	1.01	West Virginia	27
New Hampshire	1.11	Rhode Island	27
Wyoming	1.26	Pennsylvania	31
Wisconsin	1.33	Hawaii	33
West Virginia	1.35	Indiana	36
Massachusetts	1.77	Alabama	40
Alabama	2.89	Mississippi	53
Michigan	2.98	South Carolina	65
Mississippi	3.00	Vermont	70
South Carolina	3.12	New Hampshire	85
Indiana	3.46	Porto Rico	86
Porto Rico	3.72	Wyoming	86
Pennsylvania	8.47	Virgin Islands	100

A comparison of the revised unit quota system with the superseded system shows that the differences are relatively small and that the radio set-up of the country will not be unduly disturbed by the adoption of the improved system.

Concluding, it appears that the United States was 22.39 units over the maximum 400 as of July 31, 1930, averaging ½ unit per state. Twenty-five states were under quota as compared to twenty-three states over quota. This is quite a fair balance considering the many technical complications and conflicting interests.

As pointed out, the distribution of radio facilities can never be exactly proportional but it is evident that conditions will be gradually improved as some stations which are financially unsound "go off the air" and other assignments are transferred from over quota to under quota areas.

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A STUDY OF THE HIGH-FREQUENCY RESISTANCE OF SINGLE LAYER COILS*

By

A. J. Palermo and F. W. Grover

(Electrical Engineering Department, Union College, Schenectady, N. Y.)

Summary—*Existing formulas for the high-frequency resistance of single layer coils have all been obtained by making simplifying assumptions. The errors due to these assumptions are difficult to estimate and the data available are not conclusive.*

This paper gives the experimental results of a systematic study of the high-frequency resistance of single layer coils of the forms and sizes usual in radio circuits in the broadcast range of frequency.

New formulas are derived by expanding Hickman's low-frequency formulas in asymptotic series, thus making the formulas applicable to high frequencies. An empirical formula for coils of intermediate lengths is also found. These formulas are compared with the experimental results.

Introduction

It is well known that the resistance offered by a conductor to the flow of alternating current at high frequencies is much greater than that offered to direct current. For example, resistance ratios, i.e., the quotients of the high-frequency resistance by the direct-current resistance, of the value of 10 are not uncommon. The elements of a straight, isolated conductor near the center have more flux interlinkages and consequently more inductance than do those elements near the surface of the conductor. Accordingly, most of the current will flow at the surface of the conductor when extremely high frequencies are used. For this reason the phenomenon is called *skin effect*. Thus the shape of the cross section of a conductor to be used at high frequencies is of practical importance. Antenna wires made of any material subject to oxidation, should have some surface protection. In the study of skin effect, what is desired is not so much the actual resistance as the resistance ratio.

Due to the work of Maxwell,[1] Heaviside,[2] Kelvin,[3] and Raleigh,[4] the theory for the straight, isolated conductors of circular cross section is complete and tables of the resistance ratio for this case are avail-

* Decimal classification: R144. Original manuscript received by the Institute, July 7, 1930.

[1] Maxwell, *Elect. and Mag.*, 2, paragraph 690.

[2] Heaviside, *Electrician*, p. 583, 1884. p. 583, 1885; *Elect.*, papers, 1, pp. 353, 429; 2, pp. 50, 97.

[3] *Math. and Phys. Papers*, III, 491, 1889.

[4] Raleigh, *Phil. Mag.*, 21, 381; 1886.

able.[5] The theory for isolated, thin tubular conductors has been developed by Dwight.[6]

The case of two parallel, cylindrical conductors is more complicated owing to the so-called *proximity effect,* that is, the mutual effect of the magnetic field of one conductor in bringing about an asymmetrical distribution of current density about the axis of the other. The case of two such conductors forming a return circuit, has been treated by Curtis[7] by means of an integral equation method in a form applicable to low frequency and confirmed by his experiments. Butterworth[8] has treated the same problem by solving Maxwell's equations for circular conductors with the current flowing in the same or in opposite directions. His equations, which apply both to low and high frequencies, are confirmed by the very accurate work carried out experimentally by Kennelly, Laws, and Pierce.[9] The same problem has also been solved by Carson[10] and Snow.[11]

The treatment of skin effect in simple, single layer coils is still more difficult, and a rigid mathematical solution appearing out of the question, it is necessary to make certain simplifying assumptions.

Thus Sommerfeld[12] has obtained a solution for an infinitely long helical winding of wire having a rectangular cross section with the turns in contact and having a coil radius very large in comparison with the thickness of the wire. To agree with the experimental results of Wien[13] and Black[14] whose work had to deal with wire of circular cross section, Sommerfeld found it necessary to introduce into his final formula a numerical factor, 0.58. In a later paper, Sommerfeld took into account the pitch of the winding and treated the case of round wires. This latter work gave a theoretical confirmation of the numerical constant indicated in the earlier investigation and showed the influence of the pitch of the winding. It is not to be expected that Sommerfeld's formulas should apply accurately to the short coils common in radio work.

Butterworth[8] has extended his formula for the skin effect in parallel cylinders to the case of an array of them equally spaced in a plane. To this he applied a correction to take account of the curvature of the

[5] *Bureau of Standards Scientific Paper* 169, Tables xxii–xxiv inclusive.
[6] H. B. Dwight, *Trans. A.I.E.E.*, p. 1379, 1918; *Jour. A.I.E.E.*, p. 203, 1922.
[7] H. L. Curtis, *Bureau of Standards Scientific Paper* 374, 1920.
[8] S. Butterworth, *Phil. Trans. A.* 222, 57, 1921.
[9] A. E. Kennelly, F. A. Laws, and P. H. Pierce, *Trans. A.I.E.E.*, p. 1953, 1915.
[10] J. R. Carson, *Phil. Mag.*, 41, 607, 1921.
[11] C. Snow, *Proc. International Mathematical Cong.*, Toronto, 1924.
[12] A. Sommerfeld, *Ann. der Phys.*, (4) 15, p. 673, 1904; (4) 24, 1907.
[13] M. Wien, *Ann. der Phys.*, (4) 14, p. 1, 1904.
[14] T. Black, *Ann. der Phys.*, 19, 157, 1906.

turns in an actual coil, but under the assumption of a large number of turns in the coil. The approximate formula thus derived would not, therefore, be expected to apply to the case of very short coils. Butterworth showed that his expression for the resistance ratio checked the first and more important term of an empirical formula derived by Lindemann and Hüeter[15] from their measurements on short coils.

Hickman[16] has also made use of the idea of representing the coil by an array of parallel cylinders. To take account of the curvature of the turns he has assumed two parallel floors of cylinders, the current direction in one being opposite to that in the other, and has determined from the inductance formulas the distance necessary between the floors to make the mutual inductance of two pairs of go and return cylinders equal to the mutual inductance of the corresponding turns. The distance between floors was found to be 0.54 of the diameter of the coil. This assumption enabled him to consider the flux distribution between floors to be the same as that in the coil. To find the resistance ratio of the parallel floor system, Hickman has used the integral equation method of Curtis, assuming the distribution of current density to be the same in all of the wires.

Measurements at low frequencies, made by Hickman in certain long coils wound with thick wire in an open pitch, gave results confirming his formula for the resistance ratio, but differed largely from the values calculated by the formulas of Butterworth and Sommerfeld.

In view of the lack of agreement between the existing formulas for the single layer coil, all of which refer to idealized cases, further experimental evidence has seemed necessary.

The present paper has for its purpose a systematic study of the resistance ratio of the single layer coil of the forms and sizes common in radio circuits and for the broadcast range of frequency.

The cases covered are as follows:

Sizes of wire: 20, 32, 40, and 51 mil copper wire.

Diameter of winding: 3.86, 3.27, and 2.68 inches.

Number of turns: 20 to 50 turns.

Frequency range: 600 to 1500 kilocycles in steps of 100 kilocycles.

It is hoped that the curves which embody the results of the experimental part of the paper will prove generally useful in the design of coils for radio circuits.

The experimental results show that the resistance ratio depends not only upon the frequency, size of wire, and pitch of winding but also upon the ratio of the length to the diameter. The measurements

[15] R. Lindemann and W. Hüeter, *Verh. deutsch. Phys. Ges.*, 15, 219, 1913.
[16] C. N. Hickman, *Bureau of Standards Scientific Paper* 472.

have been compared with formulas obtained below by expanding Hickman's low-frequency formulas in asymptotic series applicable to the high-frequency range.

Experimental

The resistance variation[17] method was employed to obtain the high-frequency resistance. The essential parts of the set-up are shown in Fig. 1 and consist of an oscillator with a 201-A tube and a resonance circuit. This latter contained the coil to be measured, a variable air condenser of the Bureau of Standards type having a maximum capacity of 1000 mmf and a minimum capacity of 70 $\mu\mu$f, a thermocouple whose heater resistance was 1.2 ohm and made of advance wire, a galvanometer and several standard resistances also made of advance wire and sealed in glass tubes. The resistance standards had 2-mil advance wire. The resistivity for advance is 49 microhms per centimeter tube. The standards had resistances of 2.49 and 5.05 ohms and lengths of the order of 5 centimeters. For 2-mil advance wire the resistance ratio at the highest frequency used, 1,500,000 cycles, was less than 1.0003. The thermocouple heater had still smaller advance wire so that from the above it can be seen that the change in resistance as the frequency increased had no appreciable effect on the measured values either in the standards or in the thermocouple. The ends of the standards were dipped into two mercury wells to insure good contact.

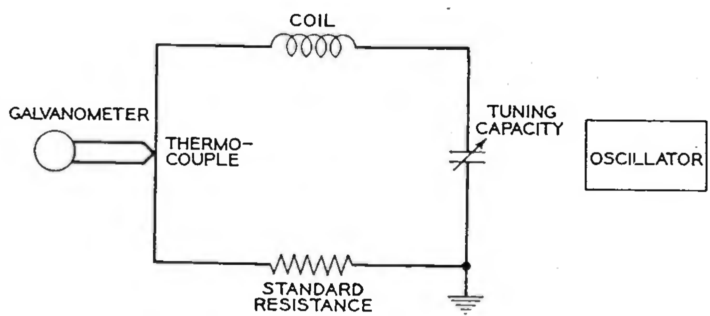


Fig. 1

The greatest deviation from the mean reading was 0.5 per cent at 600 kilocycles and 1.5 per cent at 1500 kilocycles in the measurements of this paper. These deviations compare with measurements made at the Bureau of Standards on some of the coils as follows: 0.2 per cent at 685 kilocycles and 0.6 per cent at 1375 kilocycles.

At our request the Radio Division of the Bureau of Standards

[17] Resistance variation method, *Bureau of Standards Circular* 74, page 180.

kindly made measurements on two of the coils used in this paper. The results of this comparison, which are given in the following table, show that constant errors of practical importance in the apparatus and method of measurement have been successfully eliminated.

TABLE I

Coil	Frequency in Kilocycles	Resistance B.S.	Resistance Ours
A	685	3.28	3.30
	1375	8.00	8.1
B	685	3.14	3.16
	1375	6.50	6.58

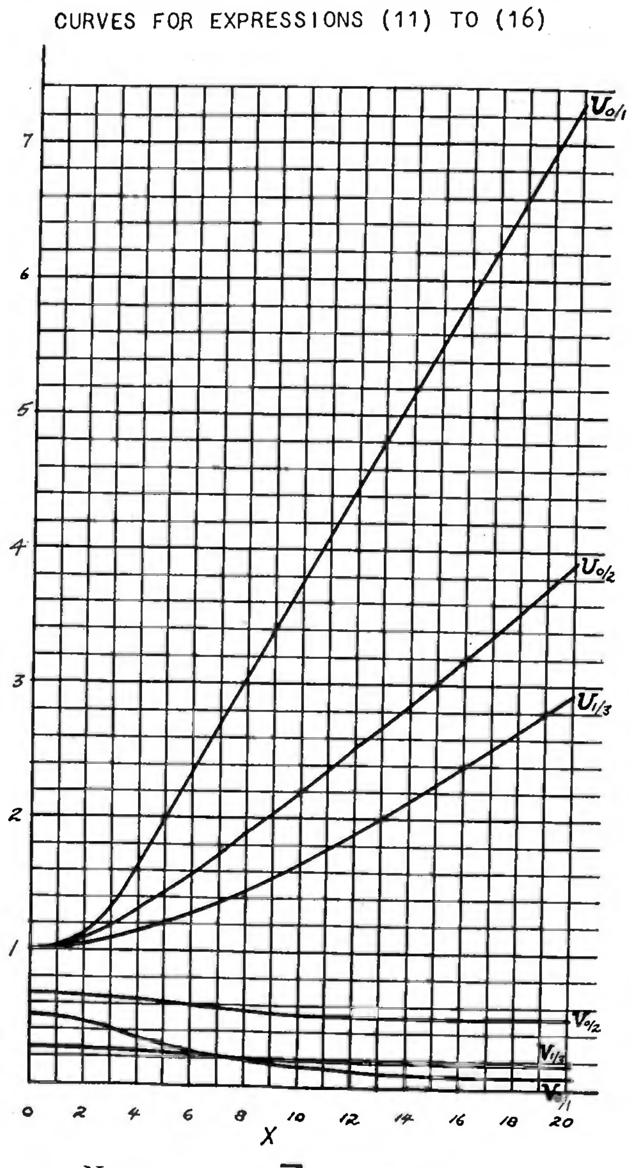
The coiling effect is plotted, rather than the resistance ratio, for greater convenience in making interpolations, and because of its interest in showing the proportionate increase in high-frequency resistance brought about by (a) the increased magnetic field due to coiling of the wire, and (b) the mutual distorting effect on the current density in the wire's cross section. The resistance ratio is found by taking the product of the coiling effect and the resistance ratio of the same wire stretched out straight. The latter may be obtained with more than requisite accuracy from the Bureau of Standards Circular, No. 74, table 17, page 309.

The measured values, given below, were obtained with the e.m.f. of the resonance circuit induced into the test coil as is customary. The effect of the self-capacity of the coil is to multiply the effective resistance of the combination of the standard resistance and the thermocouple in series by a fraction, $1/(1+C_0/C)^2$, where C_0 is the self-capacity of the coil and C is the capacity of the tuning condenser. With the small values of the self-capacity found in single layer coils and the usual large values of the tuning capacity, this factor is not important. Thus, in these measurements, C_0 did not exceed 5 $\mu\mu$f and the correction factor does not decrease the resistance by more than two per cent. If the coil is to be used in series with the coupling coil, the effective resistance, R_a, is found from the true high-frequency resistance, R, by the well known equation

$$R_a = \frac{R}{[1 - \omega^2 L C_0]^2 + \omega^2 C_0^2 R^2}.$$

For a coil of 100 mh, with $C_0 = 5$ $\mu\mu$f, the effective resistance at 1500 kc is 8 per cent less than the actual. However, at a frequency of the order of the natural frequency of the coil, R_a must become very large and the coil would act as a trap.

In the following figures the coiling effect, i.e., the ratio of the coil's high-frequency resistance to the high-frequency resistance of the same wire stretched out straight, is plotted versus the frequency in kilocycles on the right of each sheet. The curves shown on the left are rearrangements of the same data showing the coiling effect

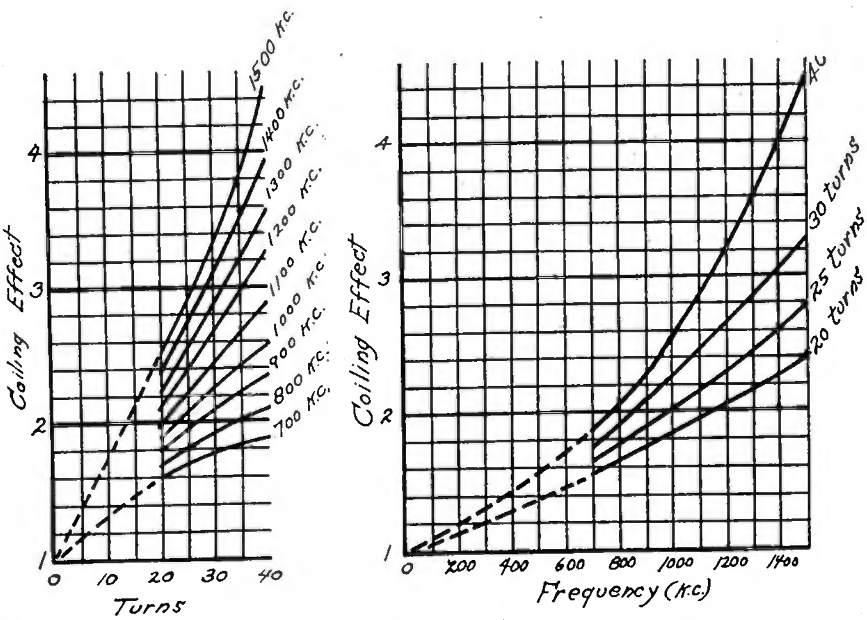


Note – $X = 2\sqrt{\lambda}$ and $\lambda = \omega\pi\alpha^2/1720$

Fig. 2

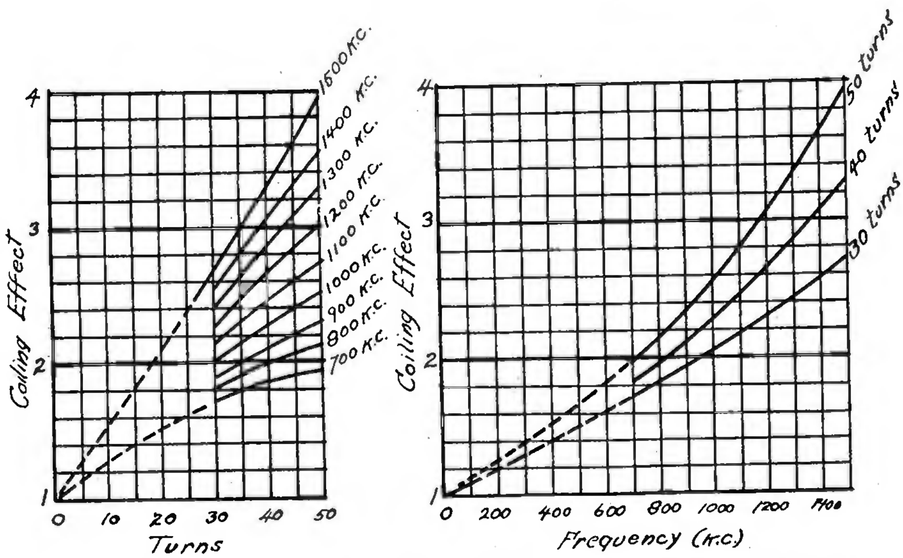
versus the number of turns at a fixed frequency. These latter curves are helpful in extrapolating for particular coils desired. All the curves shown on Figs. 4 to 9 are data curves.

These curves show some interesting facts about the coiling effect. In Fig. 4, for 20-mil wire the coiling effect increases more rapidly than for larger sizes of wire. As the wire increases in size the coiling effect increases less than linearly with the number of turns, although



20-MIL WIRE 3.82-IN. COIL DIAMETER. 33.6 TURNS PER INCH.

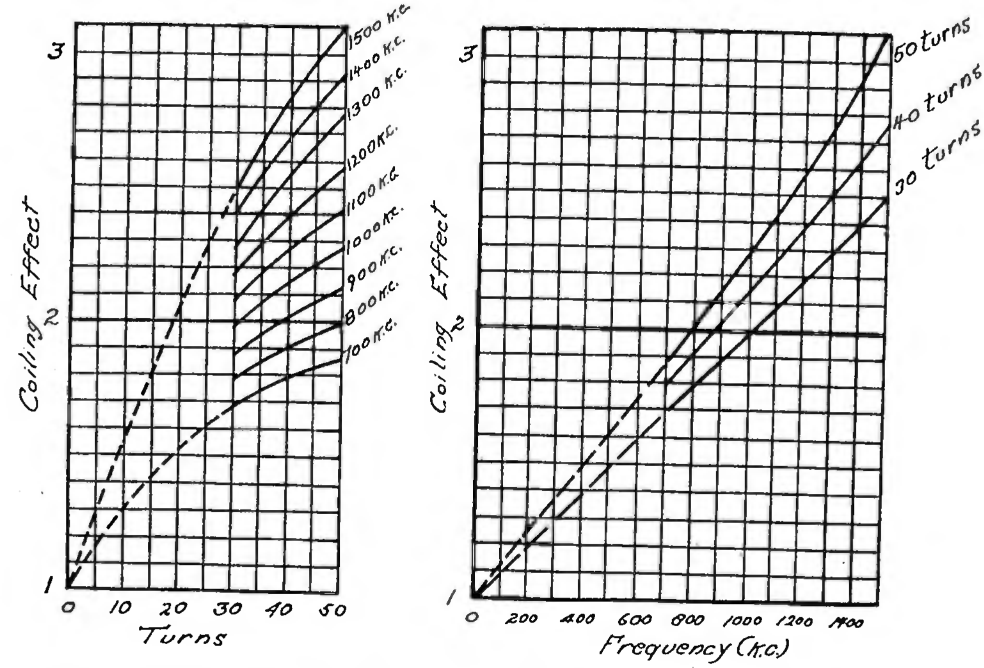
RESISTANCE RATIO FOR 20-MIL STRAIGHT WIRE	1.83	1.98	2.1	2.2	2.3	2.39	2.47	2.55	2.64
FREQUENCY (KC)	700	800	900	1000	1100	1200	1300	1400	1500



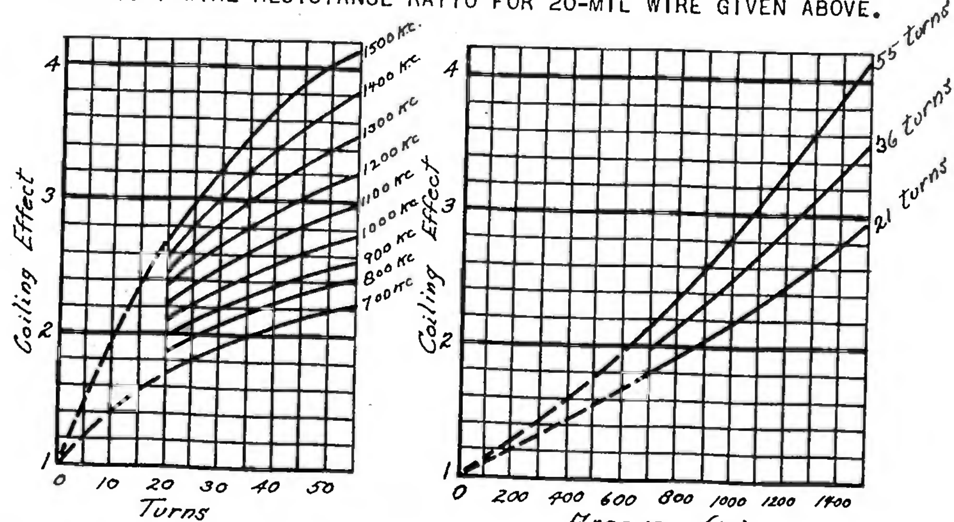
20-MIL WIRE 3.27 IN.-COIL DIAMETER. 33.6 TURNS PER INCH.
RESISTANCE RATIO FOR 20-MIL WIRE (STRAIGHT) GIVEN ABOVE.

Fig. 3

it is true that at a given frequency, two coils that are the same except for the size of wire will have the resistance ratio greater for the coil having the larger wire. It is also interesting to note that the coiling effect increases almost linearly with the number of turns for 20-mil wire at a fixed frequency of about 1,100,000 cycles.



20-MIL WIRE 2.68-IN. COIL DIAMETER. 33.6 TURNS PER INCH.
STRAIGHT WIRE RESISTANCE RATIO FOR 20-MIL WIRE GIVEN ABOVE.



32-MIL WIRE 3.82-IN. COIL DIAMETER. 22 TURNS PER INCH.

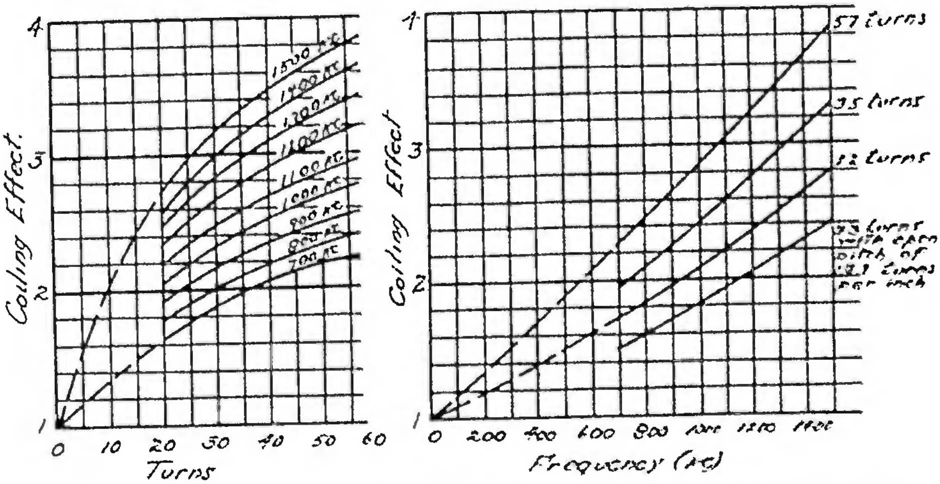
RESISTANCE RATIO FOR 32-MIL STRAIGHT WIRE	2.84	3.02	3.18	3.34	3.49	3.64	3.74	3.82	4.04
FREQUENCY (KC)	700	800	900	1000	1100	1200	1300	1400	1500

Fig. 4

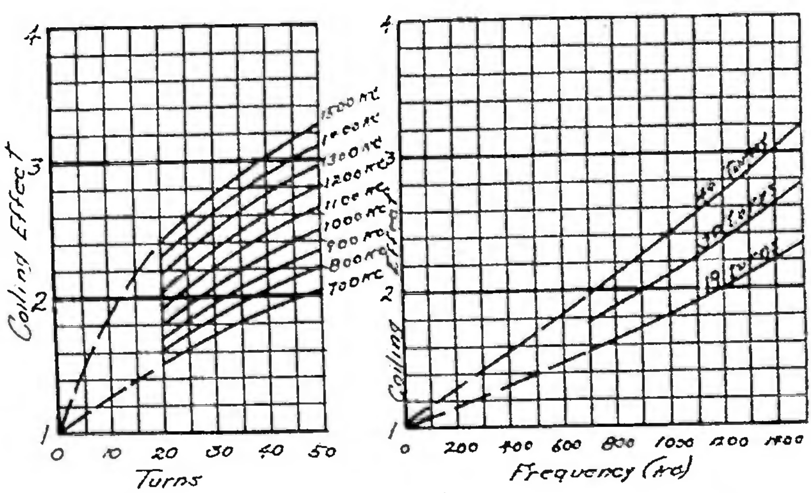
By comparison with some specific values on the curves it can be seen that the important parameters are: size of wire, number of turns, diameter of winding, and pitch of winding.

THEORETICAL

Curtis[7] sets up an integral equation for the current density to determine the resistance ratio of two parallel conductors. He considers the conductor to be made up of infinitesimal filaments. He omits the



32-MIL WIRE 3.27-IN. COIL DIAMETER. 22 TURNS PER INCH (CLOSED PITCH)
RESISTANCE RATIO FOR 32-MIL STRAIGHT WIRE GIVEN ABOVE.



32-MIL WIRE 2.68-IN. COIL DIAMETER. 22 TURNS PER INCH.
RESISTANCE RATIO FOR 32-MIL STRAIGHT WIRE GIVEN ABOVE.

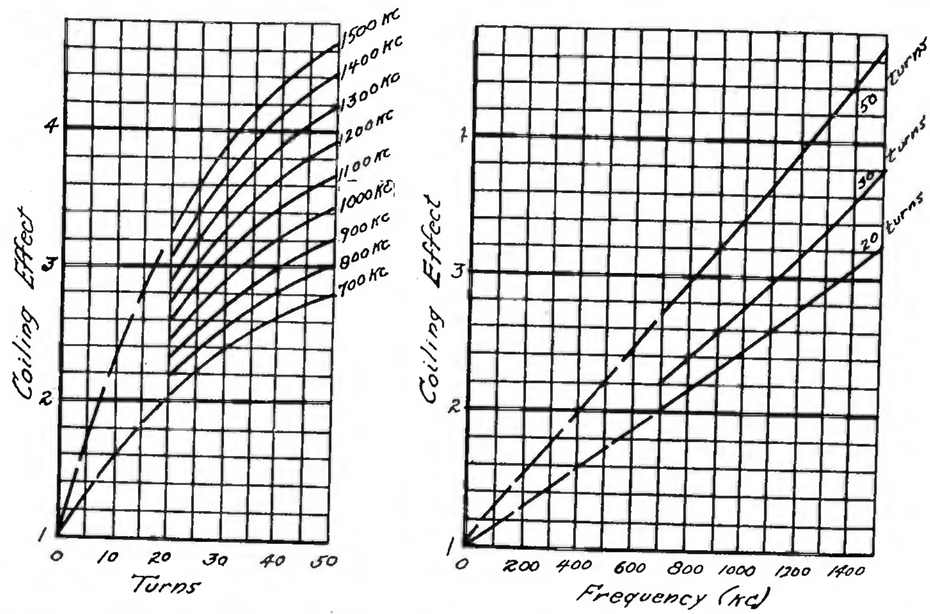
Fig. 5

voltage consumed by self-induction in his integral equation since the current and consequently the flux of the infinitesimal element, will, in the limit, approach zero. He equates like coefficients in two different equations; the one equation being his integral equation and the other

[7] See footnote 7.

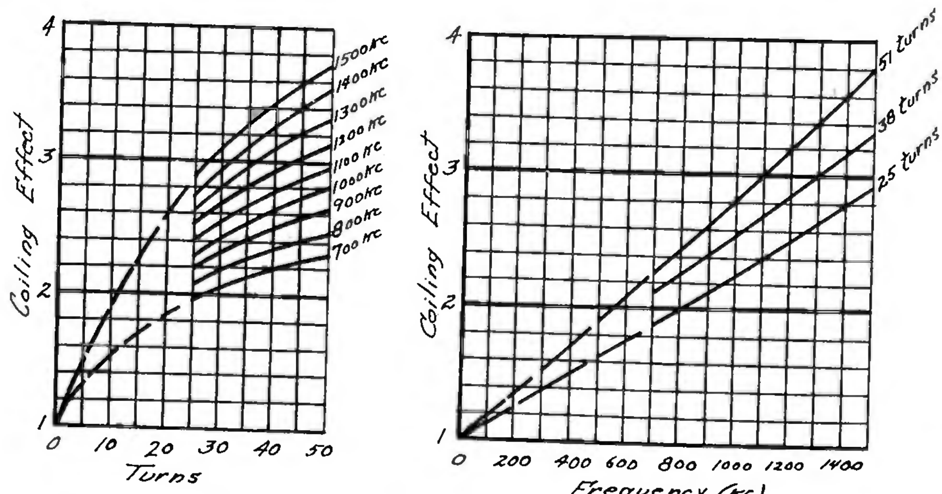
being a Fourier expansion of the sinusoidal current density function over the cross section of the wire.

Hickman[16] adapts Curtis' method to two parallel floors of wires, a



40-MIL WIRE 3.82-IN. COIL DIAMETER. 19.6 TURNS PER INCH.

RESISTANCE RATIO 40-MIL STRAIGHT WIRE	3.48	3.69	3.82	4.11	4.33	4.49	4.66	4.82	4.97
FREQUENCY (KC)	700	800	900	1000	1100	1200	1300	1400	1500



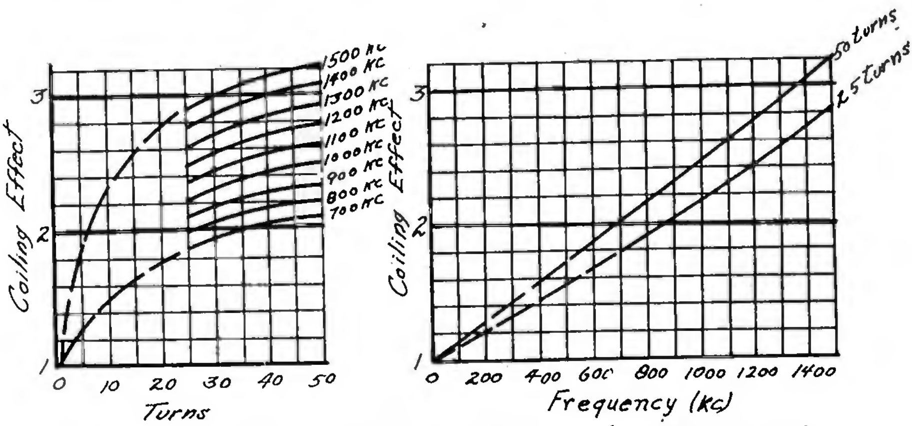
40-MIL WIRE 3.27-IN. COIL DIAMETER. 19.6 TURNS PER INCH.
RESISTANCE RATIO FOR 40-MIL STRAIGHT WIRE GIVEN ABOVE.

Fig. 6

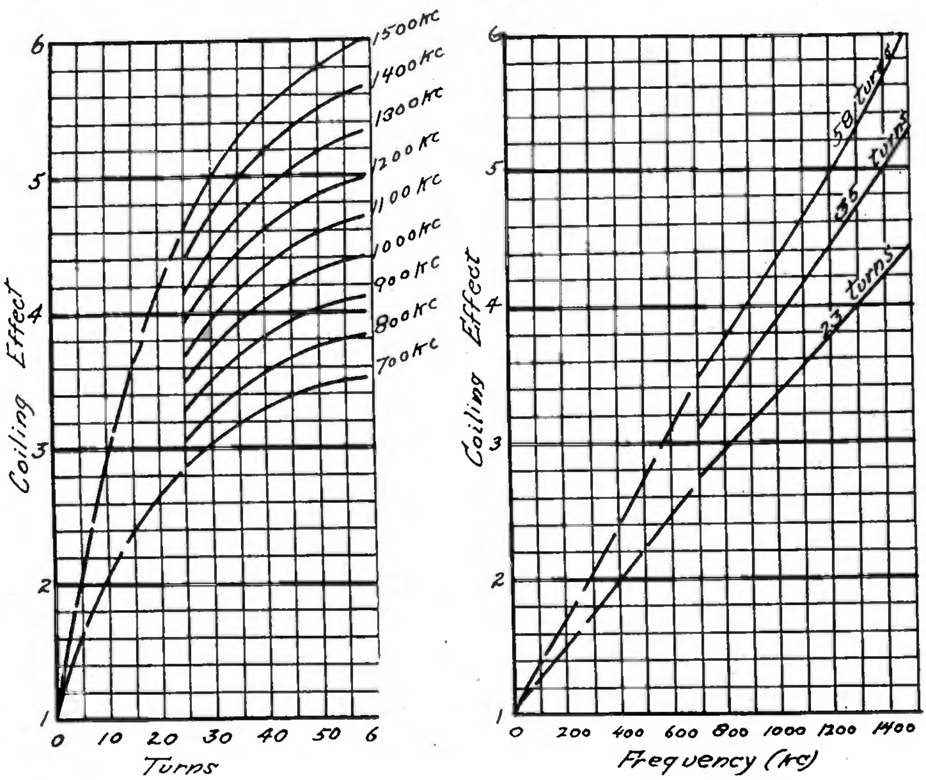
go and return circuit, and makes the distance between floors such that the mutual inductance of any two pairs of wires of the parallel floors

[16] See footnote 16.

is the same as that of two corresponding turns of the coil. This treatment allows Hickman to assume the flux distribution of the coil to be the same as that in the parallel floor system.



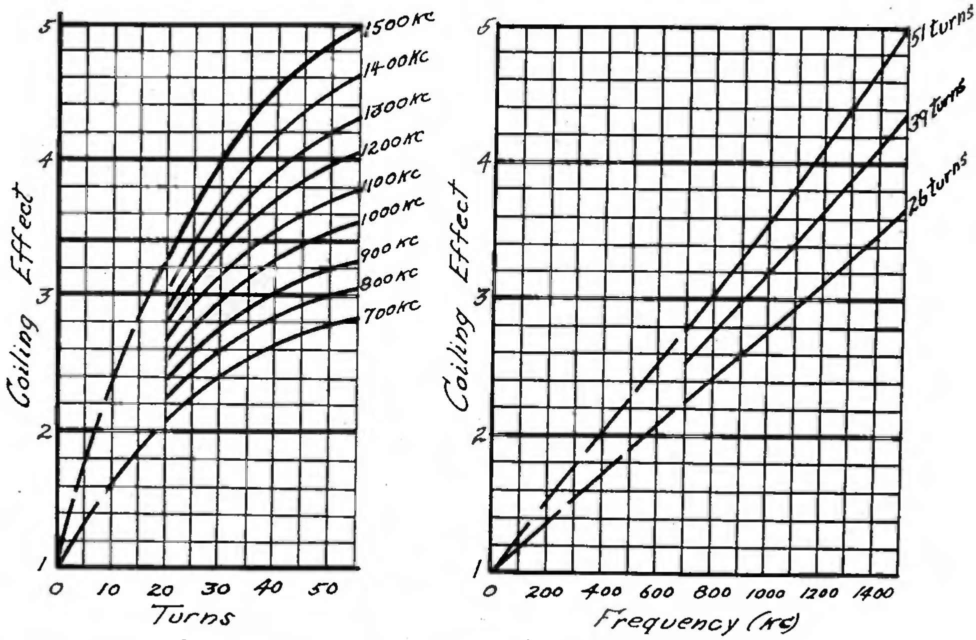
40-MIL WIRE 2.68-IN. COIL DIAMETER. 19.6 TURNS PER INCH. RESISTANCE RATIO FOR 40-MIL WIRE (STRAIGHT) GIVEN ABOVE.



S.C.C. 40-MIL WIRE ALLOWING 21.7 TURNS PER INCH ON A WINDING DIAMETER OF 3.82 INCHES SHOWING INCREASED COILING EFFECT WITH SMALLER PITCH.

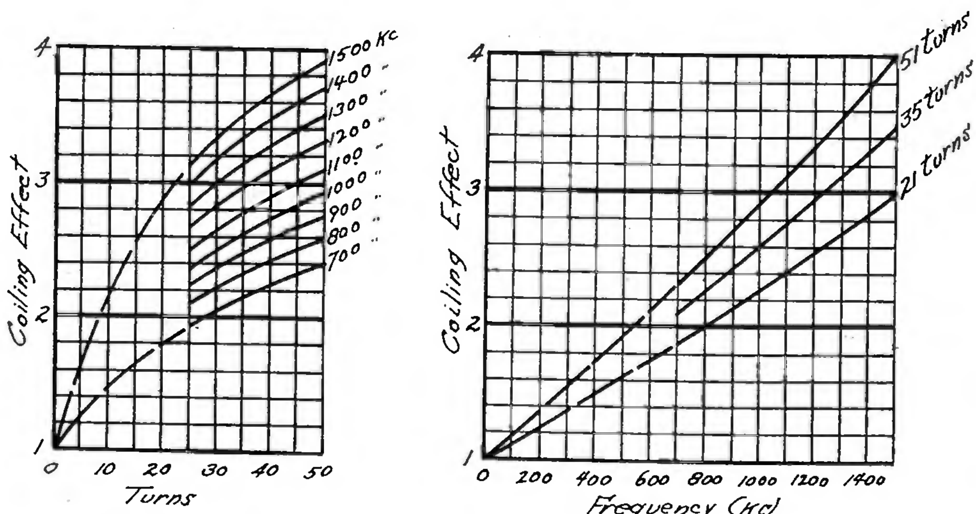
Fig. 7

The following formula obtained by Hickman holds for coils whose lengths are great compared with their diameters and are not applicable to cases where the frequency is greater than a few kilocycles, depending upon the size of wire used. The limit of appli-



51-MIL WIRE 3.82-IN. COIL DIAMETER. 16 TURNS PER INCH.

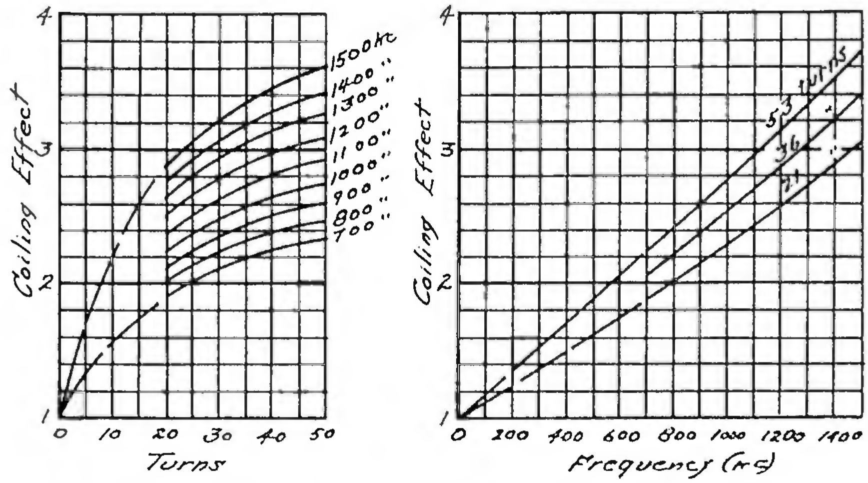
RESISTANCE RATIO 51-MIL STRAIGHT WIRE	4.37	4.66	4.91	5.17	5.4	5.63	5.87	6.07	6.27
FREQUENCY (KC)	700	800	900	1000	1100	1200	1300	1400	1500



51-MIL WIRE 3.27-IN. COIL DIAMETER. 16 TURNS PER INCH.
RESISTANCE RATIO FOR 51-MIL STRAIGHT WIRE GIVEN ABOVE.

Fig. 8

cation is for λ^2 equal to 10, whereas the value of λ^2 in the present paper goes to 5000.



51-MIL WIRE 2.68-IN. COIL DIAMETER. 16 TURNS PER INCH. RESISTANCE RATIO FOR 51-MIL STRAIGHT WIRE GIVEN ABOVE.

Fig. 9

$$\frac{R}{R_0} = U_{0/1} + \frac{G}{U_{0/2} + \frac{\lambda^2 E^2}{U_{0/2}}} + \frac{K}{U_{1/3} + \frac{\lambda^2 F^2}{U_{1/3}}} \tag{1}$$

and the formula following gives the resistance ratio for coils whose lengths are small compared to their diameters:

$$\frac{R}{R_0} = U_{0/1} + \frac{\pi^4 \lambda^2 a^4}{54 s^4 \left\{ U_{1/3}^2 + \frac{\lambda^2 F^2}{U_{1/3}} \right\}} \tag{2}$$

R = the high-frequency resistance.
R_0 = the direct-current resistance.

$$\lambda = \frac{\omega \pi a^2}{1720}; \quad a = \text{radius of the conductor (cm)}$$

s = pitch or the distance between centers of adjacent conductors (cm)
N = number of turns; D = distance between floors or 0.541 times the diameter of the coil winding.

$$K = \frac{\pi^2 \lambda^2 a^4}{9 s^2} \left\{ \frac{\pi^2}{6 s^2} + \frac{4N}{N^2 s^2 + 4D^2} - \frac{4}{s^2 + 4D^2} + \frac{1}{D^2} \right\} \tag{3}$$

$$E = V_{0/2} + \frac{a^2 \pi^2}{6 s^2} - \frac{2a^2 N}{N^2 s^2 + 4D^2} + \frac{2a^2}{s^2 + 4D^2} - \frac{0.5 a^2}{D^2} \tag{4}$$

$$F = V_{1/3} + \frac{\pi^4 a^4}{90 s^4} \tag{5}$$

$$G = 4\lambda^2 a^2 \left[\frac{1}{s} \left(\tan^{-1} \frac{Ns}{2D} - \tan^{-1} \frac{s}{2D} \right) + \frac{0.5}{D} \right]^2. \tag{6}$$

To illustrate the type of function exemplified in the U's and V's one of them is shown below in as brief a form as is possible:

$$U_{0/2} = \frac{\sum_{k=0}^{\infty} \frac{\lambda^{2k}}{\underline{|k}\,\underline{|k+1}\,\underline{|2k+1}}}{\sum_{k=0}^{\infty} \frac{\lambda^{2k}}{\underline{|k}\,\underline{|k+2}\,\underline{|2k+2}}} = \frac{1 + \frac{\lambda^2}{24} + \frac{\lambda^4}{4320} \cdots}{1 + \frac{\lambda^2}{36} + \frac{\lambda^4}{8640} \cdots} \tag{7}$$

These series are uniformly convergent, it is true, but for values of λ^2 greater than 10, the labor of calculation becomes prohibitive, if not impossible, from a practical standpoint. In the present paper λ^2 takes on values as great as 5000.

A. Russell[18] treated Kelvin's ber and bei functions and obtained the well known X, Y, Z, and W functions which are various combinations of the ber and bei functions and of their derivatives. Russell not only obtained convergent series for these functions but also obtained asymptotic series for each of them. To illustrate, the X function is given below both in its convergent form (8) and in its asymptotic form (9):

$$X = 1 + \frac{1}{2}\left(\frac{x}{2}\right)^4 + \frac{1}{4\,\underline{|4}}\left(\frac{x}{2}\right)^8 + \frac{1}{36\,\underline{|6}}\left(\frac{x}{2}\right)^{12} + \frac{1}{64\,\underline{|9}}\left(\frac{x}{2}\right)^{16} \tag{8}$$

In its asymptotic form X function is:

$$X = \frac{\text{exponential } (x\sqrt{2} + 1/(4\sqrt{2}x)}{2\pi x}. \tag{9}$$

In both (8) and (9) $x = 2\sqrt{\lambda}$.

The asymptotic form of Russell's Z function is:

$$Z = X\left(0.707 - \frac{0.5}{x} - \frac{1}{11.312x^2}\right). \tag{10}$$

The first step in the theoretical work of the present paper was to find those combinations of the convergent forms of Russell's functions that would express each of Hickman's U and V functions. The next step was to carry each of these expressions, thus found, over into their asymptotic forms. This second step involved the integration of an

[18] A. Russell, "Effective resistance and inductance of a concentric main," *Phil Mag.*, **17**, 524, 1909.

exponential of high order; the work involved integration by parts. The results obtained are as follows and these expressions enable the calculation of the U and V functions to be carried out in the high-frequency region with comparative ease:

$$U_{0/1} = 0.3535x + 0.25 + \frac{0.1325}{x} \tag{11}$$

$$U_{0/2} = 0.1767x\left\{1 + \frac{2.018}{x} + \frac{1.6384}{x^2} - \frac{6.717}{x^3}\right\} \tag{12}$$

$$U_{1/3} = 0.1178x + 0.3748 + \frac{2.633}{x} - \frac{15.91}{x^2} \tag{13}$$

$$V_{0/1} = \frac{\sqrt{2}}{x} - \frac{0.5305}{x^3} - \frac{0.75}{x^4} \tag{14}$$

$V_{0/1}$ does not occur in the resistance ratio but does occur in the inductance ratio which is outside the scope of the present paper.

$$V_{0/2} = 0.5 + 0.707x - \frac{4.007}{x^3}. \tag{15}$$

$$V_{1/3} = \frac{1}{6} + \frac{0.4701}{x} - \frac{4.28}{x^3}. \tag{16}$$

The curves for expressions (11) and (16) inclusive are shown on sheets 1 and 2. Below there is given a table of values for the same.

The values of the six U and V functions for x ranging from 4 to 20 are given in the following table:

x	$U_{0/1}$	$U_{0/2}$	$U_{1/3}$	$V_{0/1}$	$V_{0/2}$	$V_{1/3}$
4	1.670	1.24	1.07	0.342	0.646	0.24
6	2.375	1.55	1.22	0.233	0.596	0.22
8	3.081	1.86	1.41	0.172	0.582	0.212
10	3.787	2.17	1.63	0.141	0.568	0.208
12	4.494	2.55	1.82	0.118	0.555	0.200
14	5.200	2.87	2.09	0.101	0.545	0.192
16	5.907	3.25	2.35	0.090	0.536	0.190
18	6.614	3.56	2.65	0.080	0.528	0.1898
20	7.321	3.87	2.92	0.071	0.525	0.1892

Formula (1) gives the resistance ratio for a coil whose length is great compared with its diameter, and (2) gives the resistance ratio for a coil whose length is small compared with its diameter. The second term of (2) represents end effect. The coils treated in this paper have, for the most part, lengths of the order of their diameters. Therefore, not only their lengths of winding but also their end effects must be considered in any formula that is to give their net resistance ratio. Furthermore the effect of their end turns become more pronounced as the frequency is increased. For a very short coil, or a single turn of wire, the end effect is given by (2), second term, having recourse to the asymptotic series for high-frequency effects. This second term gives an end effect proportional to the square root of the frequency as the frequency becomes very great. However, as is shown by Butterworth, page 78 of the reference given above, with increase of frequency, the turns farther in from the end become more and more important in adding to the effective resistance. Thus the end effect of a coil which is not very short compared with its diameter would increase more rapidly with the frequency than indicated by the second term of (2). Formula (A), in which the frequency appears also an additional direct factor in the numerator in an attempt to combine empirically (1) and the second term of (2) for coils of intermediate lengths. This formula represents the measured values with a fair degree of accuracy as is illustrated below.

$$\frac{R}{R_0} = U_{0/1} + \frac{G}{U_{0/2} + \frac{\lambda^2 E^2}{U_{0/2}}} + \frac{K}{U_{1/3} + \frac{\lambda^2 F^2}{U_{1/3}}} + \frac{f x^4 \lambda^2 a^4}{10^5 59.4 \left[U_{1/3}^2 + \frac{\lambda^2 F^2}{U_{1/3}} \right] s^4}. \quad \text{(A)}$$

In the above formula f is the frequency in cycles and 59.4 is 54 times 1.1.

Two coils were tested by the Bureau of Standards and their measured and calculated values are given below. For convenience these coils are called Coil A and Coil B.

Coil A had the following data:

$$f = 685{,}000 \text{ cycles}, \quad N = 45 \text{ turns}, \quad a = 0.0648, \quad D = 5.24, \quad \lambda = 32.9,$$
$$x = 11.47, \quad s = 0.162$$
$$U_{0/1} = 4.3, \quad U_{0/2} = 2.45, \quad U_{1/3} = 1.77, \quad V_{0/2} = 0.558, \quad V_{1/3} = 0.2$$
$$G = 244, \quad K = 48.8, \quad E = 0.821, \quad F = 0.2277.$$

By calculation from (A) the resistance ratio was 16.47
As measured by the Bureau of Standards the resistance ratio was 15.65

At $f=1{,}375{,}000$:

By calculation from (A) the resistance ratio was 37.36
As measured by the Bureau of Standards the resistance ratio was 37.7

Coil B had the following data:

$$f = 685{,}000 \text{ cycles}, N = 50 \text{ turns}, a = 0.0406, s = 0.1113, D = 3.81,$$

$$\lambda = 12.9, \quad x = 7.2, \quad U_{0/1} = 2.77, \quad U_{0/2} = 1.73, \quad U_{1/3} = 1.32,$$

$$V_{0/2} = 0.584, \quad V_{1/3} = 0.215$$

$$E = 0.801, \quad F = 0.2342, \quad G = 34.6, \quad K = 5.44.$$

As calculated by means of (A) the resistance ratio was 7.69
As measured by the Bureau of Standards the resistance ratio was 7.61

For $f=1{,}375{,}000$ cycles:

As calculated by means of (A) the resistance ratio was 16.93
As measured by the Bureau of Standards the resistance ratio was 15.8

It is understood that all but the geometrical constants are changed with a change in frequency, i.e., the U's and V's, etc., are evaluated for the new frequency.

Conclusion

The experimental results make clear that a theoretical formula for the resistance ratio of a single layer coil must involve not only the frequency, size of wire, and pitch of winding but also the number of turns and the diameter of the winding as well. Of existing formulas only Hickman's include all of the parameters but they are not adapted to the high-frequency range. This difficulty has been removed in the present paper by the development of asymptotic expansions. In view of the fact that Hickman's expressions apply strictly only to very long or to very short coils compared with their diameters, it is to be expected that resistance ratio values measured with coils of intermediate lengths will not agree closely with values calculated from one or the other of Hickman's formulas. The empirical formula (A) given in the present paper which is a combination of the two limiting formulas of Hickman is found to give much closer agreement with the measured values of this paper.

Acknowledgment

Sincere thanks is extended to Dr. E. J. Berg, head of the Electrical Engineering Department of Union College for his inspiration and guidance in the work and to the General Electric Company and to the Bureau of Standards for the loan of apparatus and the corroboration of certain measurements in connection with the work.

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THE DEVELOPMENT OF A VISUAL TYPE OF RADIO RANGE TRANSMITTER HAVING A UNIVERSAL APPLICATION TO THE AIRWAYS*

By

W. E. Jackson† and S. L. Bailey‡

(†Airways Radio Engineer, Department of Commerce, Washington, D. C. ‡Former Airways Associate Radio Engineer, Department of Commerce, Detroit, Michigan. Now, Jansky and Bailey, Washington, D. C.)

Summary—*This paper deals with the development of a visual type of radio range which has universal application to the civil airways of the United States. Following a discussion of the relative merits of the aural and visual systems of course indication, the theory of the production of twelve courses by utilizing a three-phase radio-frequency source is presented, followed by a general description of the transmitter. The necessary requisites of goniometer design are touched upon, and performance curves of the final goniometer illustrate the effect of these factors on the results. The detrimental effects of cross-couplings between the loop antennas are pointed out, and methods given whereby these couplings may be eliminated. Further discussion covers the advantages of neutralization in the transmitter, the relative merits of various types of interstage coupling, and the method used to obtain and determine a three-phase radio-frequency supply.*

A description of the installation includes a discussion of the characteristics of the loop antennas used for transmission, together with a measured field intensity curve. A course indicator is developed which is useful in aligning and monitoring the courses. Polar space patterns of reed amplitude show the characteristics of the courses obtained with the two-, four-, and twelve-course ranges.

The installation is designed to give continuous operation with a minimum number of interruptions due to failure of the apparatus, a condition necessary to the establishment of an aid to air navigation.

Introduction

Aviation, most rapidly growing of the industries arising out of the present demand for speed of transportation and communication, depends for its successful establishment on safety of operation and maintenance of scheduled trips. One of the largest single factors in the ultimate attainment of both of these requirements is the use of radio as an aid to air navigation, both as a means of supplying weather information and marking out courses along the established highways of the air. The use of the radiotelephone as a means of furnishing weather information to aircraft has been described in a paper by H. J. Walls[1] of the Airways Division of the Department of Commerce.

* Decimal classification: R526.12. Original manuscript received by the Institute, April 28, 1930.

[1] H. J. Walls, "The civil airways and their radio facilities," Proc. I. R. E., 17, 2141–2157; December, 1929.

The development of the directive type of radio beacon, known as the radio range, has been outlined in various papers[2] which cover the fundamental principles underlying the establishment and detection of equisignal zones in space. It is the presence of these zones which provides the courses, regardless of whether the courses are indicated by aural or visual means. The aural method of indication, which presented fewer practical difficulties, underwent rapid development and, consequently, was available when the need for a radio aid to air navigation became apparent. The visual system of course indication has been developed sufficiently to warrant a trial under practical conditions which would bring out the relative merits of the two systems.

A comparison of the advantages and disadvantages of the aural and visual radio ranges follows:

AURAL

Advantages:

1. Simplicity of ground equipment insures reliability of operation and ease of maintenance.
2. Installation and maintenance is more economical than visual system.
3. Location of courses is not dependent upon transmitting tube characteristics.

Disadvantages:

1. Logarithmic characteristic of the ear tends to broaden the course.
2. Presence of static obliterates signal and reduces the useful range.
3. Requires conscious effort to determine side off course.
4. Apparent course width varies widely with different observers depending upon their experience.
5. Difficulty of maintaining a perfectly interlocked signal at the transmitter.

VISUAL

Advantages:

1. Reed indicator provides an optical comparison of the amplitudes of the received signals.
2. Deflection of the reeds is a linear function of the signal voltage.
3. Operates satisfactorily with high interference level, due to the audio selectivity.
4. Longest reed always indicates side off course, making guidance instinctive.

Disadvantages:

1. Complication of ground equipment requires more constant attention by trained personnel.
2. Cost of installation and maintenance is relatively higher than in the aural system.
3. Loss of emission of tubes in any one channel will change position of courses.

While the above tabulation does not include all of the possible advantages and disadvantages of the two systems, it serves to illustrate the major points of comparison between them.

One of the first problems which arises with any radio range system is that of fitting the available courses to the established airways. Vari-

[2] See bibliography, 2.

ous methods whereby the courses of a simple aural range transmitter may be "bent" or shifted to fit the requirements have been described by Jackson and Kear.[3] A similar study of course bending as applied to the four-course double-modulation (visual) type range has been made by Diamond.[4]

The maximum number of courses which may be served under any circumstance with either of these systems is limited by the number of existing courses, which in both cases is four. There are numerous locations, however, where it is impossible to apply all of the four courses to existing airways due to practical limitations which are encountered in course bending. There are also many cases where there are eight or ten courses emanating from a single airport. The existence of such a condition would require several complete installations of the four-course type, each operating on a different radio-frequency channel. Such a system would be economically impractical, wasteful of frequency channels, and productive of severe interference in the vicinity of the airport.

The need for a multicourse radio range was realized at an early date. Original development work on this project was carried on at College Park, Maryland, by the Bureau of Standards in coöperation with the Airways Division of the Department of Commerce. The results obtained during these tests were considered satisfactory enough to warrant the design of a polydirectional radio range which would meet the rigid requirements of the Airways Division. The theoretical and practical problems arising in the design of such a transmitter are discussed in this paper.

Theory of the Polydirectional Radio Range

The problem of producing two or four courses by means of the double-modulation system has been thoroughly discussed by various writers.[5] It should be noted that the four-course range is made possible by displacing the carriers of the two modulation channels ninety degrees (time phase). This, in effect, establishes a rotating field in space, the instantaneous pattern of the field being a figure-of-eight. Modulation of each channel by a separate audio frequency produces side bands having figure-of-eight fields fixed in space and separated by ninety degrees. These side-band fields beat with the carrier to produce in the receiver audio-frequency components of the original modulation frequencies. There are four angular positions about the transmitter at

[3] Jackson and Kear, "Applying the radio range to the airways," Proc. I. R. E., 17, 2268–2282; December, 1929.

[4] Diamond, "Applying the visual double-modulation type radio range to the airways," Proc. I. R. E., 17, 2158–2184; December, 1929.

[5] See bibliography, 6 and 7.

which the received signals from the two modulation channels are equal, thereby establishing four courses.

A full understanding of the principles outlined briefly in the preceding paragraph leads to the conclusion that a greater number of courses can be obtained by introducing a third channel in which a carrier, differing in time phase from the other channels, is modulated by a third audio frequency. It is logical to assume that in a symmetrical system the three carriers should be displaced 120 degrees in time

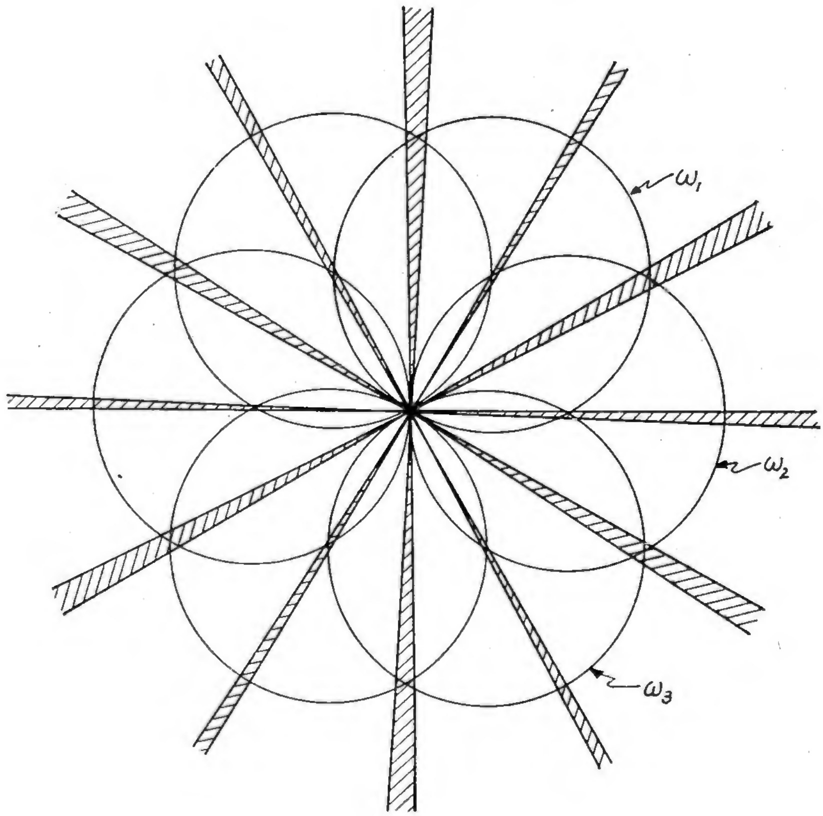


Fig. 1—Theoretical space pattern of side band fields of the triple modulation radio range.

phase. The production of a field having the characteristics of a uniformly rotating figure-of-eight, as in the four-course range, would then require that the fields due to the three carriers be displaced in space by 120 degrees. This is accomplished by a corresponding displacement of the goniometer primaries. In this case, the three side-band fields appear in space as figures-of-eight fixed at 120 degrees. Fig. 1 shows the position of these side-band fields and indicates that there will be twelve courses established, six "broad" and six "sharp." The sharp courses will have less amplitude and, therefore, less range than the broad courses.

The following section describes a transmitter which has been designed by the Airways Division for use either as a twelve-course "polydirectional" radio range or as a simple two- or four-course range.

General Description of Transmitter

A description of the transmitter is given here in order to present a clear view of the equipment to be discussed in this paper. A general view of the completed transmitter with its associated goniometer and antenna tuning unit is shown in Fig. 2, while Fig. 3 shows the constructional details of the transmitter.

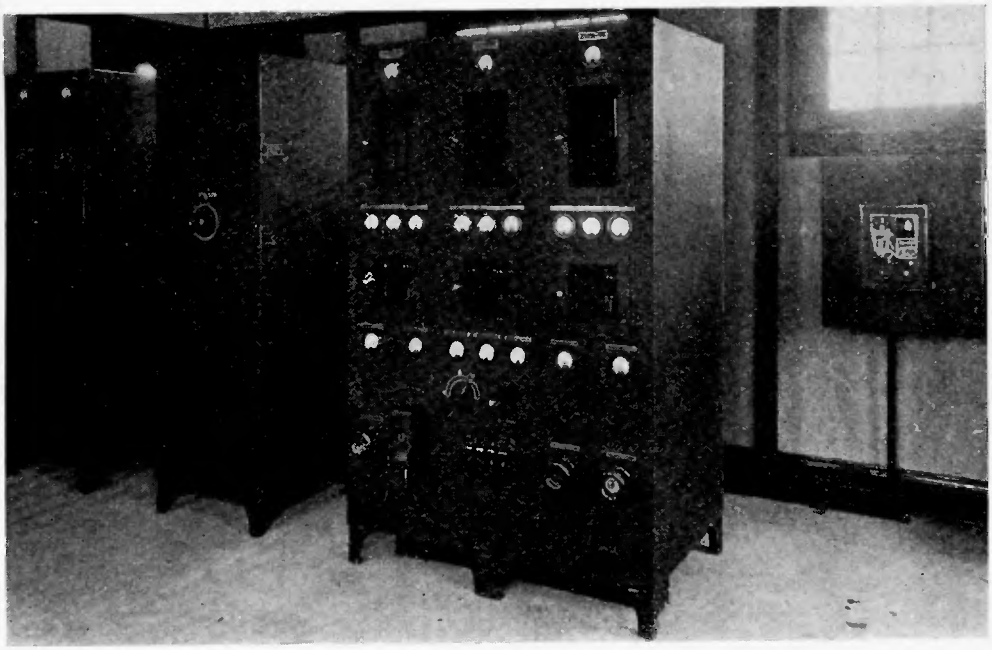


Fig. 2—General view of transmitter installation showing modulation frequency controller on the right.

The master oscillator consists of two UX-860 tubes in parallel in a conventional Hartley circuit (Fig. 4). The output of the master oscillator excites three identical chains of radio-frequency amplifiers through a phase splitting arrangement which supplies equal voltages to each chain, the time phase between them being 120 degrees. The first intermediate amplifier in each chain is a single UX-860 tube which is impedance coupled to the grid of a second intermediate amplifier made up of two UX-860 tubes in parallel. These in turn are impedance coupled to the grid of a single UV-851 tube in each channel, which acts as the final power amplifier. The output of each channel feeds into a corresponding goniometer stator; the energy in each stator being coupled independently into two loop antennas in the correct time and

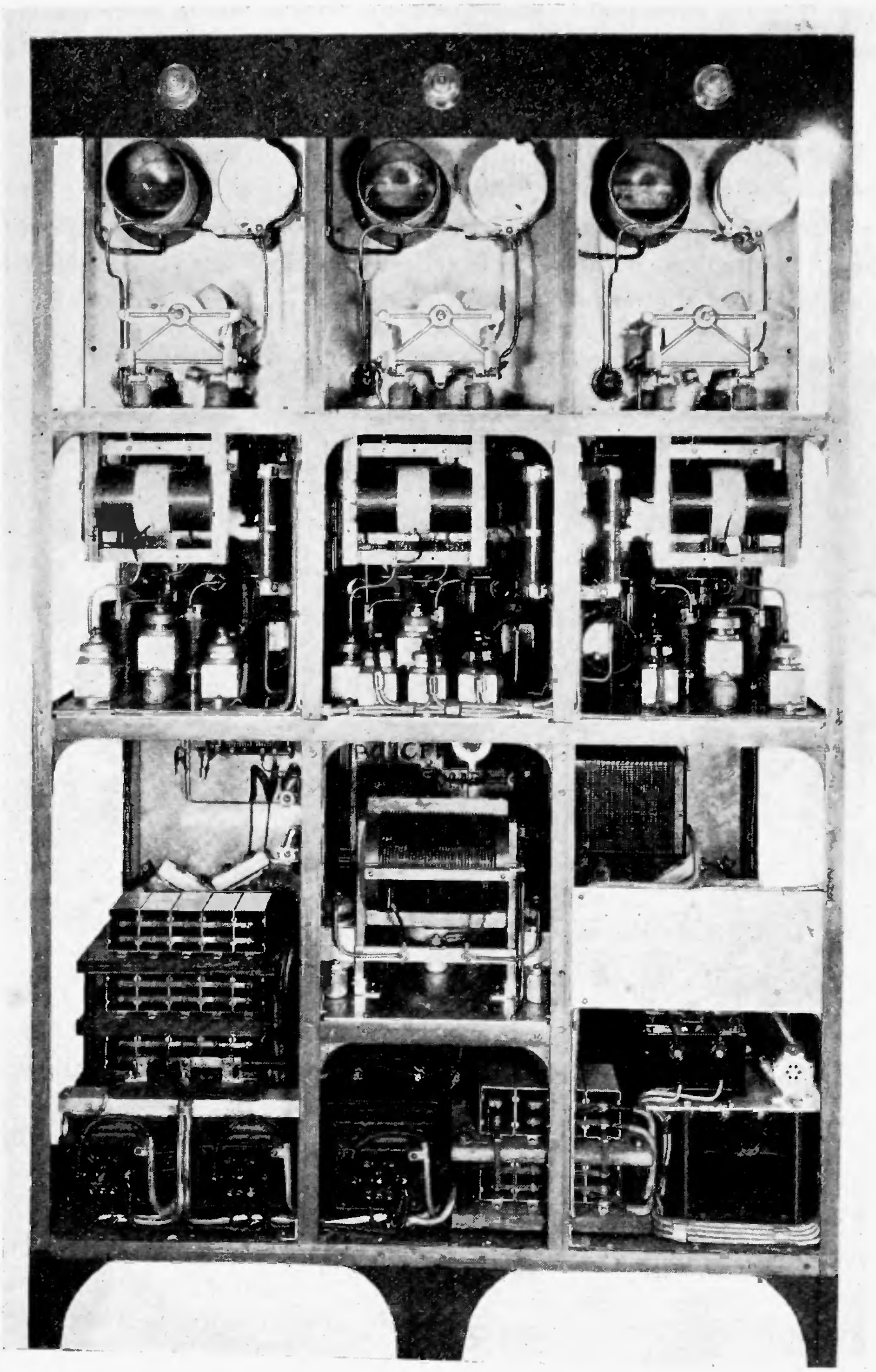


Fig. 3—Rear view of the transmitter.

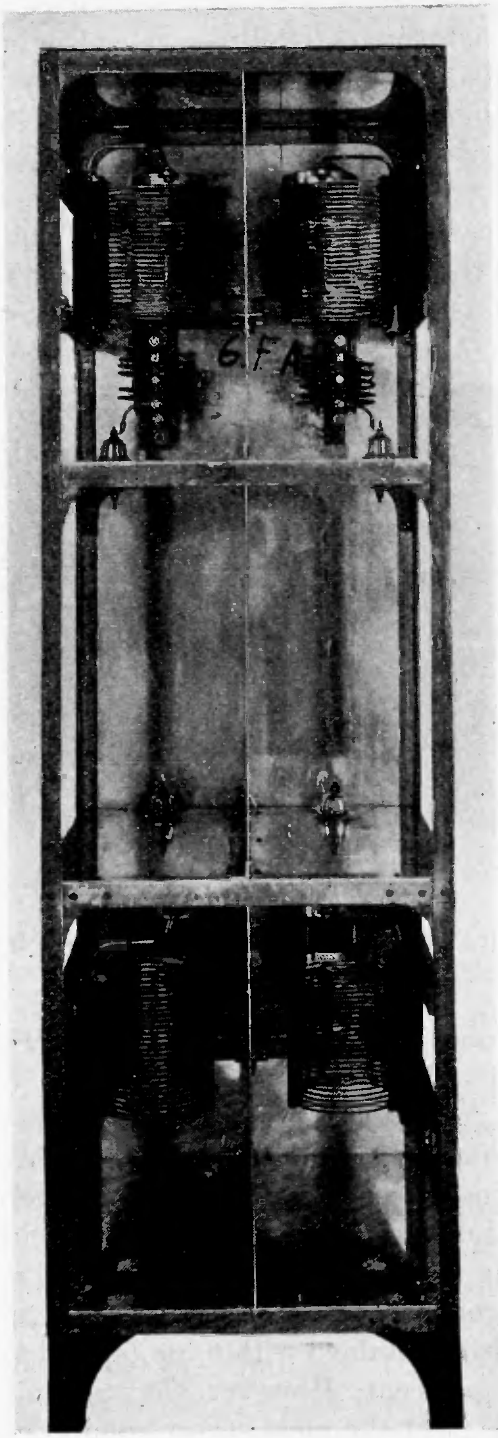


Fig. 5—Rear view of the antenna tuning unit.

space phase. An antenna tuning unit (Fig. 5) is included in the installation for the purpose of tuning the antennas to resonance.

A double commutator 2000-volt d-c generator supplies 1000 volts to the master oscillator plate and 2000 volts to the plates of the first intermediate amplifiers as well as the final power amplifiers. The plate supply of each of the three second intermediate amplifiers is completely modulated at a separate audio frequency (65, 86.7, and 108.3 cycles, respectively). These frequencies are obtained from three alternators driven through a silent chain drive by an inverted rotary converter running as a d-c motor. Both the high voltage generator

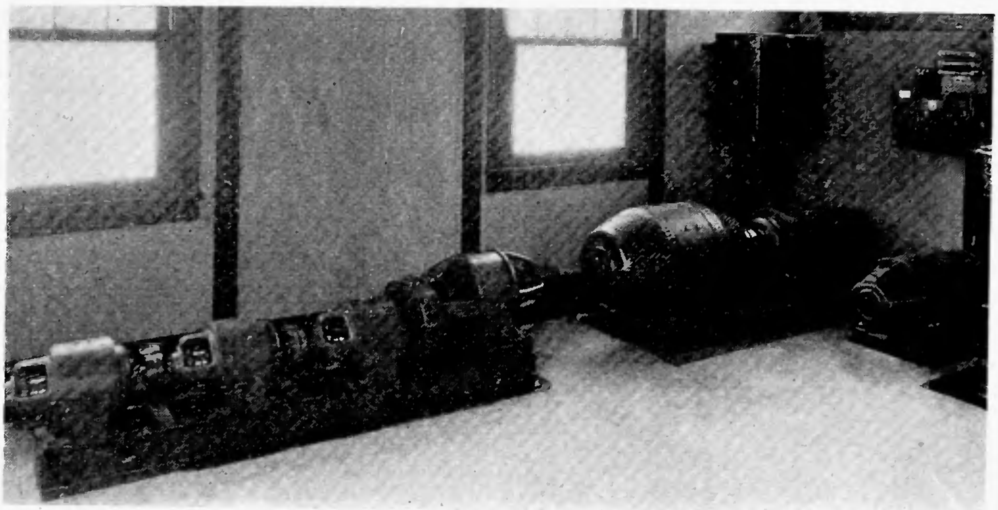


Fig. 6—View showing the high voltage generator, triple frequency machine, and low voltage d-c generator with voltage regulator.

and the triple frequency machine are shown in Fig. 6. The speed of the latter machine (and therefore the frequency of the alternators) is maintained constant by a frequency controller which operates from the a-c winding of the inverted rotary converter and controls a motor driven rheostat in the d-c field of the converter. This method holds the modulation frequencies constant to within one-tenth of one per cent of the required values. The control apparatus may be replaced by synchronous motors in locations where the frequency of the power supply is maintained within limits of plus or minus fifteen-hundredths of one per cent. However, the transmitter in its present form is designed to meet the most severe conditions which are likely to be encountered in the United States. Fig. 7a shows the frequency variation of a large commercial power network, while Fig. 7b indicates the stability obtained when using the frequency controller.

Conditions Governing the Selection of Tubes for the Transmitter

Before going into the details of transmitter development, it may be of interest to explain the reasons for the choice of tubes as outlined in the preceding paragraphs. The following tabulation sums up the requirements considered in the selection:

(1) Tubes should be air-cooled in order to simplify design and eliminate the additional equipment required with water-cooled tubes.

(2) The life of the tubes should be equal to the best tube life obtained in commercial transmitters.

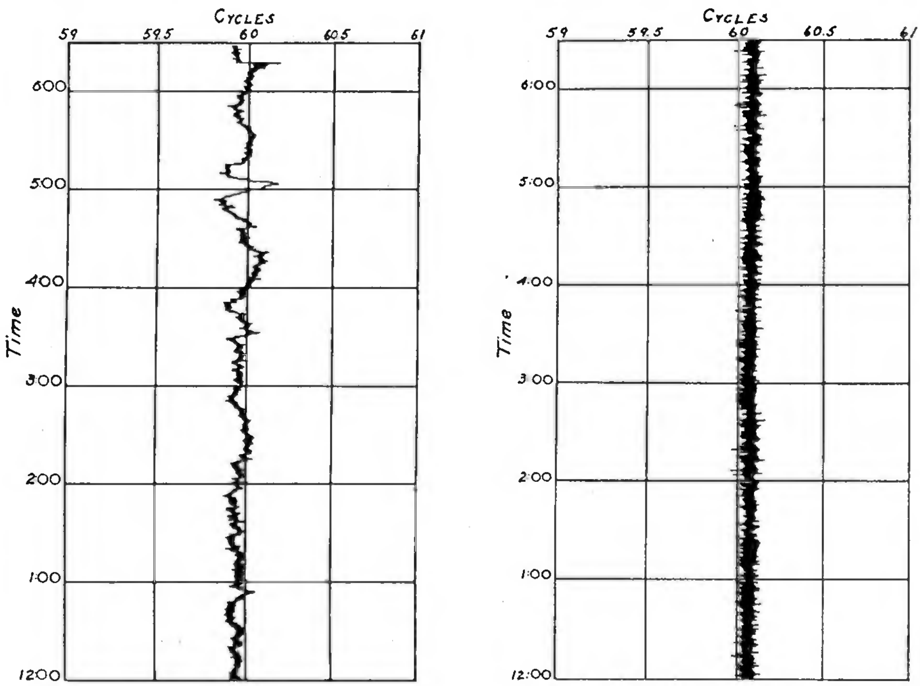


Fig. 7—(a) Frequency variation of commercial power network manually controlled. (b) Frequency variation of sixty cycles obtained from the inverted rotary converter when using frequency controller.

(3) Tubes should be selected which allow the use of the simplest practical circuits.

(4) The number of different types of tubes used should be reduced to a minimum in order to facilitate replacement from stock.

With these conditions in mind, the UX-860 (a 75-watt four-element tube) was selected for all stages except the power amplifiers. The obvious advantage of this tube is that the extremely low grid to plate capacity obtained by the use of the shield grid and associated shielding allows its use as an amplifier without the necessity of neutralization to obtain stable amplification. The UV-851 tube was chosen for use in the power amplifiers because of its exceedingly long life in normal service, as shown by its use in transmitters of the Airways Division, the Navy, and various commercial companies. One disadvantage of this tube is its high grid-to-plate capacity, which requires neutralization for stable amplification. This neutralization is supplied by the Rice method in the transmitter under discussion. Consideration was given to the use of the UV-861 (500-watt four-element tube) in place of the UV-851, but this possibility was discarded because of the short life and lower output rating of the former tube. The higher plate voltage (3000 volts) required by the UV-861 is a disadvantage only from an economic standpoint, since it does not materially increase the technical or mechanical difficulties in design.

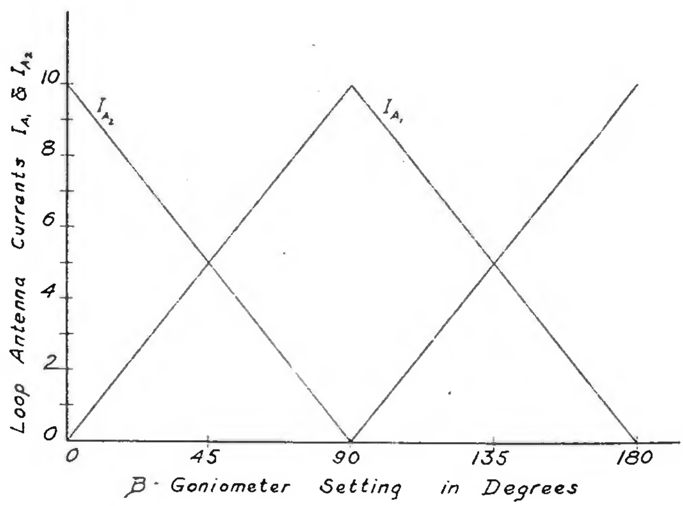


Fig. 8—Linear characteristics of a theoretical goniometer, showing the relation between goniometer setting and antenna current.

In conclusion, it might be stated that if a satisfactory one-kilowatt four-element tube were available having a life comparable to the UV-851, it would undoubtedly be used in place of the three-element tube.

Goniometer Design

There are several difficulties which are encountered in the practical design of any goniometer. These difficulties present themselves in varying degrees, depending upon the particular system in question, i.e., aural system, two-course, four-course, and twelve-course visual systems. Regardless of the system in use there are certain fundamentals which are common to all systems in question. These fundamentals are briefly as follows:

(1) The goniometer should give a sine wave variation in antenna currents when excited by only one stator coil, as the goniometer is rotated through 360 degrees.

(2) Direct magnetic and capacitive coupling between the loops should be reduced to a negligible degree.

(3) Indirect coupling by means of capacity coupling between the stators, direct magnetic coupling between the stators, and indirect coupling between the stators by way of the rotors should be eliminated or reduced to a minimum.

Each of these factors will be discussed in turn. For convenience in observing the effects derived from a nonsinusoidal goniometer characteristic, let it be assumed that the antenna current for each loop is linear with respect to angular rotation (Fig. 8).

The effect of the goniometer is to set up field patterns in space displaced any desired number of degrees from the plane of either loop, thus allowing the courses to be rotated at will. These displaced fields may be considered as being radiated from imaginary or "phantom" loops rotated in space by an angle which is a function of the angular rotation of the goniometer. It may be shown mathematically that the current in the phantom loop equals the square root of the sum of the squares of the currents in the two loops for any position of the goniometer. Also it may be shown that the angular position of the axis of the resultant figure-of-eight field with respect to one loop is equal to the angle whose tangent is the ratio between the two antenna currents for any angular position of the goniometer (β).

Mathematically

$$I_{A_p} = \sqrt{I_{A_1}^2 + I_{A_2}^2}$$

$$\text{and } \theta = \tan^{-1}\frac{I_{A_1}}{I_{A_2}} \quad \text{for any setting of } \beta$$

where I_{A_p} = the current in the phantom loop
I_{A_1} = the current in loop 1.
I_{A_2} = the current in loop 2.
θ = the angle between loop 2 and the axis of the resultant figure-of-eight in space.
β = goniometer setting.

From these two equations and the curves given in Fig. 8 it is possible to plot θ against β and I_{A_p} against β. (See Figs. 9 and 10). These two curves may be used as an indication of the relative merits of any goniometer.

From these results it will be noted that the effect of a goniometer having a linear or nonsinusoidal characteristic is to rotate the phantom figure-of-eight space diagram at a rate different than that indicated by the actual rotation of the goniometer rotors. It will also be noted that the amplitude of the resultant figure-of-eight does not remain a constant but varies depending upon the position of the goniometer rotors with respect to the stator. These two features are particularly

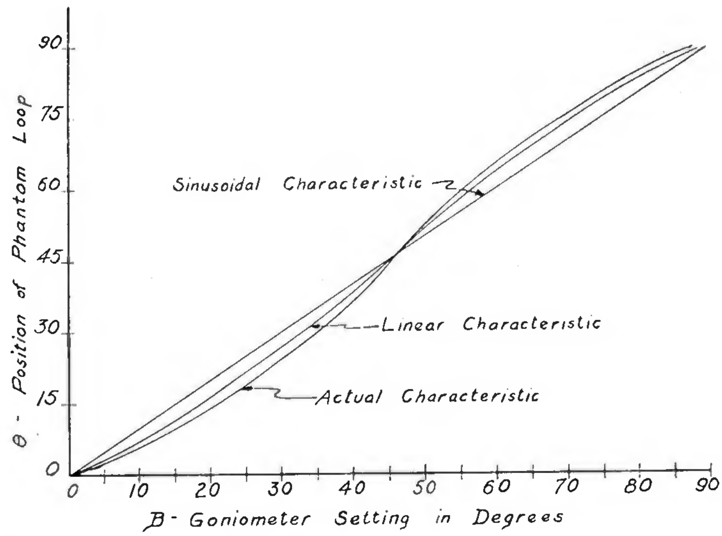


Fig. 9—Effect of the goniometer characteristic on the position of the phantom loop in space.

annoying in that the angles between courses will not remain the same with rotation of the goniometer. It is, therefore, desirable to reduce these effects to a negligible degree by designing the goniometer to follow as closely as possible a sine wave characteristic.

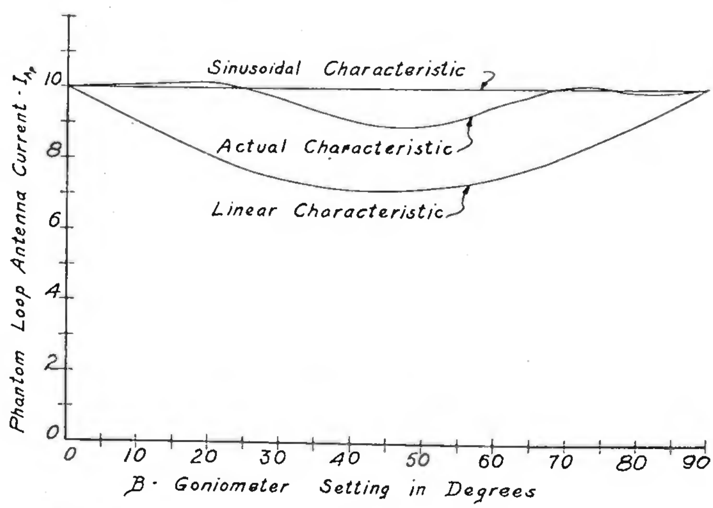


Fig. 10—Effect of the goniometer characteristic on the current in the phantom loop.

Many different goniometers have been made with characteristics of varying types. The final goniometer characteristic adopted is plotted in Fig. 11.

The effects of idiosyncrasies in goniometer design on the position of the courses of the polydirectional system are shown in Fig. 12. It will be noted that the angular position of the courses with respect to each other varies as the goniometer is rotated. The effect is eliminated only when the goniometer has a true sine wave characteristic, a condition difficult to obtain in practice.

Cross-coupling effects may be defined as the effects due to the existence of unwanted couplings between the two loop antennas. Their presence is detrimental to the proper functioning of a polydirectional radio range and due caution must be exercised to elim-

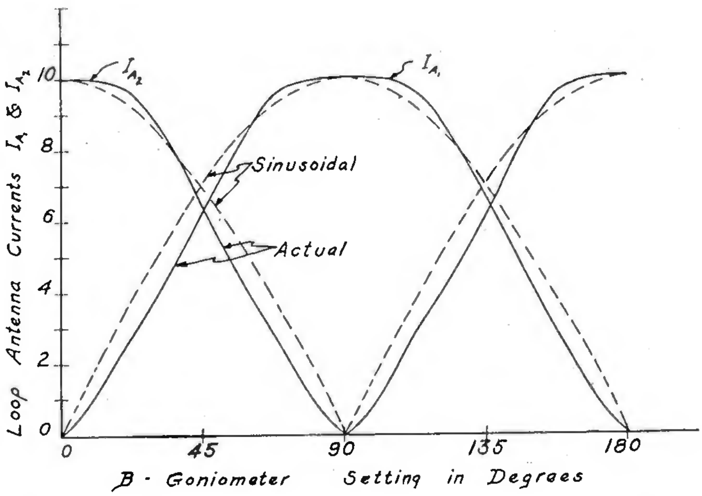


Fig. 11—Final goniometer characteristic showing deviation from theoretically perfect type.

inate them. Direct magnetic coupling between the loop antennas is reduced to a minimum by placing both the loop antennas and associated goniometer rotor coils at right angles to each other, and by shielding the loading coils in one antenna from those in the other antenna. Capacitive coupling between the two loops is minimized by adequate spacing of the loop lead-in wires and connections between the loop tuning unit and the goniometer. The crossed coils in the goniometer rotor are staggered slightly in order to reduce the capacitive coupling between them.

In addition to the direct forms of coupling already discussed, there exist various indirect couplings, the first appearing as coupling between stators by way of the rotors which induces undesired currents in both of the loop antennas. The effect of this type of cross coupling may be reduced considerably by utilizing untuned stator circuits, i.e., circuits having a fundamental outside of the band of frequencies in which the transmitter is to be worked. This method may be utilized under certain conditions, but does not prove to be entirely satisfactory since some cross coupling is always present. However, when the three stator coils are set at 120 degrees to each other, this type of coupling is almost entirely eliminated because the phase and amplitude of the undesired voltages are such that they will cancel each other.

The second of these effects occurs as a result of the direct coupling between stator coils which also has the effect of introducing undesired couplings between the loop antennas. Theoretically, the effects of

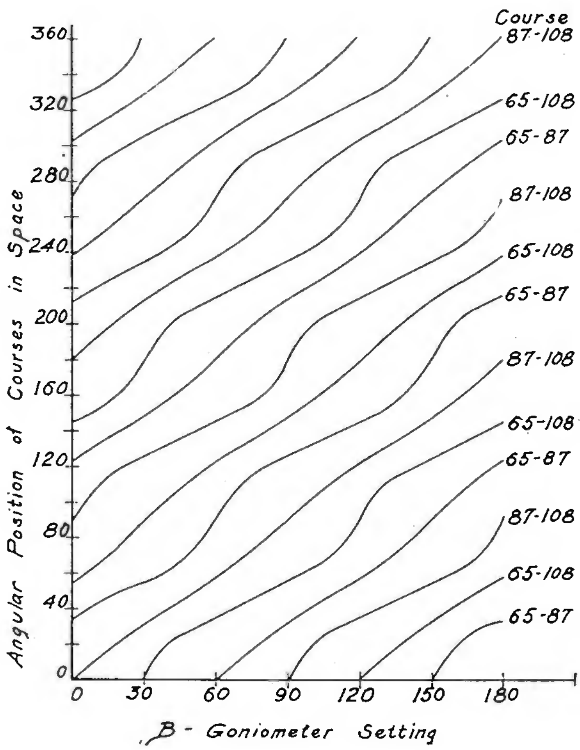


Fig. 12—Effect on course positions caused by deviation of actual goniometer from the perfect sinusoidal characteristic.

these magnetic couplings may be eliminated by coupling the stator coils at 120 degrees to each other in a single compartment. This method was not deemed practical due to difficulties which would be encountered in the electrical design and mechanical construction of such a goniometer. Considerable capacity coupling would also occur between each of the stator coils which would introduce coupling effects rather difficult to eliminate. It was, therefore, decided to simplify the construction of the goniometer and to minimize the magnetic and capacitive coupling effects by placing each stator coil in a separately shielded compartment together with each pair of cross-coil rotors. (Fig. 13).

The completed goniometer embodies all of the fundamentals pre-

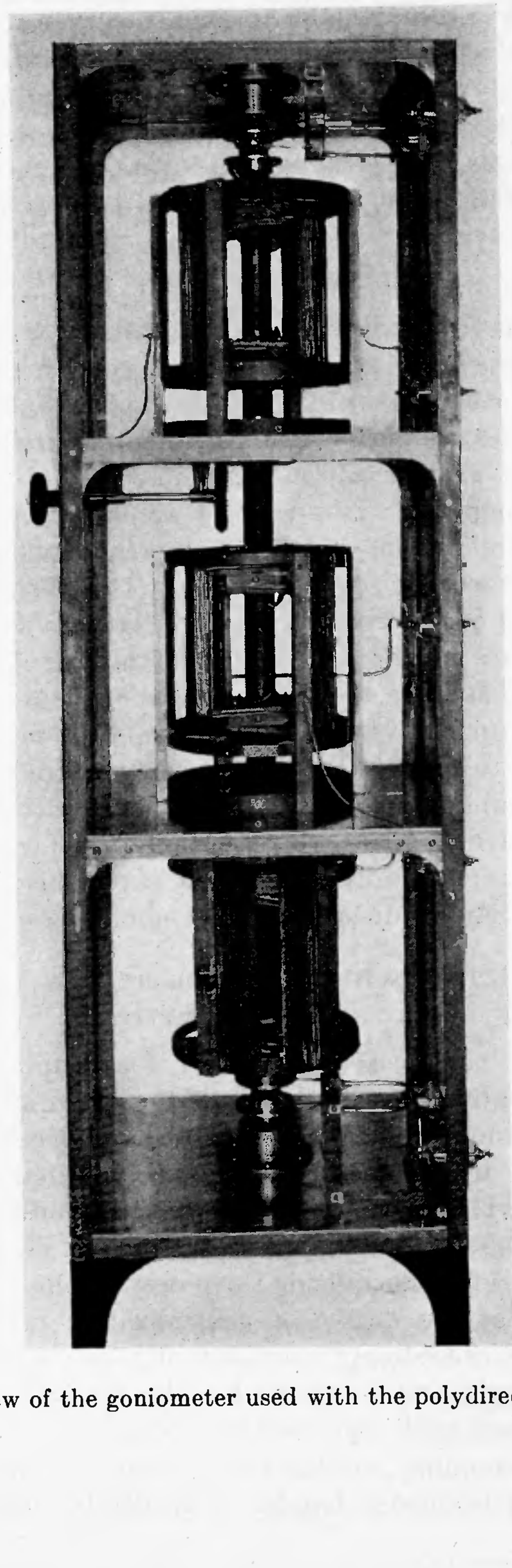


Fig. 13—Side view of the goniometer used with the polydirectional radio range.

viously discussed and although the characteristic obtained is not exactly sinusoidal, the results indicate that the approximation is close enough for satisfactory operation. It has been pointed out that small variations in the characteristic cause the angular spacing between adjacent courses to shift as the goniometer is rotated. Under certain conditions, this may prove to be an advantage rather than a disadvantage, as it will aid in the alignment of courses along established airways.

Neutralization

The two types of indirect coupling already described introduce undesired voltages in the stator coils of the goniometer, which in turn are connected to the plates of the power amplifier tubes. It is evident that if these tubes are not neutralized, the undesired voltages will be fed back to the grid circuit and amplified, thus accentuating the effects of the coupling. This type of amplification is prevented by the use of neutralization, which also increases the stability of the amplifier by preventing self oscillation. The Hazeltine method of neutralizing was considered, but was rejected in favor of the Rice method for the reason that the former method would require a plate neutralizing coil as a part of each goniometer stator, thus complicating the construction. In the Rice method the neutralizing circuit can be confined to the transmitter proper, thus reducing the effects of stray cross coupling between the amplifier branches. This method was finally adopted and has proved entirely satisfactory. The use of neutralization has practically doubled the plate efficiency of the power amplifiers over that obtained in circuits employing no neutralization.

Coupling Between the Intermediate Power Amplifier and the Power Amplifier

Theoretically the best arrangement for coupling between the second intermediate power amplifier and the power amplifier is a tank circuit tuned to the working frequency of the transmitter (Fig. 14a). This circuit may be adjusted to present the optimum output impedance to the intermediate power amplifier tube, and at the same time may be used to obtain a voltage 180 degrees out of phase with that applied to the grid for neutralizing the power amplifier. Development has indicated that the degree of neutralization possible with this type of circuit is exceedingly good, but it requires accurate adjustment for each change in frequency. For this reason it was decided to eliminate the tuned tank and use transformer coupling. (Fig. 14b). This method of coupling was tried and found to work very well for relatively narrow frequency bands. Considerable difficulty was encountered with series and parallel resonant circuit conditions in the power amplifier which gave erratic results in the form of self-oscillation, transient voltages, and unstable operation. However, these conditions were each eliminated and satisfactory results were obtained. It was observed that the output varied appreciably with wide changes in frequency (285 kc–350 kc), and that irregular changes in plate current occurred when tuning the antenna under conditions of low

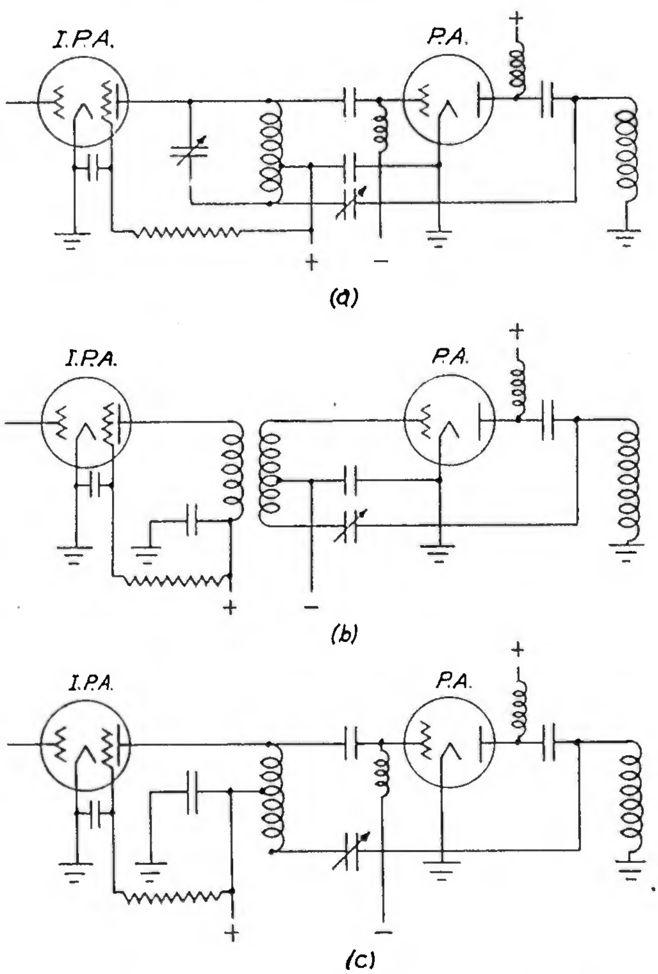


Fig. 14—(a) Schematic diagram of tuned tank coupling. (b) Schematic diagram of transformer coupling. (c) Schematic diagram of impedance coupling.

grid excitation. In order to improve these conditions, it was decided to try impedance coupling. (Fig. 14c). The coupling impedance used was a center tapped triple layer bank wound coil similar in all respects to the usual type of impedance coupling coil except that it included larger wire and extra turns for neutralizing the power amplifier. With this type of coupling many difficulties similar to those encountered when using transformer coupling were observed and eliminated. This system of coupling had the advantage over the transformer coupling in that much more uniform grid excitation and output could be obtained over the relatively wide frequency band in which these transmitters were required to work. The degree of neutralization obtainable with both the transformer and impedance coupling was not quite as complete as that obtained with the tuned tank coupling, but was found to be adequate for all requirements imposed upon it. Incidentally, the result of neutralizing this particular circuit was to reduce the amount of energy fed directly to the antenna by the intermediate power amplifier to approximately one four-hundredth the value obtained when neutralization was omitted. Thus, neutralization has the additional advantage of reducing the load and decreasing the reaction on the intermediate power amplifier.

Method for Obtaining Three-Phase Radio-Frequency Supply

The theory of establishing twelve courses is dependent upon the production of a three-phase radio-frequency supply in which the phases are symmetrically displaced. Due to the lack of accurate

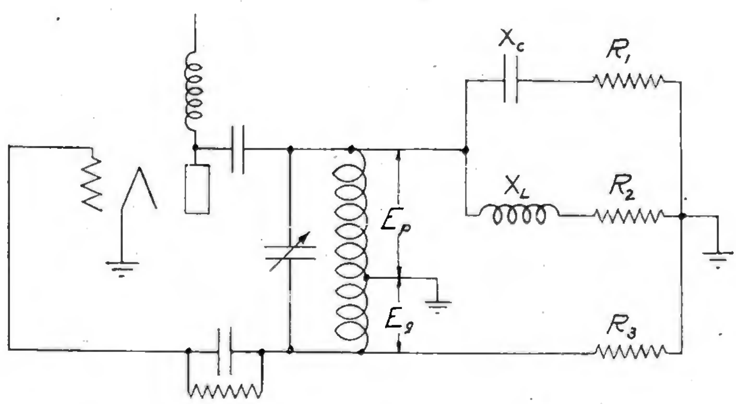


Fig. 15—Diagram showing fundamental circuit utilized in obtaining three-phase radio-frequency supply.

information on this subject, it was necessary to devote considerable time to the development of a satisfactory system. The problems encountered and the final results obtained will be discussed in succeeding paragraphs.

First attempts at the production of a three-phase radio-frequency supply involved the use of a three-phase oscillator utilizing three tubes, the tank circuit of each tube providing a separate phase of the three-phase supply for exciting each audio-frequency channel. This idea was soon discarded because of the difficulty which would probably have been encountered in keeping each of the individual plate circuits tuned to give the same frequency. Furthermore, it was realized that an individual oscillator tube for each phase was not to be desired since different tube parameters would give various outputs and phase relationships, a condition which would be most pronounced in cases of tube failure and would require trained personnel to realign the phases to their normal positions. It was finally decided to make use of the so-called phase splitting idea, utilizing a single master oscillator of the Hartley type (Fig. 15). The plate of this oscillator was supplied by a d-c source in the ordinary manner and the frequency was controlled by the usual tank condenser. Inasmuch as the radio frequency voltage applied to the plate and grid are essentially 180 degrees out of phase, it is obvious that the voltage (E_g) across the grid side of the tank is 180 degrees out of phase with the voltage (E_p)

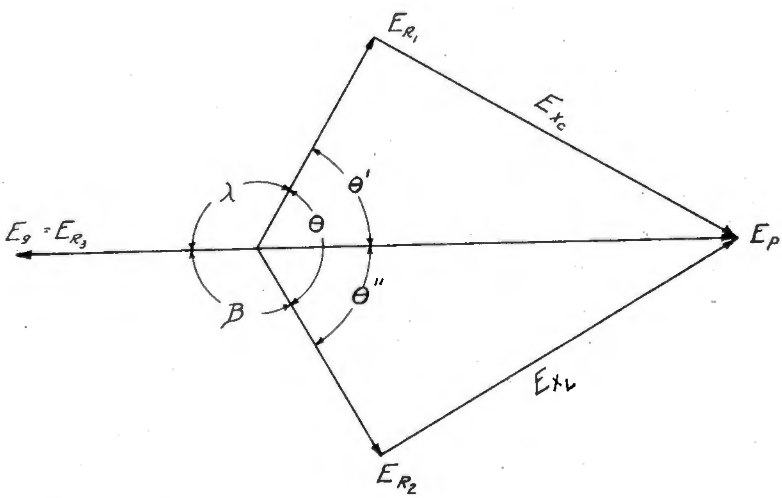


Fig. 16—Vector diagram showing fundamental phase relationships of voltages across R_1, R_2, and R_3 in Fig. 15.

across the plate side of the tank. If the voltage across the plate side of the tank circuit is applied to a condenser and resistance in series, and simultaneously applied to an inductance and resistance in series (Fig. 15), it will be noted that the phase of the voltages across R_1 and R_2 will vary in accordance with the ratio of X_L/R_2 and X_C/R_1, i.e., $\theta' = \tan^{-1} X_L/R_2$, where θ' equals the phase displacement between the applied voltage across X_L, R_2, and the voltage across R_2. In the same way $\theta'' = \tan^{-1} X_c/R_1$ where θ'' equals the phase displacement between the applied voltage across X_C, R_1, and the voltage across R_1 (Fig. 16). It then follows that the phase angle between the voltages applied across R_1 and R_2 (θ) will be equal to $\theta' + \theta''$. Also, if $X_L = 1.73 \times R_2$, $X_C = 1.73 \times R_1$, and $R_1 = R_2$, then $\theta' = 60$ deg., $\theta'' = 60$ deg., $\theta = \theta' + \theta'' = 120$ deg., and the voltage E_{R_1} across R_1 equals the voltage E_{R_2} across R_2. The voltage E_{R_3} across R_3 is equal to and in phase with voltage across the grid side of the tank coil (E_g). It then follows that

E_{R_3} is 180 degrees out of phase with the voltage E_p. Inasmuch as E_p lags E_{R_1} by 60 degrees and leads E_{R_2} by 60 degrees it follows that the voltages across the resistors R_1, R_2, and R_3 are displaced from each other by 120 degrees, i.e., $\theta=\alpha=\beta$. It will be noted that if $E_p=2\ E_g$ that the resulting voltages E_{R_1}, E_{R_2}, and E_{R_3} will be equal since $E_{R_1}=E_{R_2}=E_p \cos 60 \text{ deg.}=0.5\ E_p=E_g$. This preliminary discussion explains the fundamental principles involved in the obtaining of a three-phase radio-frequency supply. It will be seen that if the respective voltages across R_1, R_2, and R_3 are individually applied to the grids of three separate radio-frequency amplifiers, the phase relationship of the resulting a-c plate voltage generated in each plate circuit will be displaced by 120 degrees. This assumes that the phase displacements between the voltages of each grid and plate circuit are equal and that the tube input impedances are infinity. If the input and output circuits of separate amplifiers are the same, the respective phase displacements between the grid and plate circuits are essentially equal. However, the input impedance of any tube cannot be equal to infinity, therefore, the effect of this impedance must be taken into account when adjusting for the proper phase relationships. The stray capacity in the input circuit of the first intermediate power amplifier makes it difficult to determine the accurate value of input impedance. This complicates the possibility of determining the values of X_L and X_C that should be chosen in order to obtain the desired radio-frequency relationships. A mathematical analysis of this input circuit using effective measured values indicates that the resultant circuit is fundamentally a capacitive reactance in shunt with R_1, R_2, or R_3. This capacity has the effect of changing the phase relations between the voltages applied across each of the tubes. Fig. 17 shows the fundamental cir-

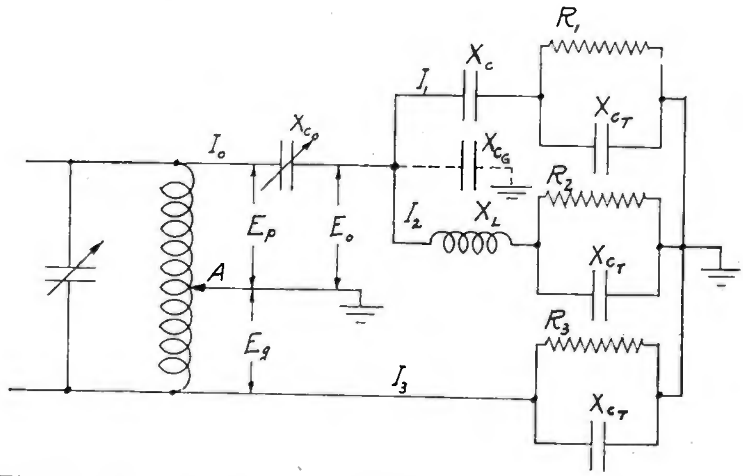


Fig. 17—Diagram showing final circuit adopted to obtain a three-phase radio-frequency supply.

cuit required in order to obtain the correct phase displacements. The introduction of X_{C_t} across R_1 and R_2 results in requiring X_C to be reduced and X_L to be increased if the voltage across R_1 and R_2 are to remain equal and displaced in phase by 120 degrees. The introduction of X_{C_t} across R_1 and R_2 has another effect which must be dealt with in the following manner. The voltage across R_1(Fig. 18) will be displaced from the voltage E_0 existing across X_C and R_1 by some angle less than 60 degrees, i.e., (60 deg. $-\phi$) and the voltage across R_2 will be displaced from the voltage E_0 existing across X_L and R_2 by some angle greater than 60 degrees, i.e., (60 deg. $+\phi$). From this it is noted that the voltages applied across R_1 and R_2 are not symmetri-

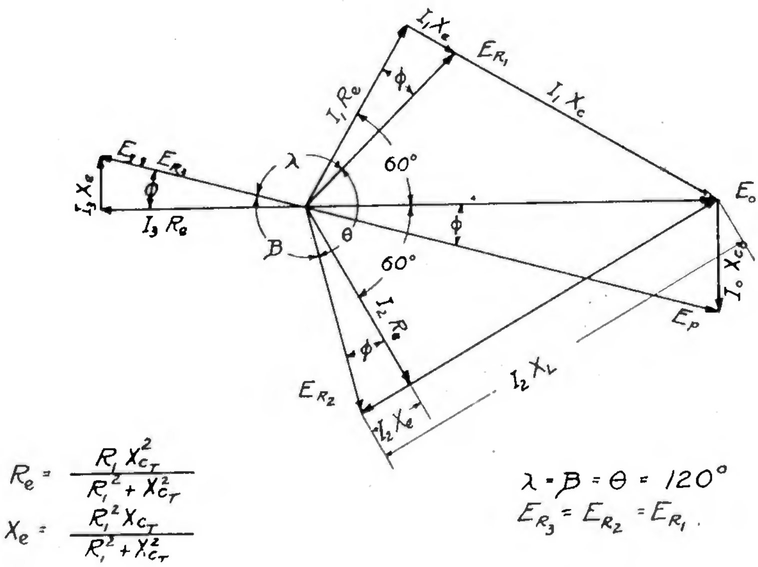


Fig. 18—Vector diagram showing phase relationships of voltages existing in Fig. 17.

cally displaced with respect to the voltage E_0, common to both the X_C and X_L branches. If X_{C_0} is set equal to zero (short circuited) it is obvious that E_p and E_0 would be identical and E_g applied to R_3 would be 180 degrees out of phase with E_0. Then E_{R_3} would be displaced from E_{R_1} by 120 deg. $+\phi$ and from E_{R_2} by 120 deg. $-\phi$. In order to compensate for this asymmetrical arrangement of phases, X_{C_0} is introduced and may be adjusted to a value which will permit all phases to be displaced from each other by 120 degrees. It is then possible by a slight adjustment of tap A (Fig. 17) to adjust the ratio of E_p/E_g so that the voltages across R_1, R_2, and R_3 are equal as well as being displaced by 120 degrees. It is of interest to observe that shifting the sliding tap A does not affect the phase relationships between E_p and E_g, but only affects the ratio of the voltages (E_p/E_g).

Grid current flowing in each first intermediate power amplifier has an important effect which has been neglected up to this point in the discussion of phase displacements. The presence of grid current reduces the tube input resistance which is otherwise essentially infinity, thus in effect placing a resistance in parallel with the fixed external shunt resistance which will vary in value depending upon the portion of the radio-frequency cycle under consideration. This variation in resistance shifts the radio-frequency phases from their normal respective displacements. In order to eliminate this effect and to predetermine the values of X_L, X_C, and X_{C_0} for different radio frequencies, each first intermediate amplifier is biased high enough to prevent flow of grid current. These tubes merely amplify the voltage across R_1, R_2, and R_3 without disrupting the existing phase displacements between the respective voltages.

Methods Used to Determine Phase Displacements

The first system used to determine when the three radio-frequency phases were displaced 120 degrees depended upon the following relationship: When each goniometer primary is excited with an individual phase of a symmetrically displaced three-phase radio-frequency source (no modulation being applied) the current in each loop antenna will remain a constant when the goniometer is rotated through 360 degrees. The restrictions are that the goniometer shall have a sine wave characteristic and the voltages applied to each goniometer primary shall be equal and remain a constant. From the following expressions it will be observed that the current in each loop antenna will remain a constant, and that these currents will be equal to each other for any setting of the goniometer, assuming that exact 120-degree time phase relationships exist between currents in each goniometer primary and that the goniometer primaries are displaced 120 degrees in space.

$$I_1 = I' \cos wt \cos \theta + I'' \cos (wt + 120) \cos (\theta + 120) + I''' \cos (wt + 240) \cos (\theta + 240)$$

$$I_2 = I' \cos wt \cos (\theta + 90) + I'' \cos (wt + 120) \cos (\theta + 210) + I''' \cos (wt + 240) \cos (\theta + 330)$$

where I_1 and I_2 represent the instantaneous current flowing in loop 1 and loop 2, respectively, for any setting of the goniometer (θ) in polar coördinates assuming the maximum currents I', I'', and I''' are each equal in amplitude when maximum mutual occurs.

It will be noted that I_1 plotted against θ will give a circle having a radius equal to I_1, and I_2 will also give a circle having a radius equal to I_2. A space pattern of field strength ϵ will also give a circular pattern in space for any setting of the goniometer since we have an expression which is similar in form.

$$\epsilon = KE_0[\cos \omega t \cos \theta + \cos (\omega t + 120) \cos (\theta + 120) + \cos (\omega t + 240) \cos (\theta + 240)].$$

It is, therefore, of interest to note that the space pattern for any setting of the goniometer is similar to the pattern obtained by plotting the loop antenna current for either loop against the goniometer setting θ as the goniometer is rotated through 360 degrees. It will be noted that the field strength space pattern or the polar diagram of current

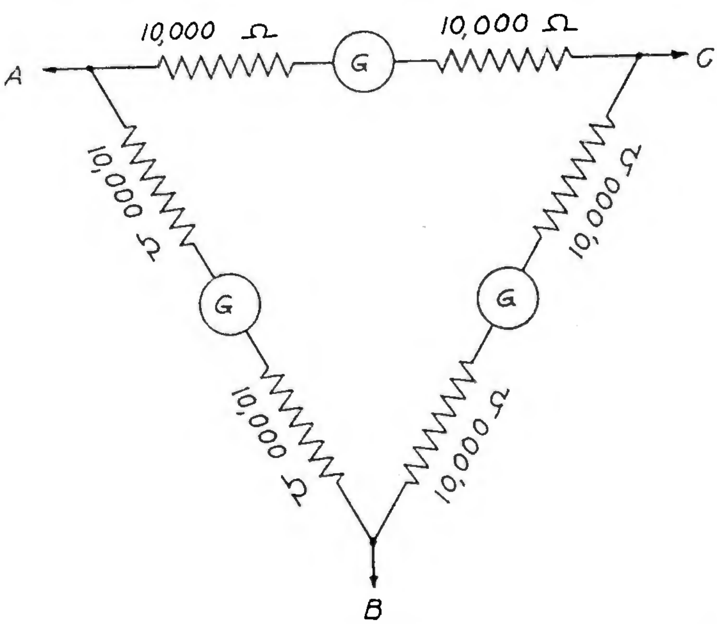


Fig. 19—Diagram showing delta arrangement used to determine the radio-frequency time phase relationship between each of the phases.

in either loop is a direct indication of symmetrical arrangement of the radio-frequency phase displacements, i.e., a circular pattern is obtained when the phases are displaced 120 degrees time phase; if the phases are not displaced 120 degrees the pattern will deviate from a circle by an amount depending upon the deviation from the 120-degree time-phase displacement. This can readily be seen by substituting values of time-phase displacement in the above equations. This system is somewhat cumbersome except as a check on the required condition, as it is a tedious cut and try process to determine exactly how the radio-frequency vectors are displaced with respect to each other. If it were possible to observe these time-phase displacements for any particular condition, considerable time would be saved in adjusting for the required 120-degree displacements, a consideration which led to the development of the following method: It is a well-known fact that the voltages across each side of a balanced delta system which is connected to a three-phase supply will be equal provided the voltages between each junction and ground are equal, and the phase displacements are 120 degrees. Any variation from this condition will cause unequal voltages to occur across the sides of the delta, thus making it possible to determine the phase angles by measurement of these voltages. This was done by placing two 10,000-ohm resistors in series across each phase at the output of the transmitter and measuring the current through the resistors by placing a current

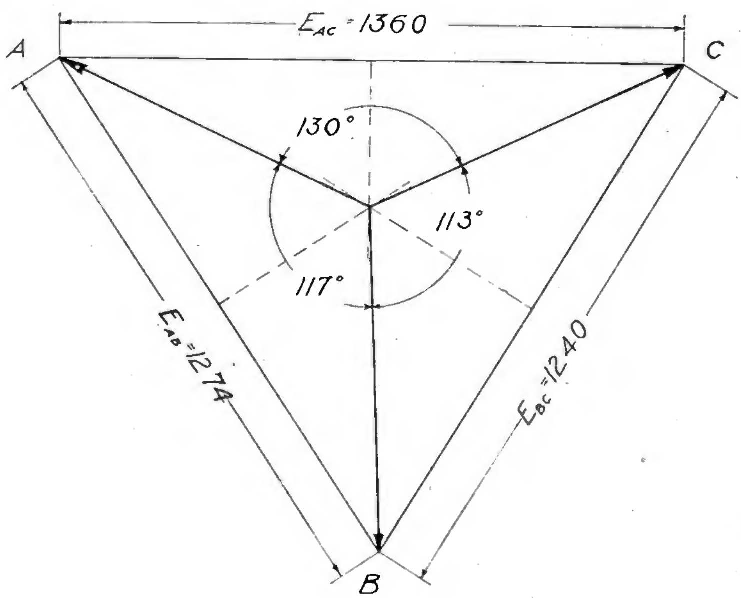


Fig. 20—Vector diagram showing method used to determine the phase relationships between each of the phases when knowing the voltages across each delta branch.

squared galvanometer in series with them. Fig. 19 shows a schematic diagram of the arrangement. An example of the results obtained using this system follows. A given setting of the phasing controls gave the following voltages across the phases:

Branch	*E*	*Phase Angle Between Voltages*
A C	1360	130 deg.
C B	1240	113 deg.
B A	1274	117 deg.

The phase angles were determined by constructing a triangle having its respective sides directly proportional to the voltages obtained across each of the branches (Fig. 20). The perpendicular bisectors of each of these sides were erected, intersecting at a single point. The

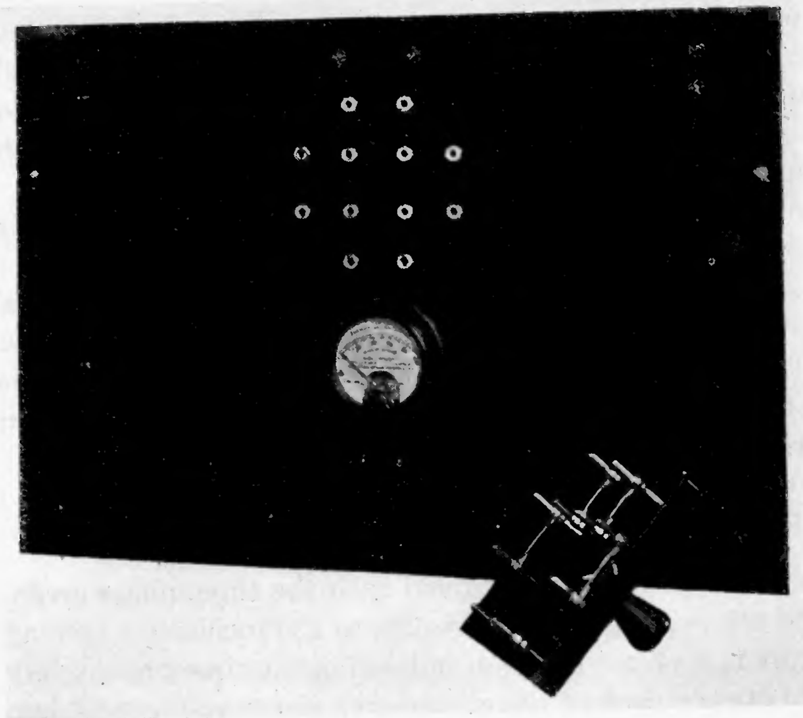


Fig. 21—View showing device used in determining the radio-frequency phase relationships.

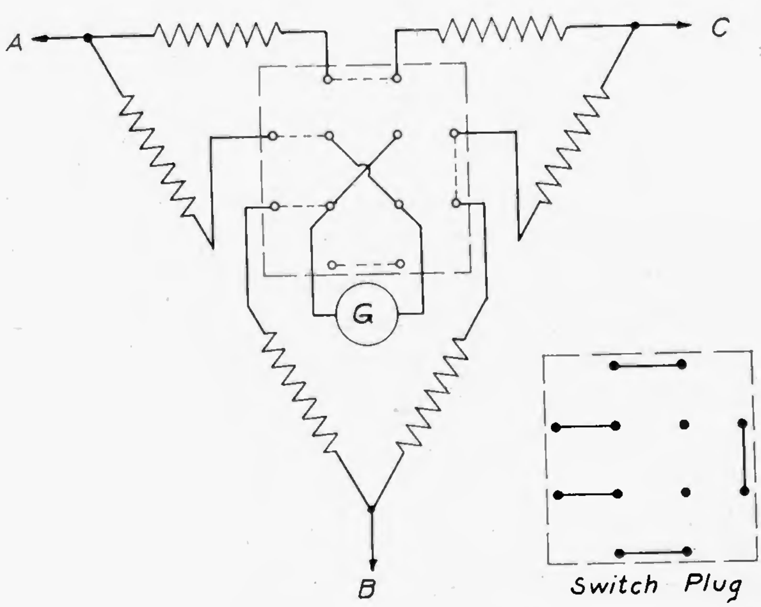


Fig. 22—Diagram showing the connections used in the instrument shown in Fig. 21.

angles between vectors drawn from this point to each corner of the triangle represent the phase displacements between the voltages applied to the junction points of each branch. This system of obtaining phase displacements has proved to be extremely helpful in adjusting X_{C_0}, X_C, and X_L to the proper values to give 120-degree phase relationships.

In order to eliminate the necessity of using three galvanometers and three calibration charts, the resistors were mounted on a panel with a single galvanometer which could be inserted in any branch by means of a rotating hand-operated plug (Fig. 21). A schematic diagram of the device is given in Fig. 22. This arrangement obviated the necessity of taking readings from more than one curve, thus permitting the voltages to be read quickly and accurately.

Results Obtained with Three-Phase Arrangement

Representative results obtained with the three-phase arrangement described above are given in Figs. 23a to 23d inclusive. Accompanying each figure is a vector diagram indicating the time-phase relationships existing between each of the goniometer stator voltages as determined by the delta-phase measuring system described. Each figure also includes a polar diagram of the current in one loop plotted against the goniometer setting in degrees. These polar diagrams have been found to coincide for all practical purposes with the field strength patterns obtained in space. Figs. 23a to 23c inclusive show clearly the results of varying X_{C_0} from *0* (short-circuited) to 6420 ohms ($C_0 = 87$ $\mu\mu$f). From these results it is quite obvious that the correct value of X_{C_0} for 285 kc is approximately 2820 ohms ($C = 198$ $\mu\mu$f). Only a few polar diagrams are given here since the material on hand is voluminous and could not be published in this paper. However, the information given herewith is sufficient to indicate that the time-phase vectors bear a definite relationship to the polar diagrams and that both methods of checking time phase relationships concur, since the three-phase vectors are approximately 120 degrees displaced when the polar diagram is nearly circular. The variations in amplitude which occur in the polar diagram are undoubtedly due to variations in the goniometer characteristic, i.e., the goniometer characteristic is not exactly sinusoidal as has been assumed.

The above results were obtained using a resistance of 5000 ohms across the tube input. Fig. 23d indicates the results obtained when this resistance is increased to 10,000 ohms. It will be observed that the value of X_{C_0} necessary to give 120-degree time-phase relationships when using 10,000-ohm resistors in parallel with each tube is rela-

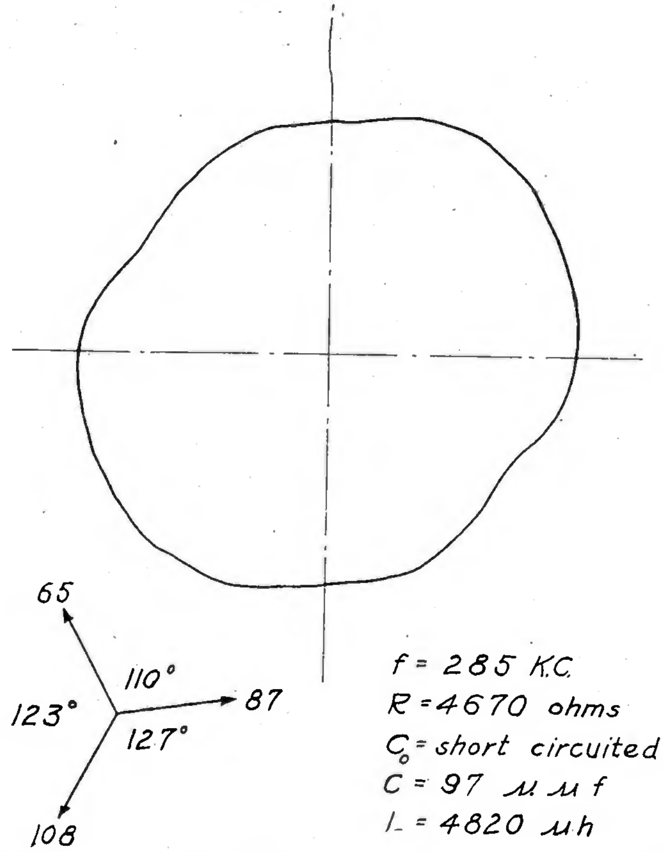


Fig. 23—Vector diagram showing time-phase relationships and the resultant polar diagram of I_{A_1} plotted against the goniometer setting. (a), (b), and (c) show the effect of varying C_0. (d) shows the effect of making R relatively large.

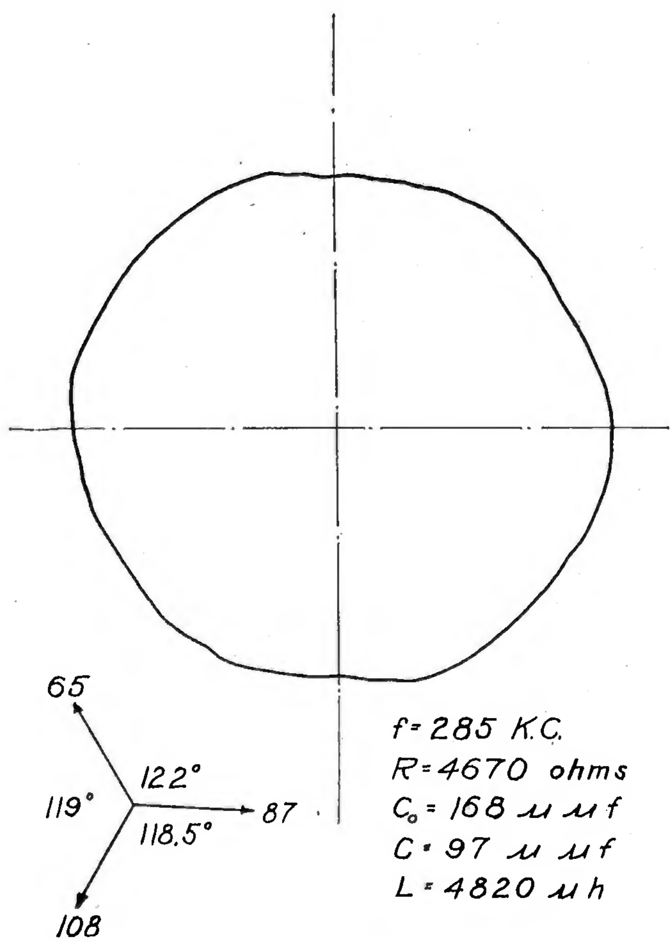


Fig.—23b.

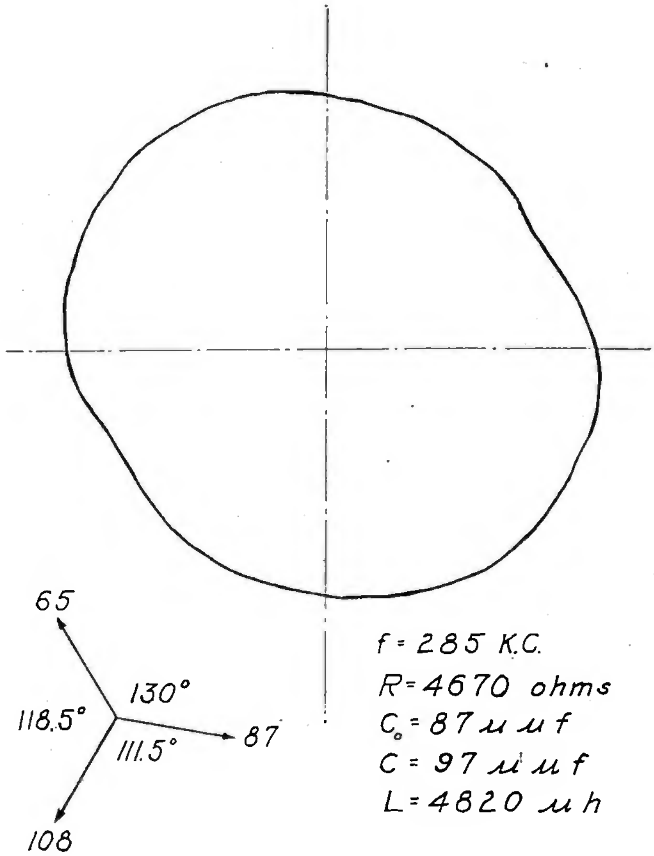


Fig.—23c.

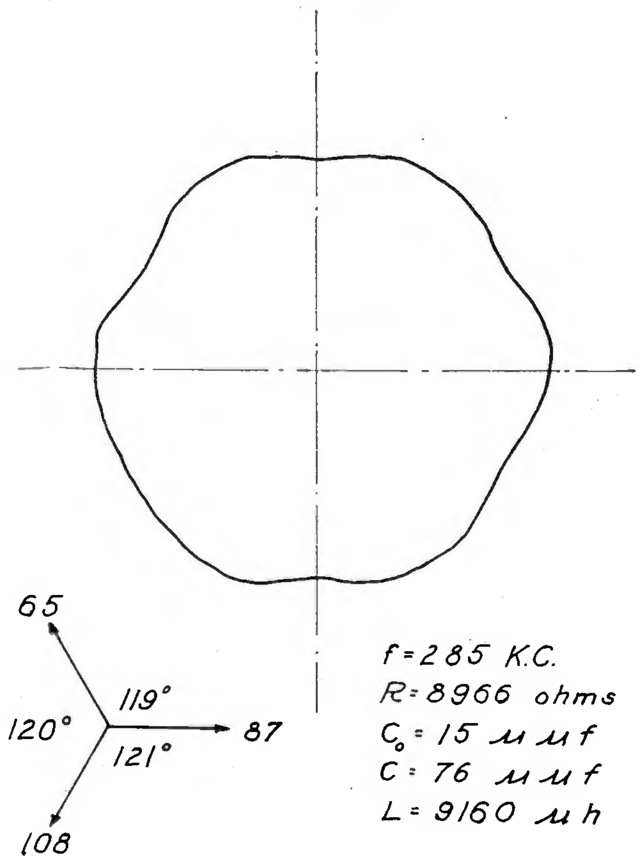
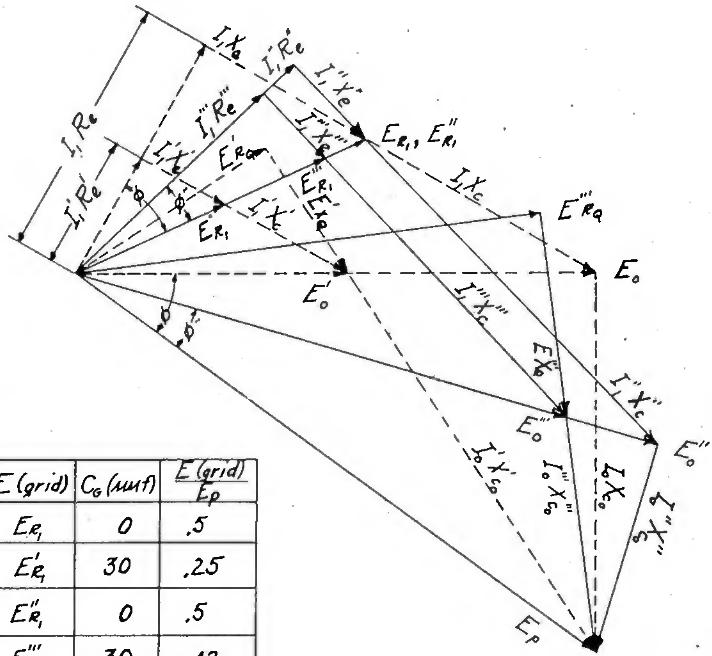


Fig.—23d.



$R_1(\Omega)$	E (grid)	C_G (μμf)	E (grid) / E_p
9340	E_{R_1}	0	.5
9340	E'_{R_1}	30	.25
4670	E''_{R_1}	0	.5
4670	E'''_{R_1}	30	.42

Fig. 24—Vector diagram showing the greatly reduced effect of C_G on the available voltage supplied to the tube when utilizing a low value of shunt resistance.

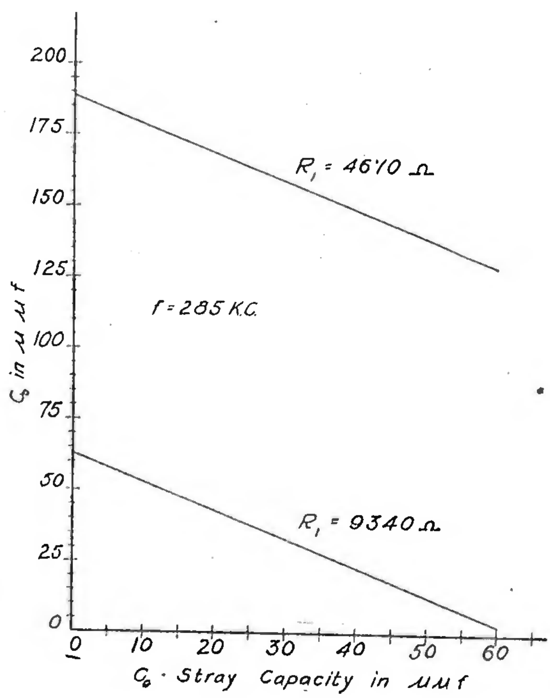


Fig. 25—Curves showing the value of C_0 required to maintain 120-degree time-phase displacements for different values of stray capacity when utilizing two different values of tube shunt resistance.

tively high ($X_{C_0}=37{,}200$ ohms $C_0=15$ $\mu\mu$f). This results in an abnormally high voltage drop across X_{C_0}, thus limiting the maximum possible voltage that can be applied to the grids of the first intermediate power amplifiers to about 140 volts; whereas about 380 volts may be obtained if 5000-ohm resistors are used with the required values of X_{C_0}, X_C, and X_L to give 120-degree time-phase relationships. The widely varying results obtained in the available voltages when the required phase relationships are obtained can be accounted for only by taking into account the stray capacity (C_G) of X_L, X_C, and X_{C_0} to ground. It will be observed in Fig. 24 that the voltage vector E_{R_1}

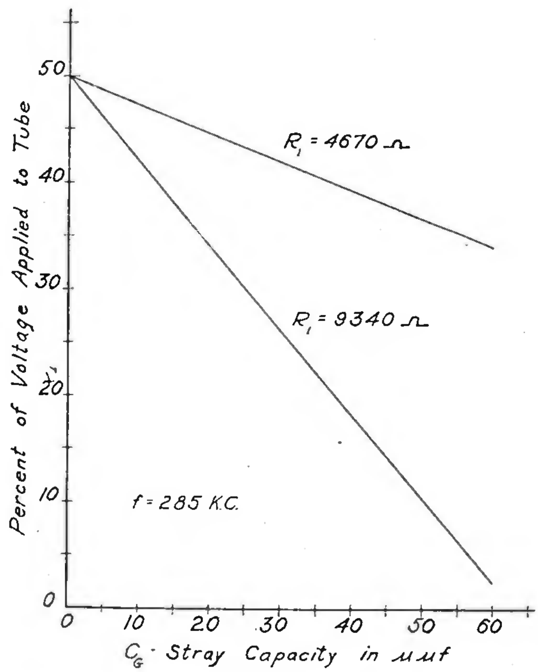


Fig. 26—Curves showing effect of stray capacity upon the per cent of total voltage E_p capable of application to the tube with different values of R_1.

reduces very rapidly with respect to the constant applied voltage E_p as C_G is increased. Fig. 25 shows the value of C_0 required in order to maintain the 120-degree phase displacements with different values of C_G and R_1, while Fig. 26 indicates the effect of stray capacity upon the voltage applied to the tube with different values of R_1. These curves clearly indicate the reason for choosing 5000 ohms instead of 10,000 ohms.

Determination of X_L, X_C, and X_{C_0} for Different Frequencies

In order to determine the value of X_L, X_C, and X_{C_0} for various frequencies, it was necessary to ascertain the absolute value of the input impedance. This was attempted in several different ways which proved to give widely varying results of an erroneous nature. From these preliminary experiments and calculations, it was eventually decided that the only method which would give the required results would be to obtain a three-phase radio-frequency supply at a given frequency and then knowing the values of X_L, X_C, and X_{C_0} used, work backwards, and determine the characteristics of the input impedance. This procedure was undertaken and is given here as there are several points of interest. The following values of X_L, X_C, and X_{C_0} were taken at 285 kc and were found to give a symmetrically phase-displaced three-phase radio-frequency supply to within the precision of measurement.

$$
\begin{aligned}
X_L &= 8730 \text{ ohms}\\
L &= 4880\ \mu\text{h}\\
X_C &= 5760 \text{ ohms}\\
C &= 97\ \mu\mu\text{f}\\
X_{C_0} &= 2820 \text{ ohms}\\
C_0 &= 198\ \mu\mu\text{f}
\end{aligned}
$$

$$\text{Ratio of voltage } \frac{E_0}{E_{R_1}} = \frac{E_0}{E_{R_2}} = \frac{Z_0}{Z_e} = \frac{Z_0'}{Z_e} = 1.90.$$

(The determination of this voltage ratio is explained in a following paragraph.)

With this data at hand it is possible to determine X_e (the equivalent series reactance of the tube input circuit).

$$X_e = \frac{X_L - X_C}{2} = \frac{8730 - 5760}{2} = 1485 \text{ ohms}.$$

This is apparent from observing the vector diagram in Fig. 27. Angle β is equal to angle α.

R_e (the equivalent series resistance of the tube input circuit) may be solved as follows:

$$\frac{Z_0}{Z_e} = 1.90$$

$$R_e^2 + X_e^2 = Z_e^2$$

$$R_e^2 + (X_e + X_C)^2 = Z_0^2$$

$$\frac{Z_0^2}{Z_e^2} = \frac{R_e^2 + (X_e + X_C)^2}{R_e^2 + X_e^2} = (1.90)^2.$$

Substituting the known values of X_e and X_C and solving for R_e

$$R_e = 4130 \text{ ohms}.$$

$$\text{Angle } \beta \text{ (Fig. 27)} = \tan^{-1}\frac{(X_C+X_e)}{R_e} \text{ and angle } \alpha = \tan^{-1}\frac{(X_L-X_e)}{R_e}$$

$$\alpha = \beta = 60.33 \text{ deg}.$$

By phase angle measurement $\alpha+\beta$ was found to equal 120 degrees or $\alpha=\beta=60$ degrees which compares very favorably with the calculated value. In order that all calculated values would be consistent, β and α were each set equal to 60 degrees and X_C and X_L were solved for, since it is apparent that X_e and R_e would remain constant for the same frequency.

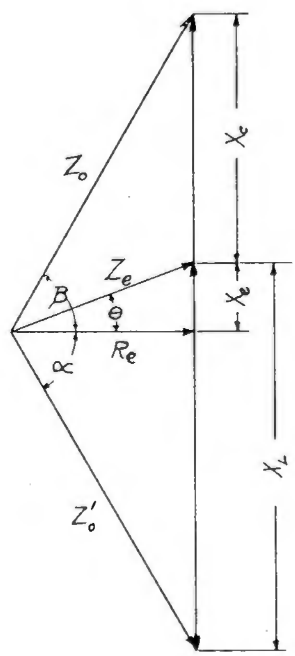


Fig. 27—Impedance vector diagram showing relaship between different vectors.

Solving for the revised values of X_c and X_L to make α and β both equal to 60 degrees, we have

$$X_C = 5665 \text{ ohms}$$
$$C = 98.6 \ \mu\mu\text{f}$$
$$X_L = 8635 \text{ ohms}$$
$$L = 4820 \ \mu\text{h}$$

$$\text{Revised ratio of impedance} = \frac{Z_0}{Z_e} = 1.88.$$

Knowing the values of X_e and R_e, it is possible to determine X_t from the following equations:

$$R_e = \frac{RX_t^2}{R^2 + X_t^2}$$

$$X_e = \frac{X_t R^2}{R^2 + X_t^2}$$

om which

$$X_t = \frac{R_e^2 + X_e^2}{X_e}.$$

Now substituting the values of R_e and X_e previously determined

$$X_t = 13,000 \text{ ohms}$$
$$C_t = 43\ \mu\mu f$$

It is now possible to determine the value of R (the effective shunt esistance) from the following simple relationship

$$R = \frac{X_e}{R_e} X_t.$$

Substituting values of X_e, X_t, and R_e.

$$R = 4670 \text{ ohms}.$$

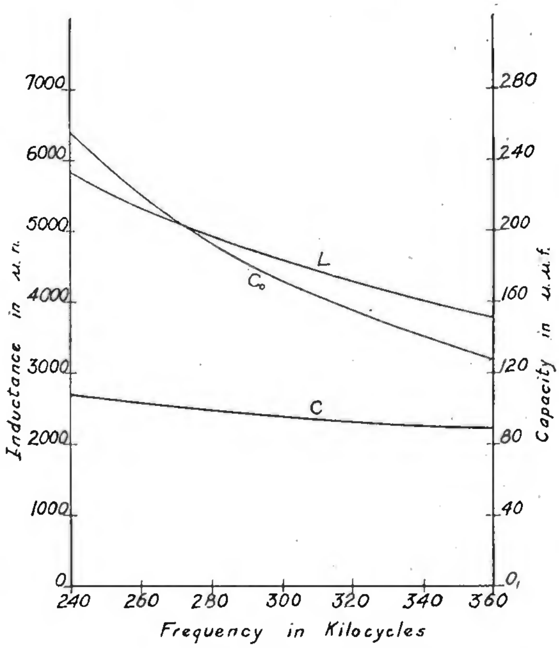


Fig. 28—Curves showing values of C, L, and C_0 required for different frequencies.

It is now assumed that the values of R and C obtained for the tube input circuit remained constant for different frequencies. With these values available it is possible to predetermine the values of X_C, X_L, and X_{C_0} required for adjusting the three time-phase displacements to 120 degrees for various frequencies. Values of C, L, and C_0 are plotted against frequency in Fig. 28. These curves made it possible to calibrate C, L, and C_0 and plot curves of their respective dial settings against frequency.

Determination of Voltage Ratio

It will be observed from the foregoing discussion that the determination of the impedance ratio Z_0/Z_c has played an important part in obtaining the absolute value of the tube input impedance. The accuracy with which R_e, X_t, and R may be determined depends directly upon the accuracy of the ratio Z_0/Z_e. The method originally used to determine this ratio was as follows: A thermogalvanometer (G_1) was placed in series with the phasing condenser as shown in Fig. 29 and a single-pole single-throw switch (S_1) was placed across the phasing condenser. The voltage E_p applied to the system was measured by the IR_M drop occurring across the resistor R_M. A given voltage E_p was applied to the system with S_1 in a closed position. The current I_1 flowing in the meter G_1 was then recorded together

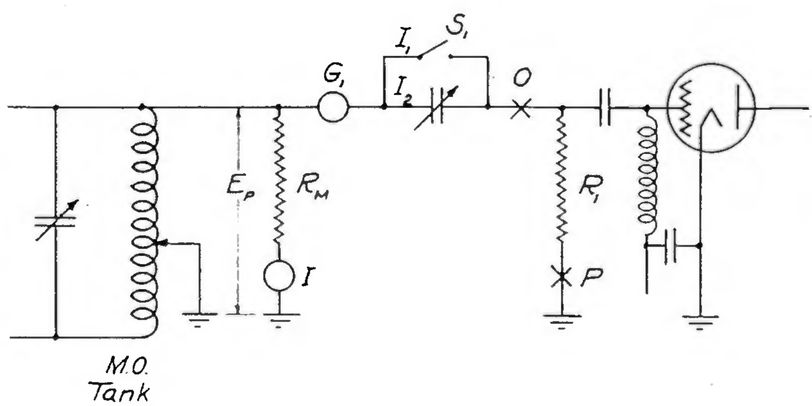


Fig. 29—Schematic diagram showing methods used in determining the voltage ratio.

with the applied voltage E_p. The switch S_1 was then opened and the voltage across R_M was readjusted to its normal value E_p by reducing the plate voltage of the master oscillator. The current I_2 flowing in G_1 was then recorded, particular care being taken to see that the voltage across R_M was identical to the reading taken when S_1 was closed It will now be seen that

$$\frac{E_P}{I_1} = Z_e$$

$$\text{and } \frac{E_P}{I_2} = Z_0$$

$$\text{Ratio of } \frac{Z_0}{Z_e} = \frac{\frac{E_P}{I_2}}{\frac{E_P}{I_1}} = \frac{I_1}{I_2}.$$

hat is, the ratio of the currents I_1/I_2 is a direct indication of the atio of Z_0/Z_e. Theoretically, this procedure appears to be valid. owever, practically this is not true, since the values of I_1 and I_2 are accurate and cannot be depended upon to give correct results be-

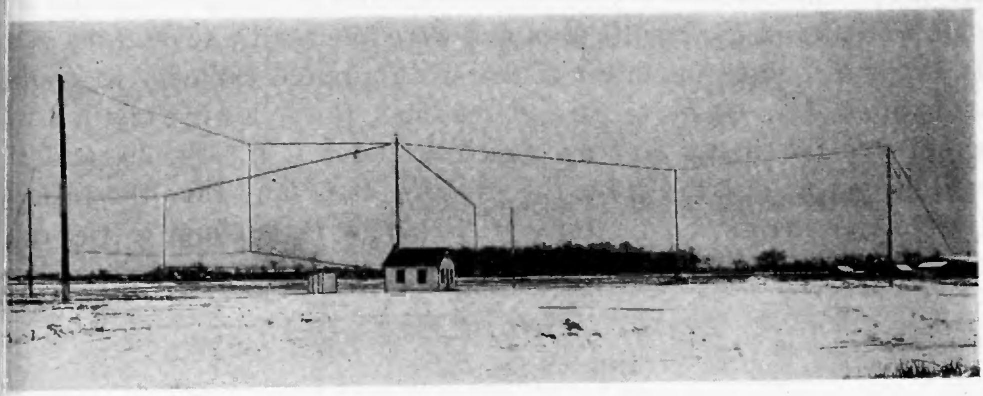


Fig. 30—Antenna system for the polydirectional radio range at Wayne County Airport, Detroit, Michigan.

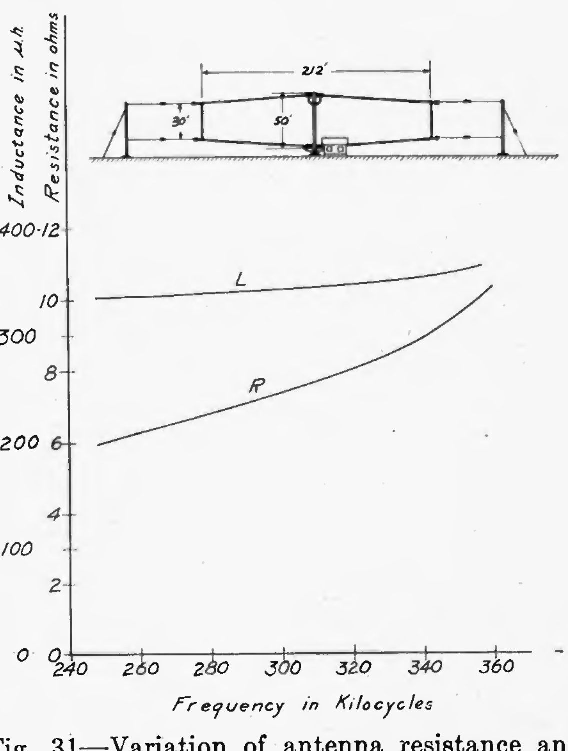


Fig. 31—Variation of antenna resistance and effective inductance with frequency.

ause of the capacity between the meter and ground. It might be xpected that by isolating the meter G_1 from ground fairly accurate esults would be obtained with this method. However, experiment ndicated that the results obtained by further isolating the meter G_1

from ground gave only a slight improvement. The meter G_1 was then removed from this position and inserted at point O Fig. 29. The results obtained with the meter in this position were a great improvement over those obtained with the meter in the original position. However, the effect of the stray capacity of the meter to ground was still present and the results obtained were in error by several per cent. In order to reduce the effect of meter distributed capacity to ground and thus obtain accurate current readings, the following method was adopted. The meter G_1 was placed at point P in Fig. 29, and the switch S_1 was omitted from the circuit. It is true that the ratio of the voltage across R_M to the voltage across R_1 is equal to the ratio of Z_0/Z_e since

$$E_{R_M} = Z_0 I_2$$

$$\text{and } E_{R_1} = Z_e I_2$$

$$\text{then } \frac{E_{RM}}{E_{R_1}} = \frac{Z_0 I_2}{Z_e I_2} = \frac{Z_0}{Z_e}.$$

This method gave results which were reasonably accurate and had the additional advantage of balancing the voltages applied to the inductive and capacitive branches.

Characteristics of Loop Antennas

Two loop antennas placed at right angles to each other are used to radiate the carrier and side bands due to the three modulation frequencies according to their respective space pattern requirements. The loops employed are 212 feet long and have an average height of 40 feet. An outside view of a polydirectional radio range station is given in Fig. 30 and should give a clear conception of the arrangement of the loop antennas. A diagram giving the dimensions of one of the loops is given in Fig. 31, together with the measured values of antenna resistance, and effective inductance for different frequencies. The distributed capacity of the above loop antenna is 57 $\mu\mu$f. Fig. 32 gives two curves of effective height for different frequencies. One curve is calculated by formula knowing the physical dimensions, while the other curve is determined by the Austin-Cohen formula knowing the field strength and attenuation constant. Particular attention should be given to the fact that the loop antenna has been reduced in height from the type formerly used in order to permit placing the polydirectional radio range as close as possible to airports without creating an obstruction, since it is essentially a homing device.

Course Indicator

In order to check rapidly the position of courses about the poly-[di]rectional radio range, it was necessary to provide the station with a [po]rtable course indicator which could be conveniently located at

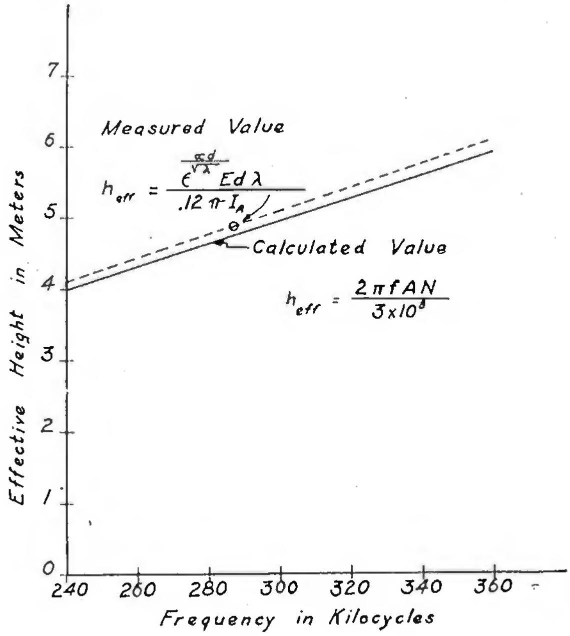


Fig. 32—Variation of effective height with frequency of antenna shown in Fig. 31.

[a]ny given point around the intersection of the loop antennas. Fig. 33 [g]ives the schematic diagram of a course indicating device which fully [m]eets these requirements and has the additional advantage of indi-

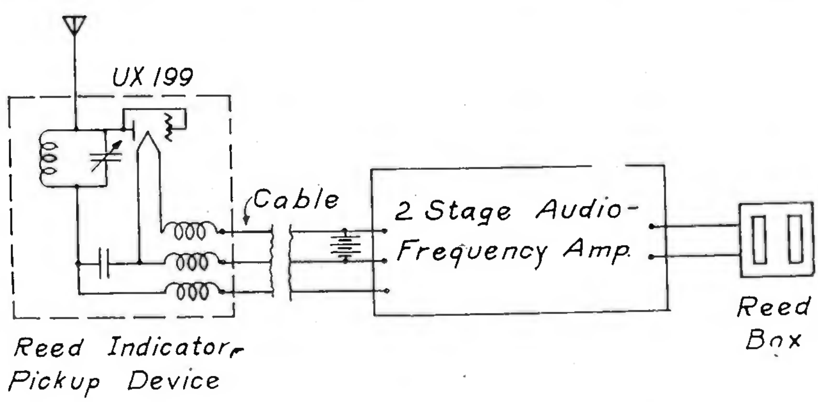


Fig. 33—Schematic diagram of course indicating device.

[c]ating the position of the courses in the radio range building. The reed [i]ndicator pick-up device shown in Fig. 34 consists of a tuned circuit [w]hich is excited by a five-foot vertical rod antenna. The voltage set [u]p across the tuned circuit is rectified by a UX-199 tube and then

delivered to the radio range building through a flexible triple conductor shielded cable. The UX-199 filament supply is furnished by a battery located in the radio range building, thus making the pickup device as portable as possible. Radio-frequency chokes are placed

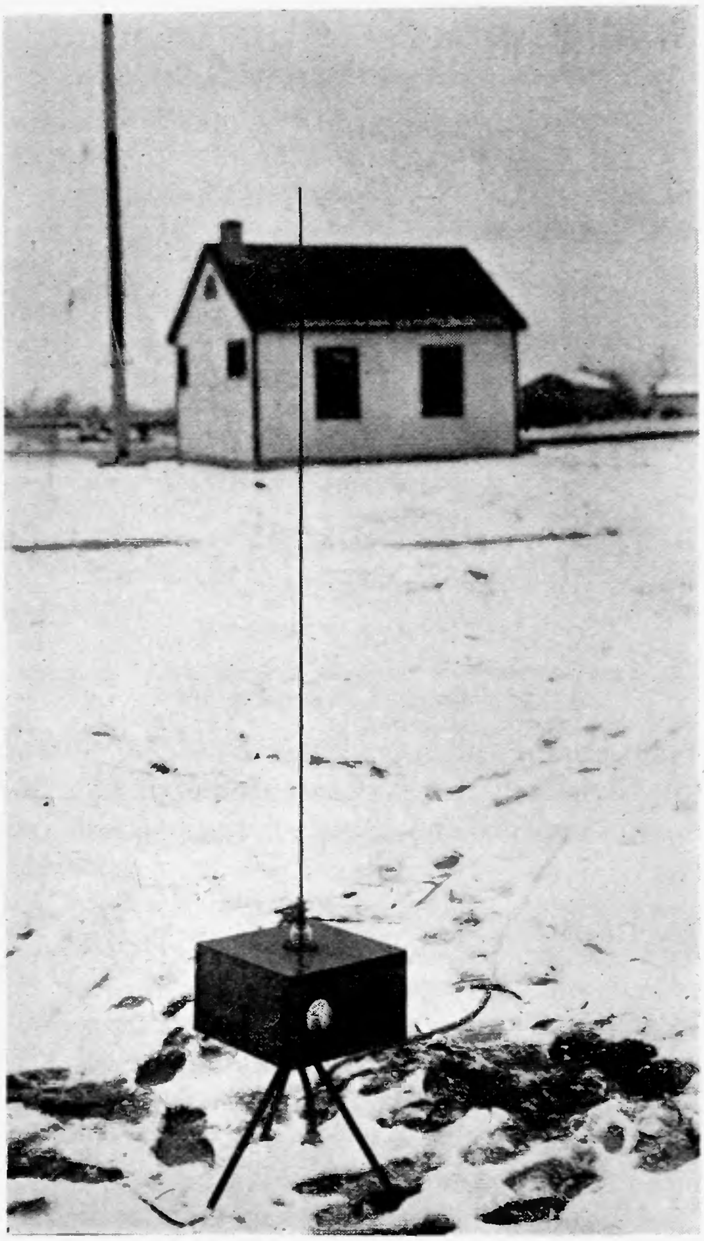


Fig. 34—Pick-up unit of course indicating device.

in each lead at the rectifier unit in order to prevent radio-frequency pickup on the line from combining with the desired voltages induced in the antenna. A conventional two-stage amplifier is used to raise the audio-frequency voltage supplied by the reed indicator pick-up device to a level suitable for application to a reed box which is located at the operator's desk. This unit has proved valuable not only as a course positioning device, but also as a course monitor.

Field Observations of Results

A brief summary of the results obtained using the equipment described above is given in the following paragraphs.

The versatility of this equipment to fit the requirements of the Airways Division is probably best demonstrated by showing the ob-

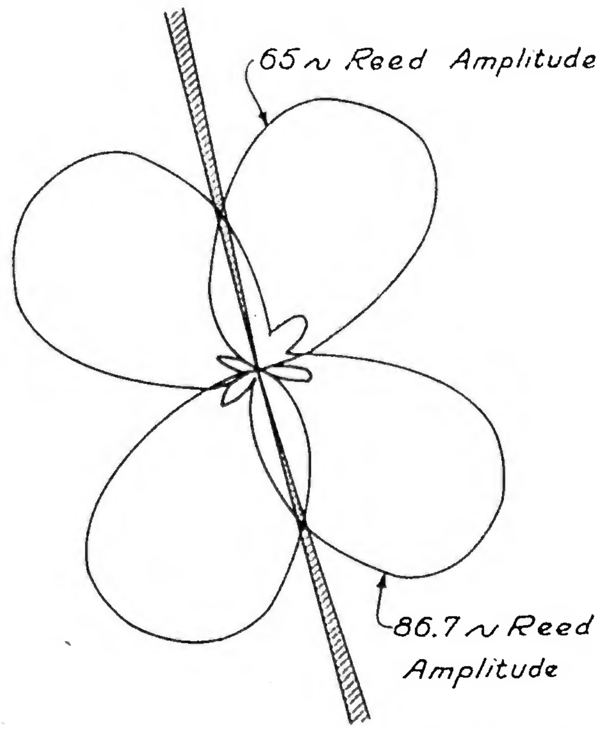


Fig. 35—Observed reed amplitude space patterns of a two-course radio range.

served field patterns which may be obtained when working the equipment under different time-phase and space-phase conditions. The three standard systems of application are the two-, four-, and twelve-course arrangements. The two-course system is established by utilizing only two of the radio-frequency amplifier channels which excite two of the goniometer stators which are placed at right angles to each other. The radio-frequency voltages applied to each of the channels are in phase (0-degree phase displacement). Observed reed amplitude space patterns of the two-course radio range are shown in Fig. 35. The four-course system is established in a similar manner except that the radio-frequency phases are displaced by 90-degrees. Fig. 36 shows the observed reed amplitude space patterns obtained when the equip-

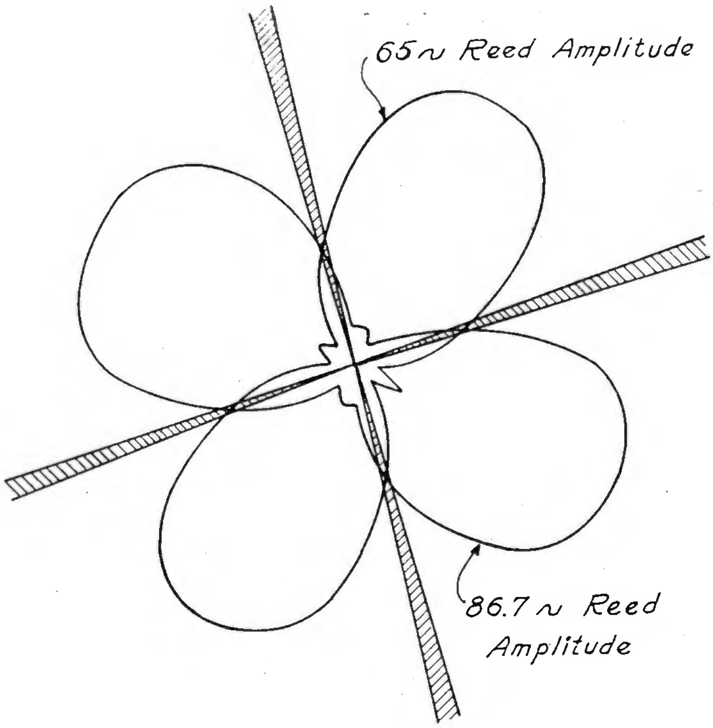


Fig. 36—Observed reed amplitude space patterns of a four-course radio range.

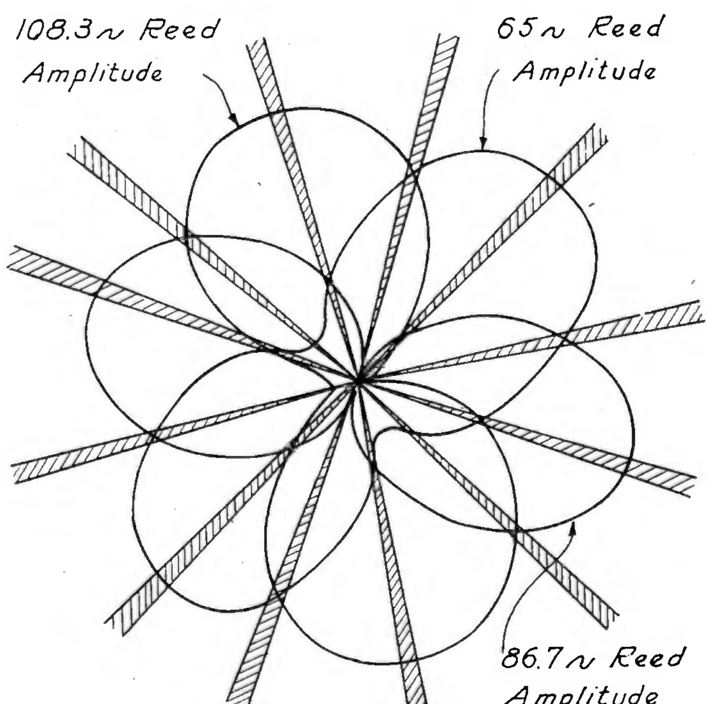


Fig. 37—Observed reed amplitude space patterns of a twelve-course radio range.

ment is used as a four-course range. As has already been explained, twelve courses exist when three channels are excited at 120-degree time-phase and the resulting outputs applied to the three goniometer stators which are placed at 120 degrees. The observed reed amplitude space patterns obtained using this arrangement are shown in Fig. 37.

From these figures it may be observed that a radio range transmitter having such an array of potential courses may be used to serve any particular case which may arise. The comparative reed amplitudes obtained when flying on the different courses are approximately as follows: (a) Both courses of the two-course range give thirty per cent

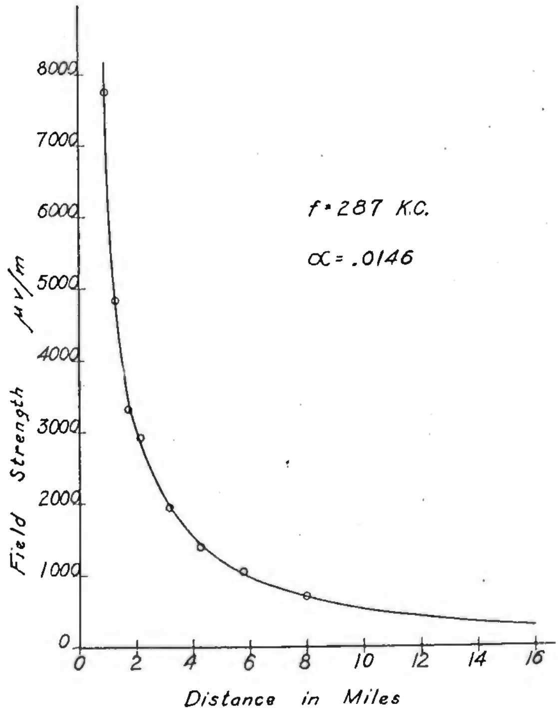


Fig. 38—Attenuation curve showing the field strength of the polydirectional radio range at different distances from the station.

greater reed amplitude than the four-course range: (b) The six strong courses of the twelve-course range give reed amplitudes equal to seventy-five per cent of those obtained with the four-course range, while the six weak courses have but half the amplitude of the strong courses. The relative amplitudes of these courses may be altered considerably by varying either the time-phase or space-phase relationships. However, these amplitudes are approximately correct, assuming normal space and time phases and a fixed plate voltage of 2000 volts in every case.

Experiment has indicated that the four-course range may be fitted to the different airways in many cases by reducing the plate voltage on one of the power amplifier branches. This may be accomplished readily by placing a resistance in series with the plate supply. This reduces the amplitude of one of the side-band figure-of-eight patterns, thus permitting the angle between adjacent courses to be adjusted to any angle between 45 degrees and 135 degrees instead of the normal 90-degree displacement. Adjusting the courses in this manner instead of varying the plate voltages on the second intermediate power amplifier has the advantage that the power amplifier grids may be worked always at high excitation, thus permitting smooth operation and high efficiency. An added advantage of varying the plate voltage of one of the power amplifiers is that the amplitude of the side-band pattern is nearly a linear function of plate voltage over the range of normal operation.

Fig. 38 shows the field strength[6] produced by a twelve-course radio range under normal operating conditions. These measurements indicate the relatively poor radiating qualities of loop antennas and also show a very high attenuation constant of 0.0146 which is considerably higher than is normally obtained at these frequencies.

Conclusion

The results outlined in the previous paragraphs indicate that the twelve-course visual radio range as developed by the Airways Division is capable of installation as an established aid to air navigation. Every effort has been made to simplify the operation of the transmitter and to reduce to a minimum the possibility of failures in regular operation. Attention has been given to details which will permit the servicing of the equipment by operators of average ability.

The design of the completed transmitter permits its use either as a multicourse radio range or as a simple two- or four-course type. This flexibility of design is particularly valuable when all conditions encountered along the airways are considered. It is true that in a majority of cases the two- or four-course range will fill the requirements, the polydirectional installation being confined to terminal locations where there is a concentration of airways.

No attempt has been made to establish definitely the relative superiority of either the visual radio range, such as described in this paper, or the aural type which is now in use. The final acceptance of either type should be based upon the opinions of pilots who have had experience with both types of ranges under all conditions encountered in commercial transport service.

[6] The Supervisor of Radio of Detroit, Michigan, made these measurements on the radio range installation at Detroit, Michigan.

Acknowledgment

The authors desire to acknowledge the parts played by the personnel of the Radio Division of the Bureau of Standards in lending their valuable coöperation toward this development. The transmitter was constructed for the Airways Division, Bureau of Lighthouses, Department of Commerce, at the shops of the Eleventh Lighthouse District, Detroit, Michigan. Special appreciation is due to Captain F. C. Hingsburg, Chief Engineer, Airways Division; H. J. Walls, Radio Engineer, Airways Division; C. A. Park, Superintendent of Lighthouses, Eleventh District; R. H. Robson, Mechanician-in-Charge, and the shop mechanics and electricians who have all assisted in making this work possible.

Authors' Note—Since completing the above paper, the authors have had an opportunity to check the theory by actual flight tests on the completed installation. These flight tests have indicated that the theoretical points discussed in this paper are valid.

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REDUCTION OF DISTORTION AND CROSS-TALK IN RADIO RECEIVERS BY MEAN OF VARIABLE-MU TETRODES*

By
Stuart Ballantine and H. A. Snow
(Boonton Research Corporation, Boonton, N. J.)

Summary—*In attempting to control the audio output of a radio receiver employing the present types of tubes by varying the control-grid bias or screen-grid voltage distortion, due to nonlinearity of the output-input voltage relation for the tube, and cross-talk are encountered at the higher signal voltages. Both effects are largely due to the rapid increase in the higher-order curvature parameters of the tube characteristic which occurs as the grid bias increases negatively, or screen-voltage decreases.*

Two tubes, designated as Types 550 and 551, have been developed to reduce these effects. They are shielded tetrodes of which the characteristic has been specially shaped to reduce the higher-order curvature in relation to the transconductance. The desired shape of characteristic is attained by a composite structure by virtue of which the tube acts as a high-mu tube at normal grid biases and automatically changes into a device of low-mu as the grid bias increases negatively.

The 550 tube will handle an input voltage of 15, and the 551, 7 volts, with negligible distortion. These represent improvements of 50 and 25 times over present 24 tubes. Cross-talk is reduced to 1/500th that obtained with the present type at input voltages of 0.1 volt.

A general discussion of the problem of distortion and cross-talk is given from both the theoretical and experimental viewpoints. Both effects are shown to be approximately proportional to the ratio of the third derivative to the first derivative of the plate current-grid voltage characteristic. A factor called the cross-talk factor is defined which represents the ratio of audio output due to the interfering signal to that due to the desired signal. This is shown to vary with the square of the interfering signal voltage. Graphical methods of analysis of distortion and cross-talk in multi-stage amplifiers are described.

The principle of the variable-mu tube is outlined and a number of structural embodiments illustrated. The paper concludes with curves and a table of characteristics of the type 550 and 551 tubes and the results of tests of reliability, reproducibility, and longevity in the manufactured product.

In RADIO broadcast receivers the control of audio output for varying antenna signal voltages is commonly effected by varying an electrode voltage, such as the control-grid bias, of the radio amplifier tubes. With ordinary triodes and tetrodes as heretofore employed, the range of antenna r–f voltage over which satisfactory control by this means can be obtained is limited, the results being unsatisfactory when the receiver is so located that the signal voltage from neighboring transmitting stations is high. In these cases several undesirable effects accompany the increasing negative control-grid

* Decimal classification: R148.1×R334. Original manuscript received by the Institute, October 20, 1930.

bias which is necessary to secure constant audio output with increasing signal voltage:

(1) Distortion due to nonlinearity in the relation between the r-f output voltage of the amplifier and the r-f input voltage;
(2) Increased intermodulation or cross-talk effects;
(3) Increased hum in a-c operated receivers due to modulation of the carrier by low-frequency hum voltages inadvertently present in the control-grid and screen-grid circuits.

The underlying cause of the first two of these effects is the increase in the higher-order curvature of the tube characteristic in relation to the effects of the first order, which takes place as the applied r-f voltages are increased.

The type 550 and 551 tetrodes have been developed with the object of reducing the difficulties enumerated above with a minimum of inconvenience and readjustment. This has been achieved by shaping the plate current-grid voltage characteristic so as to minimize the higher-order curvature over an extended range of control-grid voltage. Before discussing the structure of these tubes and the principles of their operation it will be of interest to consider briefly the underlying problems of distortion and cross-talk which they are intended to solve.

I. Problem of Volume Control by Variation of Grid-Bias

The problem to be discussed is the special one associated with the control of amplification in the radio amplifier by variation of an

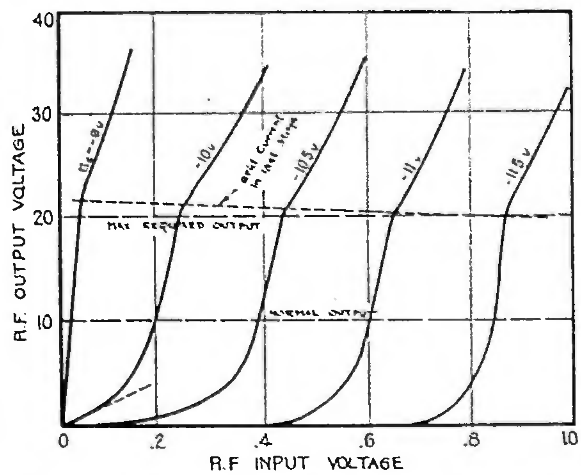

Fig. 1—Relation between input and output voltages in a 4-stage r-f amplifier employing 227 tubes ($E_b = 90$ v) showing nonlinearity at high input voltages.

electrode voltage, usually the control-grid bias. The broad problem of volume control by this and other methods, such as antenna voltage control, combination control of grid-bias and antenna voltage, etc., has been discussed elsewhere.[1] Important undesirable effects which arise in an amplifier so controlled are brought out in Fig. 1, which shows the experimental relation between the output voltage and input (or antenna) voltage in a 4-stage 227 r-f amplifier designed to supply a maximum of 20 volts for the operation of a linear detector. Each curve represents the relation for a specified value of grid bias. For values of grid bias below −9 volts the relation between output and input is linear up to the dotted curve. Beyond this point the curve is bent

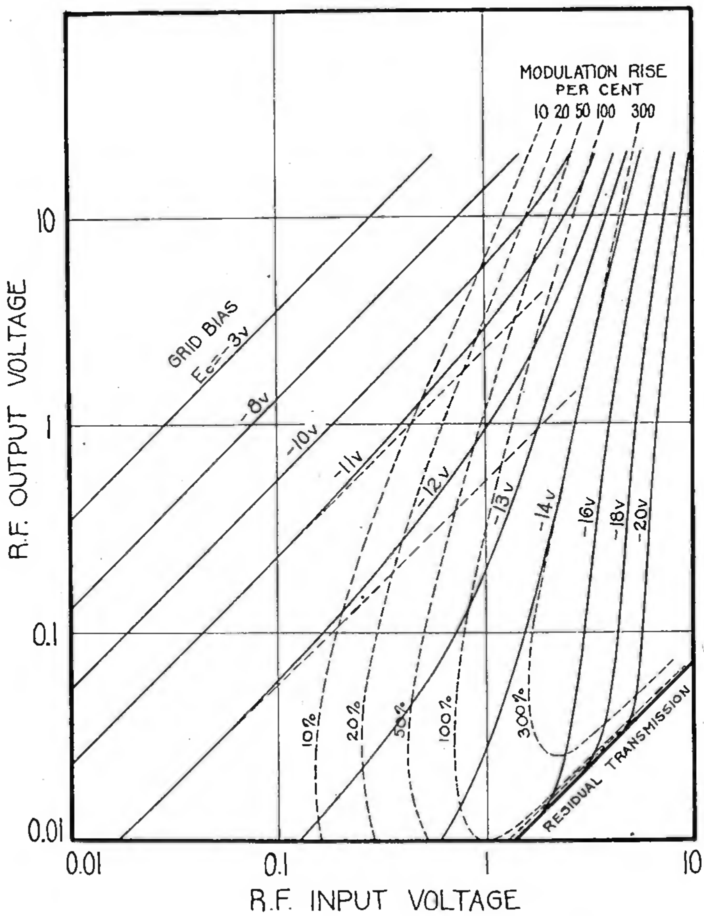

Fig. 2—Input-output voltage diagram for a single tuned r-f amplifier stage employing type 24 tetrode (E_b=180 v; E_d=90 v). The dotted lines are contours of constant modulation rise.

over by overloading due to grid current in the final tube. Above −9-volt bias, however, the output-input relations become nonlinear and an attempt to operate over this range of input voltages will result in a distorted output. This distortion is most severe for high degrees of modulation. The curved characteristics shown also have the

[1] This discussion is contained in a paper entitled "Recent developments in RFL broadcast receivers," presented at a meeting of the Philadelphia Section, Institute of Radio Engineers, held in Philadelphia, November 7, 1929.

effect, for small degrees of modulation, of effectively increasing the degree of modulation. At $E_c = -11.5$ v, for example, a signal of a certain amplitude having a modulation of 5 per cent will have its modulation increased to about 70 per cent in passing through the amplifier. This rise in modulation shall be referred to as *modulation rise.*

For a more detailed investigation and analysis of this problem the input-output relations for a single stage are of fundamental importance. Fig. 2 shows the experimental relation between the input and output voltages of a single stage employing a shielded tetrode of the 24 type ($E_b = 180$ v, $E_d = 90$ v), and a conventional tuned transformer. Logarithmic voltage scales are used in order to cover the wide ranges of magnitudes involved. Each curve represents the output-input voltage relation for a specified control-grid bias. Each curve comprises a portion over which the output varies linearly with input and a curved portion over which the output increases at a greater rate than the input. For small values of grid bias the curved portion practically disappears, but as the bias increases negatively this part continuously encroaches upon the straight portion.

The straight lines in the lower right-hand corner of Fig. 2 marked "residual transmission" represent the signal which leaks through the stage independently of the repeater action of the tube. This is of course linear. Such residual transmission may arise in several ways. In a shielded tetrode stage without provision for balancing out the coupling due to the small plate to control-grid capacity, it would arise from coupling through this capacity. In the case of an ordinary triode with a balancing network, it may arise from a slight unbalance, intentional or accidental.[2] It may also exist, or be intentionally provided, by coupling in the associated circuits.

The dotted lines on Fig. 2 represent the voltages at which various degrees of modulation distortion due to curvature, as measured by the percentage rise in modulation, occur. A series of such contours of constant modulation rise have been drawn and are useful in predicting the over-all distortion in an amplifier composed of several stages by a graphical method which we shall describe later. A reversal of the course of these contours occurs in the region in which the residual transmission is of importance due to the promotion of linearity by this linear coupling.

Inspection of Fig. 2 shows that the range over which the amplification of the stage can be varied without exceeding a given distortion decreases as the input voltage increases. For this reason, wherever

[2] A detailed discussion of the effects contributing to residual transmission and methods of rendering an amplifier stage perfectly monodic have been given by Ballantine, U. S. Patent No. 1,760,871.

possible, the burden of control in a multistage amplifier should be placed upon those stages in which the input voltages are low. We have found that a considerable extension of the range of distortionless control can be secured by means of a judicious graduation of control bias among the stages of a multistage radio amplifier and this method has been successfully used in broadcast receivers for some time. It is especially valuable where a linear detector,[3] requiring a high input voltage, is employed. In the latter case the stage preceding the high voltage detector is left out of control altogether, or the variation of its control-grid bias is considerably reduced in comparison with the ranges of variation of the earlier stages.

Fig. 3 shows the audio output versus antenna carrier voltage for an old experimental receiver employing an automatic volume control and

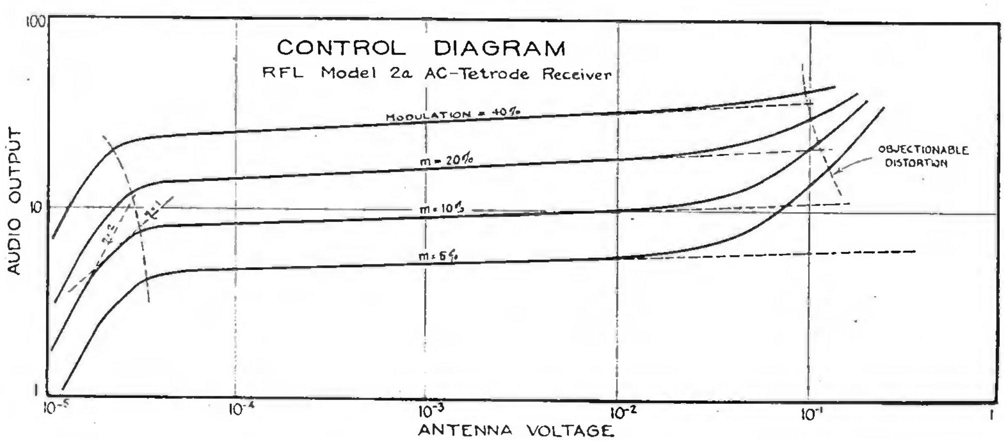

Fig. 3—Relation between audio output and antenna voltage in a receiver employing automatic volume control, exhibiting modulation rise and distortion at high input voltages.

high voltage linear detector with a fully graduated grid bias in the radio stages. The rise in audio output at the higher antenna voltages is due to modulation rise, as mentioned above, since the carrier voltage at the detector is maintained at a constant level by the automatic volume control. This rise is more marked for the lower values of modulation, as one should expect. There is no simple relation between the actual distortion produced, as measured by the generation of harmonics in the case of a pure tone modulation, and the modulation rise. At low modulations a considerably greater increase in the effective modulation can occur before the distortion (introduction of extraneous tones) becomes objectionable than in the case of high modulation. This is suggested qualitatively by the curve in Fig. 3 marked "objectionable distortion."

[3] Ballantine and Hull, U. S. Patent No. 1, 698,668; Ballantine, Proc. I. R. E., 17, 1153; July, 1929.

The modulation distortion, or rise in modulation, exhibited by Fig. 3 is a useful criterion in studying and comparing tubes for the curvature effects under consideration. Such a curve may be constructed by taking a cross-section of the stage diagram of Fig. 2, that is to say, the points of intersections of a line of constant output with the dotted contour lines for constant modulation rise. The effective modulation rises at first and then returns to the original level. This decrease is due to the effect of residual transmission through the stage which, being linear, reduces the distortion. A maximum is obtained by crossing the Fig. 2 stage diagrams at constant output, starting in the region of linear response to the left and passing through the region of curvature to the linear residual transmission curve at the right.

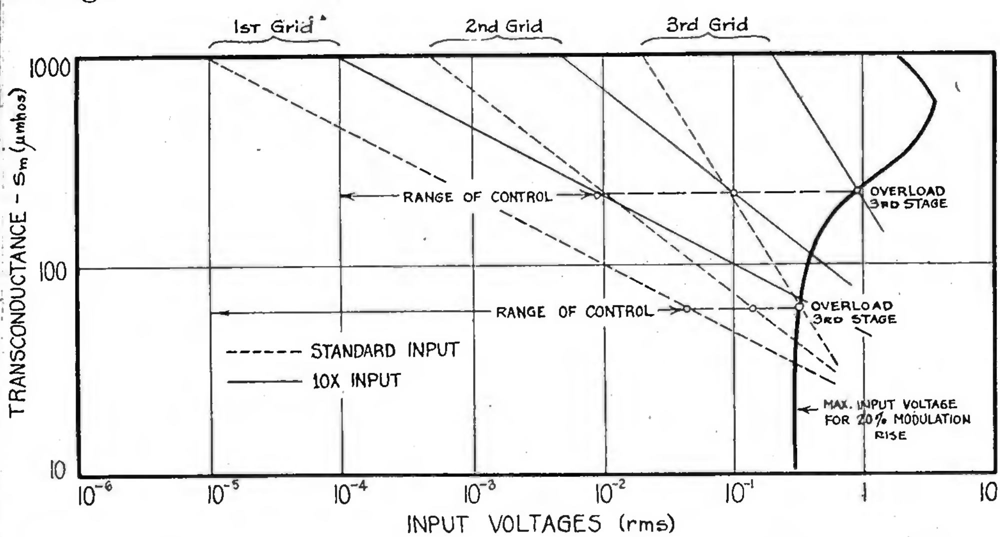

Fig. 4—Control analysis: Diagram of input voltages in a 3-stage amplifier employing 24 tetrodes; controlled by common grid bias in all stages.

A graphical method of applying the data for a unit stage to the prediction of the performance, control characteristics, overload, and selection of control bias graduation in an amplifier comprising several such stages, is shown in Fig. 4. In this diagram the values of grid bias do not appear explicitly but are represented by the corresponding values of transconductance (mutual conductance) s_m of the tube, which are plotted as ordinates. The right-hand curve, marked "maximum input voltage for 20 per cent modulation rise" represents the values of input voltage at the specified values of s_m which result in a modulation rise of 20 per cent. These values may be obtained from Fig. 2 or by direct measurement. Fig. 4 represents the case of common bias control in three stages. The voltages at the grids of the various stages under standard conditions ($s_m = 1000$ micromhos) are obtained by the conventional methods of receiver measurement using a standard signal generator. These constitute starting points at the upper edge of the diagram. Since the standard output[4] (50 milliwatts) adopted for such measurements is somewhat low in terms of average listening conditions an audio voltage output of approximately ten times (100x power) is also represented. The significance of the straight lines will be apparent. At the grid of the 3rd stage, for example, when the bias has been adjusted so that the transconductance has been decreased from 1000 to 100 micromhos, or to one-tenth of its initial value, the voltage will have been increased ten times to preserve constant input to the detector; in the 2nd stage the voltage will have been increased 100 times for standard output; and in the 1st stage to one thousand times, and so forth. It will be observed that overload occurs first in the 3rd stage and limits the range of distortionless control. In the case of standard output the overload at the 3rd stage occurs when s_m equals 60 and the input voltage equals 0.05 volt. The corresponding range of control is 5000x. For 10x output, overload in the 3rd stage occurs when s_m equals 250 at an input voltage of 9×10^{-4} v, corresponding to a range of control (input voltage) of only 90x. The unsuitability of this method of control (i.e., common grid bias) when high voltage detectors are to be operated with 24-type tetrodes is thus apparent. The method was more satisfactory when small-signal square-law detectors were employed because the voltages involved were lower.

The advantage of a graduated bias control, in which the 3rd stage is not controlled and the first and second stages are controlled by a common bias, is brought out in Fig. 5. The other conditions here are identical with those contemplated in Fig. 4. In the case of standard output the 1st stage reaches overload first at point *A* for $s_m=6$ micromhos and an input voltage of 0.3 volt. The range of control is 30,000x. At 10x output, overload is reached simultaneously in the first two stages at point *B*. Since the modulation distortion curve is vertical in this region the maximum input voltage for 20 per cent distortion is about the same for 10x as for normal output. This is enormously greater than in the case of common bias illustrated in Fig. 4, where the range was only 90x. It will be seen that the best graduation of bias depends in general upon the level of voltage at which the detector is to operate, and upon the distortion contour of the tube. A further graduation of bias between stages 1 and 2 generally results in an additional improvement.

It should be pointed out that the representation of the stage input

[4] Report of the Standardization Committee, YEAR BOOK I.R.E., p. 107, 1929.

voltages in the preceding diagrams by straight lines assumes that the transconductance is independent of the voltage. In the range of voltages over which the input-output relations are nonlinear (see Fig. 2) this is not strictly true, and curved lines would be required in a perfectly accurate representation.

We may digress for a moment to consider how nonlinearity in the input-output relations of an amplifier tube arises. A rough explanation may be based on the case of a tube, such as a tetrode with negligible external impedance in the plate circuit, and for which the $i_p - e_g$ char-

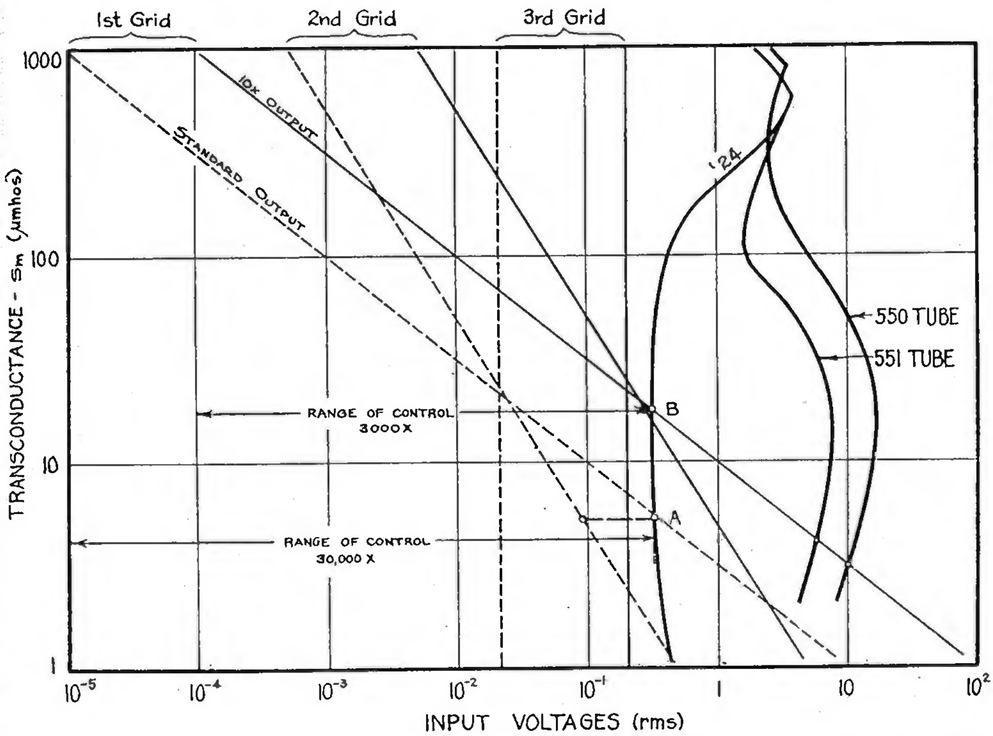

Fig. 5—Control analysis for graduated bias control; stage 3 uncontrolled; stages 1 and 2 controlled by common bias. Other conditions identical with Fig. 4.

acteristic is such that over a useful range of voltages the variable part of the plate current can be represented in a convergent series:

$$i_p = \frac{\partial i_p}{\partial e_g} e_g + \frac{1}{2} \frac{\partial^2 i_p}{\partial e_g^2} e_g^2 + \cdots . \qquad (1)$$

Now if $e_g = E_g \sin \omega t$ we find that effects of the 1st, 3rd, 5th . . . . orders contribute terms of fundamental frequency to which the tuned output circuit of the stage is resonant. Collecting these

$$i_p = I_p \cos \omega t = \left(\frac{\partial i_p}{\partial e_g} E_g + \frac{1}{8} \frac{\partial^3 i_p}{\partial e_g^3} E_g^3 + \frac{1}{192} \frac{\partial^5 i_p}{\partial e_g^5} E_g^5 + \cdots \right) \cos \omega t. \qquad (2)$$

The relation thus departs from linearity when the effects in E_g^3, E_g^5 . . . . etc., become of importance.

Assume for the moment that conditions are such that most of the distortion can be accounted for by the 3rd-order term. In these circumstances the modulation rise will be

$$D = \frac{m^1}{m} = \frac{\partial I_p}{\partial E_g} \Big/ \left[\frac{\partial I_p}{\partial E_g}\right]_0 = 1 + \frac{E_g^2}{4s_m} \frac{\partial^2 s_m}{\partial e_g^2} \tag{3}$$

approximately, for small m (degree of modulation).

It is of interest to see what sort of $i_p - e_g$ characteristic would correspond to a constant distortion for all values of E_g (grid bias). Assuming control by bias variation in one tube E_g is allowed to vary and the grid bias is to be adjusted to give a value of s_m such that the output I_p is constant. That is

$$E_g = I_p/s_m. \tag{4}$$

Substituting this in (3) we have to solve the equation

$$\frac{1}{s_m^3} \frac{\partial^2 s_m}{\partial e_g^2} = k = \text{const}. \tag{5}$$

A general solution under the condition that $\partial s_m/\partial e_y = 0$ when $s_m = 0$ is

$$s_m = \sqrt{\frac{2}{k}} \frac{1}{e_0 - e_g}$$

where e_0 is a constant. The form of this function is obvious. It differs considerably from that of ordinary tubes.

II. The Problem of Cross-Talk

The term *cross-talk* is employed here to designate that species of interference which originates in the r-f amplifier tubes by modulation between two or more signals. These modulation effects depend upon the higher-order curvature parameters of the tube and are to that extent related to the problem of distortion which has just been discussed. It is to be expected, therefore, that characteristics and applied voltage amplitudes which conduce to distortion will be commensurably troublesome from the viewpoint of cross-talk and any improvement which can be effected in a tube to reduce distortion will also reduce a large part of the cross-talk.

There are several kinds of cross-talk, of various degrees of practical importance. For our present purposes we shall distinguish between two main classes.

In the first class (Class A) we shall include effects due to beating between two signals whose frequencies differ from each other and from that to which the the receiver is tuned. Examples of this are the cases where signal frequencies f_1 and f_2 form combination frequencies of $f_1 \pm f_2$ within the tuning range of the receiver. Combinations of harmonics also occur. Effects of the $f_1 \pm f_2$ type are due to even-order effects in the tube, and depend mainly on the second derivative of the tube characteristic. Since distortion is not influenced by this derivative, (except when the effect of the external load is important) this type of cross-talk is not directly related to the production of distortion. Nevertheless it is found experimentally that the measures which we have found effective in reducing the third and other derivatives and so shape the characteristic as to reduce distortion generally reduce the second derivative and combination cross-talk also.

In a second class (Class B) we shall include cross-talk which is heard in circumstances of the following type. The receiver is tuned to a signal, which we shall call the *desired signal.* A second signal, called the *extra signal,* of different frequency, is also present at the grid of the first tube. We shall also suppose that the selectivity of the receiver and the frequency difference between the signals are such as to exclude the possibility of the extra signal being heard by simple interference in the absence of the desired signal. If the desired signal be modulated both signals are head simultaneously; if unmodulated the extra signal is heard.

This type of cross-talk depends mainly upon the *third-order* derivative of the tube characteristic, that is, upon the same derivative that causes distortion. A sufficiently good mathematical theory, accounting for the more important experimental facts, can be developed by a method previously employed in developing a theory of high voltage detection.[5]

For facility of expression we shall employ a term,called the *cross-talk factor,* which is defined as the ratio of the audio output due to cross-talk to that due to the desired signal. In many cases it will be convenient to specify further that the modulations of the two signals are equal. Such a factor should be found useful as an additional index of receiver performance.

Consider the simplified case where the desired and extra signals produce at the grid of the tube voltages of the forms

$$\text{Desired signal: } e_1 = E_1 \cos \omega_1 t \tag{7}$$

$$\text{Extra signal: } e_2 = E_2(1 + m_2 \sin at) \cos \omega_2 t. \tag{8}$$

[5] P. 1161 of footnote 3.

The mathematical theory can be developed along the lines of the following physical picture.

Fig. 6 represents the relation between the amplitude of the r-f output current of the tube and the grid bias with the desired signal (7) impressed on the grid. This relation is derivable from the input-output data of Fig. 2 and is in general nonlinear. We shall now regard the extra signal (8) as varying the grid voltage about the point E_c which represents the grid bias. In order that the output of the tube shall be audible after detection, it should be of the modulated signal type, that is, it should contain terms of the general form

$$E(t)E_1 \cos \omega_1 t \tag{9}$$

where $E(t)$ is an audio-frequency variation. An output of this type would be produced directly by the application of an audio-frequency voltage $E(t)$ to the grid. This is a familiar type of modulation, depending mainly on the *slope* of the relation shown in Fig. 6, and occurs in a-c operated receivers as "hum in the presence of carrier" when hum voltages are inadvertently present in the grid or screen-grid circuits. An output of type (9) is also produced by a voltage of the type of the extra signal (8) when the relation shown in Fig. 6 is *nonlinear*. This is the origin of Class B cross-talk.

This viewpoint may be put in mathematical form by employing the power series representation of the tube characteristic, given in (1), neglecting external impedances in the plate circuit. Inserting the signal voltages (7) and (8),

$$e_g = e_2 + E_1 \cos \omega_1 t, \tag{10}$$

in (1) and expanding we have

$$\begin{aligned} i_p = {} & D_1(e_2 + E_1 \cos \omega_1 t) + D_2(e_2^2 + 2e_2E_1 \cos \omega_1 t + E_1^2 \cos^2 \omega_1 t) \\ & + D_3(\cdots) + \cdots \end{aligned} \tag{11}$$

$$\text{where } D_n = 1/n!\, \partial^n i_p / \partial e_g^{\,n}.$$

Now collecting coefficients of e_2

$$\begin{aligned} i_p = {} & D_1E_1 \cos \omega_1 t + \tfrac{3}{4} D_3E_1^3 \cos \omega_1 t + \cdots \\ & + e_2(D_1 + 2D_2E_1 \sin \omega_1 t + 3D_3E_1^2 \sin^2 \omega_1 t + \cdots) \\ & + e_2^3(D_3 + \cdots) + \cdots . \end{aligned} \tag{12}$$

In type B cross-talk we are interested only in the part of the plate current of frequency ω_1. Collecting these terms

$$i_p = \left\{\left(D_1 + \frac{3}{4}D_3E_1^2 + \cdots\right) + (2D_2 + \cdots)e_2 + \left(3D_3 + \frac{15}{2}E_1^2D_5 + \cdots\right)e_2^2 + (\cdots)e_2^3\right\} E_1 \cos \omega_1 t. \tag{13}$$

The physical significance of this formulation may be brought out by observing that this is precisely the form of expression we would have obtained by starting with the characteristic shown in Fig. 6 and developing it in a power series in the grid voltage. The coefficients in (13) should thus be equal to the successive differential derivatives of (2), which is the power series representation of the Fig. 6 relation. This will be seen to be true by differentiating (2). A more accurate method, of course, consists in starting with the experimentally determined relation of Fig. 6, and developing it in a power series in the grid voltage.

Now let us further develop (13) by substituting the value of e_2, the extra signal, (8). Terms in e_2^2, e_2^4, etc., will yield output currents of the type (9). Collecting terms of this type we have

$$i_p = \left\{\left(D_1 + \frac{3}{4}D_3E_1^2 + \cdots\right) + \frac{E_2^2}{2}\left(3D_3 + \frac{15}{2}D_5E_1^2 + \cdots\right)(1 + 2m_2 \sin at + m_2^2 \sin^2 at) + \cdots\right\}E_1 \cos \omega_1 t. \tag{14}$$

Neglect the term in $\sin^2 at$ if m is small. It will be seen that (14) is of the form of a standard modulated signal of the type $i_p = I_p\,(1+m \sin at) \cos \omega t$.

Retaining the first terms only we have approximately

$$i_p = s_m E_1\left(1 + \frac{E_2^2}{2s_m}\frac{\partial^3 i_p}{\partial e_g^3} \sin at\right) \cos \omega_1 t \tag{15}$$

where $s_m = \partial i_p/\partial e_g =$ the transconductance. The audio output after detection of such a signal by a square-law detector will be proportional to

$$\frac{s_m E_1^2 E_2^2 m_2}{2}\frac{\partial^3 i_p}{\partial e_g^3}. \tag{16}$$

The cross-talk factor can be computed by assuming the desired signal to be modulated to the same degree of modulation. Then the ratio of cross-talk output to desired output at the detector will be

$$\text{Cross-talk factor} = \frac{E_2^2}{2s_m} \frac{\partial^3 i_p}{\partial e_g^3}. \tag{17}$$

The third-derivative of the tube characteristic again appears as a conspicuous cause of cross-talk of the type under consideration. It is interesting to note that to the order of approximation represented by this equation the cross-talk varies as the square of the extra signal and is independent of the desired signal voltage *per se*, except in so far as this determines the point of the characteristic and values of s_m, etc., at which operation takes place.

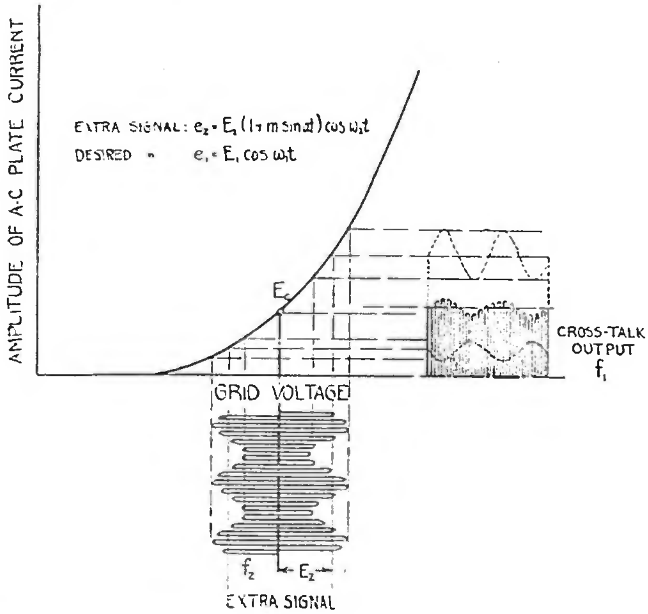

Fig. 6—Illustrating the production of Class B cross-talk. Heavy curve represents the relation between amplitude of plate current of desired-signal frequency and grid bias, with the desired signal impressed on the grid.

The similarity of the above expression to that for the distortion (3) may be noticed. The same tube parameters are involved and the cross-talk varies with the extra signal in the same way that distortion varies with the input voltage.

Turning now to the experimental aspects, Fig. 7 shows the relation between the cross-talk factor for an average 24 tetrode (dotted curves) and the new type 551 tetrode (solid curves) and the input voltage at a fixed output voltage level, and for several fixed values of extra signal. The modulations of the two signals were the same (30 per cent). The output voltage level was such that 1000 microvolts were required at normal bias (-3 v; $s_m = 1000$ micromhos) to produce it. A sufficient attenuation between this stage and the one following it was used to avoid any cross-talk effects in the second stage. The desired signal frequency was 1000 kc and that of the extra signal was 900 kc. This frequency difference was not critical at the value employed, but was sufficient to preclude direct interference on the one hand, and combination cross-talk of Class A on the other. Both desired and extra signals were applied directly to the grid of the tube. The measurement procedure was as follows: The desired signal, modulated at 30 per cent, was set at 10^{-3} volts with a bias on the tube corresponding to a transconductance of 1000, and the amplifier adjusted for a standard, or *fiducial*, audio-voltage output. The modulated desired signal was then set at a desired value and the bias of the tube

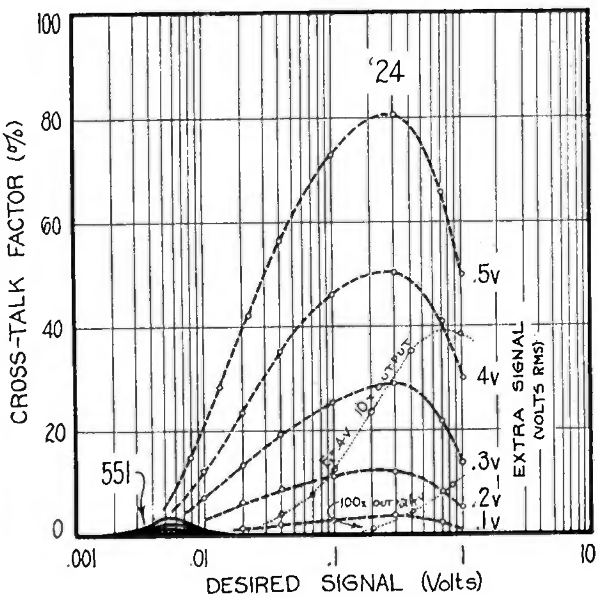

Fig. 7—Experimental curves showing cross-talk production in 24 and 551 tetrodes. Constant output maintained, corresponding to 0.001 volt desired signal at -3-volt bias.

adjusted for fiducial output. The modulation of this signal was then removed (leaving the carrier), and the modulated extra signal was then superposed. The audio output, corresponding to the cross-talk, is then noted and expressed as a percentage of the fiducial audio output, which is taken as 100 per cent.

These experimental curves confirm (17) in so far as it predicts a variation of cross-talk with the square of the extra signal voltage at a given operation point. The existence of a maximum at high desired signal voltages may be accounted for by the increase in modulation distortion and effective s_m which would produce an abnormal direct output of the desired signal, thus lowering the measured cross-talk factor.

The effect of changing the output level in the case of the 24 tube is also shown in Fig. 7, the extra signal being maintained at 0.4 volt. Two extra curves are shown for outputs corresponding to input voltages of 10^{-2}, 10^{-1} volts at -3-volt bias. The curves are roughly similar in shape and start up at points on the desired signal axis which are multiples of 10, or roughly at points at which the bias is so adjusted that the transconducance is the same. This confirms the approximate independence at low voltages of the variation of cross-talk with respect to the desired signal voltage which is expressed in (17). This independence does not persist at high desired signal voltages, as the drooping of the maxima with increasing voltage shows.

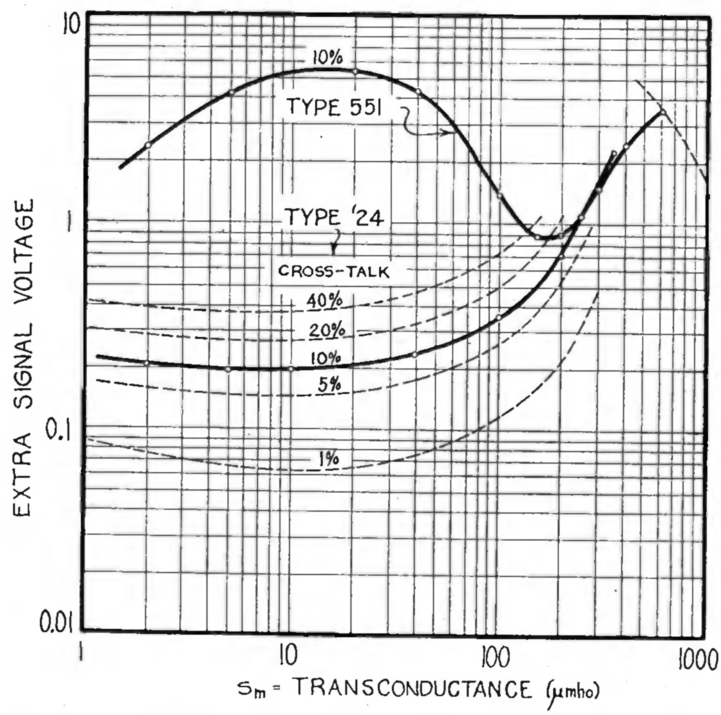

Fig. 8—Diagram showing relation between tube transconductance, produced by grid bias variation, and maximum voltage of interfering signal for various percentages of cross-talk.

For cross-talk analysis in a multistage amplifier we have employed a graphical method similar to that used for distortion analysis which was illustrated in Figs. 4 and 5. This is based on the fact that the cross-talk is approximately independent of the desired signal voltage. The basis of the method is a contour curve for the tube which shows the relation between the extra signal voltage which has to be applied to the grid to attain a given percentage of cross-talk, and the operating point on the characteristic, as represented by the transconductance at that point. A typical set of such contours is shown in Fig. 8. The solid curves represent 10 per cent cross-talk for a type 24 tube and a type 551 tube. Several other contours are given for the 24 and represent the extra signal voltages required for cross-talk factors of 1, 5, 20, and 40 per cent. The desired signal voltage varied from 10^{-3} at $s_m = 1000$ to 1 volt at $s_m = 1$. To avoid confusion only the curve for 20 per cent cross-talk is given for the 551 tube; the positions of the contours for other values are easily found by applying the square-law relationship whose approximate validity appears from the curves for the 24 tube.

The reduction in cross-talk at higher desired signal voltages which has been achieved in the type 551 tube will be seen to be considerable. At $s_m = 2.5$ micromhos, for example, the ratio of extra signal voltage for 10 per cent cross-talk amounts to about 23 to 1. This means that with a given extra signal voltage the cross-talk for the type 551 will be only 1/500th of that for the 24 type.

Figs. 7 and 8 bring out the important fact that for reduction of cross-talk attention must be paid to the tube parameters at large transconductance (low negative bias), whereas in the case of distortion the control diagrams of Figs. 4 and 5 show that their proper adjustment at lower values of s_m is of more importance.

Such limiting curves resemble very closely those which represent the relation between input voltage for a given distortion and transconductance (See Figs. 4, 5, and 16).

A typical cross-talk analysis diagram is shown in Fig. 9. The case considered is a 3-stage amplifier controlled by common variable control-grid bias in stages 1 and 2. The maximum amplification per stage is 40x and is attained at a grid bias corresponding to a transconductance of 1000 micromhos in the tubes. At the right-hand part of the diagram are plotted the cross-talk contours from Fig. 8. The output level (2.5 watts) corresponded to 10^{-3} volts at the grid of the first tube at 30 per cent modulation. The desired signal voltages are show by solid line 1. The extra signal voltages at the 2nd grid are based on the following assumptions:

(1) That the frequency difference between the signals is 20 kc; (2) that the selectivity of the two tuned circuits beyond the 2nd grid is such that a transmission of 0.5 is obtained for the extra signal; (3) that the attenuation in the tuned circuit between the 1st and 2nd stages provides a transmission of 0.2 for the extra signal.

The lines 3, 4, and 5 represent the voltages at grid 2 for extra signals of 0.1, 0.2, and 0.5 volt at the 1st grid.

Consider the cross-talk in the first tube (24) with an extra signal E_2 of 0.2 volt. As the desired signal E_1 increases cross-talk of 1 per cent is reached at point A at which point the desired signal reaches 0.03

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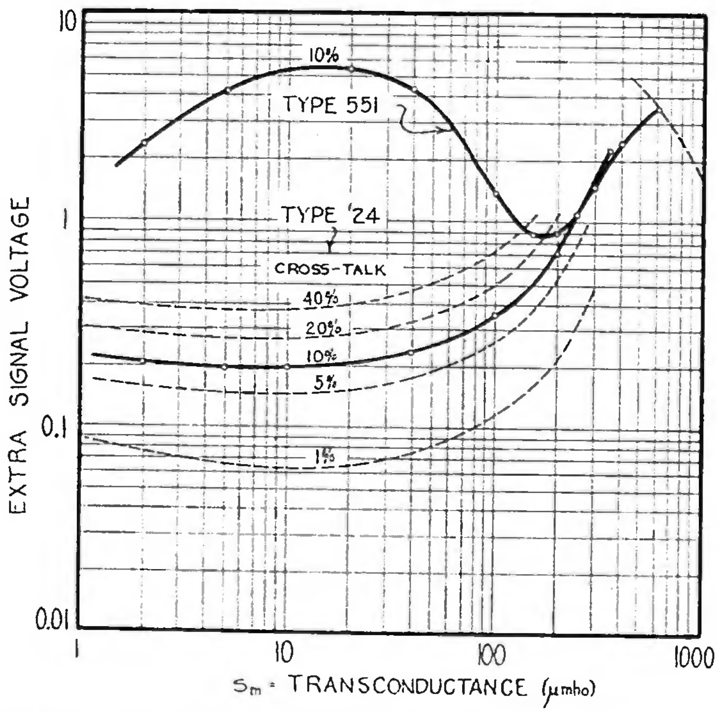

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Such limiting curves resemble very closely those which represent the relation between input voltage for a given distortion and transconductance (See Figs. 4, 5, and 16).

A typical cross-talk analysis diagram is shown in Fig. 9. The case considered is a 3-stage amplifier controlled by common variable control-grid bias in stages 1 and 2. The maximum amplification per stage is 40x and is attained at a grid bias corresponding to a transconductance of 1000 micromhos in the tubes. At the right-hand part of the diagram are plotted the cross-talk contours from Fig. 8. The output level (2.5 watts) corresponded to 10^{-3} volts at the grid of the first tube at 30 per cent modulation. The desired signal voltages are show by solid line 1. The extra signal voltages at the 2nd grid are based on the following assumptions:

(1) That the frequency difference between the signals is 20 kc; (2) that the selectivity of the two tuned circuits beyond the 2nd grid is such that a transmission of 0.5 is obtained for the extra signal; (3) that the attenuation in the tuned circuit between the 1st and 2nd stages provides a transmission of 0.2 for the extra signal.

The lines 3, 4, and 5 represent the voltages at grid 2 for extra signals of 0.1, 0.2, and 0.5 volt at the 1st grid.

Consider the cross-talk in the first tube (24) with an extra signal E_2 of 0.2 volt. As the desired signal E_1 increases cross-talk of 1 per cent is reached at point A at which point the desired signal reaches 0.03

volt. As E_1 further increases the cross-talk increases to about 5 per cent. In the second tube cross-talk of 1 per cent is reached at point B for a desired signal voltage even smaller than 0.03 volt, i.e., at E_1 about 0.01 volt. This increases to about 2.5 per cent at point C at 10^{-1} volt and this is followed by a decline. These variations in the cross-talk in the two stages are shown by the heavy curves at the bottom of the figure. It will be seen that under certain conditions the second stage of the amplifier produces more cross-talk than the first stage and its contribution is often of importance. It is also apparent that the cross-talk in the first stage increases more slowly when the

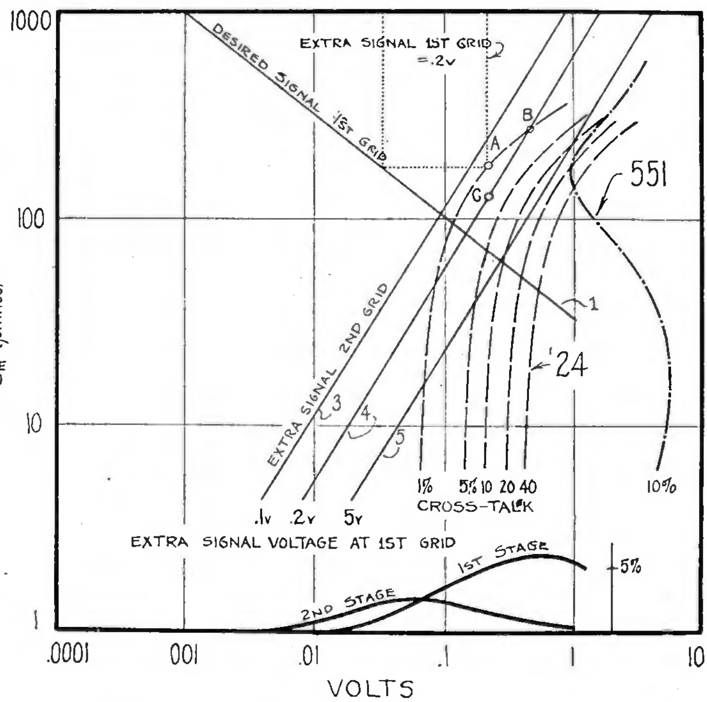

Fig. 9—Cross-talk analysis in 3-stage amplifier. Lower part of the diagram shows cross-talk contributed by 1st and 2nd stages. Volume controlled by common bias on first two tubes.

first two tubes are controlled than when only one is controlled (Cf. Fig. 7). This results in less cross-talk at high desired signal voltages because of the bending over of the cross-talk curve by the effect of modulation distortion. This example will sufficiently illustrate the use of this method in determining the cross-talk contributions of the several stages under various types of control.

Before leaving this subject a few words may be said about the practical aspects of this problem in radio receiver design. In order to study the effect in the tube we have postulated certain voltage amplitudes due to the desired and extra signals at the grid terminals. In a practical receiver at least one tuned (antenna coupling) circuit would precede the first tube, so that the selectivity of this circuit may be relied upon to a certain extent to attenuate the extra signal. It is found, however, that in practical installations this is not adequate. To achieve further reduction a number of receiver designers have had recourse to a double tuned cicuit, or double preselector. By virtue of the second-order relationship between cross-talk and extra signal voltage which has been found to hold over a wide range of voltages it may be expected that a double preselector, giving approximately geometrical selectivity, will be T^2 more effective than a single resonant circuit (T = transmission for the interfering signal). For example, if the selectivity of a single circuit is such as to give a transmission of 0.25 at a frequency difference of 20 kc, the reduction in cross-talk with two such circuits over that of a single circuit will be of the order of 16 times.

The use of such a double-resonant antenna coupling circuit is, however, attended by several disadvantages. The transmission at resonance is reduced over that of a single circuit, with the result that the sensitivity is impaired and the fluctuation[6] noise-signal ratio is increased. It is more economical to employ the tuned circuit in the radio amplifier because of the fact that when it is used in the antenna coupling system an untuned stage generally has to be employed to restore the sensitivity. In the superhetrodyne employing an intermediate frequency of an order of 200kc a certain minimum preselection before the first detector is necessary to reduce "image response;" but here also from the viewpoint of fluctuation noise-signal ratio, the additional tuned circuit is better located in the carrier-frequency amplifier than in front of it. In low-priced receivers of the "midget variety," which have lately come into vogue, and in which economy is paramount, the use of an extra tuned circuit for this purpose is quite undesirable.

The 551 tube not only renders unnecessary the use of double preselectors of this type but in most circumstances is considerably more efficacious in reducing cross-talk. This will be apparent from the experimental data which have been presented above. Over a large range of signal voltages (see Figs. 7, 8, and 9) the cross-talk in the 551 tube is of the order of 1/100th to 1/500th that of the ordinary 24 tube. In order to obtain this same improvement with 24 tubes by means of a preselector comprising a chain of resonant circuits of normal selectivity, with an extra signal 20 kc from the desired signal, at least *three*

[6] F. B. Llewellyn, Proc. I. R. E., 18, 243; February, 1930. Stuart Ballantine, *Contributions from the R.F.L.*, No. 21; Proc. I. R. E., 18, 1377; August, 1930. P. O. Farnham and A. W. Barber, *Contributions from the R. F. L.*, No. 20; Proc. I. R. E., 18, 1338; August, 1930.

tuned circuits would be required; or to achieve the same reduction with a preselector comprising two tuned circuits, a frequency displacement amounting to 100 kc would be required. Comparative tests between receivers employing biresonant preselectors and ordinary tetrodes and receivers employing but one tuned antenna coupling circuit (and in one or two cases, none) and equipped with 551 tetrodes have been made in many localities where cross-talk is particularly troublesome and have added practical confirmation under actual operating conditions.

III. Description and Characteristics of Type 550 and 551 Tetrodes

Our solution of the problems of distortion and cross-talk in vacuum tubes consists essentially in reshaping the plate current-grid bias characteristic in such a way as to minimize the higher-order curvature, particularly at points where the transconductance is low. Of the many experimental tubes which have been tested in the course of this development two have been selected for commercial production and have been designated as Types 550 and 551. Both tubes are designed to operate at the same plate and screen voltages as the ordinary 24 type. Their characteristics, such as plate resistance and transconductance, are the same at normal operating voltages as those of ordinary tetrodes of present commercial types.

The Type 550 tube is capable of handling an input voltage of approximately 15 with less than 20 per cent modulation rise, or about 50 times that of the standard 24 type; the Type 551 tube is capable of handling 7 volts, or 25 times that of the 24 type. Both tubes are fundamentally of the shielded tetrode type, although the principle by which the reduction of distortion has been achieved can be applied equally well to triodes and other types of structure. This principle may be explained as follows:

Fig. 10 shows the grid-plate transfer characteristic of an ordinary tetrode of the 24 type. To raise the input voltage at which distortion appears the design could be changed so as to decrease the mu-factor of the tube, thus spreading out the range of grid voltages and obtaining a characteristic of the type shown at (A). This, however, entails a sacrifice in transconductance, and amplification, at the initial bias if the plate current at this point is adjusted to the same value. If the transconductance at the initial bias is to be maintained at the same value instead, then an increase in plate current must be tolerated as shown by the characteristic (B).

A way out of this difficulty is illustrated in Fig. 11. The amplifier stage shown in Fig. 11a comprises two tubes in parallel. Of these tubes (A) is of the high-mu type, and (B) is of the low-mu type. A greater number of tubes, of properly graduated mu's may also be employed. The effective $i_p - e_g$ characteristics of tubes A and B are

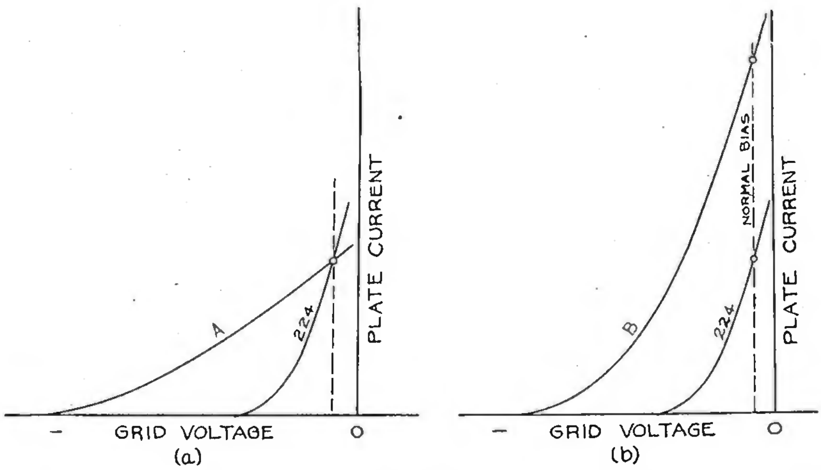

Fig. 10—Showing (a) decreasing transconductance and (b) increasing plate current accompaning attempts to decrease distortion by decreasing the mu-factor.

suggested by the dotted lines in Fig. 11b and their combined effect by the solid line. The high-mu tube A yields high amplification, but can handle only small input voltages; the low-mu tube B yields low amplification but can handle high input voltages. In combination the two

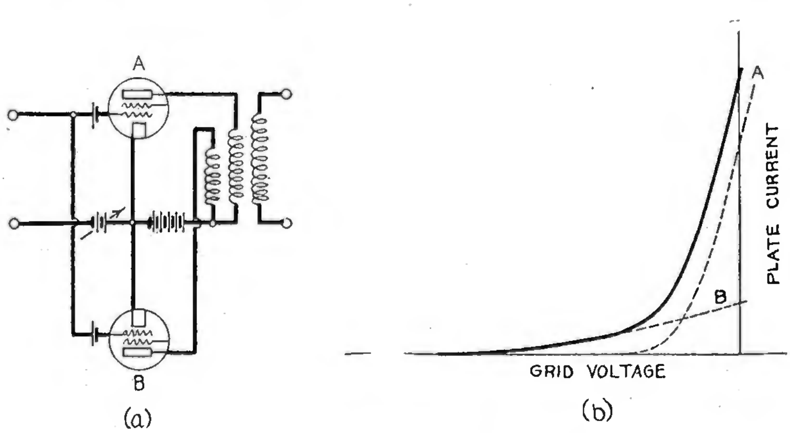

Fig. 11—Method of decreasing distortion and cross-talk.

tubes complement each other to produce a characteristic of the desired type. At low biases, where amplification is required and the applied voltages are low, both tubes are active, but due to its higher transconductance, most of the amplification is produced by A. As the bias increases negatively tube A is automatically cut off and the operation is gradually shifted to tube B, which is capable of handling the increasing input voltages. The arrangement may be regarded as one in which the mu-factor decreases continuously as the grid bias increases negatively to keep step with the increasing input voltage. The resulting $i_p - e_g$ characteristic (Fig. 11b) resembles that of ordinary tubes at low negative grid bias, but is extended, or "tailed" at the higher grid biases.

For reasons of economy it is desirable to incorporate this principle

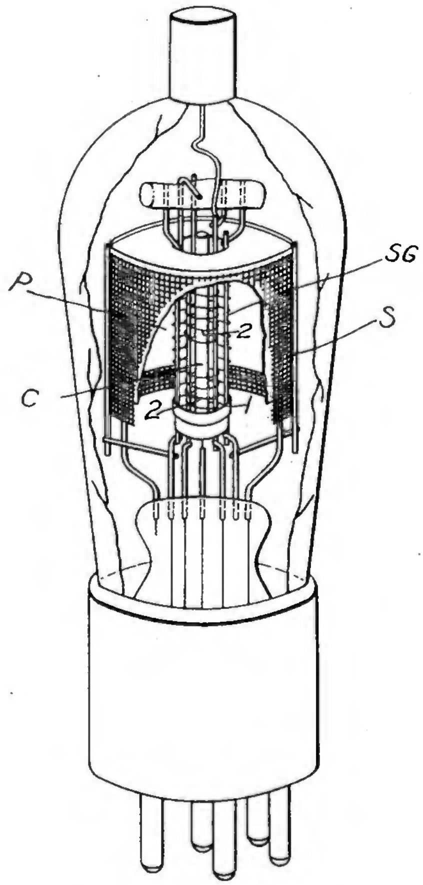

Fig. 12—Structure of tetrode of variable-mu type.

in a single tube structure. One way of accomplishing this is shown in Fig. 12, as applied to a shielded tetrode with an equipotential cathode. The control-grid is divided into two sections which are mounted with a gap between them. At low negative biases the entire cathode is operative and the tube has about the same characteristics it would have if the gap were not present; as the grid bias increases negatively the electron current through the upper and lower parts of the control-grid are cut off leaving a low-mu control through the gap. At these bias voltages the tube acts as if the upper and lower sections of the conrol-grid were formed of solid metal and controlled the current through he gap in the ordinary manner. Gaps may also be placed at the ends f the cathode (See Fig. 13g) instead of in the middle.

Some of the other structural embodiments of this principle which ave been contemplated and tested are shown in Fig. 13. (a), (b), c), and (d) show electrodes of variable diameter; (e) and (f) show

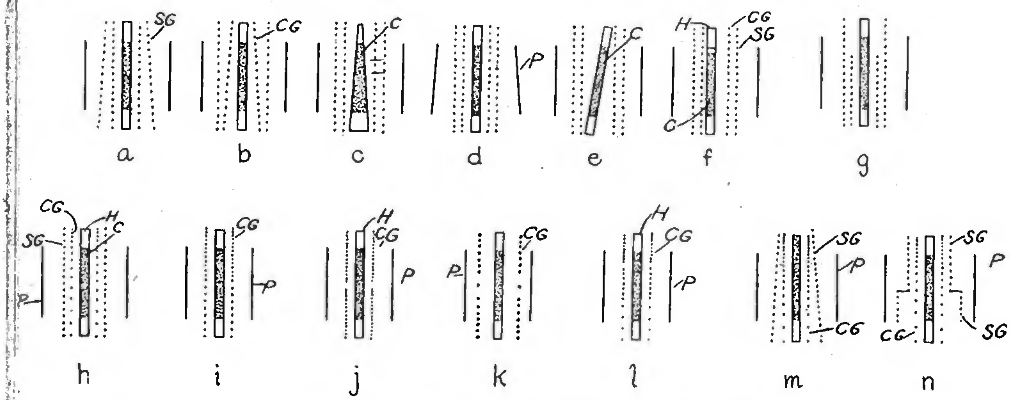

Fig. 13—Illustrating several types of variable-mu tube structure.

cathodes in tilted and eccentric positions; (h) and (i) show control grids of variable pitch in structures of the tetrode and triode types; (j) shows the mid-cathode control-grid gap of Fig. 11 applied to the triode; (k) shows a double gap and (l) a combination of gaps and variable grid pich; structure (m) is a combination of variable pitch and

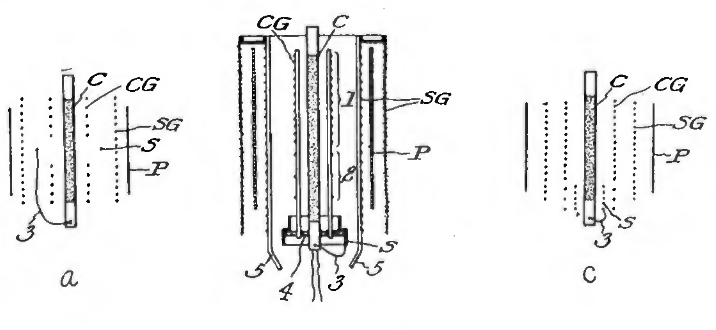

Fig. 14—Several types of variable-mu structure.

variable screen-grid diameter and (n) illustrates an abrupt change in electrode diameter as contrasted with the continuously variable diameters shown in some of the other structures. All of these structures provide a tube having a mu-factor which decreases continuously with increasing negative grid bias.

Somewhat more complicated types of structure having better economy of plate current at low grid bias are shown in Fig. 14. In these structures an additional low-potential element is provided for reducing the plate current at the portion of the electron stream which is controlled by the relatively open or coarse section of the control grid. The shield-ring S in Fig. 12 is connected to the cathode as shown at 3.

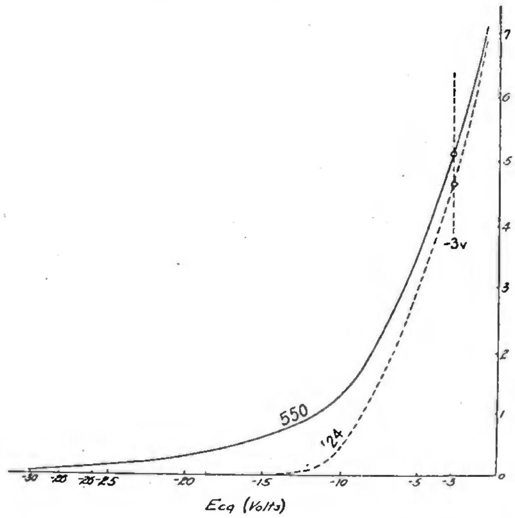

Fig. 15—Plate current-grid voltage characteristic of variable-mu tube showing reshaped characteristic for reduction of distortion and cross-talk.

The transfer characteristic ($i_p - e_g$) of the type 550 tetrode is shown in Fig. 15 together with that of the ordinary type 24 tube for

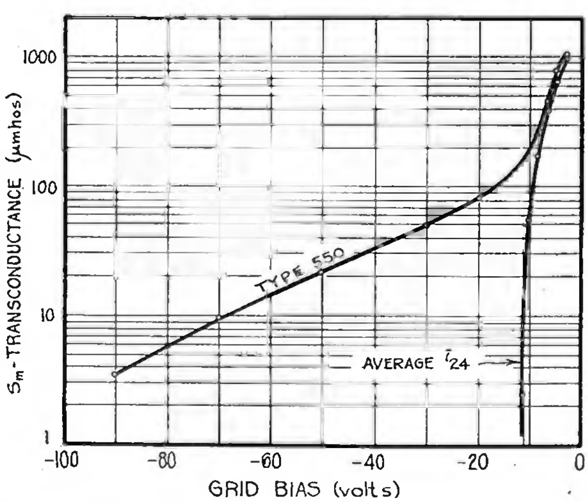

Fig. 16—Transconductance-grid voltage characteristic of Types 550 and 24 tubes.

comparison. The transconductance (mutual conductance) curves for the two types are shown in Fig. 16, as a function of the control-grid bias voltage. It will be observed that the transconductance of the type 550 tube is about the same as that of the type 24 at low grid bias. This result has been attained with an increase of plate current at this voltage of only 10 per cent. The other characteristics of the tube (plate resistance, etc.), are also approximately the same as those of the type 24 at this point ($E_c = -3$ v.).

The superior volume control characteristics of the 550 tube at high signal voltages have been shown in Figs. 4 and 5. Fig. 17 represents the input voltage which can be applied before the modulation rise reaches 20 per cent. It will be seen that the new tube is capable of handling input voltages approximately 50 times those which can be applied to the 24 type for the same distortion. An input voltage of approximately 15 volts is necessary to produce 20 per cent rise in modulation at $s_m = 15$ micromhos.

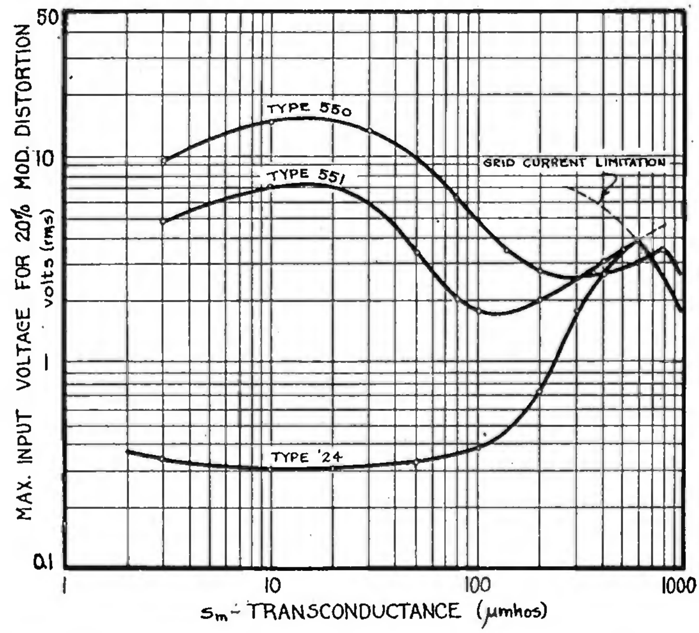

Fig. 17—Distortion limits for Types 550, 551, and 24 tubes; representing maximum input voltage which can be applied for 20 per cent modulation rise.

Fig. 16 shows that the grid bias required for control with the type 550 tube is several times as large (*ca.* 80 volts) as that required by the 24 (*ca.* 15 volts). It many receivers the provision of such a range of bias may be inconvenient, especially in a-c operated receivers, in view of the desirability of maintaining the screen-grid and plate voltages at their proper values with respect to the cathode. In broadcast receiver operation, the number of localities in which voltages as high as this are encountered is rather small. We have therefore designed a second tube, designated as type 551, having characteristics intermediate to those of the 550 and 24 types. This tube is especially suitable for use in radio broadcast receivers. Its structure is somewhat simpler than that of the 550 type. Approximately 25–30 volts grid bias are required for full control (i.e., control for carrier voltages up to the distortion point) as compared with 80 volts for the type 550. This is easily provided.

The distortion characteristics of the Type 551 tube are shown in Figs. 4 and 5. Fig. 17 shows the input voltage required to produce 20 per cent modulation rise. A maximum of 7 r-f volts can be applied at $s_m = 15$. The control diagram of Fig. 5 is of interest as indicating the improvement to be expected in a 3-stage amplifier with common bias control of the first two stages.

The cross-talk characteristics of the 551 tube have already been discussed and are shown in Figs. 7 and 8.

A comparative summary of the important characteristics of the types 550, 551, and 24 tubes is given in the following table:

TABLE I

Characteristic	Type 24	Type 550	Type 551
Plate voltage recommended	180 v	180	180
Screen voltage recommended	90	90	90
Normal grid bias	−3	−3	−3
Normal transconductance (micromhos)	1,000	1,000	1,000
Normal plate resistance (ohms)	400,000	300,000	400,000
Normal plate current (ma)	4	5.6	5.3
Grid bias for $s_m = 10$ micromhos	−13	−70	−30
Max. input voltage for 20 per cent distortion	0.3	15	7
Relative cross-talk factor (max.)	100%	0.5%	3%
Relative cross-talk at $s_m = 10$ micromhos	100%	none	0.12%

In each case the values are the averages obtained from a large number of tubes of regular commercial production.

This tabulation indicates the interchangeability of the 551 tube with the present 24 types in the radio-amplifier stages of a broadcast receiver. The only change necessary is an increase of the range of control-grid bias from about 15 to 30 volts. The ordinary method of inserting the grid bias in series with the cathode may be retained.

This tube is therefore not a "new tube" in the sense that it requires substantial readjustment of present types of receivers. It will make possible, in future receivers of both the ordinary and superhetrodyne types, a number of important economies not realizable heretofore on account of the necessity of protecting the present tubes from high voltage amplitudes and interfering signals. One of its important uses will be in receivers employing automatic or remote volume control, where it extends the range of control by a factor of 25. In such receivers the use of a "Local-Distance" switch would be contrary to the object of the control system.

IV. Manufacturing Aspects

A sufficient number of these tubes have been manufactured during the past year in three tube factories by regular production methods to verify that no extra cost or new manufacturing problems are introduced, and to secure evidence of reliability and longevity of the special characteristic. We were particularly anxious to check the longevity of the emission in the low-mu area, where the electron current density is somewhat higher than in the other parts. No evidence

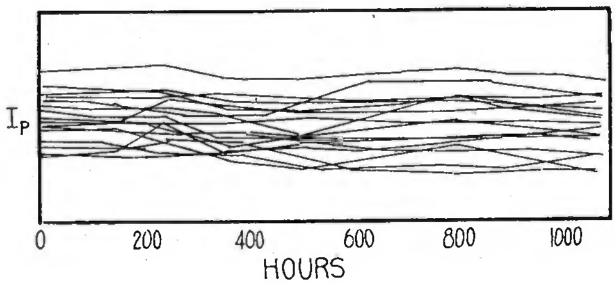

Fig. 18—Typical results of 1000-hour life test on current from low-mu area of Type 551 cathode.

of premature failure of the current in this area was observed, however, in life tests extending to 3000 hours. A typical 1000-hour run is shown in Fig. 18. This represents the current at a negative grid bias sufficiently high to restrict the current to the area under suspicion.

We wish to acknowledge our indebtedness to engineers of Radio Frequency Laboratories for receiver tests in the field; and to the Arcturus, Raytheon, and Grigsby-Grunow tube companies for coöperation in manufacturing studies.

Proceedings of the Institute of Radio Engineers
Volume 18, Number 12 *December, 1930*

SUMMARY OF PIEZO-ELECTRIC CRYSTAL CONFERENCE HELD BY U.S. NAVY DEPARTMENT, DECEMBER 3-4, 1929*

A CONFERENCE of representatives of various departments of the Government, of various radio manufacturers, and scientists interested in radio, was convened on December 3rd, 1929, in Washington under the auspices of the Bureau of Engineering, Navy Department, to discuss methods of obtaining accurate radio-frequency emissions from Naval radio transmitters by the use of piezo-electric crystals.†

It was pointed out that accuracy of emitted frequency in Naval radio transmitters is not only required to reduce interference with stations operated by other organizations, but also to obtain reliability and speed in communications within an organization, the various units of which are assigned the same frequency. The latter is a phase of the general problem of radio communications which is more particularly applicable to Naval communications and the demands are more severe than in commercial radio service.

For the purpose of bringing about a uniform understanding among the conferees, representatives of the Naval Research Laboratory offered for discussion:

(a) A resumé of the crystallography and piezo-electricity of quartz.
(b) A description of the methods of cutting and testing quartz used at the Naval Research Laboratory.
(c) An account of the experience and practices of the laboratory relative to the two principal "cuts" used in the art.

* Decimal classification: R060. Original manuscript received by the Institute July 23, 1930.

† Those present were: Rear Admiral H. E. Yarnell, USN, Captain A. I. Bass, USN, Commander E. C. Raguet, USN, Lt. Comdr. S. D. McCaughey, USN, Lt. Comdr. T. A. M. Craven, USN, Lieut. H. C. Rodd, USN, Lieut. W. B. Goggins, USN, Lieut. J. B. Dow, USN, H. Graf, G. W. Kelley, and J. W. Wright of the Bur. of Engineering, Navy Dept.; Comdr. Chas. H. Maddox, USN, Lt. Comdr, H. E. Fischer, USN, and Lt. E. E. Stone, USN, of DNC Office, Navy Dept.; Captain C. C. Cole, USMC, Capt. J. H. Gardner, USA, W. W. Reynolds, W. W. Ostein, Coast Guard; C. W. Hansell, RCA; V. M. Lucas, P. D. Andrews, I. F. Byrnes, A. L. Ellis of G. E. Co.; R. C. Hitchcock, J. B. Coleman, C. M. Hobart, D. G. Little of W.E.M. Co.; L. E. Whittemore, A.T. & T. Co.; F. R. Lack and E. L. Nelson, Bell Tel. Labs.; Dr. K. S. Van Dyke, Dr. W. G. Cady, Wesleyan University; W. E. Downey, Commerce; Dr. E. D. Tillyer, Amer. Optical Co.; J. K. Clapp and J. W. Horton, General Radio Co.; Dr. C. B. Jolliffe, R. B. Wright, Bur. of Standards; Comdr. M. A. Libbey, USN, Lt. Comdr. A. D. Douglas, USN, Dr. A. H. Taylor, Dr. L. P. Wheeler, Dr. L. H. Dawson, C. B. Mirick, L. A. Gebhard, R. B. Owens, J. E. Leahy, L. C. Young of N.R.L.; C. C. Kolster, I. L. Weston, L. C. Herndon, G. E. Sterling, A. Batcheler, and C. T. Manning of Dept. of Commerce.

It is unnecessary in this place to repeat the matter presented under (a) since more adequate descriptions of these topics are easily accessible in the literature on the subject.[1] It was pointed out that there is considerable discrepancy in the literature in matters of terminology and the hope was expressed that this conference would come to some agreement as to nomenclature in at least some of the instances.

The matter presented under (b) may be summarized as follows: A crystal as shown in the idealized drawing Fig. 1 is first mounted with the *Z* or optic axis horizontal. This is determined by so mounting

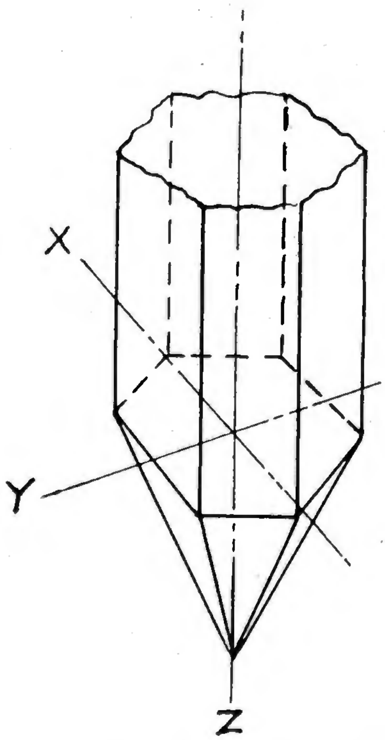

Fig. 1

the crystal that gang-saw blades will pass through the crystal in a plane determined by the "growth" lines which are usually found upon the faces of the crystal. Upon completion of this operation we have hexagonal slabs such as are shown in Figs. 2 and 3. These slabs are then examined under the polariscope for twinning and to determine how nearly the faces of the slabs are perpendicular to the optic axis.[2] They are then again mounted under gang-saws so that slabs such as those shown in the shaded portions of Figs. 2 and 3 are cut out. If

[1] See e.g., Sosman, The Properties of Silica, Chem. Cat. Co., 1927.
[2] A detailed description of these operations and of the polariscope used will be found in an article by Dawson, *Jour. Opt. Soc. Amer.*, 13, 517, 1926.

the slabs are mounted so as to yield the cut shown in Fig. 2, we have what is known variously as the "zero degree," the "face perpendicular,"

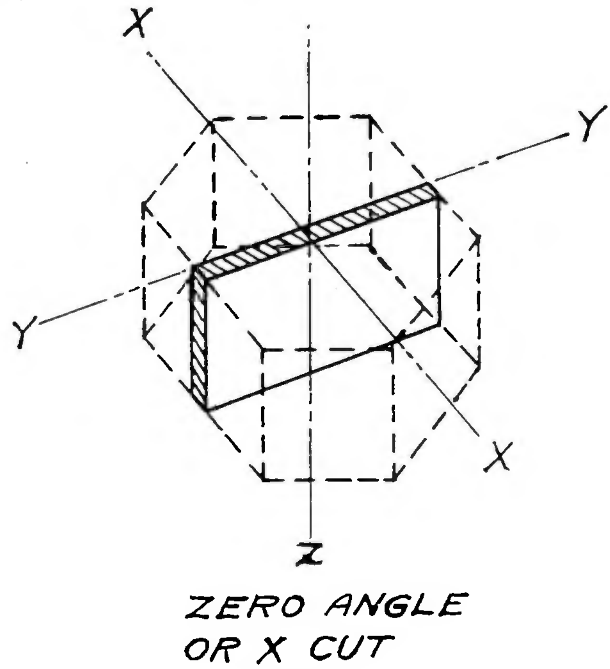

Fig. 2

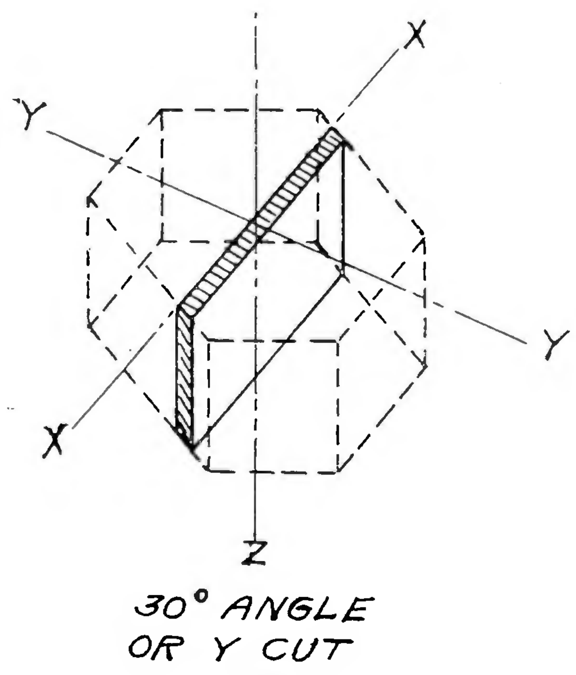

Fig. 3

or the "Curie" cut. If they are mounted so as to yield the cut shown in Fig. 3 we have what is known as the "30-deg." or the "face parallel" cut.

The next operation consists in cutting these thin slabs into rectangles of the desired size (approximately 1 inch square) after which the edges are trimmed by gang-grinding. The final machine operation consists of gang-grinding the surfaces of the plates to the desired thickness. The plates are then ready for the finishing room where they are hand ground to the specified frequency in a close duplicate of the circuit in which they will be used in service.

The variations occurring in the literature as to the nomenclature of the two cuts was emphasized and the desirability of the conference reaching some agreement as to a uniform nomenclature was pointed out.

The matters presented under (c) may be summarized as follows:

First, although it is possible to obtain good oscillating plates of either cut from partially twinned material, the results obtained from such plates are so variable as to output that the Naval Research Laboratory discards all twinned material.

Second, our experience with the cut shown in Fig. 2 shows that the constant of the material which has been known as the "meters per millimeter" varies from about 103 to 107 when the circuit in which the plate is used is tuned to the frequency of the standing wave system parallel to the X axis. When it is tuned to the standing wave system parallel to the Y axis we find the meters per millimeter to vary from 110 to 118. Average values in the two cases are 104.8 and 113.5, respectively. When the cut used is that shown in Fig. 3 our experience is that the meters per millimeter range from 130 to 175 with a mean value of approximately 153. It was pointed out that a name would be desirable for this important empirical constant of oscillating plates. It was further pointed out that the observed variations in the value of this constant are probably due to three causes: first, inaccurate orientation of the plate with reference to the crystal axes; second, improper dimensioning of the plates which results in improper mechanical coupling between the possible standing wave systems existing in a plate; and third, the use of partially twinned material.

Third, in the experience of the Naval Research Laboratory the temperature coefficient of frequency for the cut shown in Fig. 2 is negative and in the neighborhood of from 20 to 30 parts in a million, depending principally upon the kind of crystal holder used. In the case of the cut shown in Fig. 3 the temperature coefficient is positive and from 2 to 3 times as great in numerical magnitude as in the case of the other cut.

Fourth, it has been found at the Naval Research Laboratory that in the case of the cut shown in Fig. 2 it is possible so to dimension the plates that but a single frequency exists for a considerable variation in the tuning capacity of the plate circuit of the tube in connection with which it is operated. If the dimensions are not proper, so-called parasitic or extra frequencies exist in the neighborhood of the desired frequency, and these can easily be made effective in the circuit by slight changes in the setting of the tuning condenser. The cause of these extra frequencies is to be attributed to mechanical couplings between the possible standing wave systems existing in the plate. Proper dimensioning results in bringing these different standing wave systems into a proper phase relation.

It has further been the experience of the Naval Research Laboratory that it is not so simple a matter to eliminate by edge-grinding these parasitic oscillations in the case of the 30-deg. cut shown in Fig. 3. The mode of vibration which results when this cut is used is much more complicated than in the case of the zero-degree cut and hence it is a more complicated matter to insure the proper phase relations between the various standing wave systems. In order to be sure of freedom from parasitic oscillations it is the experience at this Laboratory that the accuracy of cutting must be such that the optic axis does not penetrate *both* surfaces of the plate. Since for a given frequency the 30-deg. cut is approximately two-thirds of the thickness of the zero degree cut it follows that a higher degree of accuracy in the orientation with respect to the optic axis is required in the case of the 30-deg. cut than in the case of the zero-degree cut.

Fifth, it has been the experience at the Naval Research Laboratory that when the parasitic frequencies have been eliminated at one temperature they are liable to reappear if that temperature is very much altered, and that this effect is much more likely to occur in the case of the 30-deg. cut shown in Fig. 3 than in that of the zero-degree cut shown in Fig. 2.

The conference agreed that the foregoing expressed generally their understanding of the principles involved in the piezo-electric properties of quartz crystals, and adopted, for the purposes of the conference, the following terminology:

(a) The zero-degree cut be hereafter referred to as the X-cut and the 30-deg. cut be called the Y-cut (See Figs. 2 and 3).

(b) The terms "X-waves," "Y-waves," and "Z-waves" be applied to waves of whatever type, whose directions of propagation are parallel to the X, Y, and Z axes, respectively. Thus, for example, an X-wave in an X-cut plate would mean that which has hitherto been called the longitudinal effect vibration, or the thickness vibration, of a Curie-cut plate; the Y-wave in an X-cut plate would be the same as the logitudinal vibration caused by the transverse piezo-electric effect or, as it has sometimes been called, the lateral vibration; and the Y-wave in a Y-cut plate would be the new term for the thickness vibration for this cut.

(c) The term "wave constant" be used to designate the quantity hitherto known as the "meters per millimeter."

The conference next discussed standards for cuts, sizes, and shapes for piezo-electric quartz crystals used for frequency control. The Navy stated that its present practice is to use X-cut crystals having dimensions of approximately one inch by a little more than one inch for power crystals in the band 1000 to 6000 kilocycles. The reason for using this size is that it appears to be best for directly controlling the output of a fifty-watt vacuum tube consistent with good operation and economical production of crystals. The particular dimensions chosen seem best adapted to large scale production while permitting the crystal to be edge-ground to eliminate spurious frequencies. The Navy has hesitated to use Y-cut crystals for this type of service because of the tendency, in this type of crystal, to develop spurious frequencies, and the difficulty encountered in edge-grinding to eliminate these spurious frequencies.

The X-cut is also used by the Navy for direct and monitor control and for precision measuring instruments, but the dimensions are not standard in every case. That form of crystal known as stabilized crystal control is not approved for use in Naval transmitters. The Navy uses the direct contact type of holder.

Statements from representatives of companies present at the conference indicated that the Y-cut in rectangular shapes, used in the air gap type of holders, was the common commercial practice. One company voiced the opinion that insurmountable manufacturing difficulties are not encountered in the production of Y-cut crystals having the same length and breadth as the X-cut used in the Navy, and recommended loading the crystal lightly and adding the necessary stages of amplification to secure the required output.

Another company stated that they used Y-cut crystals as do the others but, unlike them, they use for aircraft circular or square crystals, depending on the frequency, clamped tightly between special electrodes. These electrodes are designed to leave the edge of the crystal free. For high accuracy, they load the crystal lightly, as do the others, because the small amount of energy dissipated in a lightly loaded crystal simplifies the problem of maintaining the oscillating crystal at a constant temperature. This company prefers the circular crystal plate for crystals less than one to two millimeters in thickness as the production of such crystals is considered to be a better manufacturing proposition. Moreover, the circular plate, when of the Y-cut, lends itself to rigid clamping, for when so mounted the crystal can be clamped between the upper and lower electrodes with as much pressure as can be applied without actually crushing the crystal. A crystal so mounted can withstand severe shock without changing frequency.

The conference agreed to recommend that rectangular crystals approximately one inch square be considered standard for the present for direct frequency control in Naval transmitters, excepting in aircraft. It was also agreed that it is necessary to investigate further the various cuts, shapes, and sizes for future use.

It was agreed that a fixed operating temperature of 50 deg. C was satisfactory. This temperature was chosen because it is higher than any surrounding temperature likely to be encountered and yet not so high as to affect adversely the operation of the crystal or require too much energy to maintain the crystal compartment at this temperature under average conditions.

Certain manufacturers of broadcast transmitters preferred to have the temperature subject to variation in order to make the final adjustment of frequency by temperature control; but for Naval equipment, using several crystals in a single compartment, this method is not applicable.

In the discussion of grinding tolerance and tests, it appeared that the general consensus of opinion was that specifications should require performance in a specified circuit rather than tests for mechanical tolerances. It was pointed out that an error of two degrees in determining the optic axis or a departure from uniform thickness of only one ten-thousandth of an inch may lead to trouble. It was brought out that a factor of importance equal to any other in obtaining accurate frequency emissions from a transmitter was the characteristics of the circuit in which the crystal was placed. It was asserted that it is practically useless to grind a crystal for one type of circuit and use it in another and still to expect the emissions to be accurately on the desired radio frequency. It was recommended, therefore, that in order to obtain accuracy of 0.01 per cent in the frequency of emissions it is essential that manufacturers of crystals for the Naval Service be provided with an actual physical reproduction of a Navy standard-control circuit including temperature-control compartment, the associated circuit and proper provision for connecting with the first amplifier stage without its associated output circuits. And, further, if uniformity of accuracy in results is required in an operating organization faced with communication problems such as are encountered in the Naval Service, the conference deemed it advisable to consider the adoption of a standard circuit for all Naval transmitters employing crystal control, the crystal manufacturers to be furnished with a reproduction of this circuit.

The various types of crystal holders were discussed and it was agreed not to recommend a particular holder as a standard but to require all holders to be capable of being mounted and operated in Naval equipment using existing mounting devices.

In connection with the foregoing, the following was adopted by the conference:

(a) Crystal holders for Naval aircraft frequency indicators in the band 545–1000 kilocycles shall be made to fit the existing type of round socket.

(b) It is not necessary to specify what parts of a rough crystal should be used, nor is it necessary to specify the methods of determining the useful parts of a crystal. Specifications outlining performance are more desirable. It is unwise at the present time to specify either the X- or Y-cuts. Instead, there should be specified a crystal in a standard holder using a temperature coefficient not to exceed 100 parts in a million, to be used with a thermostatic control accurate to within 0.1 of a degree Centigrade at 50 deg. C. There should also be specified tolerances in respect to frequency drift and parasitic frequencies with given variations in operating voltages, circuit tuning, and operating temperature, in a standard circuit when delivering a specified voltage to the grid of the succeeding tube.

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PIEZO-ELECTRIC TERMINOLOGY*

By
WALTER G. CADY
(Wesleyan University, Middletown, Conn.)

Summary—*The following familiar terms are first recapitulated: direct and converse piezo-electric effects, longitudinal and transverse effects, longitudinal, transverse, flexural, and torsional vibrations. It is suggested that when the bending of a bar or plate is involved, the term "flexural" rather than "transverse" be adopted.*

With reference to quartz crystals, the terms "X-cut" and "Y-cut" are recommended in place of "Curie cut" and "thirty-degree cut" for plates perpendicular to the X and Y axes respectively. It is suggested that the terms "X-waves," "Y-waves", or "Z-waves" be applied to waves of mechanical vibration the direction of propagation of which is parallel to the X, Y, or Z axis respectively, whatever the mode of vibration. For the quantity "meters per millimeter," the name wave-constant, *to be designated by the symbol h, is recommended.*

Brief mention is made of the simpler types of vibration of Rochelle salt plates.
A piezo-electric resonator *is defined as any device that may be excited piezo-electrically into resonant vibration at one or more frequencies. In a more restricted sense the term is also applied to such a device when so connected as to exert no appreciable controlling effect upon the applied frequency through its reaction. A* piezo-electric stabilizer *controls over a narrow range the frequency of a circuit which already oscillates, while a* piezo-electric oscillator *controls the frequency of an oscillating circuit which in the absence of the crystal would not oscillate. The same crystal may of course function in any one of the three ways, depending upon circuit conditions. Since the distinction between a stabilizer and an oscillator lies largely in the amount of regeneration, it follows that the transition from one to the other may be gradual.*

A crystal monitor *or* piezo-electric monitor *consists of a piezo-electric preparation in an independent circuit of low power (resonator, stabilizer, or oscillator), serving as a frequency standard to which a generator may be tuned. A* piezo-electric calibrator *consists of a piezo-electric preparation of one or more crystals, so connected as to serve as a frequency standard for the calibration of other apparatus.*

For a plate cut with its edges parallel to the X, Y, Z axes of the crystal it is recommended that the respective lengths of the edges be denoted by x, y, z. Or in general, the symbols l, b, e may be used for length, breadth, and thickness.

The need of complete explicitness in specifying the dimensions and orientation of the crystal preparation, direction of electric field, frequency, and, so far as possible, the mode of vibration, is emphasized.

* Decimal classification: 537.65×R030. Original manuscript received by the Institute, Septmer 3, 1930. This paper was prepared in practically its present form prior to the Conference on Crystals which was held under the auspices of the U. S. Navy in Washington, December 3 and 4, 1929. Its publication has been delayed in order that it might appear in the same issue of the PROCEEDINGS with the report of the Conference. (Page 2128.) It now includes all of the recommendations respecting terminology that were adopted at the Conference.

THE AMOUNT of published material on piezo-electric crystals and their applications in radio has reached a stage where the need of greater uniformity in symbols and terminology seems urgent. No less urgent is the need of all possible explicitness in specifying (1) the directions of the dimensions of the crystal preparation with respect to the crystal axes, (2) the direction of the applied electric field or fields, and (3) the type of vibration and direction of wave propagation in the crystal so far as they are known.

The present suggestions are offered in the hope that, whether they win general acceptance or not, they may at least serve as a step toward the goal.

This discussion is concerned primarily with quartz crystals. Nevertheless the basic ideas are applicable to any piezo-electric material.

At the risk of repeating what is already sufficiently familiar (although even here some inconsistencies occur in the literature) the following fundamental definitions will first be given.

Piezo-Electric Effects

Direct Effect, called by the Curies the "primary" effect.—Electric polarization produced by mechanical strain, changing its sign with reversal of the strain.

Converse Effect, (sometimes called the "reciprocal" effect)—Mechanical stress produced by the application of an electric field, changing its sign with reversal of field.

All piezo-electric crystals necessarily exhibit both the direct and the converse effect.

Longitudinal Effect—This term is commonly applied only to those cases where a dilatation in a given direction is accompanied by an electric polarization in the same direction.

Transverse Effect—As commonly employed this term refers to a dilatation at right angles to the associated electric field.

Types of Mechanical Vibrations

Longitudinal Vibrations—This term may be applied either to rods or to more extended masses in which the motion of the vibrating particles is parallel to the direction of propagation of the wave, that is, normal to the wave-front. This use of the word "longitudinal" has nothing to do with the longitudinal *effect*. Vibrations of this type are also called "compressional" and "extensional." Longitudinal vibrations may be produced in either fluids or solids.

Transverse Vibrations—This term is properly related to transverse (distortional) *waves* in the same manner in which the term "longitudinal vibrations" is related to longitudinal waves. The vibrating particles move in a direction parallel to the wave-front and normal to the direction of propagation. Familiar examples are electromagnetic radiations, vibrating strings, membranes, and thin plates.

With piezo-electric crystals transverse vibrations may occur when the direction of the electric field is such that the field produces a shearing stress about some axis. If this axis is parallel to one of the principal dimensions of the parallelepiped, the wave propagation may be expected to take place in a direction parallel to another of the dimensions. Such vibrations may also be called "shear vibrations."

Flexural Vibrations—These usually occur in elongated plates or bars and are frequently called "transverse" or "lateral" vibrations. In order to distinguish them from the transverse vibrations described above, it would seem better to use the word "flexural." They are associated with a bending of the specimen in a certain plane, hence it is best to refer, for example, to "flexural vibrations in the YZ plane."

Torsional Vibrations are those in which a relative angular displacement (shearing strain) about the axis of figure, usually a cylinder or prism, takes place between adjacent cross-sections. For example, we speak of torsional vibrations "about the X-axis."

From what has been said it is evident that it is ambiguous to refer to the "direction of vibration in a crystal," unless the type of vibration is also made clear.

Crystal Preparations

It is in general important for the proper understanding of any paper on piezo-electricity or its applications that the various dimensions of the preparation be clearly specified with respect to the crystal axes, and that the values of the dimensions be stated. Exceptions may, of course, be made when one of the universally recognized cuts is used and the dimensions of the plates are not essential.

If the preparation has the form of a parallelepiped (plate, rod, etc.,) with edges parallel to the X, Y, and Z axes of the crystal, it is suggested that the symbols x, y, and z be employed to denote the lengths of the three edges parallel to the three crystal axes, respectively.[2] Or for any

[2] For definitions of the X, Y, and Z axes for the various crystal systems, as well as for general piezo-electric data and equations, see International Critical Tables, New York, VI, 208, 1929. In conformity therewith (see also R. B. Sosman, The Properties of Silica, New York, 1927, p. 556), it is recommended that henceforth the term *piezo-electric modulus* be applied to that coefficient, characteristic of the material, the product of which by a component of impressed electric intensity gives a mechanical stress-component; and that the related constants be termed the *piezo-electric strain-constants*. These constants are denoted by Voigt by e and d respectively. Although this is a reversal of Voigt's terminology, it is consistent with English usage with respect to the elastic coefficients. In the case of quartz crystals, it is customary to use the symbol X for an electric axis, and Z for the optic axis.

orientation of edges the symbols *l*, *b*, *e* may be used for length, breadth, and thickness, respectively.[3] Furthermore, unless it is already sufficiently obvious, the direction and frequency of the applied electric field should be made clear.

Quartz Crystals

Coming now more specifically to quartz, we consider first the manner of indicating the orientation of the more common "cuts." We have to do here with flat plates, the normal to the faces of which (direction of thickness, *e*) is either parallel or perpendicular to the electric axis of the crystal.

In the first case, we have the cut variously referred to in the literature as "Curie cut," "zero-angle cut," "perpendicular," or "normal"

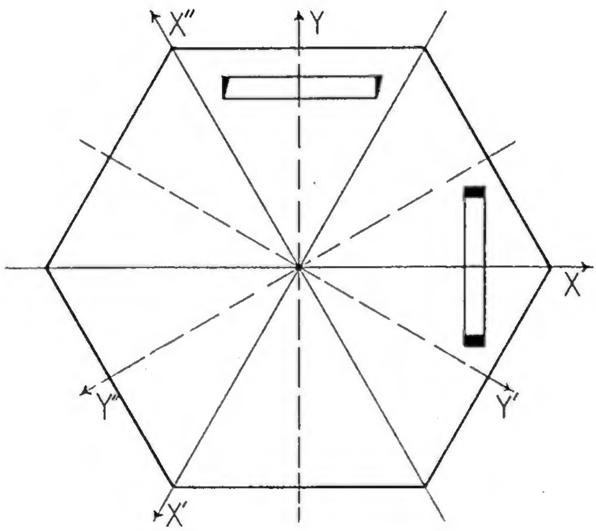

Fig. 1

cut. Owing to the evident ambiguity in the use of any of the last three, the term "Curie cut" is preferable. However, a still more concise term would be the "X-cut," denoting a plate the normal to whose face, and hence for which the applied electric field, is parallel to an *X*-axis.

Similarly, the term "Y-cut" would apply to the second type of quartz plate, which has hitherto been referred to as the "30-deg. cut" or "parallel cut." The author ventures to urge that the general adoption of the terms "X-cut" and Y-cut" would at once meet the need for definiteness, brevity, and consistency.

Fig. 1 illustrates the X-cut (at the right) and the Y-cut (above) with reference to the crystal axes. The thickness dimensions are parallel to the *X*- and *Y*-axes, respectively, while in each case the

[3] The use of the symbol *e* for thickness, first introduced by the Curies (*e*= *épaisseur*) is preferable to *t*, since in some equations *t* may be confused with the symbol for *time*.

breadth of the plate is perpendicular to the diagram (parallel to the optic axis).

As is well known, it is exceedingly difficult to cut quartz plates with such precision as to avoid very complex types of vibrations, so that in any given case, owing partly to the effects of elastic reaction and lack of uniformity of the elctric field, a wild medley of both longitudinal and transverse waves may be present. Fortunately, the frequencies commonly employed in practice are usually found to have values in fair agreement with those calculated for one or other of the simple vibration modes. In what follows only these simple modes need be considered.

X-Cut—The chief mode of vibration is longitudinal, in the direction of the *X*-axis, employing the longitudinal piezo-electric effect; the *Y*-axis, employing the transverse piezo-electric effect; or the *Z*-axis, in which case the vibrations are produced by elastic reaction. The use of the term "transverse vibrations" of quartz plates as applied to longitudinal vibrations which are "transverse" with respect to an electric axis is inconsistent and likely to lead to serious confusion. The characteristic elongation of the plate through the transverse effect is illustrated in Fig. 1.

One of the most important characteristics of a piezo-electric resonator is that quantity known as the "meters per millimeter," that is, the number of meters of electromagnetic wavelength for the fundamental mode of vibration along any dimension, divided by that dimension expressed in millimeters. For this quantity the term *wave-constant* is recommended.[4] The term is of course applicable to resonators formed from any kind of crystal. It is suggested that for the wave-constant the symbol h be adopted.

In order to avoid the confusion that arises from attempts at specifying modes of vibration, it is recommended that the term "X-waves," "Y-waves," or "Z-waves" be applied to waves of whatever type whose direction of propagation is parallel to the X, Y, or Z axis respectively.[5] For example, instead of the phrase "longitudinal vibrations of a Curie-cut plate in the direction of the *Y*-axis" we would now write "Y-waves in an X-cut plate."

Y-Cut—The only mode of vibration that has received much attention hitherto appears to be a shear vibration, the shearing strain taking place about the *Z*-axis, and the direction of wave propagation

[4] This term was adopted at the Washington Conference on Crystals (see p. 2133, this issue), upon Professor Van Dyke's suggestion.
[5] Terminology adopted at the Washington Conference.
[6] According to the terminology adopted at the Washington conference, we therefore use the term "Y-wave" in this case.

being parallel to the Y-axis.[6] The wave velocity, and hence the wave-constant for the fundamental frequency, calculated on the assumption of shear vibrations from the accepted values of the elastic constants of quartz, are in fair agreement with observation.[7] In Fig. 1 the nature of the shearing strain produced in a Y-cut plate is indicated.

Rochelle Salt

This crystal when in an electric field is subjected primarily only to a shearing stress, hence in the case of plates cut with all edges parallel to the axes, shear vibrations are the type to be expected in an alternating field, although, through elastic reactions, the possibility of longitudinal vibrations is not excluded. As is well known, plates may also be cut from a Rochelle salt crystal in such a manner as to exhibit resultant extensions and contractions in directions perpendicular to the electric field, so that in this sense one may speak of longitudinal vibrations produced by the transverse effect in Rochelle salt. No longitudinal effect exists with this crystal.

Resonator, Stabilizer, Oscillator

Strictly, all piezo-electric preparations commonly used in radio are resonators. Nevertheless, in order to avoid the confusion which has already begun to appear in various publications, it is suggested that the term "resonator" be used, as a rule, in a more restricted sense, and that the following definitions be adopted.

Piezo-Electric Resonator—Any device that may be excited piezo-electrically into resonant vibration at one or more frequencies. In a more restricted sense the term is also applied to such a device when so connected as to exert no appreciable controlling effect upon the applied frequency through its reaction.

Piezo-Electric Oscillator—A circuit containing a resonator and possessing too little regeneration to oscillate of itself, but which oscillates through the reaction of the resonator when the latter is vibrating near one of its normal frequencies with energy derived from the circuit. Such a circuit is often called a "crystal-controlled" or "quartz-controlled" circuit, also a "piezo-oscillator."

Piezo-Electric Stabilizer—A stabilized circuit is one which oscillates without the resonator, but the frequency of which is, usually over a rather narrow range, stabilized when the resonator is connected to the circuit. The resonator itself may in this case be referred to as a "piezo-electric stabilizer."

[7] W. G. Cady, "A shear mode of crystal vibration, *Phys. Rev.*, 29, 617, 1927.

Since the distinction between a stabilizer and an oscillator lies largely in the amount of regeneration, it follows that the transition from one to the other may be gradual.

A *Crystal Monitor* or *Piezo-Electric Monitor* consists of a resonator in an independent circuit of low power (resonator, stabilizer, or oscillator) serving as a frequency standard to which a generator may be tuned.

A *Piezo-Electric Calibrator* is a resonator, or set of resonators, so connected as to serve as a frequency standard for the calibration of frequency meters, etc.

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AVIATION COMMUNICATION*

By

J. Stuart Richardson

(Northern Electric Company, Ltd., Montreal, Quebec, Canada)

Summary—*This paper deals with the different functions of radio in the field of aviation. Requirements encountered in the design of suitable equipment are indicated and details given of equipment now operating commercially.*

WITHIN the last two years great progress has been made in the field of aviation, particularly in its application to the requirements of commerce, and nowhere has progress been so rapid as on this continent. The United States and Canada are covered with a network of airways which stretch from coast to coast, and from the Gulf of Mexico to the Arctic Circle. Even with the great developments which have been made, however, it cannot yet be said that air transportation has become completely a part of modern life.

The contribution of aviation to this era is speed, and this factor alone is a big recommendation to the commercial world. The use of airplanes for the transportation of mail and important freight is growing apace, but the growth of passenger traffic is slow.

The development of the aviation industry as a means of passenger transportation will depend on the achieving of "air-mindedness" by the general public, and this in turn will depend on a number of other factors. Air schedules will have to be as dependable as railway schedules. There will have to be more airdromes, more facilities for the repair of aircraft, more manufacturing concerns, and above all more organization on the routes than we have at present. Furthermore, every possible step will have to be taken to ensure the safety of aircraft in flight. The average person still regards a trip in an airplane as something of an adventure, and as long as this state of mind continues he will naturally give his custom to the railroad companies when it comes to a question of transportation. The air-travel age will have arrived when the man in the street takes to the air as readily and with as little thought of danger, as he does to travel by rail.

So it can be said that an important factor in the growth of aviation, especially with regard to passenger transportation, is that of safety. Anything which can give a further measure of safety to aircraft in flight will affect greatly the future of aviation.

* Decimal classification: R520. Original manuscript received by the Institute, July 10, 1930. Presented before Fifth Annual Convention of the Institute, August 19, 1930.

Experiments have been taking place for some time, and the results of these experiments are now being adopted commercially, in the adaptation of radio for use by aircraft. It is believed that radio can bring a measure of safety to aviation, and the aim of this paper is to show in just what manner this may be accomplished.

Reference will be made in this paper to the developments in aviation communication both in Canada and the United States, and to the requirements of the Federal Radio Commission. Although in Canada we are not governed by the requirements of the Federal Radio Commission, the developments here will proceed along similar

Fig. 1—Trimotored Ford plane used as a laboratory for the development of the equipment described in this paper.

lines to those in the United States, in order to facilitate air travel between the two countries.

Communication has always been of paramount importance to systems of transportation. The safe operation of our railroads depends absolutely upon reliable communication. The same can be said of marine transportation, to which radio communication has given a factor of safety that it did not hitherto enjoy. Communication is of even greater importance in aviation, because of the relatively higher speed, and because of the greater dependence of aircraft upon weather conditions. Radio can give to aircraft three important things; weather information, guidance, and communication with the ground.

To the pilot the advantage of being able to receive weather reports and other pertinent information in flight is immeasurable. With the mileage per hour of planes ever increasing, a great change of weather may be experienced by a pilot in the space of one hour. With the high speed at his command a pilot is able to make a detour around a reported storm center, providing he has sufficient fuel, unless he decides that an immediate landing is more expedient.

The second important factor is guidance. Any instrument which will give specific guidance between point of take-off and point of landing is obviously a great advantage, and nothing yet produced can compare with the radio beacon for the fulfilling of this function. An experienced pilot relying on his instruments alone to guide him over a familiar course through haze or cloud, may by a combination of circumstances be put a long way off his course. But by the aid of a radio-beacon receiver his guidance from point to point is specific and accurate, no matter what the state of the weather.

A radio receiver in a plane further constitutes a definite economic asset. It assists a pilot in keeping to his time schedule, enables him to follow a set course through any weather conditions, and so saves time in flight, and reduces the consumption of gasoline considerably. Furthermore, the emergency supply of gasoline may be reduced, thus making room for more pay load.

The addition of transmitting equipment in the plane will give a still higher degree of safety. It will enable the pilot to communicate with stations on the ground, and to request any particular information he may require. He will be able to communicate with planes flying in the opposite direction over the same section of airway, and so obtain advice on the weather immediately ahead of him, as well as being able to report his own weather conditions to the ground.

Further, it has been demonstrated experimentally on numerous occasions that speech from aircraft may be put over the ground telephone system, and communication established with any point. It is unlikely, however, that such a service could be made available to passengers, without interfering in some way with the factor of safety which radio is intended to supply.

Fog is undoubtedly the most dangerous weather condition that a pilot may meet, a condition frequently met with near the coasts, and as landing in fog is one of the problems that must be solved if aviation is to progress, interesting experiments have been conducted towards this end. The future development of radio equipment of a special nature will go a long way towards reducing the danger of "flying blind", and landing in fog.

Should a plane get into difficulties in the air, radio can be of very little assistance, but the claim made for radio, is that it will enable a plane to avoid hazardous flying conditions.

The requirements encountered in the design of radio equipment for aircraft use have perhaps been more numerous and rigorous than that for any other field of radio communication. It has had to be superdependable, simple, sensitive, and light. It must also be compact, and furthermore, robust.

In the first place it must be dependable, because in the majority of cases no adjustment is possible in flight. It may be necessary to operate the equipment for a few seconds at a time only, or on the other hand for the whole period of flight, yet it must function 100 per cent all of the time if it is to fulfill its purpose. Radio equipment that develops a fault in flight constitutes just so much useless weight.

Then the equipment must be simple to operate. In all but the larger types of passenger craft the pilot will also be the operator, and it is important that the operation of the radio equipment shall not interfere with his main business of piloting.

The equipment must also be sensitive. The gain in the receiver must be sufficient to build up the strength of signals in the earphones to overcome the combined noise of wind, engine and propeller. It must be sensitive again, because of the poor antenna system which is available, for the best type of plane antenna is relatively inefficient compared with that of a ground station. The trailing wire is the most efficient type of antenna for aircraft, but this offers certain mechanical disadvantages, so that the strut type is in more common use, although less efficient electrically.

The equipment must also be light, since the weight a plane may carry is necessarily limited, and every pound of navigating equipment displaces a like amount of pay load. As the main business of aircraft is transportation, great consideration must be given to this problem, with due regard at the same time to the main business of the radio equipment,—safety.

It must also be compact, since space also is at a premium in aircraft. In general the equipment is placed in the fuselage to the rear of the cabin. Here it is out of the way of the pilot and passengers, and as far as possible from the high tension system, which may cause trouble from induction. It must also be robust, to be able to withstand the shocks of taking off and landing without suffering mechanical damage.

Reference has been made to the high tension system as a source of induction. A large type of plane may have as many as 72 spark plugs operating within a few feet of the radio equipment, and unless precautions are taken the consequent induction will overcome all but the loudest signals. This may be reduced to a negligible amount by adequate screening. This means that each plug must be covered by a metallic cap, the magneto encased in a close-fitting metal box, all wires throughout the system covered by copper braiding, and this braiding in turn must be grounded to the frame of the plane at frequent intervals.

Power for the operation of aircraft radio equipment may be obtained in a number of different ways. A wind driven generator may

Fig. 2—Receiving dynamotor. This is driven by a 12-volt storage battery and supplies 200 volts for the plate circuit of the receiver. Weight 8 pounds.

be used, a generator coupled to the engine, or a dynamotor. A wind driven generator is the lightest source of obtaining power, but this has the disadvantage of rendering the equipment inoperative except during flight. On the occasion of a forced landing this might prove a serious handicap.

In the following paragraphs reference is made to particular pieces of equipment. It should be understood that all the equipment referred to has been developed by the Bell Telephone Laboratories for the Western Electric and Northern Electric Companies. There are, however, a number of manufacturing companies engaged in studying the radio requirements of aircraft, and a number of different types of aircraft radio equipment are now on the market.

One type of wind driven generator designed for supplying power to the receiver weighs only seven pounds. This is a double voltage generator supplying ten volts for the heater circuit of the tubes, and 200 volts for the plate circuit. The action of the wind on a small propeller drives the generator at a speed of 6500 r.p.m. The propeller is of the constant speed, self-regulating type. It has a starting torque

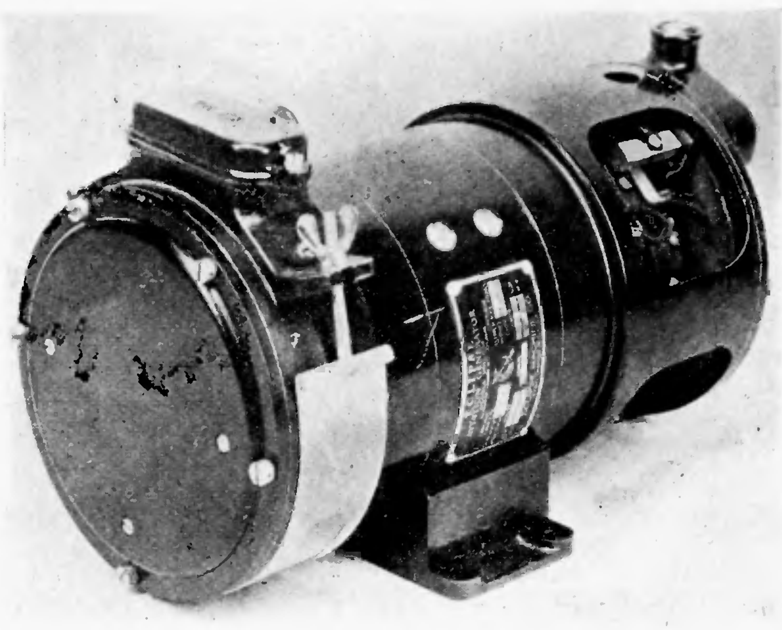


Fig. 3—Transmitting dynamotor. This is driven by a 12-volt storage battery and supplies 1050 volts for the plate circuit of the transmitter. Weight 30 pounds.

such that any airplane speed in excess of seventy miles an hour is sufficient to bring the generator up to the speed at which it will deliver its rated output.

An alternative method of obtaining power for the plate circuit of the receiver is by a dynamotor, which operates from the same 12-volt storage battery which is used to heat the filaments. This weighs eight pounds, and its output is 50 ma at 200 volts. The output circuit is equipped with a filter to eliminate the noise caused by the sparking of the commutator brushes of the dynamotor. If the plane is already equipped with a storage battery for other purposes than the radio, then the weight of this power system is very little more than that of the wind driven generator. This system has the added advantage of operation either in flight or on the ground.

For the operation of the transmitter similar types of generators of greater capacity are required. The wind driven generator weighs thirty pounds, and has two output circuits, 20 amperes at 12 volts for the filaments; and 400 ma at 1050 volts for the plate circuit. The transmitter dynamotor weighs 30 pounds and requires an input of 65 amperes at 12 volts. The battery for its operation also heats the tube filaments, and its plate output is the same as that of the wind driven generator. On planes where the dynamotor is used a 12–15-volt, 50-ampere generator connected to the engine of the plane is necessary in order to keep the battery charged. A generator coupled to the engine is an alternative method of obtaining power for the transmitter. Special control is provided to maintain constant voltage automatically at all engine speeds. 1050 volts are taken off for the plate supply, while the regular airplane battery, floated across the low potential supply of 12 or 14 volts, is used for the filaments and the heater element of the quartz-crystal thermostat control.

Before equipment may be designed for any field of radio communication it is first necessary to decide what frequency shall be employed.

By international agreement it was decided that the band of 230 to 353 kc be reserved exclusively for radio beacon stations.[1,2] It has been found that stations 250 miles apart will give satisfactory guidance to a plane in flight, and also that it is practicable to transmit weather and other information by telephone to aircraft from these stations. Reception of signals from a beacon station may be either aural or visual, and their operation is briefly as follows: Either type of station has two large loop antennas placed at right angles to each other. The signals from each loop are strongest in the plane of the loop, and in a direction at right angles to the plane of the loop the signal strength is zero. By means of a goniometer it is possible to orient the courses in space to any predetermined angle, even though the loops proper are fixed. With the aural type, each loop emits a different signal and at a point midway between the lines of greatest signal strength, the two signals interlock into a long dash, which is the indication to the pilot that he is on his course.

This can be more easily understood by the consideration of a beacon station using the Morse characteristics A(.—) and N(—.) on its two loops. The dot of the A is transmitted between the dash and dot of the N signal, and the dash of the A follows the dot of the N. From this it can be seen that at a point where the signal strength

[1] H. Diamond, "Applying the visual double-modulation type radio range to the airways," Proc. I. R. E., 17, 2158–2185; December, 1929.

[2] Kear and Jackson, "Applying the radio range to the airways," Proc. I. R. E., 17, 2268–2283; December, 1929.

from each loop is the same, the two characteristics will blend together in the form of a long dash.

With the visual type, the radio frequency from each loop is modulated continuously at different low audio frequencies. These signals operate a small indicator composed of two reeds, which is placed on the instrument board in front of the pilot. The degree of amplitude of each reed is governed by the position in space of the airplane from the line of true course. When the airplane is on the true course both reeds vibrate at equal amplitude. The visual indicator is so arranged that the reed vibrating at the larger amplitude always indicates the side to which the airplane is off its course. This is accomplished by a plug-in arrangement of the reed-box proper which is manipulated by the pilot as he changes from one course to another.

The majority of the beacon stations now in operation provide four courses. However, there are numerous airports where more than four air routes converge. To take care of this requirement experiments are now being conducted for the provision of twelve radio courses from one beacon station.

The aural method of reception proves fatiguing to a pilot if extended over a long flight, especially at times of heavy static, so that development in this field is proceeding with a view to still further improvements in the method of visual indication. Headphones will still be required however, for the reception of weather reports and other messages. The usual type of headphone is now being replaced by a headset employing an individually molded earpiece carrying a small receiver originally developed for the hard of hearing. These receivers weigh less than an ounce, are snapped into the earpiece, and may be worn for long periods without discomfort.

It might be mentioned that in Europe the guidance of planes is conducted on somewhat different lines to that in general use in North America. All passenger-carrying planes are equipped for two-way communication. Radio direction finding stations are established at strategic points, and upon request from a plane two or more of these stations determine the direction of travel of the radio waves from the plane. These bearings are communicated to the base station, the position of the plane calculated and supplied to the plane by radiotelephone. The disadvantage of this system lies in only one plane being able to obtain a bearing at any one time, and although the whole operation may be completed in a minute and a half, it does not make for satisfactory working where several planes are converging on one airport at the same time.

The accompanying illustration is of an aircraft radio receiver developed for the reception of either aural or visual signals from beacon stations. Three stages of radio-frequency amplification are employed, and one stage of audio frequency. The input stage is not tuned, but is provided with a special input filter to avoid interference from unwanted stations. Means are also provided for adapting the receiver to antennas of various dimensions. Thorough shielding is provided for the circuit elements, and the tuning coils are mounted in individual copper-shielded containers. These coil assemblies plug into sockets similar to those provided for the vacuum tubes. With this

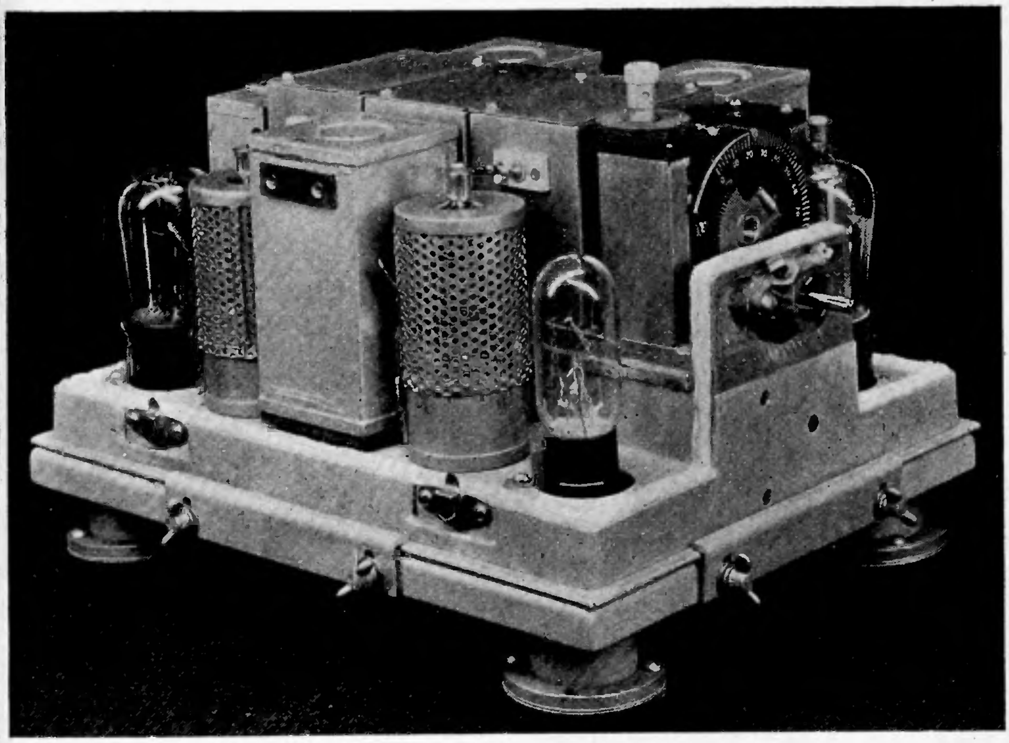


Fig. 4—Aircraft radio receiver with cover removed showing cushioned mounting base.

receiver only one set of coils is required to cover the frequency band of 250 to 500 kc. A welded sheet-aluminum cover protects the apparatus from dirt and moisture.

A cushioned mounting base is provided for these receivers. The base is permanently installed in the plane, and the receiver may be readily removed therefrom. A single plug connector provides for all power supply and output leads from the receiver. This is quickly detachable from the receiver. Remote volume control is accomplished by mounting the potentiometer which controls the shield-grid potential of the radio-frequency amplifiers in a small unit which is located within reach of the pilot. Remote tuning is accomplished by the use of a flexible shaft operated at a speed 264 times that of the condenser shaft. The radio receiver may be located as far as 35 to 40 ft. from the pilot, and the tuning accomplished with practically no backlash.

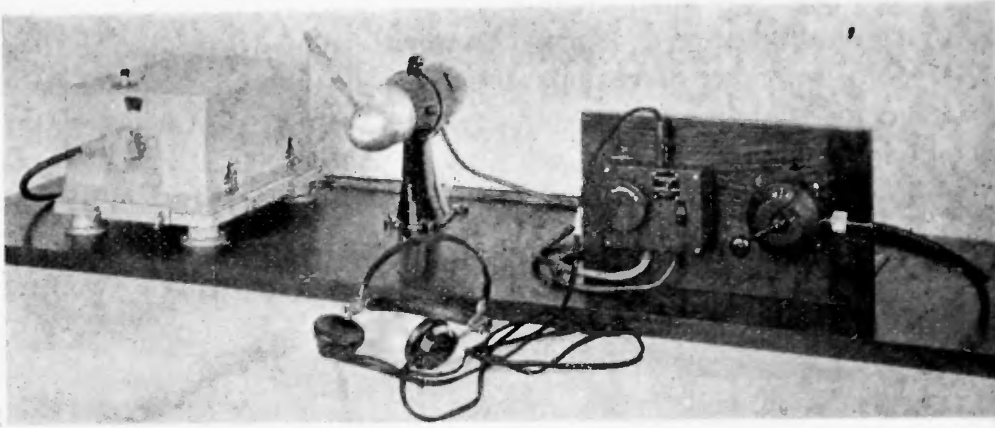


Fig. 5—Demonstration set-up of aircraft radio-receiving equipment showing type of wind driven generator.

This shafting connects with the condenser driving head of the receiver, and may be easily detached, as well as the power and output plug. The tubes used are of the equipotential cathode type, which

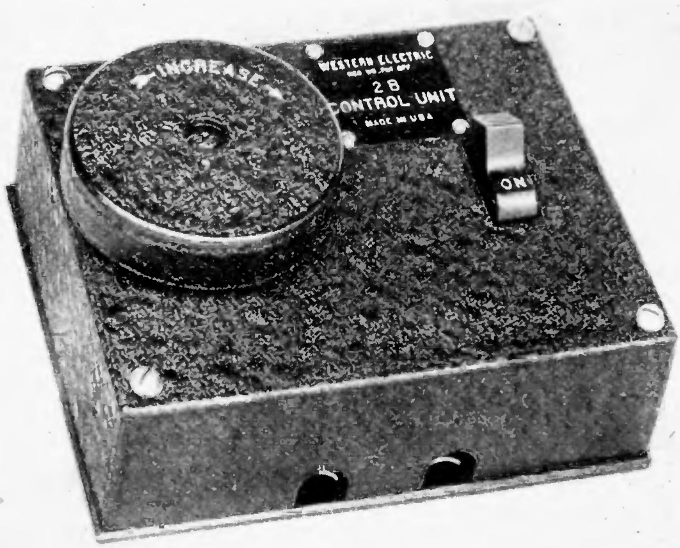
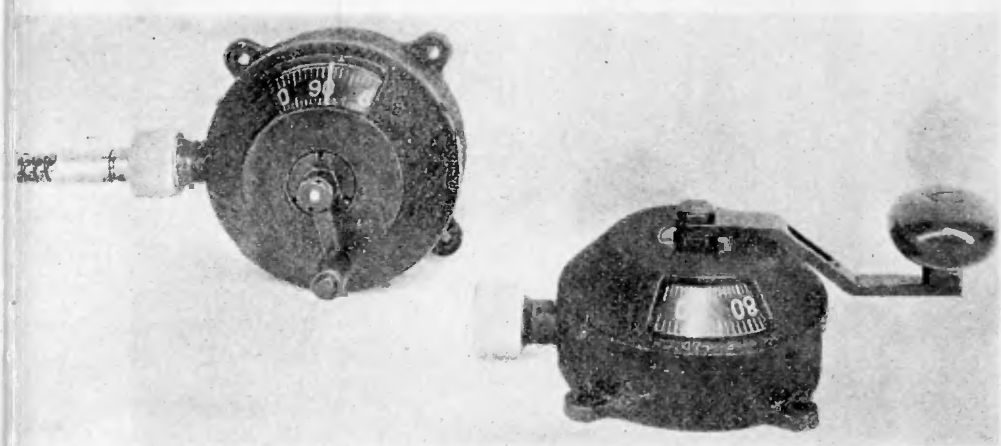


Fig. 6—Remote volume control. This includes a switch for applying power to the filament and plate circuits, and a potentiometer for controlling the shield-grid potential of the radio-frequency tubes.

tends to prevent the introduction of noise from the filament supply. The elements are solidly supported, making the tubes nonmicrophonic and not susceptible to external mechanical jarring and vibration. The total weight of this receiver with accessories is only 30 pounds.

A number of bands in the frequency range of 1500 to 6000 kc vas also allocated to mobile services, and for purposes of transmission, ttention was turned to these frequencies. Extensive tests were carried out on this band, and it was found that communication of a ature sufficient to fulfill the requirements of commercial aircraft ould be obtained. In general, the lower frequencies of this band ppear to give the most satisfactory results, however, satisfactory ransmission from a plane below a frequency of 3000 kc is only obained when a trailing-wire antenna is used. The use of the strut ntenna being far more satisfactory from a mechanical point of view, t may be on the whole more desirable to use this in conjunction with he higher frequencies of the band, despite the fact that this method



ig. 7—Two types of tuning control. These may be placed within reach of the pilot and are connected with the condenser drive spindle of the receiver by means of a flexible shaft.

omewhat restricts the range of communication. "Skip distance" s more noticeable on the higher frequencies, but a change of frequency vill, under normal conditions guarantee satisfactory communication oth by day and night.

As a result of the tests conducted by one company, a 50-watt ransmitter has been developed, which has been found to function atisfactorily under all conditions of flight. It consists essentially of temperature-controlled quartz-crystal oscillator, a frequency doubler, modulating power amplifier, and an audio-frequency amplifier. It s arranged for 100 per cent modulation, so that the peak power is 200 watts. The operating frequency is maintained to less than plus or minus 0.025 per cent under all conditions by the use of the quartz-crystal oscillator. This transmitter weighs 32 pounds complete with tubes, and is 17 in. high, 16 in. wide, and 12 in. deep.

A three-point master switch is used to control the operation of the receivers and the transmitter. In the first position everything is "off". The second position supplies current to the tube heater elements of the receiver, and operates the dynamotor supplying plate current. In this position, which is normal while in flight, the heater circuit of the quartz-crystal chamber is also energized. The third position energizes the filaments of the transmitter, and operates the transmitting dynamotor. In this position, everything is in readiness for transmission. However, no oscillations occur in the transmitter, and reception is possible. In order to transmit, it is then only necessary to press a push button which starts the oscillations in the transmitter. In some installations this push button is located on the hand microphone. In others it is located on the "stick", so that the pilot may operate it without moving his hand. During a conversation, this button is pressed while talking and released while listening. Relays perform all the necessary switching functions. On account of the very

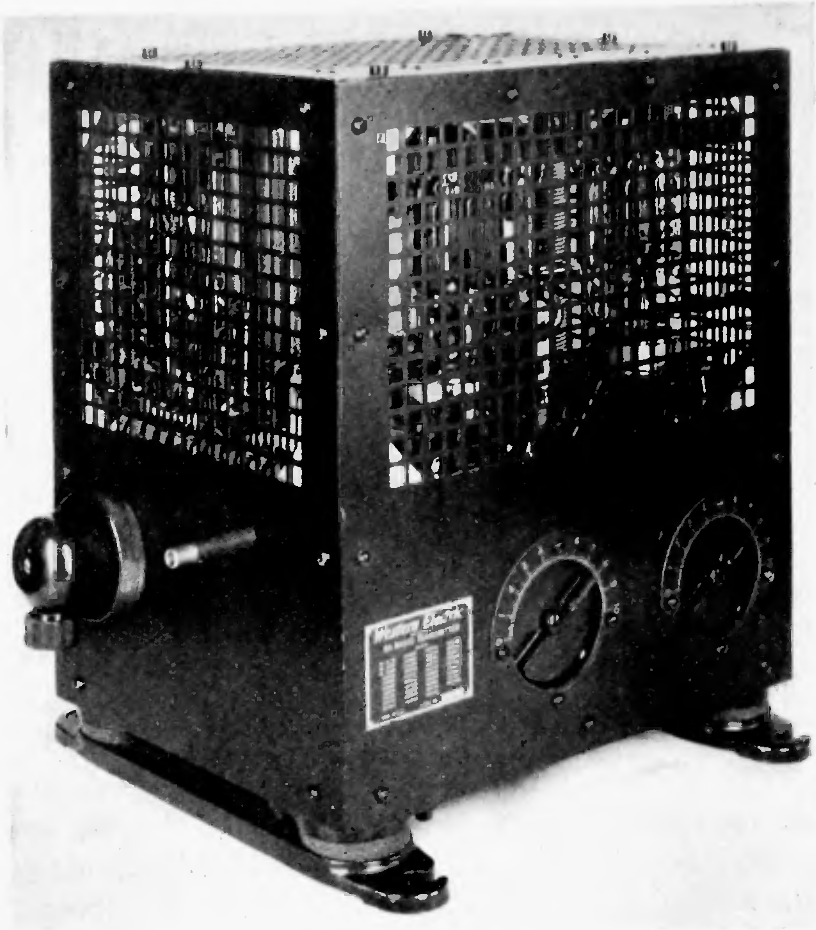


Fig. 8—Fifty-watt aircraft radio transmitter.

high noise level in many planes it is necessary to employ a telephone transmitter with a rubber mouthpiece. This fits close to the lips and practically all noise is excluded from the transmitter. The closed cavity into which the speaker talks is so shaped as to avoid serious distortion of the speech. A special transmitter has also been developed for use in planes where the noise level is not so high. This is held in position in front of the pilot's mouth by supports which are secured to both sides of his helmet. This transmitter may be dropped below the pilot's chin or turned over his head when not in use.

A separate receiver is required on a plane equipped for two-way communication. This is similar in design to the beacon receiver previously described, except that three sets of coils are required to cover the frequency band of 1500 to 6000 kc.

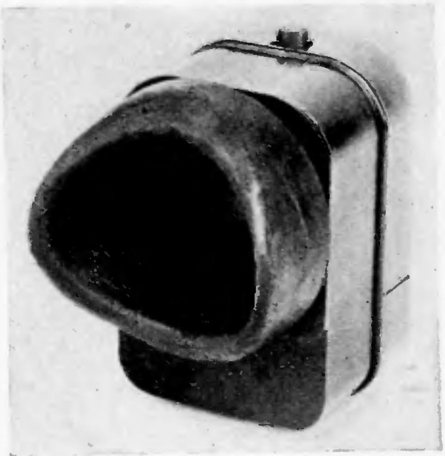


Fig. 9—Type of microphone for use in aircraft where the noise level is high. The push button on the top is pressed for speaking and controls the oscillations in the transmitter.

The weight of two-way communication equipment in a plane would be less than 150 pounds.

The transmitter and receivers mentioned above have been designed for easy removal from a plane for any necessary servicing. This facility recommends itself to aircraft operating companies, which might wish to equip all their planes with the remote control and other accessories, and have just sufficient transmitters and receivers to take care of their daily requirements. This arrangement will obviate having valuable equipment lying idle in a hangar during the period of overhauling a plane.

For purposes of aircraft communication, telephony, as against telegraphy, will undoubtedly come into general use. Apart from the extra speed of message sending and receiving, it reduces the need for carrying a radio operator. There are, however, certain air routes where telegraphy will be preferable. Particularly where long stretches of water have to be covered, or where the combined effect of distance and high static level has to be met, communication by telegraphy would be more reliable.

Fig. 10—Pilot's headset with close-talking microphone and phonette-type earpieces attached. This arrangement leaves both hands free for the controls.

The respective Governments of the United States and Canada are anxious to assist in every possible way in the developing of commercial aviation, and towards this end have spent large sums of money in building and maintaining ground stations for providing weather information, and for the guidance of aircraft. Messages of an urgent nature may be sent through these stations to any particular airplane, but in the future it is most likely that each large aircraft operating company will own and operate ground stations along its routes using

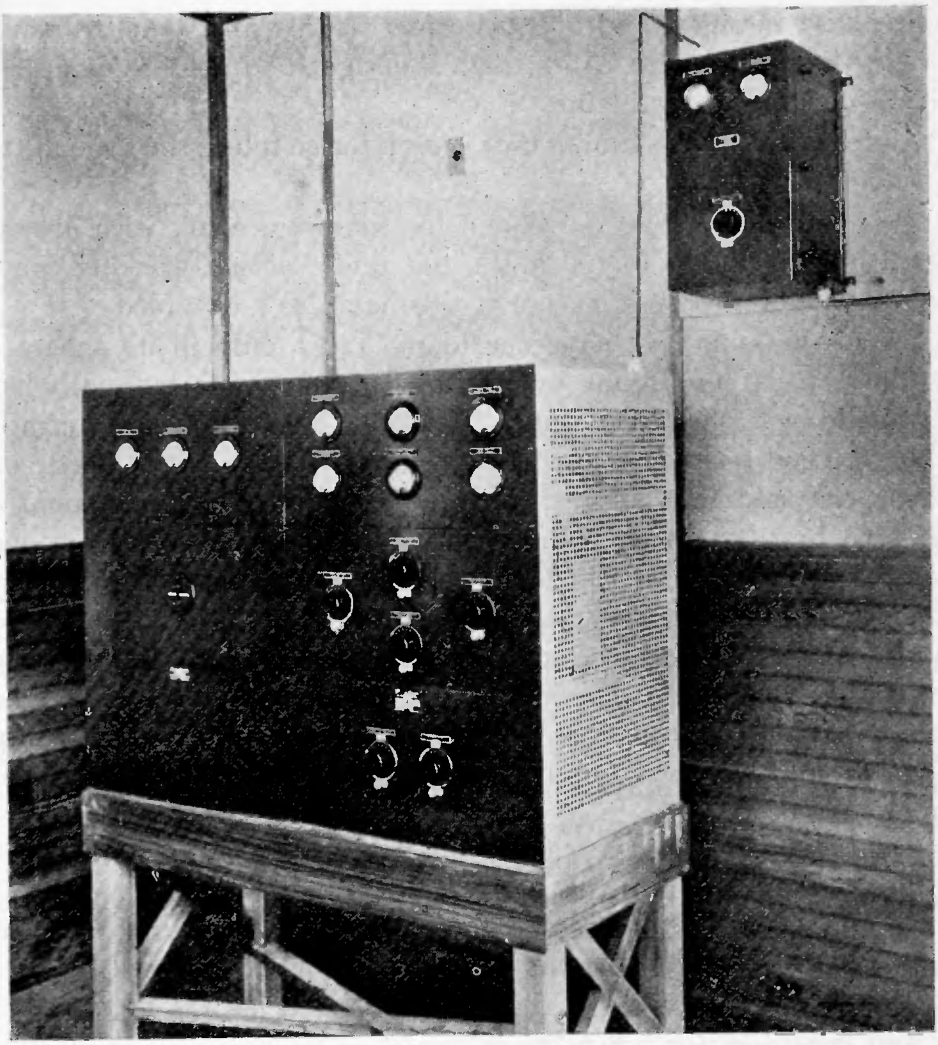


Fig. 11—400-watt ground transmitter developed for communication with aircraft. This shows the power-supply unit, the transmitter, and the antenna tuning unit.

the frequency band of 1500 to 6000 kc for communication with its planes.

For this purpose a ground transmitter has been developed, which has a carrier power of 400 watts, and which is capable of complete modulation. This requires a plate potential of 1000 to 2500 volts for operation, and consists essentially of a temperature-controlled quartz-crystal oscillator, audio-frequency amplifier, a frequency doubler, a modulating amplifier, and power amplifier. The circuit arrangement is similar to that of the 50-watt aircraft transmitter, except that a power amplifier has been added to increase the output supply to the antenna.

The power-supply unit consists of two three-phase rectifiers, one single-phase rectifier, and a transformer. This unit employs seven mercury-vapor rectifier tubes, and is designed to operate from 220-volt, 3-phase, 60-cycle mains.

A separate antenna tuning unit is used for coupling the transmitter to the antenna.

This transmitter was developed for communicating with aircraft over distances of 100 to 150 miles, but under actual working conditions distances greatly in excess of this have been obtained.

Up to this point we have considered radio only in its sphere of usefulness to aircraft flying over scheduled routes, but especially here in Canada will there be other functions which it may admirably fulfill. A good portion of our north country is still undeveloped, and it has been said that aircraft will put forward the development of that area by fifty years. Whatever methods of transportation may later follow, aircraft is the most economical method of surveying. Furthermore, this country is studded with lakes and rivers, which make ideal landing places for seaplanes and flying boats. In the winter time planes fitted with skis may land on the frozen surfaces of these lakes and rivers. In this area too, are many mines and lumber camps which are making great use of aircraft for the transportation of personnel and equipment.

From time to time we read in the newspapers of a plane having made a forced landing in this north country, and so becoming "lost". The organizing of a flying search party is a costly undertaking. One commercial company has estimated that the locating of one man who was lost in the bush last fall cost fifty thousand dollars. Cost is not the first consideration when life is involved, but had this plane been fitted with radio equipment, it is more than likely that the cost of locating this man would have been the cost of one plane making one trip.

A further use to which aircraft radio is being extensively put in Canada is in the field of forestry protection. Planes for this purpose have been used for a considerable time, but the addition of radio on the planes has greatly extended their sphere of usefulness. A radio-equipped plane may report the location of a fire without returning to the base station, which ensures personnel and fire-fighting equipment being at the location of the fire with the least possible loss of time. In the provinces of Manitoba, Saskatchewan, and Alberta, radio-equipped planes working in conjunction with a chain of ground radio stations are maintained for this purpose by the Department of National Defence. The province of Ontario has organized its own flying service for the protection of its forests. This in connection with a highly efficient ground radio system has meant an incalculable conservation of the timber wealth of the Province.

In Canada radio is further extensively used for the collection and dissemination of weather information for use by aircraft, over areas where no other means of communication are available.

Aircraft is still in course of development, and so also is radio for use in aircraft. It has been shown that radio is the only possible means of communication for aircraft, and because communication is so essential, great developments are to be expected. Reliable radio communication can give to aircraft a factor of safety which nothing else can give. Whereas radio has been a late addition to other methods of transportation, it will grow up side by side with aviation, and fulfill relatively an even more important function.

In connection with the assistance that radio can render aviation, it would be topical to make mention of the recent transatlantic flight of Major Kingsford-Smith in the "Southern Cross". Had this plane not been equipped with radio it is extremely doubtful whether the flight would have had a successful termination. Kingsford-Smith has stated that at a certain distance from Newfoundland he was flying an erratic course owing to the unreliability of his compass. It was due to the bearings obtained from the radio direction finding stations at Cape Race and Belle Isle that a safe landing was made at Harbour Grace. Whatever developments take place in transatlantic flying, it is certain that radio will take an important place.

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SOME PROPERTIES OF GRID LEAK POWER DETECTION*

By

Frederick Emmons Terman and Nathaniel R. Morgan

(Stanford University, California)

Summary—*It is shown theoretically that grid leak power detection can be obtained without undue distortion provided the grid condenser reactance X to the modulation frequency, and the grid leak resistance R are related to the degree of modulation m in such a way that* $X/R \geqq m/\sqrt{1-m^2}$. *The load limit of the grid leak power detector is the point where plate rectification becomes appreciable, i.e., when the straight line part of the plate characteristic is exceeded. Theory is developed which indicates that the maximum carrier voltage a grid leak power detector will handle is slightly less than one-half the input voltage that the tube will take as a properly adjusted amplifier operating at the same plate voltage; and that the maximum undistorted audio-frequency output voltage of the detector is in the neighborhood of one-third the output voltage developed by the corresponding amplifier. The equivalent input resistance of a grid leak power detector is proved to be greater than one-half the grid leak resistance.*

Experimental results are given which verify the theoretically determined point at which distortion becomes excessive. These results indicate that the grid leak power detector is more sensitive than the C bias power detector, and will ordinarily give at least as much undistorted output when operated at the same plate voltage.

THE unsatisfactory results generally obtained with grid leak grid condenser detection of large signals are not inherent properties of this method of rectification, but rather are the result of its improper use with large input voltages. Under suitable conditions the grid leak detector will give less distortion when the signal is large than when it is small.

The theory of grid leak power detection is exactly the well known grid-condenser-charge-and-leaking-off-of-charge analysis that has been so often incorrectly applied to weak signal detection. The way the details actually work out can be seen from Fig. 1. At each positive crest of the signal the instantaneous grid potential goes slightly positive, resulting in an impulse of grid current that charges the grid condenser negatively, thereby putting a bias on the grid. During the interval between the grid current impulses some of the accumulated charge on the condenser leaks off through the grid leak, to be replenished by the next spurt of grid current. The voltage applied to the detector grid as a result of the charge on the grid condenser follows the path of the heavy line in Fig. 1 (c) and neglecting the

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radio-frequency components introduced by the leaking-off of charge between cycles varies in almost exactly the same way as does the modulation envelope of the signal, with a resulting modulation frequency variation in plate current.

The difference between the action taking place in Fig. 1 and that occurring with grid leak detection of weak signals is that in the latter case the change of grid bias produced by the signal is so small that the grid current never entirely stops flowing. The grid condenser under these conditions discharges through the relatively low resistance grid-filament path rather than through the high resistance grid leak. The

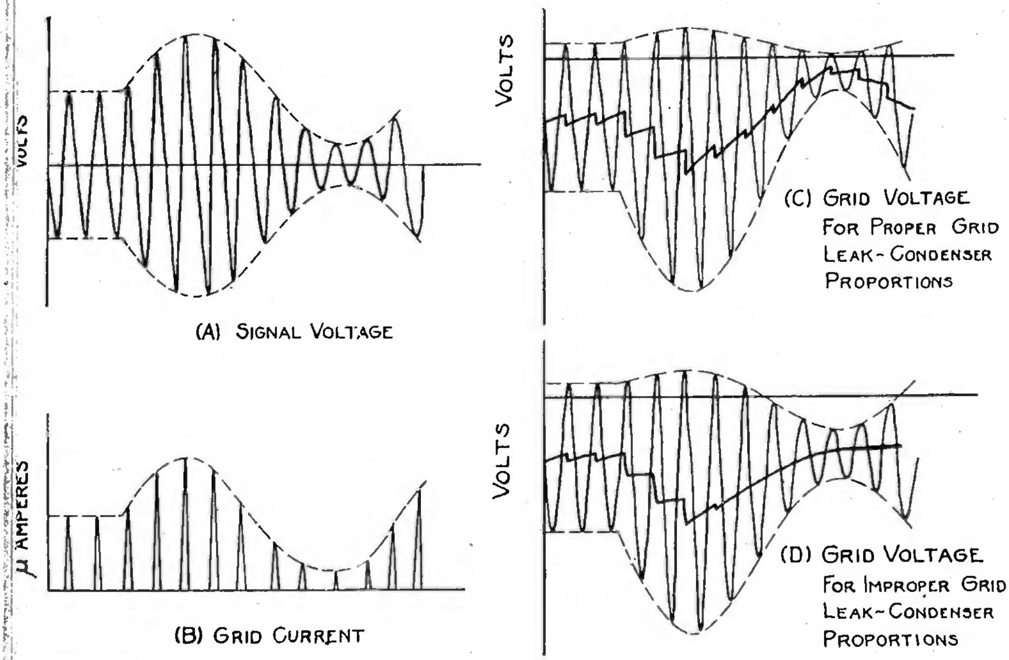

Fig. 1—Details involved in the detection of large signals by means of grid leak and grid condenser.

result is that the proper sizes of grid leak and grid condenser for weak signal detection are entirely different from values suitable for handling large signals.

The voltage across the grid condenser of the power detector can decrease only as fast as the condenser charge leaks off through the grid leak. If this leakage is slower than the rate at which the modulation envelope decreases, then the condenser voltage cannot follow the modulation envelope, as is the case in Fig. 1 (d), and the result is distortion coupled with loss of sensitivity. Experiments showed that as long as the grid condenser charge could decrease as fast as the modulation envelope the distortion was small. The rate of decay of the grid condenser charge is determined by the grid leak-condenser combination, while the rate of decrease of the modulation envelope depends upon the modulation frequency f and the degree of modulation m of the signal. Taking these factors into account, it is shown in Appendix I that the distortion will be small provided the grid leak-condenser proportions satisfy the relation.

$$\frac{X}{R} \geqq \frac{m}{\sqrt{1 - m^2}} \tag{1}$$

where R is the grid leak resistance, m the degree of modulation of the signal, and X is the reactance of the *effective* grid condenser capacity

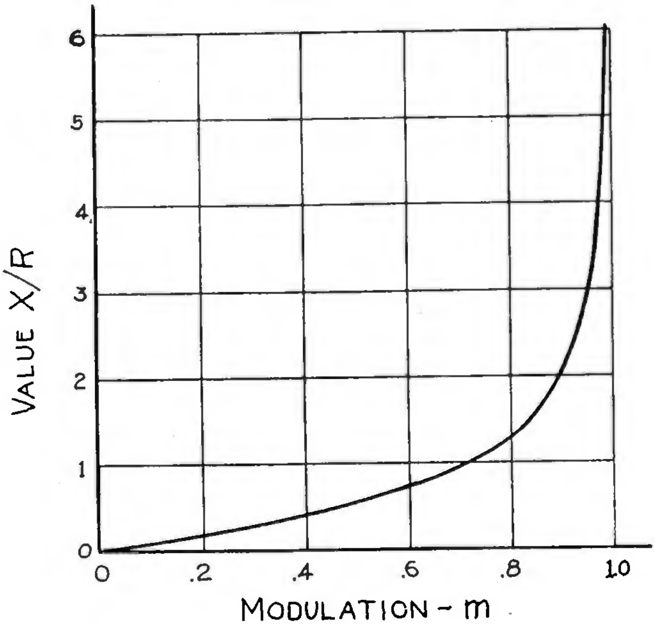

Fig. 2—Graph of the equation $(X/R = m/\sqrt{1-m^2})$ showing the relation of X/R to m at which distortion begins to become serious.

at the modulation frequency in question. The effective grid condenser capacity is the actual capacity plus the input capacity of the tube, taking into account the plate load reaction. In order to permit ready visualization of the conditions at which distortion begins to increase rapidly, (1) has been plotted in Fig. 2. It is apparent from this figure that for every modulation frequency there is a maximum degree of modulation beyond which distortion begins to be large, but that the permissible value of m increases as the frequency is reduced.

Although a small grid condenser and a low resistance leak make for low distortion because of the resulting high value of X/R, selectivity and sensitivity requirements limit the extent to which one may go in this direction. Since the available signal voltage is applied to the grid of the detector through the grid condenser capacity, it is necessary for this capacity to be somewhat larger than the effective grid-filament tube capacity offered to the signal frequency. With ordinary receiving tubes a grid condenser capacity of 50 to 100$\mu\mu$f is about right. With the grid condenser capacity fixed in this way, the grid leak resistance should be made just low enough to prevent undue distortion under the operating conditions to be encountered, which for high quality reproduction of speech and music represents values from 0.20 to 0.50 megohms. The degree of modulation at which distortion begins as computed from Fig. 2, is given by the solid line in Fig. 3 for the case of 0.25-megohm leak, 100-$\mu\mu$f grid condenser and

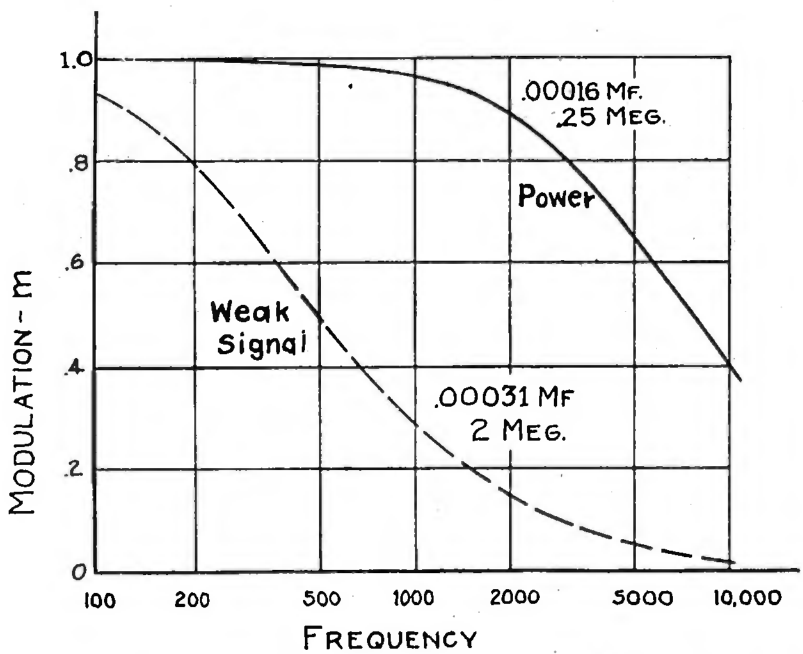

Fig. 3—Allowable *m* as a function of modulation frequency for grid leak power detection for proportions of grid leak and condenser suitable for weak signal detection compared with those recommended for power detection.

50-$\mu\mu$f audio-frequency tube input capacity. Since most of the power of speech and music is on the frequencies below 1000 cycles, it is apparent that the degree of modulation encountered at the higher frequencies will be much less than 100 per cent, and a characteristic such as shown in Fig. 3 will give practically no frequency discrimination in ordinary reception. For purposes of comparison the allowable modulation as a function of frequency is also shown in Fig. 3 for a grid leak-condenser combination suitable for weak signal detection. The potential of the grid return lead is not nearly as important with the grid leak power detector as when handling weak signals. For ordinary conditions the adjustment that gives best results on moderate strength signals will also be satisfactory with the larger inputs, and so is to be preferred.

The ratio of modulation frequency voltage applied to the grid of the detector by the grid condenser charge to the modulation frequency voltage contained in the signal envelope is a measure of the completeness of rectification being obtained from the detector. This ratio, (which can also be expressed as a percentage and called the percentage rectification) is approximately constant for any particular tube type over a wide range of signal amplitude and degree of modulation, so is a useful constant of grid leak power detectors. When a carrier voltage of E_s modulated to a degree m is applied to a power grid leak detector having a completeness of rectification β, the modulation frequency voltage that the detection process causes to be applied to the detector grid is $\beta m E_s$ which is then amplified in the usual manner. Values of β as a function of signal amplitude are given in Fig. 8 for a number of representative tubes. The completeness of rectification depends upon the steepness of the grid voltage-current characteristic, and so is greater as the cathode voltage drop of the detector is reduced.

The input resistance of the grid leak power detector is shown in Appendix II to be slightly more than the grid leak resistance divided by twice the completeness of rectification. The grid leak resistance should therefore be no lower than is necessary to prevent undue distortion. With a grid leak resistance of 0.25 megohm the equivalent input resistance amounts to about 150,000 ohms, which is slightly higher than the input resistance of the usual weak signal grid leak detector.

The maximum signal voltage that may be applied to a properly adjusted grid leak power detector without producing distortion is the input at which plate rectification begins to be appreciable. Such plate detection follows a square law, with the resulting production of second harmonic distortion as well preventing the undistorted portion of the output from being proportional to the signal amplitude and degree of modulation. Plate rectification takes place only at high signal amplitudes when the grid bias produced by the grid condenser charge is sufficiently great to cause the operating region to extend into the curved part of the grid voltage plate current characteristic, or in other words, when the radio-frequency signal is distorted while being amplified in the plate circuit. The load capacity of the grid leak power detector therefore varies with plate battery voltage in almost the same way as does the load capacity of an amplifier, and it is just as impossible to handle large inputs to a grid leak power detector without high plate voltages as it is in the case of an amplifier.

It is convenient to express the maximum carrier voltage that may be applied to a grid leak power detector in terms of the voltage that may be applied to the same tube when acting as an ordinary amplifier at the same plate battery voltage. It is shown in Appendix III that if the plate load impedance to the modulation frequency is low and 100 per cent modulation is allowed for, then

$$\frac{\text{Allowable carrier voltage}}{\text{Allowable amplifier input}} = \frac{k}{1+\beta} \qquad (2)$$

where β is the completeness of rectification, and usually lies between 0.7 to 0.9, while k represents the ratio of proper grid bias for amplifier operation with no load impedance to proper grid bias for amplifier operation with normal load impedance. If the allowable amplifier input is taken as the value recommended by the tube manufacturers for amplifier operation, k can be expected to be very close to 0.70.

When the plate circuit of the grid leak detector contains an impedance to the modulation frequency the carrier voltage that may be applied to the detector is somewhat greater than the value given by (2). This is because the effective plate voltage as far as the radio-frequency signal currents are concerned is slightly greater than the battery potential during the crest of the modulation cycle as a result of the voltage developed across the load impedance by the modulation frequency currents in the plate circuit. Such a load impedance may increase the allowable detector input by as much as one-third if the completeness of rectification is great and the load has a very high impedance, but will ordinarily have somewhat less effect.[1]

With ordinary tubes the allowable carrier voltage that can be handled by the grid leak power detector is at least one-third and not more than one-half of the maximum input that the same tube will handle as an amplifier at the same plate battery voltage. This is evident when reasonable values of β and k are substituted in (2), and estimates are made correcting for the effect of plate load impedance. The ratio of maximum undistorted modulation frequency voltage that can be delivered by the grid leak power detector to the output for amplifier operation is β times the ratio of inputs, given by (2) after correction for plate load. With ordinary tubes this ratio of outputs ranges from 0.30 to 0.40 with sufficient certainty to enable these figures to be used for tentative design purposes.

[1] This statement is based on a quantitative analysis that is not included in this paper but which can be obtained by an extension of Appendix III.

Experimental Results on Power Detection

The method followed in making measurements of power detection was to apply a known carrier voltage modulated a predetermined amount to the detector under test, and then to make an analysis of the audio-frequency voltage developed across a 10,000-ohm resistance in the plate circuit. A shunt feed was used in all tests. In order to simplify the experimental procedure, advantage was taken of the fact that it is the ratio X/R which determines the action of the grid leak-condenser combination at a particular modulation frequency. It was consequently possible to work at a constant modulation frequency of 60 cycles and simulate the effect of different frequencies by changing the grid condenser capacity in such a way as to give X the appropriate value that would result with a constant condenser and a varying frequency.

Analysis of the output voltage of the detector was made possible by the "power circuit harmonic analyzer" which is part of the equipment of the Ryan High Voltage Laboratory. This piece of apparatus, which was developed by the American Telegraph and Telephone Company for inductive interference investigations, operates by selecting the desired frequency component with suitable tuned circuits and measuring its potential with a vacuum tube voltmeter. In the detector investigation it was found that under ordinary conditions the principal distortion frequency was the second harmonic, which could therefore be used as a measure of the output quality.

A typical set of experimental results selected from a large number is given in Fig. 4. The curves for grid leak-condenser detection are for a value of X/R well above the minimum allowable value given by (1) and Fig. 2, while the one for plate rectification is for a grid bias found to give maximum output with large input voltages. It is apparent that under the conditions used in obtaining Fig. 4 the grid leak power detector is not only several times as sensitive as the plate rectifier when both have the same plate voltage but also produces less distortion, and gives more undistorted output. With all except the very lowest signal amplitudes the grid leak power detector is practically a *linear rectifier*, and is somewhat superior to plate detection in this respect. The effect of the plate voltage on the load capacity of the grid leak detector is clearly brought out in Fig. 4. It is just as futile to expect power from a detector operating with 22 or 45 volts on the plate as it is from an amplifier under similar circumstance.

The effect of varying the ratio X/R of the grid leak-condenser combination is shown in Fig. 5. As X/R becomes less than the critical value given by (1) and Fig. 2 the distortion increases rapidly, accompanied by a loss of sensitivity, exactly as would be expected by theory. The small residual distortion remaining with adequate values of X/R is due to the fact that the curve of voltage across the grid condenser due to condenser charge is not quite symmetrical about the

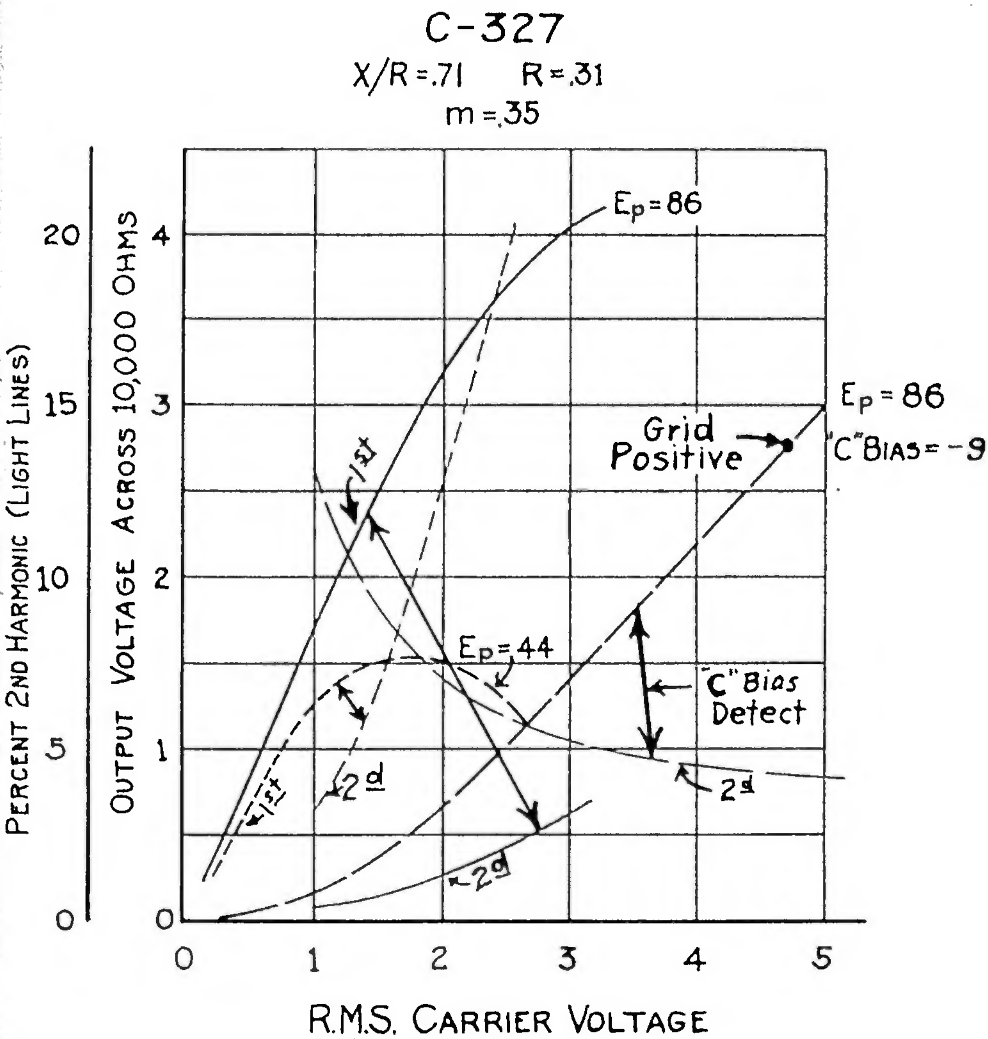

Fig. 4—Performance of grid leak and "C" bias power detectors under favorable conditions. The solid and dotted lines are for grid leak detection, while the dash line applies to plate rectification.

point of minimum signal amplitude as can be seen in Fig. 1, and to distortion in the plate circuit introduced when amplifying the rectified signal from the grid.

Investigation also showed that as long as the ratio X/R was kept constant, the value of leak resistance R was of secondary importance in influencing the sensitivity and distortion. The only effect observed was a slight reduction in sensitivity at low values of leak resistance caused by the increased energy consumed by the detector under those conditions.

Since the grid leak detector operates at approximately zero grid potential when no signal is being received, the maximum plate voltage

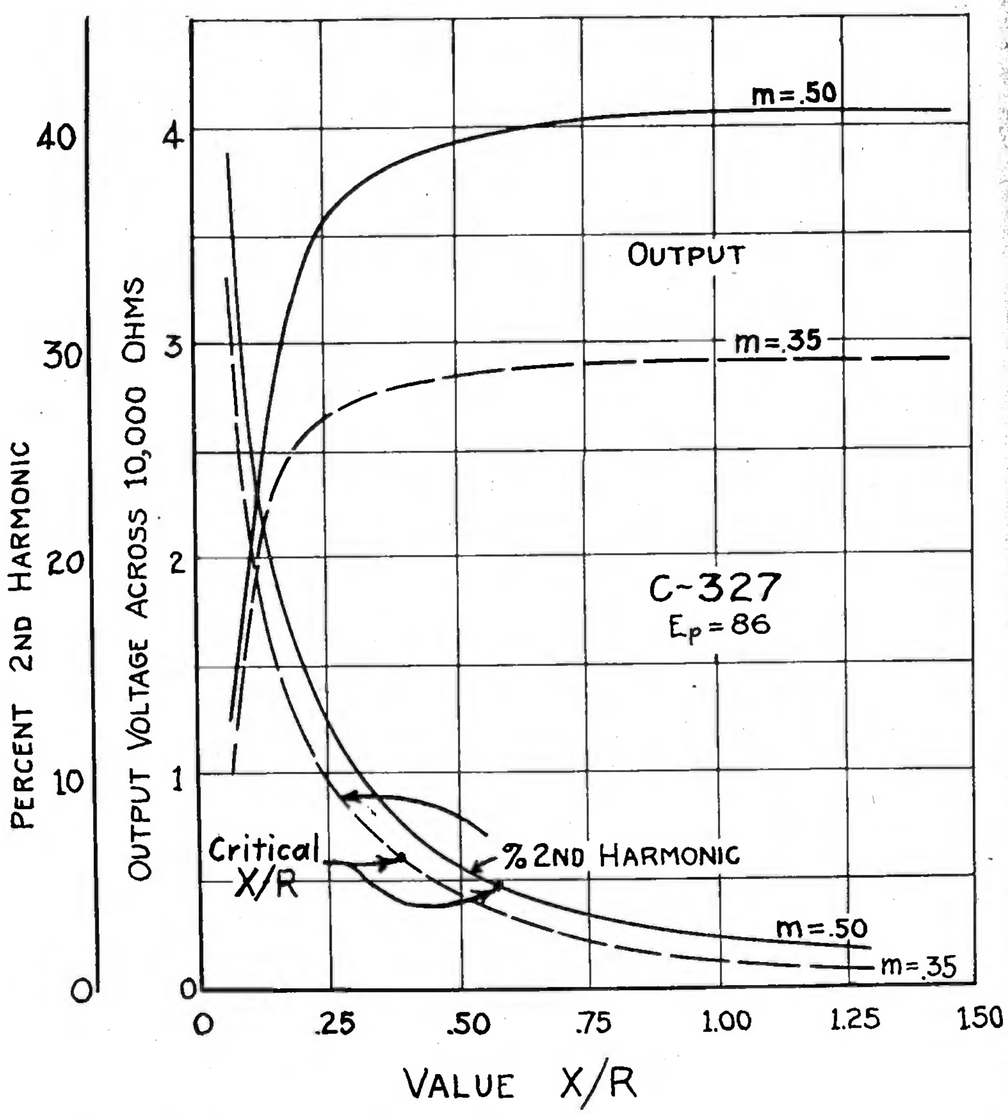

Fig. 5—Effect of X/R on sensitivity and quality, showing how both rapidly become poorer when the critical value of X/R as determined from Fig. 2 is passed.

that may be used with such a detector is fixed by the allowable plate current of the tube. In a general way it may be said that low-mu tubes are not satisfactory for grid leak detection at high power levels because of their high plate currents with zero grid bias, and that a plate potential of 80 to 90 volts is about the limit with tubes such as the '27 type. It is possible to increase the load capacity of the grid leak power detector without the use of excessive plate currents by employing the expedient shown in Fig. 6 where the high battery voltage is cut down to a safe plate voltage by a series resistance that is thoroughly by-passed to all audio frequencies. When no signal is applied to the tube

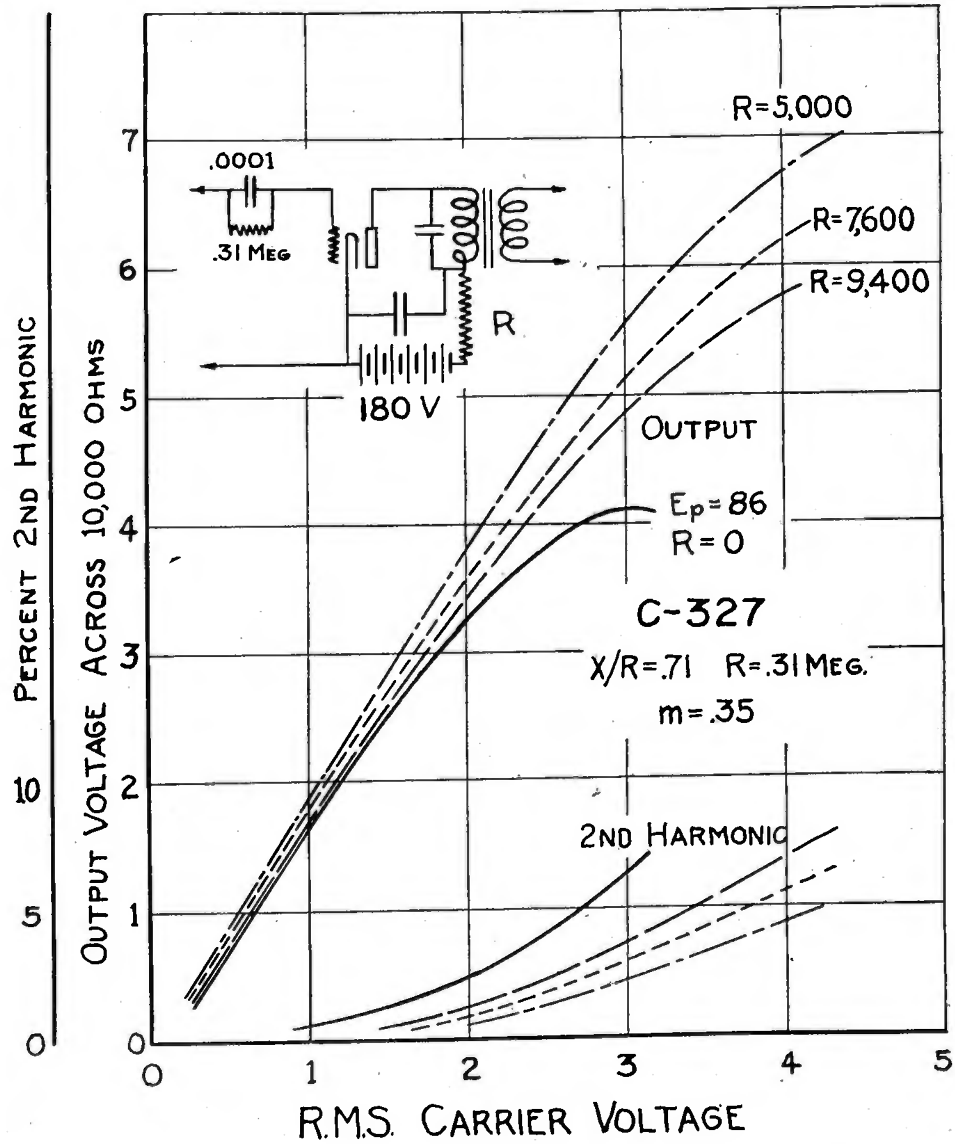

Fig. 6—Performance of grid leak power detector when operated with a high plate battery voltage that is reduced to a safe value by a series resistance, showing increased undistorted output that may be obtained with the aid of this expedient to keep the plate current to a safe magnitude.

the grid is at zero potential, but the resistance keeps the plate current to a reasonable value. Upon the arrival of a signal, the carrier wave produces a grid bias by charging the grid condenser, which cuts down the plate current, and hence the voltage drop in the resistance. The result is that the voltage on the plate of the tube is low when no signal is present, but increases when the presence of the signal makes an increased potential allowable. The net effect is to increase the undistorted power that can be safely obtained from a given tube.

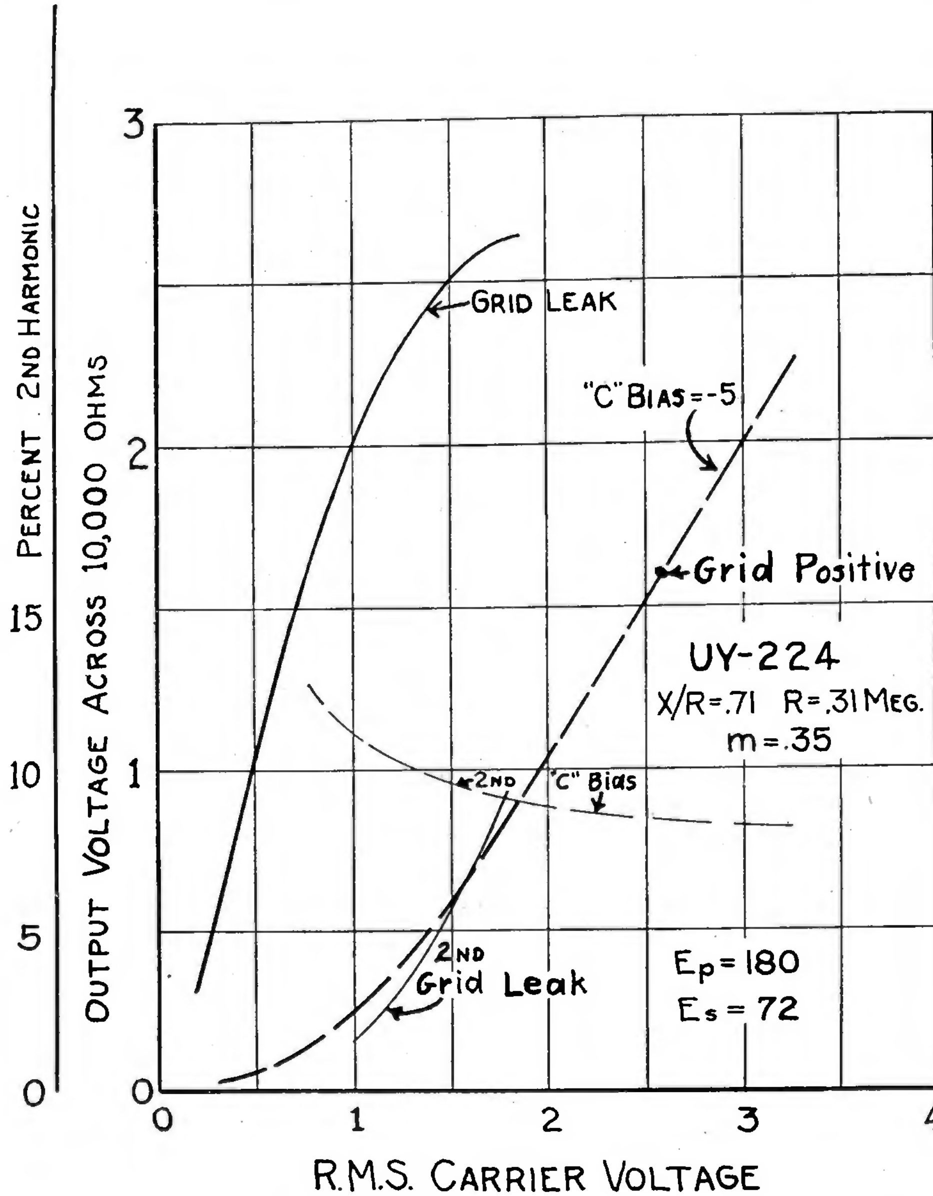

Fig. 7—Comparison of grid leak and plate detection with the screen-grid tube operating at high power levels.

In view of the work that has recently been done on screen-grid detection[2] the results shown in Fig. 7 are of interest. With this par-

[2] F. E. Terman and Birney Dysart, "Detection characteristics of screen-grid and space-charge-grid tubes," Proc. I. R. E. 17, 830; May, 1929. J. R. Nelson, "Detection with the four-element tube," Proc. I. R. E. 16, 822; June, 1928.

ticular screen-grid tube, power detection with grid leak and grid condenser is more sensitive, gives more undistorted output, and less distortion than does plate circuit power detection.

The completeness of rectification was measured in a number of cases and some of the results are shown in Fig. 8. Different tubes of

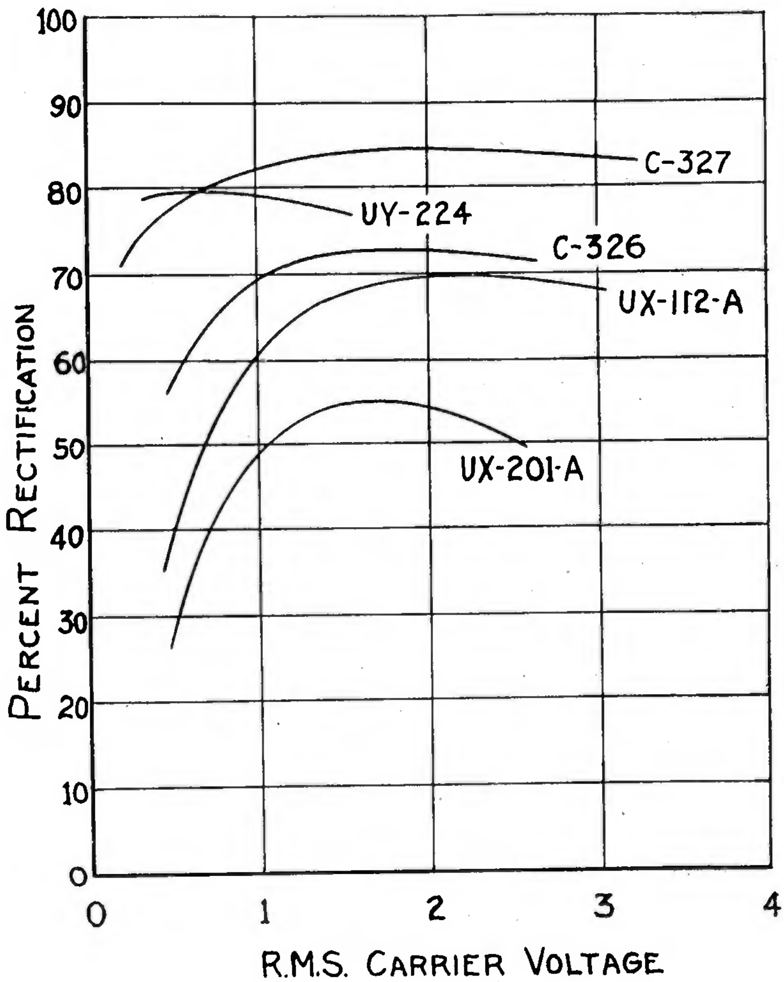

Fig. 8—Completeness of grid leak power detection as a function of signal amplitude for a number of representative tubes.

the same type were found on test to be very similar so only one curve for each kind of tube is shown. It is apparent that the completeness of rectification for any tube type is sufficiently constant over the normal signal range as to make it possible to associate a particular variety of tube a particular completeness of rectification. Thus the 327 tube can be considered as giving 80–85 per cent rectification as a grid leak power detector.

Appendix I

Derivation of Equation (1)

This equation gives the relation between modulation frequency f, degree of modulation m, grid leak resistance R, and effective grid condenser capacity C which must exist if the grid condenser charge is to be able to leak off through the grid resistance fast enough to follow the variations in the modulation envelope at all times.

The rate of change of voltage across a condenser C charged to a voltage E' as a result of leakage through a resistance R is $-E'/RC$. If the equation of the envelope of the modulated signal wave is

$$e = E(1 + m \cos 2\pi ft)$$

then at time $t = t_0$ the envelope magnitude e_0 is

$$e_0 = E(1 + m \cos 2\pi ft_0)$$

while the rate of change of envelope is

$$de/dt = -mE2\pi f \sin 2\pi ft.$$

The grid condenser rate of discharge can follow the rate of change of modulation envelope at any time t_0 of the modulation cycle provided $-E'/RC \geqq de/dt$, where de/dt is evaluated at $t = t_0$ and $E' = e_0$, or when $-mE2\pi f \sin 2\pi ft_0 \leqq -E(1+m) \cos 2\pi ft_0/RC$. This can be reduced to

$$X/R = 1/2\pi fCR \geqq m \sin 2\pi ft_0/(1 + m \cos 2\pi ft_0). \qquad (3)$$

The point on the modulation cycle where it is most difficult for the grid condenser charge to keep up with the envelope change is at a value to that makes the right-hand member of (3) a maximum which is when $\cos 2\pi ft_0 = -m$. The maximum value of the right-hand member of (3) is therefore $m/1\sqrt{-m^2}$, and the charge on the grid condenser can decrease at least as fast as the modulation envelope changes when $X/R \geqq m/\sqrt{1-m^2}$.

Appendix II

Equivalent Input Resistance of Grid Leak Power Detector

The grid current in the grid leak power detector flows only when the signal voltage is at or near its crest value, as is clearly shown in Fig. 1(B). The power absorbed by the detector input is accordingly slightly less than the product of the crest signal voltage and the average grid current. Since the average grid current is equal to $\beta E/R$, where β is the completeness of rectification, E the crest value of signal voltage, and R the grid leak resistance, one can write

$$\text{Grid power loss} = \beta E^2/R = \frac{(\text{Effective signal})^2}{(R/2\beta)}. \qquad (4)$$

The denominator of this last term represents the equivalent input resistance to the signal, which is accordingly $R/2\beta$.

Appendix III

Allowable Input Signal

When a signal of carrier amplitude e_s modulated to a degree m is rectified with a completeness β, the maximum potential across the grid condenser is $\beta(1+m)e_s$, and takes place when the signal is at the crest of the modulation cycle with an amplitude $(1+m)e_s$. The most negative instantaneous potential on the grid is the sum of bias and signal, which sum must not exceed twice the proper grid bias for amplifier operation without plate load if distortion is to be avoided. Since the proper grid bias depends upon plate load impedance, being less for lower impedances, the bias E_c recommended for amplifier operation at the same plate potential employed in the detector must be multiplied by a factor k, having a value approximately 0.70, to give the bias for amplifier operation with no plate load. The maximum values of e_s that will not produce distortion must then satisfy the relation

$$2kE_c = (1+\beta)(1+m)e_s.$$

Remembering that E_c is the allowable input for normal amplifier operation, this equation can be rewritten as

$$\frac{\text{Allowable carrier } e_s}{\text{Allowable amplifier input}} = \frac{2k}{(1+\beta)(1+m)}. \qquad (5)$$

This reduces to (2) when $m=1.0$.

Appendix IV

Experimental Technique

A circuit diagram of the experimental set-up is shown in Fig. 9. The modulated oscillator was a conventional oscillator with an a-c voltage superimposed upon the d-c plate battery voltage to produce the modulation. Linearity of modulation was insured by keeping the plate voltage swing to the limits within which d-c tests showed the oscillator output to be proportional to plate voltage. The output of the oscillator was amplified by a broadly tuned radio-frequency amplifier neutralized to prevent self-oscillation. The amplifier and oscillator were inclosed in separate galvanized iron boxes which contained filament and plate batteries as shown.

The voltage delivered to the detector under test was obtained from a potentiometer having a total resistance of 2500 ohms. This low value

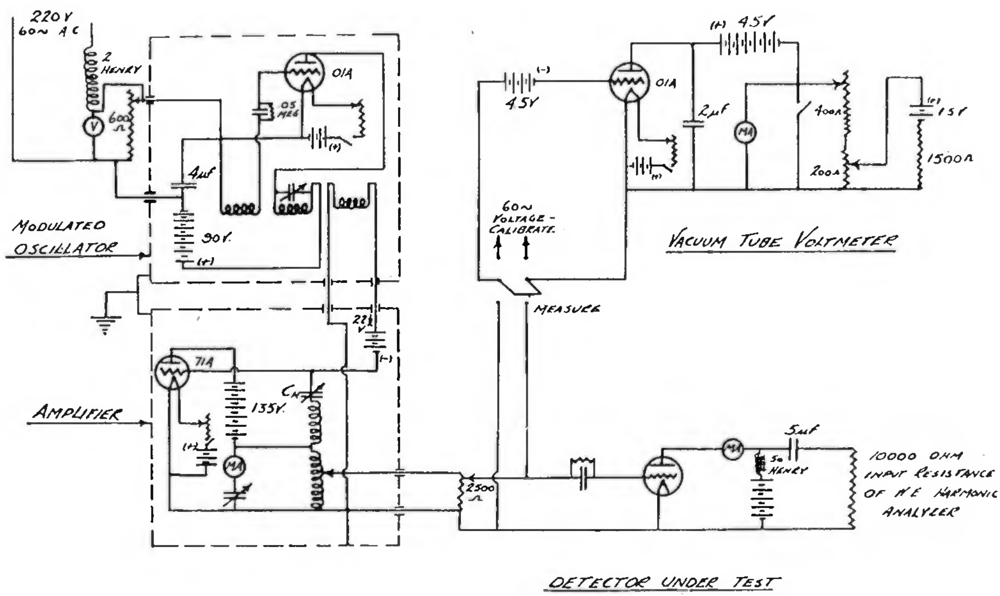

Fig. 9—Circuit diagram of experimental set-up used in making measurements on grid leak power detectors.

was used to prevent the asymmetrical grid losses in the detector from producing wave-form distortion. The voltage applied to the detector was read by a vacuum tube voltmeter which could be calibrated from the 60-cycle line voltage. The rectified output of the detector was delivered to the 10,000-ohm input resistance of a Western Electric "power circuit analyzer" as indicated in the figure.

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ADVANCES IN TRANSOCEANIC CABLE TECHNIQUE*

By

Hobart Mason
(Western Union Telegraph Co., New York City)

Summary—*Progress in the construction of transatlantic cables is described from the earliest to the most recent type, which operates at 1400 letters per minute. Without technical detail the development of cable operating mechanisms is sketched indicating the improvement which resulted from the use of relays, magnifiers, and regenerators. The application to the cable art of landline multiplex telegraph practices is shown. Operating procedures to secure both economy of plant utilization and at the same time good speed of service are noted. The handling of those loaded cables which may not be duplexed and the limitations imposed thereon by traffic and by other plant conditions is discussed. Various cable codes are compared. Service requirements are outlined and usages noted leading up to a study of the economics and the balancing of speed of service against plant and operating costs. Trends of growth are given with data covering the fall of file as to hours of the day, direction, and class of business.*

The subject of transoceanic cable technique is so broad that even though only the most important phases of it be touched, as is the aim of this paper, yet many points pertaining to that subject title must be passed by. Although the whole art is relatively young, barely exceeding the allotted human span of years, yet the most interesting strides have been those of the past decade. Therefore, it seems natural and reasonable to discuss it from its inception.

For convenience in presentation we shall consider cable technique under four heads: cable, apparatus, operation, service.

Cable

From the laying of the first Atlantic cable in 1858 up to 1923 there had been no marked change in the design originally adopted for transatlantic service. Such changes as were made had to do chiefly with the proportions of the various parts: conductor, insulation, and sheathing.

The first transatlantic cable was designed with a conductor of seven number 22-gauge copper wires laid up to form a single strand weighing 107 lbs. per nautical mile. The insulation was composed of three coatings of gutta-percha weighing 260 lbs. per nautical mile. The core (conductor and insulation together) was served with an ample wrapping of jute yarn to act as a protective cushion and also to give a sufficient bulk to allow the sheathing wires to lie evenly around the

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circumference. The sheathing of armor wires served the dual purpose of providing a protective wall and adequate tensile strength. The sheathing was composed of eighteen strands, each containing seven galvanized iron wires of number 22 gauge. The completed cable was treated with a preserving compound of tar, pitch, and linseed oil. It was flexible and mechanically strong but the many small wires in the sheathing were not only difficult to galvanize but presented a large surface for corrosive action.

The next Atlantic cable was sheathed with ten single wires of comparatively large cross section thus exposing a smaller surface to the action of rust. As an additional protection each individual wire was served with manila hemp saturated with a mixture of tar, India rubber, and pitch. This was called open-type sheathing as it did not form a closed wall of metal about the core. Its bulk decreased the specific gravity of the cable and its rough surface increased the surface friction during laying. These were desirable characteristics for laying purposes as they reduced the strain on the paying-out brakes.

The design was used up to 1880 when it was decided that a closed-sheath type forming a solid metallic wall about the core was more desirable. It was found that with the open-sheathed type the cable tended to flatten out while being picked up during repairs. This caused the sheathing wires to take the strain unevenly and frequently caused breakage. Another disadvantage was that when the serving decayed and fell off, the core, surrounded only by a loose cage-like structure, was especially exposed to attacks by marine life.

In the modern closed-type sheathing each armor wire is covered with a preservative compound and a prepared tape, and the sheathed cable has an outer covering of two servings of jute yarn steeped in "freed" coal tar, laid on spirally in opposite directions, alternating with three coatings of compound applied hot. This design is now used almost exclusively with the exception that individual taping of sheathing wires is sometimes omitted.

To protect the core from the borings of teredos and other marine organisms a wrapping of brass tape which entirely encases the core is provided on cables laid in situations warranting it, especially in tropical waters and within certain depth limitations.

The sheathing is varied to conform to conditions existing along the route; thus, in deep water where the liability to mechanical injury is remote, the armor is of the lightest type consistent with sufficient tensile strength to enable the cable to be successfully laid or lifted for repair. In shallow water the nature of the sea bottom, the prevalence of rocks, the occurrence of large ice masses, the risk of damage by ships' anchors or by trawls, and other local conditions determine the amount and kind of protection needed. Quite generally a long ocean cable has four or five degrees of armor protection. While the core remains the same throughout, the over-all size tapers from about the size of a wrist at the shore ends to about the size of a thumb at the deep-sea portions.

Apart from improvements in apparatus design, improved cable speeds were at first achieved only by increasing the size of the core. Step by step this course was followed until one transatlantic cable was laid with a core having 1100 lbs. of copper and 450 lbs. of gutta-percha per nautical mile.

The advantages to be derived from increasing the inductance of cable telegraph circuits had long been recognized but it was not until 1923 that certain nickel-iron alloys possessing special magnetic properties were developed and made the inductive loading of long submarine cables practicable. This loading consists of wrapping the copper conductor throughout its length with a narrow tape or a wire of one of these high permeability alloys. The insulation and mechanical protection of the cable are precisely the same as in nonloaded cables.

The first continuously loaded cable, laid in 1924, extends from Hammel, Far Rockaway, New York, to Horta on the island of Fayal in the Azores. The conductor is composed of a solid copper wire closely surrounded by six helically disposed copper tapes. The loading, applied immediately outside of the copper, is a helical wrapping of "permalloy" tape. Over the gutta-percha insulation is a serving of jute yarn. Then comes the armor, which in the deep-sea section consists of 18 galvanized steel wires, number 13.4 B.W.G. each covered with a prepared tape and the whole served with two layers of jute yarn steeped in "freed" coal tar. In the deep-sea section the physical constants are as follows.

Finished diameter	1.01	in.
Weight of copper	565	lbs. per nautical mile
" " permalloy	85	" " " "
" " gutta-percha	375	" " " "
" in air	2.249	tons " " "
" in water	1.325	" " " "
Breaking stress	10.4	"

The total length of main core is 2328.5 nautical miles, the westerly sea earth is 112.5 miles, the easterly sea earth 2.0 miles, the resistance (main core) is 4460 ohms, capacitance 853 μf and operation is conducted at 1560 letters per minute simplex (65 cycles, five-unit code).

The most recent loaded cable laid extends from Bay Roberts, Newfoundland, to Horta. It is shorter and therefore of smaller core than the one above detailed but is more interesting because it is capable of being operated duplex. To achieve this the cable is loaded throughout its middle portion but as the ends are approached the loading becomes progressively lighter and finally is omitted altogether. This tapering effect makes possible balancing by an artificial line which is of unusual design in that in addition to matching the inductive loading of the cable various other factors had to be taken into account which are neglected when balancing for low speed operation. The total length of the main core is 1341 miles, the sea earths adding 58 miles. The entire weight of the copper conductor is 481,471 lbs., the mu-metal loading 59,886 lbs., and gutta-percha 376,064 lbs. In the deep-sea section the finished cable is 0.82 inches in diameter weighing in air 1.636 tons per nautical mile and in water 0.945 tons. This cable was designed to operate 2000 letters per minute simplex or 1000 letters per minute duplex. Because of the novelty of the design certain margins were allowed. It is interesting to note that it is now duplexed with a balance satisfactory for operation at 1400 letters per minute (58 cycles, five-unit code) and there is evidence that still higher speeds may be obtained. Since higher speeds require closer attention on the part of the regulating attendants it presumably will not be developed until the volume of traffic makes it economical to do so.

Apparatus

The design of cable equipment has undergone more definite changes than has cable design. On the first cable sending was by a double-current key, depressing one lever put negative battery to line, depressing the other lever put positive battery to line. Reception was by a mirror galvanometer. In 1867 Lord Kelvin invented the siphon recorder with its substantial advantage of a permanent record, which device is still largely used. Transmission was improved by designing a transmitter fed with a perforated tape which insured uniform dot spacing and better utilization of line time than direct hand sending. The tape for this transmitter was perforated by a mallet punch, buttons for dot, space, and dash, respectively, being struck by rubber-tipped handles held in the operator's two hands. The mallet punch was superseded by a power-driven machine, generally called a "Klein" in contraction of the manufacturer's name, still widely used, which perforates the tape from an ordinary typewriter keyboard. Evidently the operator's output is much increased with this machine as he makes a single stroke for a single letter as against as many strokes as there are dots and dashes in the letter under the previous methods.

On the receiving side and still prior to the advent of the loaded cable, substantial changes were made. Among the many mechanical kept constantly rotating in order to overcome the friction of rest. This periphery has a middle insulated portion, "no-man's land," flanked on either side by conducting surfaces. The contactor, then, may make slight excursions, such as those due to unwanted parasitic currents within the field of no-man's land, without actuating the sending-on relays. When, however, the contactor is deflected sufficiently to reach one or the other of the conducting surfaces a suitable relay is closed, sending on positive or negative currents.

Of considerable importance is the Heurtley magnifier, one of a number of magnifiers invented and which is still largely in use. In this device the original received impulse, instead of moving a siphon tube or a contacting pointer, slightly displaces a pair of extremely fine wires with reference to adjacent similar fixed wires. This system of wires, having a substantial temperature coefficient of resistance, is connected to form two of the arms of a Wheatstone bridge, and is continuously heated by a battery current. The motion imparted by the receiving coil causes an increase in the separation between two of the wires forming one arm of the bridge with the natural result of lowering their temperature, while at the same time the two wires forming the other arm of the bridge are brought even closer together with a resultant increase in temperature. Thus changes are caused in the bridge balance which are made to actuate more rugged apparatus. Customarily this magnifier works into a drum relay.

The success of the rotary repeater developed for landline multiplex operation led to the development of a synchronous regenerator for ocean cables. The first such regenerator was put in service during 1921 and since that time all Western Union cables have been operated on through the use of this device.

The regenerator consists of a circular distributor face plate with eight concentric rings, the outer four being segmented and the inner four solid. On these rings move brushes mounted on a common insulating frame which is rotated by a La Cour phonic wheel motor brushes mounted in pairs constitute rotating electrical connections between the appropriate solid and segmented rings.

The receiving relays are actuated by but a small part, about 20 per cent in length, of the incoming signal chosen at that part of the signal where the amplitude is the greatest and the likelihood of error due to induced currents the least. As the name regenerator implies, the signal is sent on from a suitable segment with all of the volume and the definition with which it was started out from the office of origin.

The use of vacuum tubes on nonloaded cables, while tried, never reached commercial importance partly for reasons discussed below, and partly because of the fact that it is desirable for the receiving apparatus to have a natural periodicity about the same as that of the incoming signals. Mechanical apparatus could be made to approximate this condition with the result that while sensitive to the desired frequency it was relatively insensitive to unwanted induced currents of other frequencies. The application of the printer to the cable service was a rather natural adaptation of landline practice after other equipment had made possible the securing of substantial five-unit signals at the receiving end.

As a means of increasing the maximum frequency obtainable over a cable when two-element synchronous operation is used, a synchronous vibrating system is applied to interpolate single impulse signals which, due to the attenuation of the cable, are received with so small an amplitude that they will not, of themselves, operate the receiving device. This circuit, based upon the so-called "dot maker" originally suggested by Judd and Davies, supplies locally the incoming signal which, so to speak, might be said to be lost in transmission. If left uncontrolled it indicates alternately plus and minus impulses, but when two or more successive impulses of like polarity are transmitted then the received impulse is sufficiently strong to actuate and override the local mechanism and to cause indications according to the transmission. Through this device the frequency capacity of the cable is in effect doubled for a recognizable signal need be received only when two similar pulses have been sent.

The advent of the loaded cable completely changed the technique of the cable equipment. Until then virtually all receiving equipment used a fundamental receiving unit based upon the principle of the d'Arsonval galvanometer. The speed of the loaded cables is comparatively so great that this type of receiving unit is not capable of following the high frequency of the incoming signal.

Vacuum-tube amplifiers were rather difficult to apply to the very low frequencies of the old-style cables due to the fact that it was difficult to build coupling transformers or interstage coupling networks that would pass the low frequencies associated with these cables, but this type of amplifier lends itself nicely to the conditions encountered in the new high-speed loaded cables.

The cable shaping networks, consisting of magnetic shunts and shunted receiving condensers, were not discarded when the amplifier was put on the cable. In fact the use of the amplifier made it possible to increase the use of these input shaping networks. Shaping could be carried further as only a small voltage change on the grid of the first tube is necessary with this type of amplifier, whereas with the galvanometer type enough energy must be left in the signal after shaping to operate the galvanometer properly.

In addition to the input networks used with the amplifier, it becomes possible to couple the amplifier with shaping networks so that refinements can be achieved through a certain amount of shaping applied between each stage.

The output of the amplifier has sufficient energy to operate a high-grade polar relay which in turn works into the active recording equipment used in connection with the cable.

Because the limits of operation are, in general, set by parasitic induced currents it is best to consider the received impulses as measured by their voltage level rather than according to current or energy values. In landline telegraphy we like to receive signals at a level of perhaps 25 volts. On long ocean cables satisfactory operation suggests 30 mv and if the Heurtley magnifier is used, 10 mv. With the vacuum-tube amplifier we can well work down to levels of 3 mv. These figures indicate the increasing care which must be given in high-speed working in protecting from stray currents and especially in refining the artificial line where duplex balances are attempted.

Receiving equipment quite similar to landline multiplex printing equipment is used with loaded cables. The principal modification necessary to adapt landline multiplex to loaded cable working is to provide means for automatically reversing the direction of transmission since most loaded cables are not designed for duplex operation.

Each of the simplex loaded cables has connected with the multiplex distributor a timing train arranged so that it can close a contact periodically once every half minute or multiple thereof. An adjustment provides for varying this time from one-half to eleven minutes. When the timing contact closes, an automatic motor-driven cam switch is brought into operation. Since the circuit is running in synchronism the switches at each station are actuated at the same instant and at their respective ends each switch performs the following operations: at the sending station—stops the transmitters, switches the cable from sending to receiving, removes short circuit from amplifier input, and cuts in printers; at the receiving station—cuts off the printers, short-circuits the amplifier input, switches the cable from receiving to sending position, and starts transmitters.

These functions are all performed with proper time intervals between each operation and although the switching may take place in the middle of a work, when transmission is resumed in the same direction it will be taken up exactly where left off, without errors. The time needed to perform these functions and reduce the cable to a condition suitable for operation in the opposite direction is about 6 sec., hence a turnaround every two minutes results in a capacity loss of 5 per cent.

In order to decrease the time interval between the perforating of a message and its actual passage through the transmitter it is desirable to shorten the loop of tape which necessarily exists between the perforator and transmitter, yet some storage must be provided if the economic waste of having the cable run idle during pauses in the operator's performance is to be avoided. Toward this end a storing transmitter is used. Instead of perforating holes in a tape a free keyboard sets up pins in a cylindrical machine which are scrutinized by appropriate sending contacts which put the proper electrical impulses to the line. The transmitting members may catch up to the pin-setting members, or the latter may get ahead to the extent of the number of pins provided by the periphery of the cylinder as the relation between operator performance and cable transmission varies.

Operation

Prior to 1871 transoceanic cables were operated in one direction only but subsequent to that date most of the nonloaded cables have been operated duplex. The requirements for an artificial line for a cable are vastly more rigorous than those for landline use. The capacities are, of course, very much greater than anything known in landline service. Recently temperature-controlled rooms in which to store them have been developed in order to reduce changes in electrical constants due to temperature variations.

To avoid trouble from other cables landing at the same station and, too, from ground potentials engendered by electrical enterprises on shore, it is necessary when duplexing a cable to connect the ground side of the artificial line not directly to the ground at the office but to an extra conductor called a "sea earth" which is a separate core identical with the main core and included with it inside of the protective armor. At a point which varies from a few miles in some old-type cables to 112 miles in one instance of a modern loaded cable, the sea earth is brought out, the conductor is exposed, and electricially bonded to the cable sheath wires.

Prior to 1918 all transatlantic cables were operated in sections with manual relay stations at least at each end of the long ocean sections. The development of the repeaters previously mentioned and especially the rotary regenerator (which, along with synchronous channelization of the faster old cables, somewhat preceded the loaded cables) made it practical to operate from terminal to terminal without any intermediate handlings. Today direct circuits are operated between New York and London, Liverpool, Berlin, Paris, Havre, and between Boston and London. A similar circuit has been worked between San Francisco and London but merely for demonstration purposes (as traffic conditions do not warrant such a set-up). That circuit had repeaters at some 17 points along its length.

An operator can perforate cable business at about 300 letters per minute, a rather lower rate than is customary for landline operators because of the greater difficulty due to code words in cable traffic. With cables now capable of going 1500, 2000, or 2400 letters per minute it is evident that channelization must be used and various numbers of channels on cables both old style and loaded, are in use or planned for up to a present maximum of eight. Channelization of course means the distribution of line time between various sets of operating equipments through synchronous distributors at the two ends of the circuit. As synchronism has to be preserved through the relays, it follows rather naturally that the rotary regenerative repeater should be used. With channels not exceeding 300 letters per minute transmission and reception can be carried on quite as in landline multiplex practice and with the same equipment of perforator, transmitter, and printer.

The development of the regenerator has greatly improved operating efficiency in several directions. A few years ago the operating lost time on cables might be as much as 15 per cent whereas now it is customarily held below 4 per cent. Prior to 1918 each intermediate cable station required a complete complement of both sending and receiving operators in both directions, now only the testing and regulating attendants remain.

Where a simplex loaded cable constitutes the only circuit between two points it must, of course, be turned around at frequent intervals both for the sake of the speed of service of the regular business, and to permit expeditious handling of inquiries and corrections regarding errored messages. When, however, such a cable is one of a group of circuits between two given termini then it may be worked for substantial periods in one direction only. The difference in the fall of business eastbound and westbound as indicated on curves subsequently shown, makes it especially desirable to take advantage of this facility. In such case it is not necessary for the cable to be automatically switched because the loss of time in ordering it turned around by message is so small a fraction of the whole time as to be unobjectionable. In actual practice the loaded cable which extends from New York to London normally operates about as follows, the hours of turnaround and even the actual number of turnarounds varying somewhat from day to day according to load conditions. It must be remembered that this performance is possible only because of the several paralleling cables that work in both directions all the time.

Start Turn	
West	5:00 A.M.
East	9:00 A.M.
West	9:45 A.M.
East	10:00 A.M.
West	10:25 A.M.
East	10:45 A.M.
West	11:15 A.M.
East	1:45 P.M.
Close down	10:30 P.M.

Figs. 1 and 2 show respectively the eastward and westward fall of traffic by hourly periods. The effect of the five-hour time difference between London and New York is clearly evident. So, too, is it evident that customers handle their cable file much as they do their mail. That is to say, they tend to allow their cables to accumulate and file them all at the hour of closing the business office just as they post most of their letters at that hour. In American landline telegraph practice this is not the case, the load curve, like the telephone curve, showing a major peak shortly after 10 A.M. and an almost equal peak around 3 P.M. Perhaps the difference in clocks can partially explain the peculiarity of the cable curves. Eastward any cable not filed before noon will await the next morning for delivery anyhow unless night reception is provided for. Westward anything transmitted prior to 2 P.M. is too early for delivery in New York. The public's attitude is hardly logical and round-trip communications could well be handled within a given day. Arbitrage business frequently is completed with a round-trip time of but two or three minutes. The fact seems to be that customers are not educated to look for and avail themselves of the service offered.

Of the many signaling codes that have been developed for telegraph purposes only two find extensive use in cable operation. Notable for their absence are the two methods using Morse code common in domestic manual handlings, (a) for short and way circuits, the open and closed

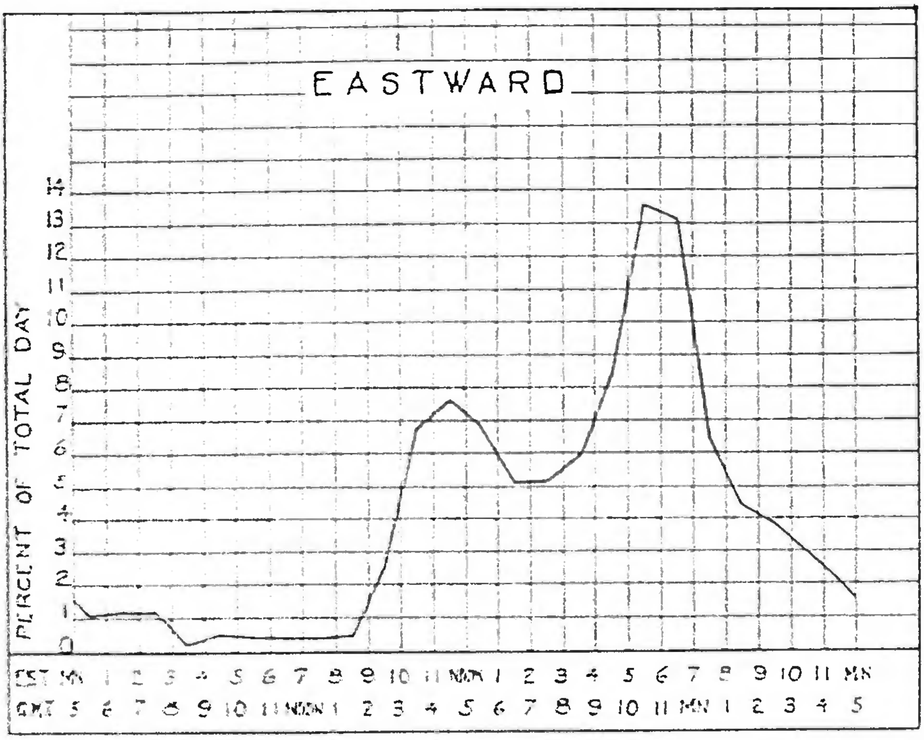
EASTWARD
PERCENT OF TOTAL DAY
0 1 2 3 4 5 6 7 8 9 10 11 12 13 14
EST MN 1 2 3 4 5 6 7 8 9 10 11 NOON 1 2 3 4 5 6 7 8 9 10 11 MN
GMT 5 6 7 8 9 10 11 NOON 1 2 3 4 5 6 7 8 9 10 11 MN 1 2 3 4 5

Fig. 1

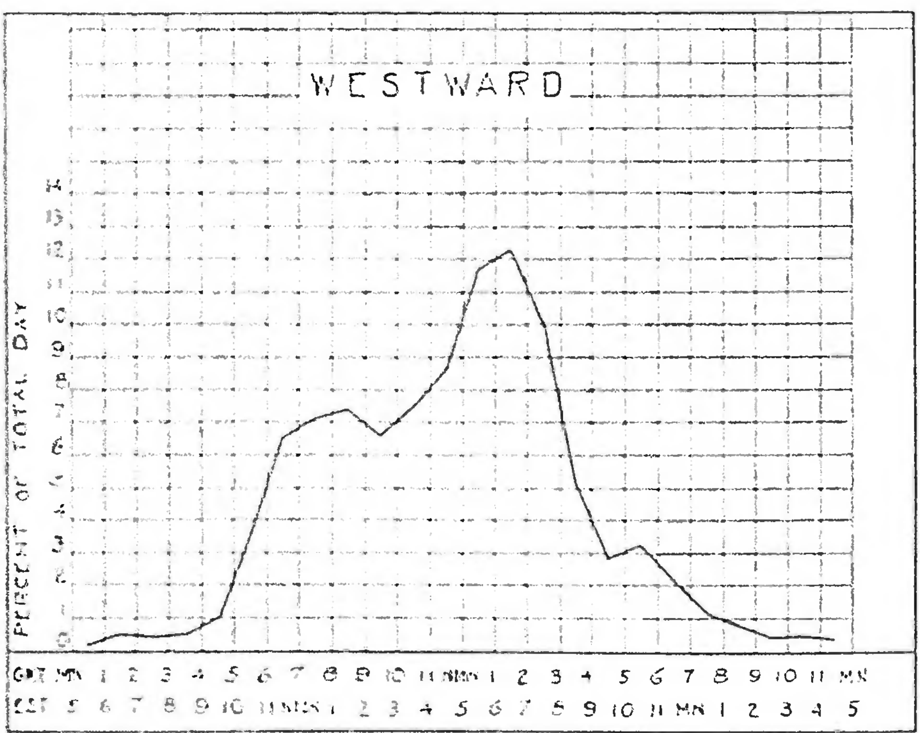
WESTWARD
PERCENT OF TOTAL DAY
GMT MN 1 2 3 4 5 6 7 8 9 10 11 NOON 1 2 3 4 5 6 7 8 9 10 11 MN
EST 5 6 7 8 9 10 11 NOON 1 2 3 4 5 6 7 8 9 10 11 MN 1 2 3 4 5

Fig. 2

circuit where a short contact is a dot, a long contact a dash, and the open position making a space which by its length denotes space between elements of a character, space between characters, or space between words; (b) for longer trunk circuits, the polar operation, where positive polarity to line is "marking," negative is "spacing," the length of marking and of spacing denoting as in (a). In important landline working it is impracticable to operate with significant periods of no current on the line due to the action of induced parasitic currents. On cables, however, a no-current interval may be made significant and so, because of the shorter code possible through the permutation of three rather than two elements, the "cable code" of three elements was, and except for printer operation, still is generally used. Here positive polarity to the line is a dot, negative a dash (dots and dashes being the same length) and no current provides the spaces between elements, letters and words. For printer operation it was found desirable to revert to the more rugged two-element five-unit code used on landline multiplexes, no reliance being placed on the no-current interval. This makes all characters five units long and, failing to avail of short signals for frequently occurring letters, would have considerably increased the line time had it not been possible to use the interpolator synchronous vibrating system or "dot-maker" previously described which device is not applicable to three-element codes. An analysis of traffic as it occurs indicates that the relative lengths of these codes (including spaces) are as follows.

Continental Morse	2 element	9.27	time units per character
Cable code	3 "	3.71	" " " "
Five-unit printer	2 "	5.00	" " " "
Five unit with synchronized vibrator		2.50	" " " "

The cable code, being transcribed from recorder tape by typewriters, permits of certain short cuts not tolerable where printer reception produces the delivery copy. Contractions, particularly of place names, the use of "short figures" (e.g., *A* for 1, *U* for 2, etc.) in cable code, and the necessity for carriage shift signals in printer code together make about 16 per cent difference in favor of cable code so that the length of cable code might be considered to that of printer code as 3.12 is to 2.50.

The ratio just mentioned leaves out of consideration two factors which are of significance in considering the volume of business which can be moved under the two methods. With a given voltage limitation, say 50 volts to ground, evidently the swing from 50 volts minus to 50 volts positive in the two-element system is twice as great and more likely of proper reception than is a swing from zero to 50 volts in, say, the plus direction. Thus the two-element signals are more rugged than the three-element signals and higher speeds resulting in greater attenuation may be tolerated. On the other hand, in favor of the cable code is the fact that within limits an operator can read through a faulty signal or will at least know that it is questionable, while a printer with a faulty signal will boldly print something wrong which may not be discoverable if in a code message. The numerical values to be attached to these two facts are not readily determinable but practice indicates that a substantially increased volume of traffic will be handled by the two-element code.

Service

As generally with concerns providing service to the public, all progress is headed toward two goals, one the improvement of the service, the other a reduction of the cost.

The major item of improvement in service has undoubtedly been the reduced over-all time from origin to destination of cable messages due to the elimination of intermediate manual retransmissions. Prior to the introduction of cable repeaters such manual handling was necessary at least at the ends of the ocean sections, in Newfoundland on the one hand, and Ireland, or the southwest extremity of England, on the other. Generally, too, a handling in Nova Scotia was involved and over certain routes all four places mentioned were involved. Really a part of the same improvement is the increased celerity in securing corrections to mutilated or obviously errored messages.

Accuracy is another desideratum and evidently the elimination of manual retransmissions results in improvement in this particular. It is possible to develop machines that will have fewer than any stated number of errors per thousand words simply by spending enough money in perfecting the apparatus. It is not possible, however, to improve the operator indefinitely; the best performance of which we have record covering an extended period, showing 0.628 operator errors per thousand paid words handled. Errors due to mechanical failures are much less numerous than this.

Base Rates Between New York and London

Date	Full Rate
July 28, 1866	$100.00 for 10 words or less
Nov. 1, 1866	50.00 " " " " "
Dec. 1, 1867	25.00 " " " " "
Sept. 1, 1868	16.41 " " " " "
June 1, 1869	10.00 " " " " "
Aug. 10, 1869	7.50 " " " " "
May 1, 1872	1.00 per word
May 1, 1877	0.75 " "
Oct. 1, 1880	0.50 " "
Sept. 1, 1888	0.25 " "
Apr. 20, 1923	0.20* " "

* Note: An "XUR" rate of 25 cents for a limited amount of extra-fast traffic is still maintained.

Perhaps properly classed as improvements in service are the reduction of rates from time to time and the introductions of various classes of cable traffic. The above table interestingly shows reductions that have occurred in the rates for regular service.

On December, 6, 1911, a "Deferred Rate" of just half the normal rate was put into effect and at the same time the cable letter and week-end letter rates were established. The "Deferred Rate" has remained at half the normal (except when entirely suspended under pressure of wartime traffic) ever since, but the latter rates have varied as follows:

Date	Cable Letters	Week-End Letters
Dec. 6, 1911	$1.50 for 20 words	$1.50 for 30 words
Jan. 1, 1913	.75 " 13 "	1.15 " 25 "
Apr. 1918	suspended	account war
Apr. 20, 1923	1.20 for 20 words	1.00 for 20 words
Mar. 4, 1929	1.00 " 25 "	0.75 " 25 "

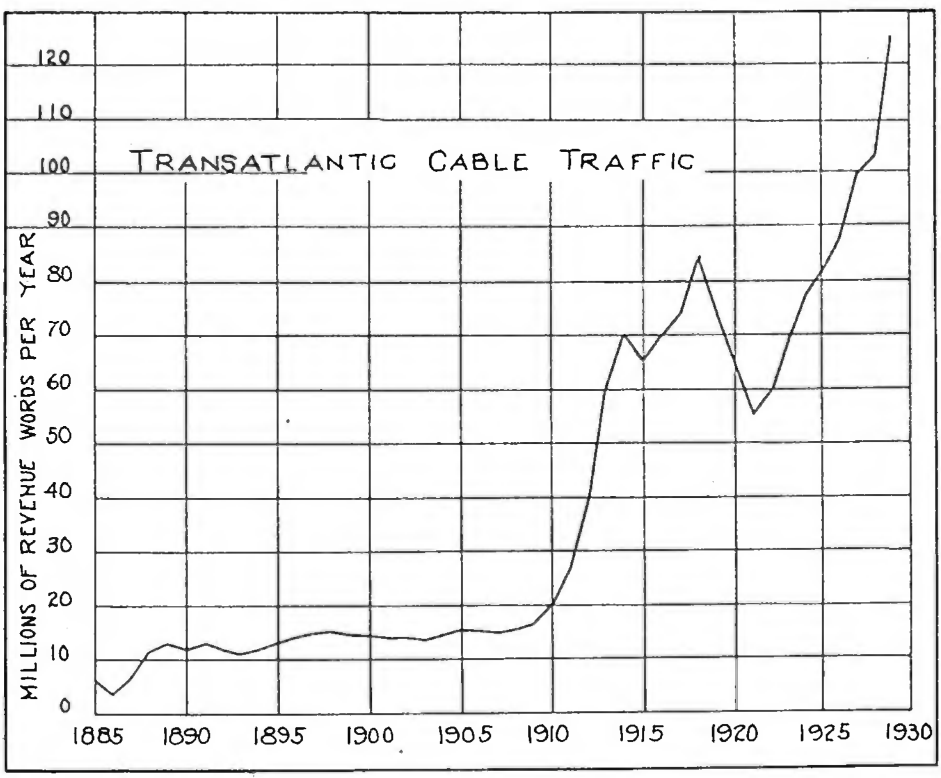

Fig. 3

On the cost side the two big items of improvement are, of course, the reduction of manual retransmission expense and the vastly increased carrying capacity per cable due to loading. As the circuit cost chargeable against a message is reduced through increased carrying capacity per cable so it progressively becomes increasingly economical to provide more adequate facilities for the volume of the file. The more adequate these facilities are in proportion to the file the more nearly complete "readiness to serve" for every filed message will be found. It has not been found economical as yet to engineer the cable plant on an almost 100 per cent "readiness to serve" basis but a much better proportion of business is handled within short time limits now than was formerly the case. As in any communications industry the cable file is extremely uneven throughout the twenty-four hours as is indicated by Figs. 1 and 2 which show respectively the fall of traffic eastward and westward in transatlantic cable service. That the improvement in the accuracy and the speed as well as the reduction of the cost, both absolutely and through the introduction of deferred services, has greatly stimulated transatlantic business as indicated in Fig. 3 which shows for one of the larger cable systems the growth in transatlantic cable traffic for the past forty years.

That the popularity of the letter services is increasing and doubtless is in good measure responsible for the considerable growth in total traffic is well shown by the following comparison of only four years' spread.

Classification of Paid Words Transmitted

	1926		1930	
	East	West	East	West
Full Rate	40 per cent	42 per cent	24 per cent	28 per cent
Press	7	18	8	24
Deferred	28	19	16	14
Letters	25	21	52	34

It may be interesting to note that the entire traffic handled divides up about as follows:

Letters and spaces in paid-for matter 74 per cent

Unpaid letters and spaces in paid business (place from, data, message number., etc.) 18 per cent

Unpaid business (services, deadheads, inquiries for corrections and replies) 5 per cent

Operating functions (shift from upper to lower case or vice versa) 3 per cent

A recent analysis of cable traffic shows the following regarding the length of paid cable messages, spaces being counted as characters.

	Characters per Word	Words per Message
Full rate, code	9.35	9.9
Full rate, plain language	7.20	16.0
Deferred rate, plain language	6.79	20.9
Press	7.13	81.7

Some curious and not readily explained fluctuations occur in the ratios of eastward and westward cable traffic. However, the general trend toward a greater file east than west can be associated with a preponderance of westward transatlantic radio business and this, at least in part, is explained by the political, commercial, and service conditions on the respective sides of the ocean. The figures for the past decade are

PER CENT OF REVENUE WORDS

	East	West
1921	50	50
1922	47	52
1923	51	49
1924	55	45
1925	55	45
1926	53	47
1927	59	41
1928	58	42
1929	59	41
1930 (part)	57	43

As necessary in covering a topic so broad, this paper touches only lightly on the many high spots of advance, both technical and operating, in the cable art. It is hard to imagine in the near future further advances comparable with those of the recent past. But the continually increasing use of cable services by the public in the face of increasing competition from radio and from improved mail services, together with the spirit which animates the cable industry, makes the author believe that a survey taken, say ten years hence, will show changes and improvements perhaps rivaling in importance those of the years herein reviewed.

BOOK REVIEWS

The National Physical Laboratory Collected Researches, Vol. XXI, 1929. Published by His Majesty's Stationery Office, London, 448 pp. 10 x 12 paper. Price £1 2s. 6d.

This is a collection of 22 papers describing work done by the British National Physical Laboratory. All of these papers are on electrical or magnetic subjects and 12 are on radio. With the exception of a paper by Dye and Hartshorn, "On a primary standard of mutual inductance for presentation to the Imperial Government of Japan," the papers of this collection have been published elsewhere between the years 1924 and 1928. Of the radio papers one is by Wilmotte, two are by Hartshorn, two by Dye, five by Smith-Rose and Barfield, and two by Hollingworth. Of the remaining papers, one is by Dye and Hartshown, one by Webb, one by Hartshorn and Wilmotte, two are by Wilmotte, and five by Hartshorn. Of this latter group some papers such as Hartshorn's, "A method of measuring very small capacities," may apply to radio. It is a great convenience to have these otherwise scattered papers collected into one volume, though it is difficult to understand why publication in this form is not kept more nearly up-to-date.

S. S. KIRBY*

Radio Operating Questions and Answers, by Nilson and Horning. Third edition, 267 pp. Published by McGraw Hill Book Company, Inc., New York, N. Y. Price, $2.00.

This book has been written especially for students and operators about to take the government examination for a radio operator's license. It contains questions and answers covering the field of commercial, broadcast, and amateur radio operating. This material serves to bring out the salient points and to show the general form of answering questions. "Questions and Answers" should be supplemented with a radio textbook as it is not a complete text in itself.

S. S. KIRBY*

Radio and its future, edited by Martin Codel. Published by Harper & Brothers, New York. 349 pp. 9 illustrations. Price $4.00.

What is the present status of radio and what is going to be its future place in American life? These questions are discussed in the book, "Radio and Its Future," by prominent men of the radio field. How did broadcasting start? H. P. Davis, vice president of the Westinghouse Company, tells us of the early days of KDKA. Who pays the radio bill? This question is discussed by the editor in his chapter on the radio structure. What are the facts about television? Dr. Herbert Ives of the Bell Laboratories reviews this subject for us. These questions and many more on broadcast communications, industry, regulation and scientific considerations are discussed in language which a layman can understand.

The book closes with a chapter on the future of radio by Dr. Lee deForest who has seen "his own inventions develop from crude experiments to world-wide institutions beyond the wildest flights of fancy," and who is not afraid of making rash predictions.

This is not a technical book but an interesting summary of the present status of radio.

S. S. Kirby*

*Bureau of Standards, Washington, D. C.

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed on this page may be obtained gratis by addressing a request to the manufacturer or publisher.

Bulletin 932 of the General Radio Co., 30 State St., Cambridge, Mass., is a 52-page catalog of apparatus, accessories, and parts published for service men, amateurs, and experimenters.

A loose-leaf folder containing several "Technical Bulletins on RCA Radiotrons" is available from the Commercial Engineering Department of the RCA Radiotron Co., Inc., Harrison, N. J. These technical bulletins describe the UX-171-A, UX-222, UY-224, UX-226, UY-227, UX-245, UX-280, and UX-281, tubes. In most cases a family of plate-voltage, plate-current curves are supplied for each tube. A chart giving average characteristics of Radiotrons is also included in the binder.

A four-page folder describing Federal anti-capacity switches is available from the Federal Anti-Capacity Switch Corp. 129 Dearborn St., Buffalo, N. Y.

The Webster Co. of 850 Blackhawk St., Chicago, has recently issued a catalog in the form of a loose-leaf folder describing its complete line of amplification equipment.

Vacuum thermocouples are described in Bulletin 3130 of the Thermal Instrument Co., 38 Woods Ave., Somerville, Mass.

Puncture Proof Filter Condensers is the title of a 60-page booklet describing various electrolytic condensers made by the Amrad Corporation, Medford Hillside, Mass.

The advantages of aluminum die castings is the subject of a 24-page booklet recently issued by the Aluminum Company of America, of Pittsburgh, Pa.

A 32-page booklet recently published by the Aerovox Wireless Corp., 72–80 Washington St., Brooklyn, N. Y., describes the complete line of dry electrolytic condensers made by this concern.

Bulletin No. 6D of Jenkins and Adair, 3333 Belmont Ave., Chicago, describes their type C condenser microphone and accessories. Bulletin No. 10A describes a variable attenuator for broadcast or sound picture purposes. Transformers, gain and mixing controls, amplifier accessories and other audio frequency apparatus for public address, sound picture, or broadcast equipment are listed in Bulletin No. 1D.

Proceedings of The Institute of Radio Engineers
Volume 18, Number 12 *December, 1930*

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," a revised edition of Bureau of Standards Circular No. 138 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 (first edition of Circular 138) 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.

R100. RADIO PRINCIPLES

R113 Ratcliffe, J. A. and White, F. W. G. The electrical properties of the soil at radio frequencies. *Philosophical Magazine* (London), 10, 667–680; October, 1930.

Giving the results of laboratory measurements of the effective conductivity and the effective dielectric constant of the soil and their variation with frequency. The bearing of these results on some problems of the propagation of radio waves is discussed.

R113.5 Stetson, H. T. The influence of sun spots on radio. *Jour. Franklin Institute*, 210, 403–419; October, 1930.

The author attempts to correlate sun spot phenomena, magnetic storms, and radio reception. Curves, drawn from data taken over a period of years, indicate that magnetic effects on the earth vary directly with sun spot numbers, and that the intensity of radio signals varies inversely with numbers of sun spots. A description of the recording apparatus used in taking the data is given.

R113.6 De Groot, W. Some remarks on the analogy of certain cases of propagation of electromagnetic waves and the motion of a particle in a potential field. *Philosophical Magazine* (London), 10, 522–40; October, 1930.

The conditions are investigated for which a wave-equation of the type $\Delta\Phi+(\omega^2/\mu^2)\Phi=0$ may be such that a group of waves move as a mass-point in a field of force which is independent of the frequency of the group. The movement of a group of electromagnetic waves in a medium containing charged particles may also be made to correspond to that of a particle in a potential field of force. The formulas derived are applied to the case of radio waves reflected by the Kennelly-Heaviside layer.

R120 Wilmotte, R. M. General formulas for the radiation distribution of antenna systems. *Jour. I. E. E.* (London), 68, 1174–90; September, 1930.

Formulas are given for the distribution of radiation in different directions. The effects of the antenna elements, of polarization, of current distribution, and ground images are discussed. A number of special cases are deduced from these general formulas, and curves are given to facilitate computation.

R120 Wilmotte, R. M. The radiation distribution of antennas in vertical planes. *Jour. I. E. E.* (London), 68, 1191–1204; September, 1930.

The radiation distribution of an antenna in a vertical plane was obtained by measuring in an aeroplane the strength of a received signal from an excited antenna on the ground. The position of the aeroplane relative to the antenna was obtained by means of a theodolite on the ground, the signal strength being recorded on a photographic film. Substantial agreement was obtained between theory and experiment, the discrepancies increasing with the frequency.

R125 Bäumler, M; Krüger, K; Plendl, H; Pfitzer, W. Strahlungsmessungen an Kurzwellen-Richtantennen der Grossfunkstelle Nauen. (Field intensity measurements of the radiation from the directional antenna system of the Nauen transmitting station). *Zeits. für Hochfrequenztechnik*, 36, 1–13; July, 1930.

A report on the experimental investigation of the radiation characteristics of the 64-element short-wave directive antenna ($\lambda = 16.92$), set up at Nauen for communication with Japan. The experimental results agreed closely with results calculated from the theory.

R133 Okabe, K. On the magnetron oscillation of new type. Proc. I.R.E., 18, 1748–49; October, 1930.

A report of the experimental results attained with a new type of magnetron oscillation. Comparatively intense oscillations with a wavelength of the order of 10 cm were produced.

R134 Weiller, P. G. A new power detector system. *Radio Engineering*, 10, 33–34; October, 1930.

A power detector circuit wherein the grid leak and condenser are omitted, grid bias being obtained by inserting an automatic, nonlinear resistance in the grid-plate return.

R140 Wigge, H. Die Frequenzabhängigkeit des Widerstandsverstärkers. (The frequency characteristic of resistance-coupled amplifiers.) *Zeits. für Hochfrequenztechnik*, 36, 24–27; July, 1930.

A discussion of the principles involved leads to a method of determining the constants of a resistance-coupled amplifier having a linear frequency characteristic over a desired frequency band and having a maximum amplification under these conditions.

R143 Butterworth, S. On the theory of filter amplifiers. *Experimental Wireless & W. Engr.* (London), 7, 536–41; October, 1930.

In this work the problem of filtering is attacked from a new angle in which use is made of systems of simple filter units separated by tubes so that the property of filtering with that of amplification is combined in one amplifier. The filter units may be made very small and compact even for band-pass work and the latter exhibit the added advantage over the orthodox type in that since the resistance is under complete control, uniform sensitivity can be obtained in the pass region.

R145.3 Loos, C. Experimentelle Untersuchungen an Spulen mit leitenden Kernen und Hüllen. (An experimental investigation of coils having conducting cores and covers). *Zeits. für Hochfrequenztechnik*, 36, 13–24; July, 1930.

Nonmagnetic, conducting cores and covers were found to have the following effects on the capacity, self-inductance and resistance of a coil. (1) The capacity of the coil is appreciably increased only when the clearance between coil and cover or core is very small. (2) The self-inductance of the coil is reduced by certain cores and covers in which case the resistance increase of the coil is small; again the self-inductance may be little affected and in this case the resistance of the coil is considerably increased.

R145.3 Scott, K. L. Variation of the inductance of coils due to the magnetic shielding effect of eddy currents in the cores. Proc. I.R.E., 18, 1750–64; October, 1930.

An analysis is made of the shielding effect of eddy currents on the flux in the interior of cores of cylindrical or flat sheet material. It is shown that the counter voltage of self-inductance of an iron-cored coil is due only to the component of flux in the core which is in phase with the flux at the surface of the core. Expressions are obtained and curves plotted showing the variation of inductance of a coil with frequency, or with the conductivity and permeability of the core material. Sample calculations and some experimental results are given. The results show that the inductances at high frequencies are actually less than the predicted values, which leads to the suspicion that some factor other than eddy currents causes the flux in the interior of the core to decrease with increasing frequency.

R200. Radio Measurements and Standardization

R201 McLennan, J. C. and Burton, A. C. Method of obtaining a visible spectrum of waves of radio frequency. (Letter). *Nature* (London), 126, 130; July 26, 1930.

It was found that for a given wavelength there is a maximum heating effect produced in a medium, the specific conductivity and dielectric constant of which are connected with the frequency by a simple law. A glass tube filled with agar-agar, tetriodomercurate of silver, and a few drops of an electrolyte, when placed in a radiation field acts as a crude spectrograph.

R213 Hall, E. L. Accurate method of measuring transmitted wave frequencies at 5000 and 20,000 kilocycles. *Bureau of Standards Journal of Research*, 5, 647–52; September, 1930.

The measurement of the frequency of a radio transmiting station offers a convenient means of intercomparison of frequency standards. While there have been several articles published with station frequency measurements, these have, for the most part, considered frequencies in the broadcast band or lower. This paper describes a method of measuring station frequencies applicable to any frequency, but dealing primarily with frequencies of approximately 5000 and 20,000 kc. A high order of precision is obtained by the use of harmonics and audio-frequency-beat notes. Most of the units of the equipment employed have been previously described in connection with other applications.

R231 Illgen, H. Einfluss der Oberflächenbeschaffenheit von Drähten auf die Selbstinduktion bei hohen Frequenzen. (The influence of the surface nature of a wire on its self-inductance at high frequencies). *Zeits. für Hochfrequenztechnik*, 36, 50–62; August, 1930.

Using an interference method for measuring the self-induction of straight wires of uniform length and cross section but having different coverings, it was found within the frequency range of 1000–15,000 kc: (1) that oxidation of thin insulating coverings had no measurable effect; (2) that the equations of Försterling are reliable for calculating the self-inductance of wires with conducting covering, and (3) that electrolytic coverings have a great influence which is, however, mostly capacitative.

R241 Benz, F. Dämpfungsmessungen an Induktivitäten bei Hochfrequenz. (The measurement of damping in inductances at high frequencies). *Zeits. für Hochfrequenztechnik*, 36, 41–49; August, 1930.

Ordinary methods of measuring the damping resistance were found to be unreliable at high frequencies. Accordingly a satisfactory thermal method was developed, which the author describes.

R242 Raghava, Rao, S. V. and Watson, H. E. The measurement of small currents. *Experimental Wireless and Wireless Engr.* (London), 7, 552–556; October, 1930.

A study of the grid current characteristics of different types of tubes and their suitability for measuring small currents. It was found that the main requirements of a tube for this work are: extremely good insulation of the grid (a conducing factor being low filament temperature), high mutual conductance and low impedance.

R242.1 Schlesinger, K. Ein neuer Strommesser für Hochfrequenz. (A new ammeter for high frequencies). *Zeits. für Hochfrequenztechnik*, 36, 62–65; August, 1930.

Describing a new type of thermoammeter for measuring small high-frequency currents. A small taut filament is mounted in a vacuum tube and set vibrating by an audio oscillator. Current passing through the filament increases its temperature and thereby changes its natural frequency. The condenser control of the oscillator is calibrated in amperes. The instrument has an internal resistance of one ohm, is accurate to within 1 per cent and may be overloaded 1000 per cent without injury.

R243.1 Hayman, W. G. A compensated vacuum-tube voltmeter with balanced bridge output. *Experimental Wireless & Wireless Engr.* (London), 7, 556–59; October, 1930.

The author describes a balanced bridge output or indicator circuit for the thermionic voltmeter. The advantages over the conventional type of output circuit are: high input impedance, stability, and immunity of the galvanometer from accidental current unbalance.

R270 Fassbender, H; Eisner, F; and Kurlbaum, G. Untersuchungen über die Ausbreitungsdämpfung elektromagnetischer Wellen und die Reichweiten drahtloser Stationen im Wellenbereich 200–2000 m. (The attenuation of electromagnetic waves and the range of radio stations using 200-to 2000-meter wavelengths). *Elektrischen-Nachrichten Technik*, 7, 257–76; July, 1930.

The results of a large number of field intensity measurements taken during 1928–1929, were recorded and assimilated in such a manner as to facilitate the mathematical determination of electromagnetic wave propagation over land, within the wavelength range of 200 to 2000 meters.

R300. Radio Apparatus and Equipment

R339 Stevens, H. B. Electron tube relays. *Radio Engineering*, 10, 23–25; October, 1930.

A discussion of the characteristics and industrial applications of electron-tube relays, including grid glow tubes, photoelectric cells, and phototube amplifiers.

R355.21 Kaar, I. J. and Burnside, C. J. Some developments in broadcast transmitters. Proc. I.R.E., 18, 1623–60; October, 1930.

After briefly summarizing the present development of broadcast transmitters, the authors give a detailed description of several types of transmitters, ranging in power from 100 watts to 50-kw, that have been developed in the laboratories of the General Electric Co.

R387.1 Baerwald, H. G. Das Amplitudensieb, eine Anordnung zur Amplitudenstatistik unregelmässiger Vorgänge. (An amplitude filter, an arrangement for determining the amplitude characteristics of irregular wave trains). *Elektrische-Nachrichten Technik*, 7, 362–68; September, 1930.

A special type of push-pull circuit feeding into a linear output tube thence to an integrating instrument, is used to examine peak values of an irregular wave.

R400. Radio Communication Systems

R400 Lee, A. G. The radio communication services of the British Post Office. Proc. I.R.E., 18, 1690–1731; October, 1930.

A presentation of the main outline of the services and varied activities of the British Post Office in the fields of radiotelephony, telegraphy, and broadcasting, with a more detailed reference to special features of equipment.

R500. Applications of Radio

R521.1 Kruse, R. S. A multi-range receiver with four tuned circuits. *QST*, 14, 21–25; October, 1930; *Electronics*, 1, 336–37; October, 1930.

A multi-range (235–8000kc), single control, four-circuit aircraft receiving set, incorporating latest advances in receiving set design, is described in detail by the author.

R530 Isbell, A. A. The RCA world-wide radio network, Proc. I.R.E., 18, 1732–42; October, 1930.

An outline of the development of the RCA communication system during the past ten years and a description of the present world-wide network. Some reference is made to the type of equipment used.

R600. Radio Stations

R600 Quäck, E. and Mögel, H. Dichte der Kommerziellen Kurzwellenstationen. (The congestion of commercial short-wave stations). *Elektrische-Nachrichten Technik*, 7, 277–79; July, 1930.

A short discussion of the problems to be solved if serious congestion in the commercial short-wave bands is to be prevented.

R612.1 Eckersley, P. P. and Ashbridge, N. A wireless broadcasting station for dual programme service. *Jour. I. E. E.* (London), 68, 1149–73; September, 1930.

The British Broadcasting Co. plans to build a number of twin transmitting stations capable of radiating two programs simultaneously at different frequencies, and each at a power of 50 kw in the antenna. The first one has been completed at Brookmans Park, Hertfordshire, England. The problems encountered during its construction are discussed and a description of the completed station is given.

R614 Mögel, H. Betriebskontrolle von Kurzwellensendern. (Operating control of short-wave transmitters). *Elektrische-Nachrichten Technik*, 7, 334–48; September, 1930.

A description of the large short-wave station at Nauen, Germany, with special attention given to the methods of control used.

R800. Nonradio Subjects

551.5 Chapman, S. and Ferraro, V. C. A. A new theory of magnetic storms. *Nature* (London), **126**, 129–130; July 26, 1930.

An attempt to infer the course of events when a neutral ionized stream of particles from the sun is directed towards the earth has led to results which are believed to indicate how magnetic storms are produced.

621.314.3 Laub, H. Beitrag zur Theories des Resonanztransformators. (A contribution to the theory of resonance transformers). *Elektrische-Nachrichten Technik*, 7, 348–62; September, 1930.

A comprehensive mathematical and experimental investigation of the mutual effects of two circeuits, one (the primary) excited by an a-c source and the other (secondary) being closed by an output impedance; the circuits being coupled through a resonance transformer.

621.374.6 Turner, H. M. and McNamara, F. T. An electron-tube wattmeter and voltmeter and a phase shifting bridge. Proc. I.R.E., **18**, 1743–47; October, 1930.

This paper describes a wattmeter, for measuring power of a few microwatts or more, a phase shifting bridge for controlling the angular relation of two potentials without changing their amplitudes, and a voltmeter for measuring potentials of a few millivolts without amplifying.

621.385.96 Goldsmith, A. N. and Batsel, M. C. The RCA photophone system of sound recording and reproduction for sound motion pictures. Proc. I.R.E., **18**, 1661–89; October, 1930.

The general considerations governing the selection of a sound-on-film recording and reproducing system are analyzed, and the variable width track for recording and the dynamic cone with directional baffle for reproduction are described as best suited for studio and theater applications. The recorders, sound heads for projectors, amplifiers, switching and control devices, power supply, and loud-speaker systems of various types of RCA Photophone equipment, are described in detail.

CONTRIBUTORS TO THIS ISSUE

Austin, L. W.: See PROCEEDINGS for January, 1930.

Bailey, S. L.: Born October 7, 1905 at Minneapolis, Minnesota. Received B.S. degree in E. E., University of Minnesota, 1927; M.S. degree, 1928. Research in field intensity measuring apparatus. Assistant radio engineer, Airways Division, Department of Commerce, 1928; associate radio engineer, Airways Division, Department of Commerce, 1929. Consulting engineer, Jansky and Bailey, 1929 to date. Associate member, Institute of Radio Engineers, 1928.

Ballantine, S.: See PROCEEDINGS for July, 1930.

Cady, W. G.: See PROCEEDINGS for July, 1930.

Clapp, J. K.: See PROCEEDINGS for April, 1930.

Edes, N. H.: Born October 22, 1898 in Bermuda. Royal Military Academy, 1916–1917; commissioned, Royal Engineers, 1918; Royal Corps of Signals, 1921. Received B.A. degree in Mathematics and Physics, Clare College, Cambridge, 1926. Instructor, School of Signals, Catterick, 1929 to date. Associate member, I. E. E. Associate member, Institute of Radio Engineers, 1922.

Grover, F. W.: Born September 3, 1876 at Lynn, Massachusetts. Received B.S. degree, Massachusetts Institute of Technology, 1899; M.S. degree, Wesleyan University, 1901; Ph.D. degree, George Washington University, 1907; Ph.D. degree, University of Munich, 1908. Instructor, Wesleyan University, 1899–1901; Lafayette College, 1901–1902; Laboratory assistant, Bureau of Standards, 1902–1904; assistant physicist, 1904–1907, 1908–1911; professor of physics, Colby College, 1911–1920; assistant professor of electrical engineering, Union College, 1920–1922; associate professor since 1922. Engaged in educational work for U. S. Signal Corps, summers of 1917 and 1918. Consultant, Bureau of Standards, 1918 to date. Member, Institute of Radio Engineers, 1917.

Jackson, W. E.: Born May 8, 1904 at Bridgewater, Massachusetts. Received B.S. degree in E. E., Brown University, 1925. Westinghouse Electric and Manufacturing Company, 1923; New England Telephone and Telegraph Company, 1924; radio engineer, General Electric Company, 1925–1927; radio engineer, Department of Commerce, Airways Division, Bureau of Lighthouses, 1927 to date. Associate member, Institute of Radio Engineers, 1929.

Judson, E. B.: Born December 3, 1898 at Washington, D. C. In Naval radio service, 1917–1919. Laboratorian at U. S. Naval Research Laboratory, Bureau of Standards, 1919–1923. Radio engineer, Radio Transmission and Research Laboratories, Bureau of Standards, 1923 to date. Associate member, Institute of Radio Engineers, 1926.

Kenrick, G. W.: See PROCEEDINGS for September, 1930.

Mason, Hobart: Born 1879 at Yonkers, New York. Received B.S. degree in E. E., Polytechnic Institute of Brooklyn, 1900. Instructor, Polytechnic Institute of Brooklyn, 1901. Engineering department, New York and New Jersey Telephone Company, 1902–1909; traffic department, Western Union Telegraph Company, 1912–21; engineering department, 1921–1927; traffic engineer, Western Union Telegraph Company, 1927 to date. Member, A. I. E. E. Nonmember, Institute of Radio Engineers.

Morgan, N. R.: Born December 17, 1903 at Penryn, California. Received A. B. degree, Stanford University, 1928; E. E. degree, 1929. Engineer in charge of reports, Federal Telegragh Company, 1929 to date. Associate member, Institute of Radio Engineers, 1929.

Palermo, A. J.: Born May 16, 1899 at Schenectady, New York. Received B.S. degree in E. E., Union College, 1922; M.S. degree in E. E., 1924; instructor in electrical engineering, Union College, 1922 to date. Associate member, Institute of Radio Engineers, 1929.

Richardson, J. S.: Born 1901 at Nottingham, England. Received education, Nottingham University College. Technical staff, English Marconi Company, 1918; engineering staff, Northern Electric Company, 1928. Associate member, Institute of Radio Engineers, 1930.

Terman, F. E.: See PROCEEDINGS for January, 1930.

Wymore-Shiel, I. J.: Born in Mahaska County, Iowa. Received B.S. degree, Drake University, 1918; M.S. degree, George Washington University, 1925. With Division of Metallurgy, Bureau of Standards, 1919–1924; Laboratory for Special Radio Transmission Research, 1924 to date. Nonmember, Institute of Radio Engineers.

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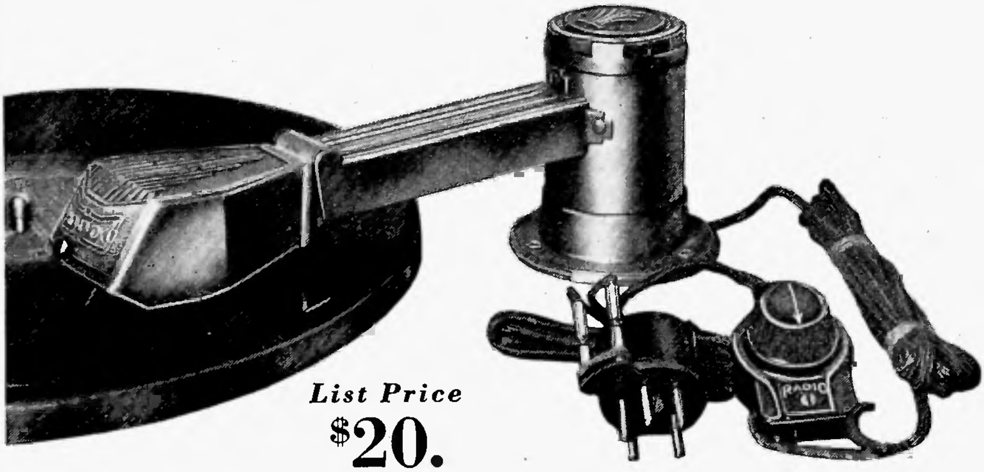

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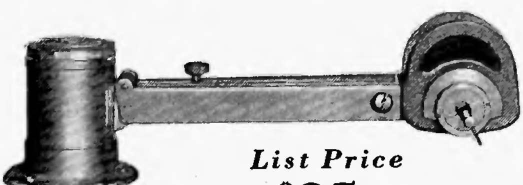

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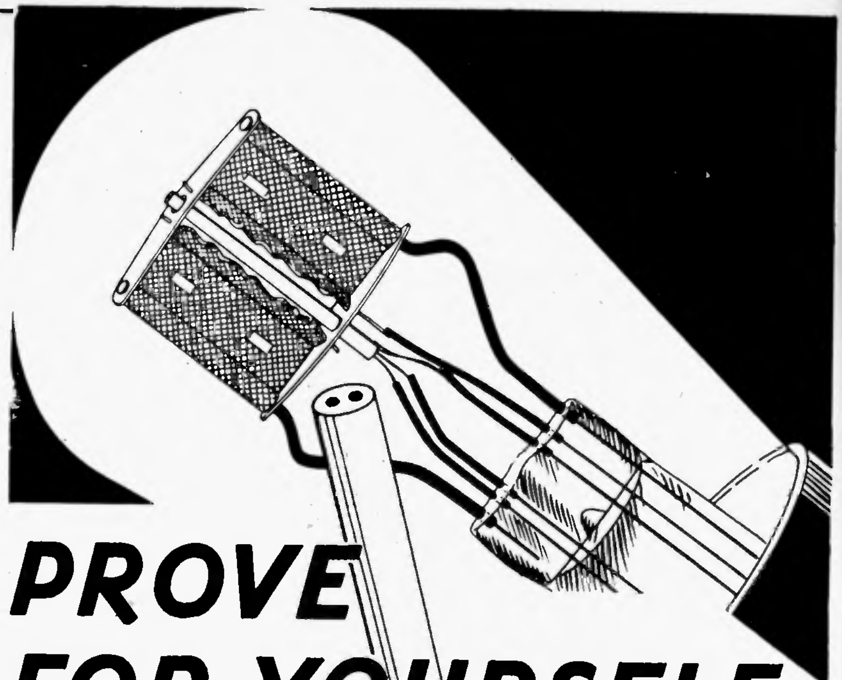

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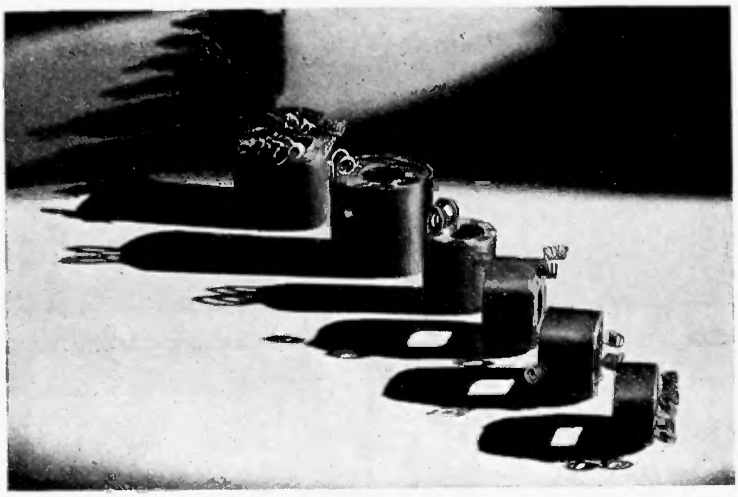

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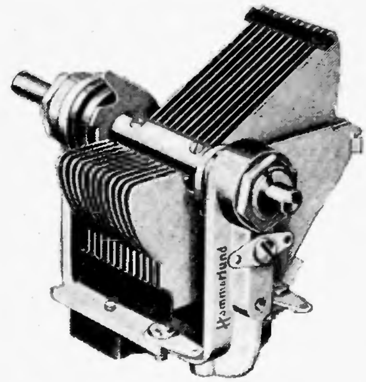

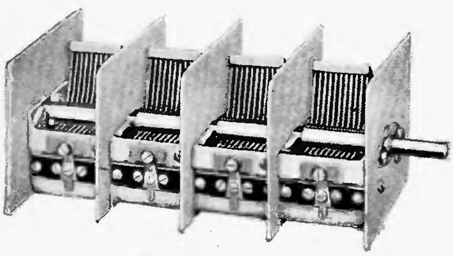

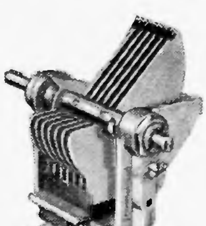

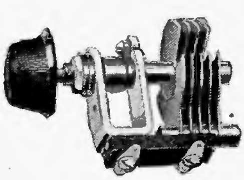

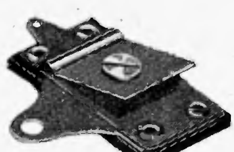

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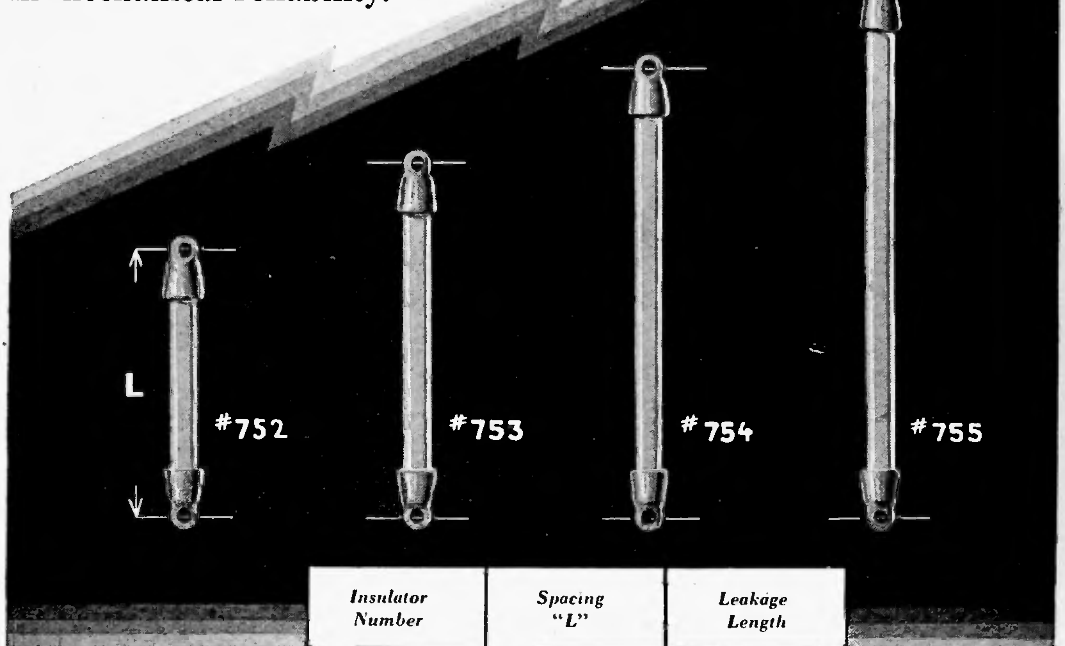

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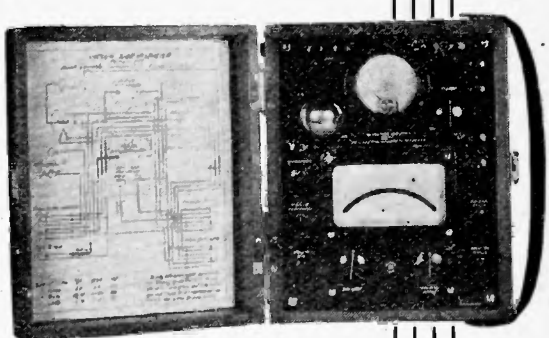

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To the Board of Direction
Gentlemen:

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)

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(Date) (City and State)

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(Signature of references not required here)

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Mr. ....Mr. ....

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Mr. ....

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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...

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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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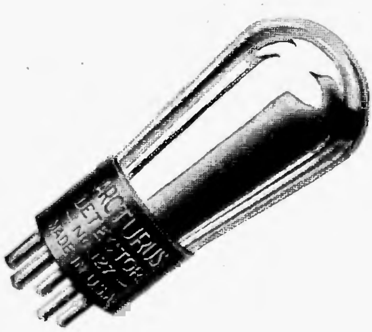

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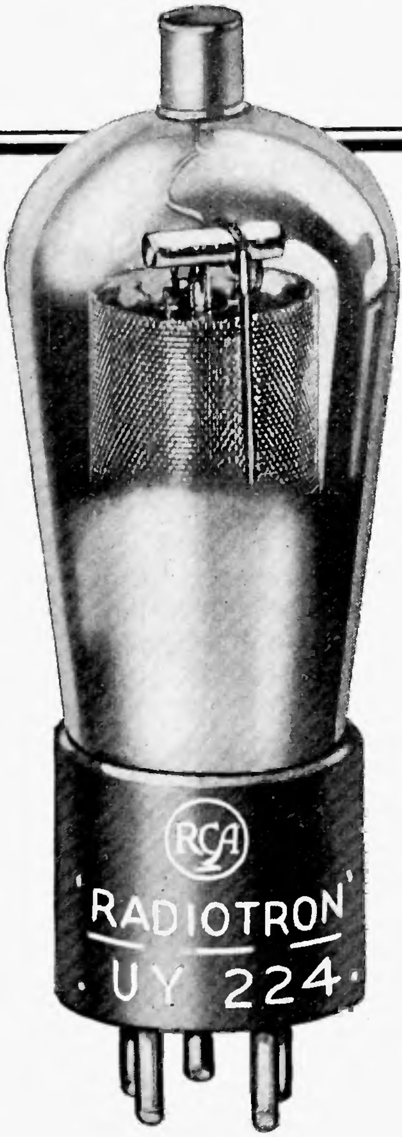

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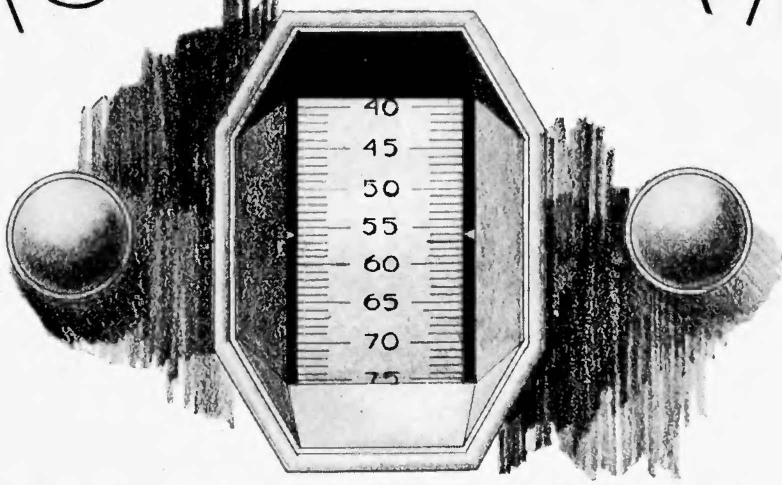

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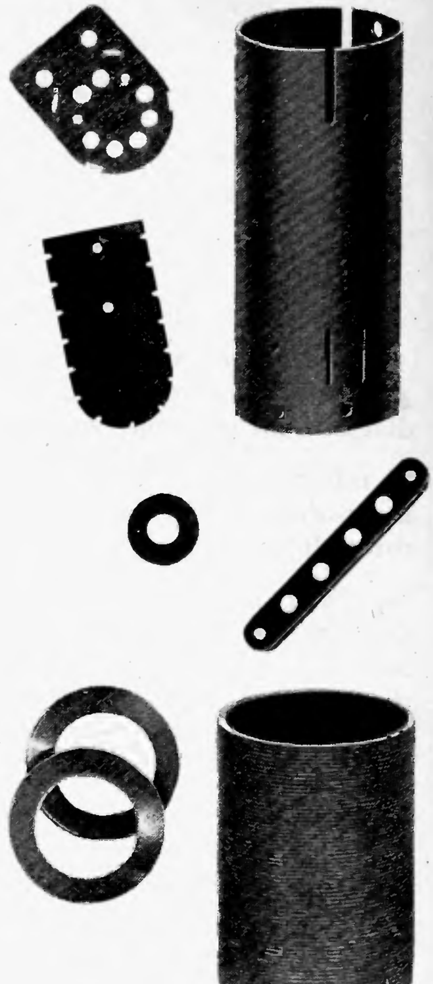

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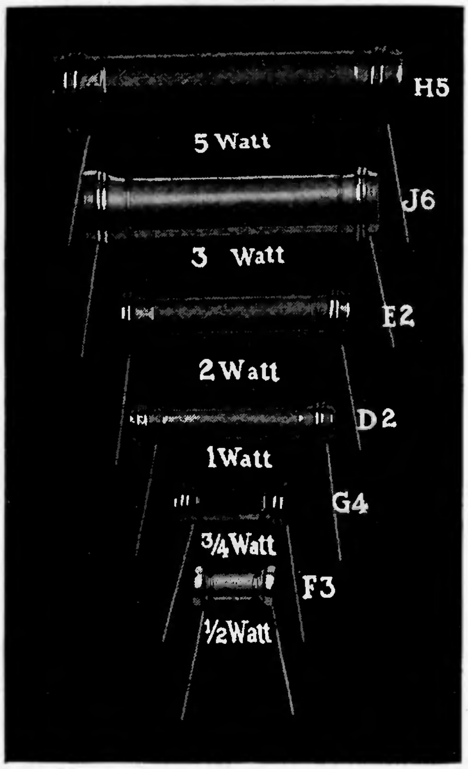

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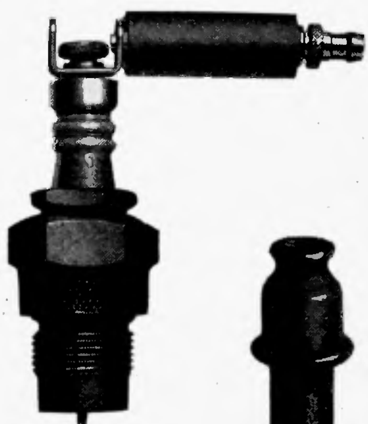

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Alford, Andrew	708 N. Kenmore Ave., Los Angeles, Calif.
Anderson, Donald T.	United Reproducers Corp., 410 S. Harwood, Dallas, Tex.
Annett, Earl	3041 Gladstone Ave., Apt. 209, Detroit, Mich.
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Banditson, Harry	Modern Radio & Equipment Co., 191 Franklin St., Buffalo, N.Y.
Batchelder, S. P.	25 Windemere St., East Springfield, Mass.
Belock, Harry D.	678 Pennsylvania Ave., Brooklyn, N.Y.
Biar, L. E.	512 W. South St., Angola, Ind.
Blackwell, George C.	General Motors Bldg., Rm. 724, 1775 Broadway, New York City.
Bonell, Ralph K.	463 West St., New York City.
Brattain, Walter H.	Bureau of Standards, Washington, D. C.
Brightman, R. B.	U. S. Navy, Eyak, Mile 7, Alaska.
Burchard, John C.	Radio Corporation of America, 100 W. Monroe St., Chicago, Ill.
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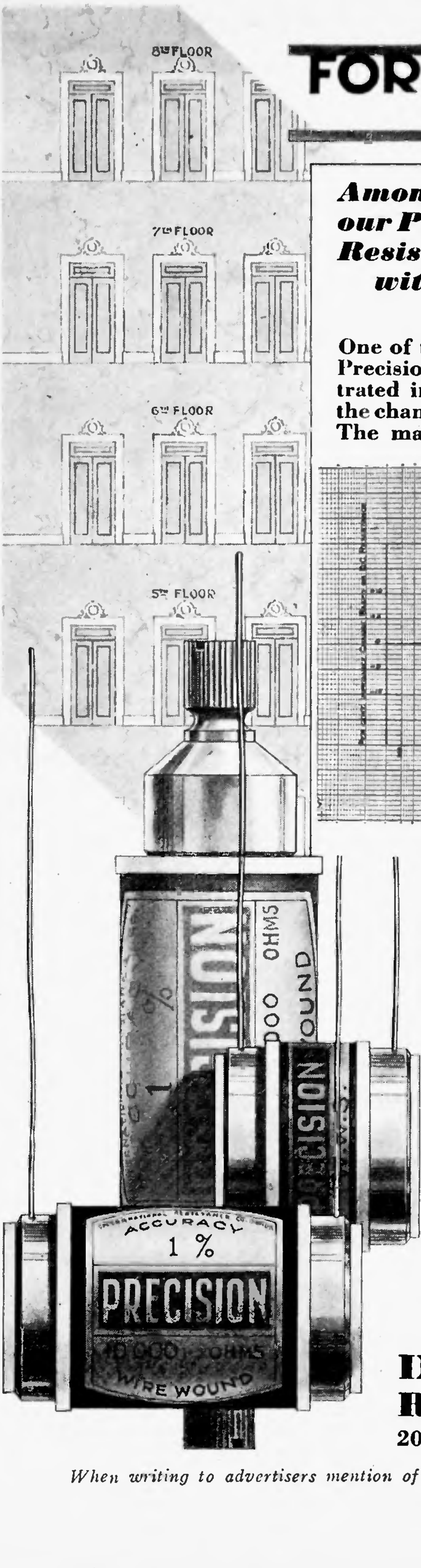

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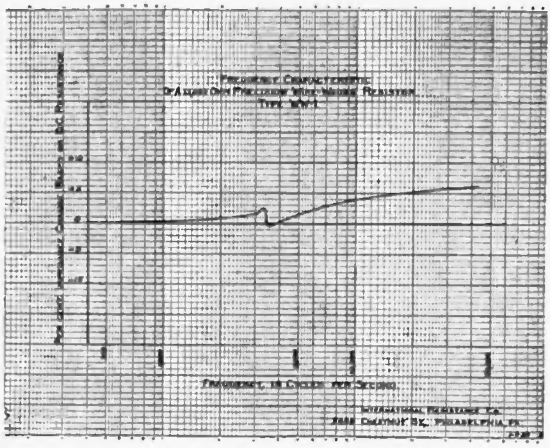

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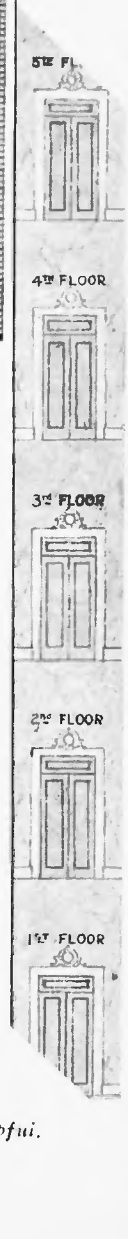

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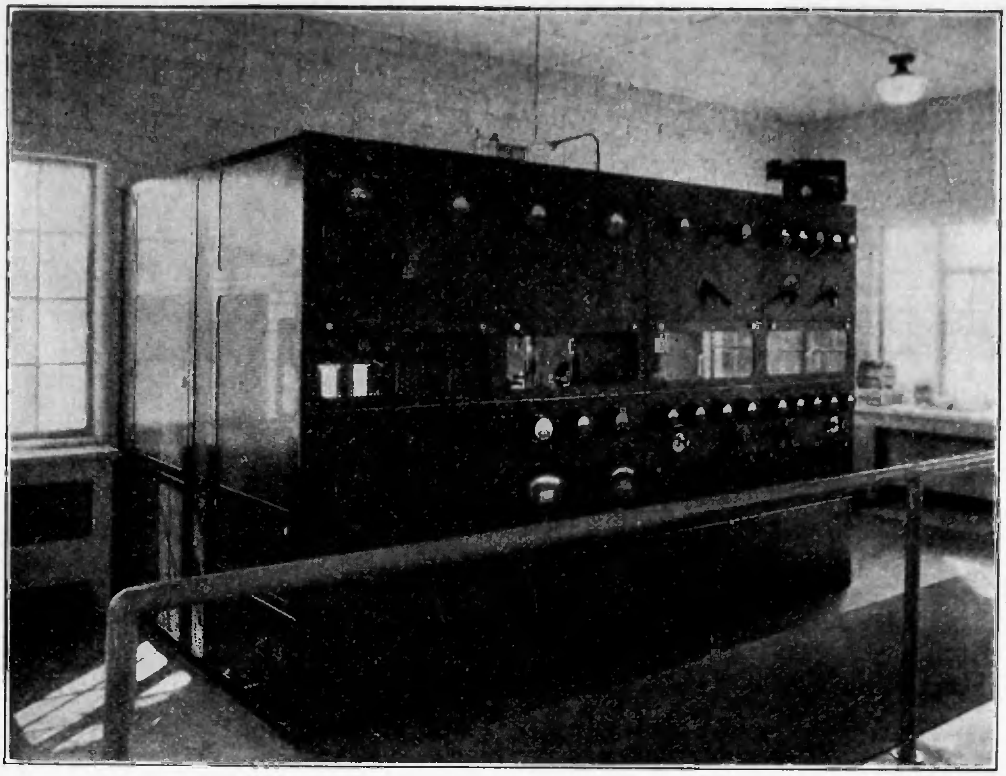

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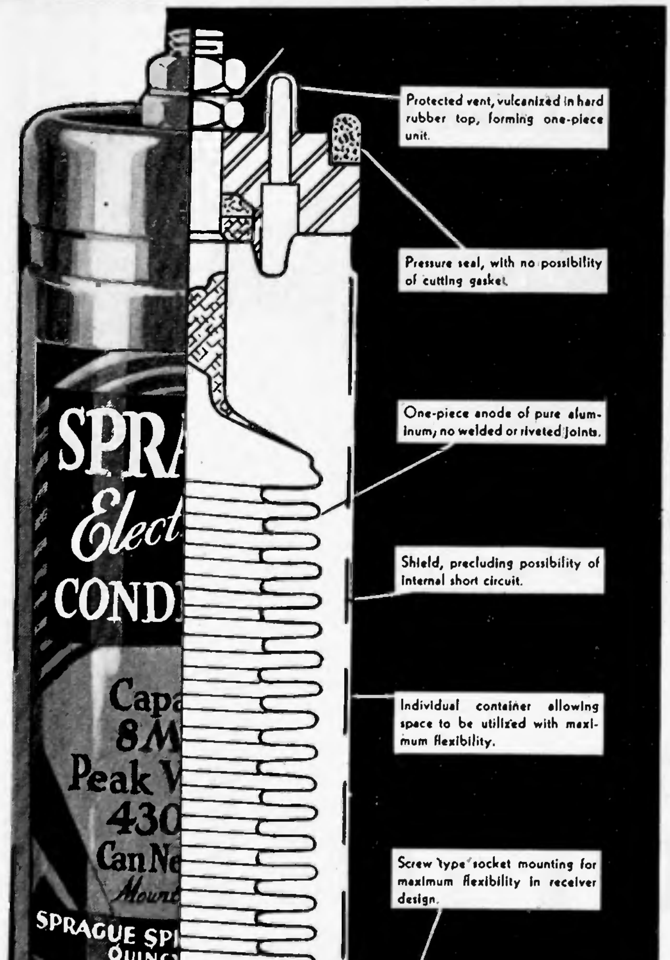

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