VOLUME 23 APRIL, 1935 NUMBER 4

PROCEEDINGS
of
The Institute of Radio Engineers

Tenth Annual Convention
Detroit, Michigan
July 1, 2, and 3, 1935

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Institute of Radio Engineers Forthcoming Meetings

TENTH ANNUAL CONVENTION
DETROIT, MICHIGAN
July 1, 2, and 3, 1935

JOINT MEETING
American Section, International Scientific Radio Union and Institute of Radio Engineers, Washington, D. C.
April 26, 1935

CINCINNATI SECTION
April 16, 1935

DETROIT SECTION
April 19, 1935

LOS ANGELES SECTION
April 16, 1935

NEW YORK MEETING
April 3, 1935
May 1, 1935

PHILADELPHIA SECTION
April 4, 1935
May 2, 1935

PITTSBURGH SECTION
April 16, 1935

TORONTO SECTION
April 17, 1935

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 23 April, 1935 Number 4

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PART II

Technical Papers

The Institute of Radio Engineers

GENERAL INFORMATION

INSTITUTE. The Institute of Radio Engineers was formed in 1912 through the amalgamation of the Society of Wireless Telegraph Engineers and the Wireless Institute. Its headquarters were established in New York City and the membership has grown from less than fifty members at the start to several thousand.

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

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

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Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Admissions Committee. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before April 30, 1935. These applications will be considered by the Board of Directors at its meeting on May 1, 1935.

For Election to the Associate Grade

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INSTITUTE NEWS AND RADIO NOTES

March Meeting of the Board of Directors

The regular monthly meeting of the Board of Directors was held in the Institute office on March 6 and those present were: Stuart Ballantine, president; Melville Eastham, treasurer; Arthur Batcheller, O. H. Caldwell, Alfred N. Goldsmith, Virgil M. Graham, R. A. Heising, George Lewis, E. L. Nelson, H. M. Turner, A. F. Van Dyck, L. E. Whittemore, William Wilson, and H. P. Westman, secretary.

Forty-three applications for Associate membership, three for Junior, and eight for Student grade were accepted.

Representation on a newly organized American Standards Association Sectional Committee on Electric and Magnetic Magnitudes and Units was accepted, and J. H. Dellinger, chief of the Radio Section of the National Bureau of Standards was appointed.

Because a large proportion of the PROCEEDINGS editorial space has recently been devoted to papers of a theoretical nature, a policy of encouraging increased publication of practical papers, covering the design and general description of equipment and manufacturing methods was adopted.

A committee comprised of Virgil M. Graham as chairman, L. F. Curtis and E. T. Dickey was appointed to review the "Standards for Power-Operated Radio Receiving Appliances" which have been submitted to the American Standards Association for approval by the Underwriters Laboratories.

Ten new registrations were received by the Emergency Employment Service during February bringing the total to 711 of whom 528 are members. Four jobs were handled and two filled, one of which is considered permanent and one temporary.

Because of a change in position, L. H. Larime resigned as chairman of the Tenth Annual Convention Committee and H. L. Byerlay was named to replace him.

An invitation from the American Institute of Physics for the Institute to be represented on a Committee on Applied Physics was accepted and William Wilson asked to serve.

Joint I.R.E.–U.R.S.I. Meeting

As previously announced, a joint meeting of the American Section of the International Scientific Radio Union and the Institute will be held on April 26, 1935. There will be two sessions at the National

Academy of Sciences Building, 2101 Constitution Avenue, Washington, D.C., beginning at 10 A.M. and 2 P.M. Papers will be limited to fifteen minutes each to allow time for discussion. The following papers are listed at the time of going to press:

"The London General Assemby of the International Scientific Radio Union," by J. H. Dellinger, National Bureau of Standards.

"Further Results of a Study of Ultra-Short-Wave Transmission Phenomena," by C. R. Englund, A. B. Crawford, and W. W. Mumford, Bell Telephone Laboratories.

"Experiments with Ultra-High-Frequency Transmitting Antenna in Close Proximity to the Ground," by H. Diamond and F. W. Dunmore, National Bureau of Standards.

"Ionosphere Measurements during the Partial Eclipse of the Sun of February 3, 1935," by J. P. Schafer and W. M. Goodall, Bell Telephone Laboratories.

"The Graphical Analysis of a 10,000-Hour Kennelly-Heaviside Layer Record," by Harry R. Mimno, Harvard University.

"Recent Ionosphere Measurements in the Southern Hemisphere," by L. V. Berkner, H. W. Wells, and S. L. Seaton, Carnegie Institution of Washington.

"Some Continued Observations of Ultra-High-Frequency Signals over Long Indirect Paths," by Ross A. Hull, American Radio Relay League.

"Terrestrial Magnetism and Its Relation to World-Wide Short-Wave Communications," by Henry E. Hallborg, RCA Communications, Inc.

"Radio Propagation over Spherical Earth," by C. R. Burrows, Bell Telephone Laboratories.

"Direction Finding of Atmospherics," by John T. Henderson, National Research Council of Canada.

"Theoretical Explanation of Published Measurements of Vertical Plane Radiation Characteristics of High Vertical Radiators," by K. A. MacKinnon, Canadian Radio Broadcasting Commission.

"Some Developments in Low Loss Inductances," by F. E. Terman, Stanford University.

"Measurement of High-Frequency Impedance with Networks Simulating Lines," by W. L. Barrow, Massachusetts Institute of Technology.

"The Accuracy of Low Voltage Cathode Ray Tube for Oscillographic Radio Measurements," by L. E. Swedlund, Westinghouse Electric and Manufacturing Company.

"The Detection of Frequency Modulated Waves," by J. G. Chaffee, Bell Telephone Laboratories.

"A Novel Modulation Meter," by H. N. Kozanowski, Westinghouse Electric and Manufacturing Company.

"On the Nature of Transmitter Key Clicks and Their Suppression," by A. Hoyt Taylor and L. C. Young, U. S. Naval Research Laboratory.

"Grid Dissipation as a Limiting Factor in Vacuum Tube Operation," by I. E. Mouromtseff and H. N. Kozanowski, Westinghouse Electric and Manufacturing Company.

"Application of Secondary Emission," by V. K. Zworykin, RCA Victor Division, RCA Manufacturing Company.

Supplementary Program

(Papers to be Presented if Time Permits)

"A Graphical Aid in the Design of Networks for Distortion Correction," by E. A. Guillemin, Massachusetts Institute of Technology.

"The Directive Antenna of KYW Station," by R. N. Harmon, Westinghouse Electric and Manufacturing Company.

"Industrial High-Frequency Generators Using Vacuum Tubes," by H V. Noble, Westinghouse Electric and Manufacturing Company.

Massachusetts Institute of Technology Graduate Studies

Three new subjects for graduate students have been offered by Massachusetts Institute of Technology during the current year and the fourth is announced for next year. These are:

Vibration phenomena and oscillations
Engineering electronics
Mathematical analysis by mechanical methods
Super high voltage engineering and vacuum electrostatic machinery

If sufficient demand exists, these subjects may be offered in the 1935 summer courses. Application should be made before May 1 to C. E. Tucker, Associate Professor, Department of Electrical Engineering, Massachusetts Institute of Technology, Cambridge, Mass.

Committee Work

Membership Committee

A Membership Committee meeting was held in the Institute office on March 6 and was attended by I. S. Coggeshall, chairman; W. F. Cotter, W. G. Ellis, H. C. Humphrey, E. W. Shafer, and C. E. Scholz.

Objectives for the current year were discussed. A mailing piece setting forth information on membership matters was revised and approved. Plans were made for digesting information to be obtained from questionnaires circulated at the New York meeting on the same evening.

STANDARIZATION

Technical Committee on Electro-Acoustic Devices—IRE

The Technical Committee on Electro-Acoustic Devices operating under the Institute's Standards Committee met on March 1 in the Institute office and those present were H. F. Olson, chairman; Sidney

Bloomenthal, L. G. Bostwick, W. B. Goggins, Knox McIlwain, Hans Roder, V. E. Whitman, Harold Zahl, and H. P. Westman, secretary.

A general discussion was held of several portions of the existing report being considered for revision. Various portions of the work were divided among the members of the committee and preliminary reports are to be prepared for consideration at the next meeting.

Technical Committee on Electronics—IRE

The Technical Committee on Electronics operating under the Institute's Standards Committee met on March 8 in the Institute office and those present were B. J. Thompson (acting chairman representing B. E. Shackelford), E. A. Lederer, G. F. Metcalf, O. W. Pike, P. T. Weeks, and H. P. Westman, secretary.

Progress reports submitted by the various subcommittees which are operating under this technical committee were reviewed. Various problems which the subcommittees have encountered were discussed and instructions issued where necessary.

Subcommittee on Electron Beam and Miscellaneous Tubes

The Subcommittee on Electron Beam and Miscellaneous Tubes operating under the Technical Committee on Electronics of the Institute met on March 7 at the Institute office. Those present were G. F. Metcalf, chairman; A. B. DuMont, M. S. Glass, B. J. Thompson, and H. P. Westman, secretary. The committee devoted its time chiefly to a consideration of terminology on cathode ray tubes and similar devices and prepared a number of proposals.

Subcommittee on Gas-Filled Tubes

This Subcommittee of the Electronics Committee of the Institute met on March 7 at the Institute office and those present were: O. W. Pike, chairman; D. V. Edwards, H. E. Mendenhall, Dayton Ulrey, P. T. Weeks, and H. P. Westman, secretary.

It reconsidered certain material on graphical symbols and definitions which was prepared at its previous meeting. It extended its work in these two divisions of the field and completed a report on them which will be submitted to the Technical Committee on Electronics at its next meeting.

Subcommittee on Small High Vacuum Tubes

The Subcommittee on Small High Vacuum tubes operating under the Institute Technical Committee on Electronics met in the Institute office on March 7. P. T. Weeks, chairman; M. Cawein, George Lewis,

H. A. Pidgeon, E. W. Shafer, and H. P. Westman, secretary, were present.

A number of proposals prepared at the previous meeting of the committee were reconsidered. A report of the Electronics Subcommittee of the Sectional Committee on Electrical Definitions of the American Standards Association was compared with existing Institute standards and a number of modifications made to make these two sets of standards identical.

TECHNICAL COMMITTEE ON TRANSMITTERS AND ANTENNAS—ASA

The Technical Committee on Transmitters and Antennas operating under the Sectional Committee on Radio of the American Standards Association met at the Institute office on February 13. Those present were: Haraden Pratt, chairman; H. A. Chinn, A. A. Oswald, E. G. Ports, G. H. Shannon (representing J. L. Finch), and H. P. Westman, secretary.

The committee devoted its time to consideration of the material on safety standards included in the 1933 Standards Report of the Institute.

Institute Meetings

BOSTON SECTION

The Boston Section met on December 21 at Harvard University and the meeting was presided over by H. R. Mimno in the absence of Chairman Chaffee. Fifty members and guests were present and twelve attended the informal dinner which preceded the meeting.

H. W. Fletcher of the New England Telephone and Telegraph Company presented "A Description of the Boston-Provincetown Radio and Wire Circuit." He described the design and operation of a 65-megacycle radio link through which regular telephone communication is carried on between Provincetown and Green Harbor which is connected by wire to Boston. It is used regularly as a direct Boston-Provincetown toll circuit. At each terminal of the radio link the receivers are operated constantly and the insertion of a telephone plug into the transmitter jack which is located with jacks for other toll circuits, starts the transmitter in operation. The transmitter is placed in operating condition so rapidly that no time delay is required before the ringing signal is transmitted. This is in the form of a 1000-cycle tone interrupted at twenty cycles. Equipment similar to that used on the transatlantic telephone circuits are supplied to insure privacy of conversations. The paper was discussed by the chairman and Messrs. Bowles, Dallin, and O'Neill.

Buffalo-Niagara Section

The February 13 meeting of the Buffalo-Niagara Section was held in the ballroom of the Hotel Lafayette. One hundred twenty-five members and guests attended the meeting and fifty-four were present at the informal dinner at which B. T. Simpson of the Gratwick Laboratories spoke. The meeting was presided over by L. E. Hayslett, chairman.

"Some Applications of Ultra-High-Frequency Communication" was the subject of a paper by L. C. F. Horle, consulting engineer. He outlined the history of the development of ultra-high-frequency communication. He then discussed the physical and electrical limitations of design of equipment for ultra-high-frequency work due to the peculiarities of those waves as compared with lower radio frequencies. Recent developments employing very small tubes and special materials having very small coefficients of thermal expansion were described as were the problems of antenna design and the coupling of the equipment to the antenna. The design and advantages of directional antennas were stressed.

Various feasible applications including their use for police, amateur, local broadcasting, portable, and mobile services were outlined. A police system employing a main 500-watt transmitter installed at Newark, N. J., for two-way communication with police cars was discussed. The 65-megacycle telephone link between Boston and Provincetown was described.

A considerable amount of ultra-high-frequency equipment was displayed and a portable set was demonstrated with the operator carrying it to various parts of the ballroom. A general discussion followed the presentation of the paper.

Chicago Section

The November 9 meeting of the Chicago Section was held at the Medinah Athletic Club and was presided over by H. S. Knowles, chairman. The attendance was 102 and twenty were present at the dinner which preceded the meeting.

A paper on "Small Set Design Considerations" and another on "Recent Advances in High Fidelity Receivers" were presented by D. E. Harnett of Hazeltine Laboratories. The first paper dealt with a typical small receiver design prepared by Hazeltine Laboratories. Concrete examples of the technical and economic problems involved in this design were described and general conclusions drawn from them.

In the second paper, the general problems and standards for high fidelity receivers were described and included methods of obtaining a wide frequency range of reproduction with satisfactory selectivity. The papers were discussed by Messrs. Church, Crosley, Knowles, and Wilcox.

The December 21 meeting of the section also held at the Medinah Athletic Club was in charge of Alfred Crosley, vice chairman. The attendance at the meeting was eighty-five and at the dinner, eighteen.

"High Fidelity Radio Transmitters" was the subject of a paper by H. C. Vance, district manager of the Engineering Products Division of the Radio Corporation of America. In it he outlined the complete design of the latest RCA high fidelity broadcast transmitter with particular reference to the possibilities which such transmissions offer the set designer. The improvement in power supply, simplicity, and operating cost was pointed out.

The annual meeting of the Chicago Section was held at the Medinah Athletic Club on February 15 with H. S. Knowles, retiring chairman presiding. In the election for officers Alfred Crosley, consultant, was made chairman; Harold Vance, of the RCA Victor Company, vice chairman; and J. Kelly Johnson of Wells-Gardner and Company, secretary-treasurer.

The newly elected chairman then presided and introduced Mr. Dameron of P. R. Mallory and Company who spoke on the subject of "Dry Electrolytic Condensers." The history of the development of electrolytic condensers was outlined and emphasis placed on troubles encountered in securing materials of adequate chemical purity. The more important construction methods were described, special attention being given to the three more commonly used processes; hand pasting, dipping, and continuous processing. The advantages and disadvantages of gauze, cellulose, and cellophane separators were outlined and the formation, temperature, corrosion, capacity and life characteristics of the various types discussed. The use of these condensers in industries outside of radio was described. Several of the 135 members and guests who attended the meeting participated in the discussion of the paper and twenty five were present at the dinner which preceded the meeting.

H. C. Brown was appointed chairman of the Meetings and Papers Committee, L. S. Starrett to Membership, and E. Kohler to the Publicity Committee.

Cincinnati Section

The Cincinnati Section met on February 12 at the University of Cincinnati. A. F. Knoblaugh, chairman, presided and eighteen members and guests were present.

"Thermionic Emission in Vacuum Devices" was the subject of a paper by G. W. Bain, chief engineer of the Kenrad Corporation. Colonel Bain delivered the paper in the absence of R. H. Matlock who was originally scheduled for it. In it he outlined the development of thoriated tungsten filaments, discussing their characteristics and describing rejuvenating processes. He then treated oxide-coated types, speaking of methods for aging and explained some of the peculiarities of this type of emitter.

Cleveland Section

A meeting of the Cleveland Section was held on January 24 at Case School of Applied Science. Karl Banfer, chairman, presided and the attendance was sixty two, eleven of whom were present at the dinner which preceded it.

A paper on "Ship-to-Shore Two-Way Telephone Communication" was presented by R. A. Fox, engineer in charge, Lorain Telephone Company. He reviewed the problems of transmitter, receiver, shore antenna, and connections with land telephone systems. An unusual and practical method of automatic volume control for audio-frequency circuits to prevent overmodulation was outlined. Fading of signals was attacked by using horizontal and vertical antennas on separate receivers. Tests were made last summer with combination transmitters and receivers on freight boats on the Great Lakes. Under favorable conditions they were able to maintain two-way telephone communication with land via a relay station near Lorain. Dialing and calling systems are now being developed. A demonstration of equipment used in this work was presented and two-way communication established with the relay station.

Connecticut Valley Section

J. A. Hutcheson, chairman, presided at the January 24 meeting of the Connecticut Valley Section held at the Hotel Garde in Hartford. Seventeen were present.

A paper on "The Control of Radiated Field Patterns" was presented by J. L. Reinartz of RCA Radiotron. He first reviewed his early experiments in the field of high-frequency radiation and his observations while on an arctic exploration trip in 1925. From these data, he developed a theoretical antenna which would permit control of the radiated field pattern in so far as the angle of radiation was concerned and, to a lesser extent, of the polarization. Experimental work in the ultra-high-frequency region and at frequencies as low as seven megacycles confirms the theory. The antenna consists of a quarter-wave horizontal section comprising half a current-fed Hertz. The other quarter-wave section is vertical downward and is connected through a tuned circuit to ground. Adjustment of the tuned circuit provides control of the angle of radiation.

The Hotel Charles in Springfield was the place at which the Connecticut Valley Section met on February 21. Thirty members and guests were present and J. A. Hutcheson, chairman, presided.

"New Developments in Electrolytic Condensers" was the subject of a paper by Dr. Robinson of the Sprague Condenser Corporation. In it he treated the construction and performance of both wet and dry electrolytic condensers. He described the production of the oxide film on the electrode and the way in which borate ions penetrate the film resulting in its being increased to permit operation at higher potentials. The operation of these condensers on alternating current was described and it was pointed out that their capacitance and power factor changes with frequency. How various mechanical features of design can be used to develop condensers suitable for operation at the lower frequencies was outlined. A substantial number of those present participated in the discussion which occupied about as much time as the presentation of the paper.

Detroit Section

A meeting of the Detroit Section was held on February 22 in the *Detroit News* Conference Room. A. B. Buchanan, chairman, presided and forty-five members and guests attended. Twelve were present at the dinner which preceded the meeting.

A paper, by E. E. Jaenicki of the Traffic Department of the Associated Press, was presented on the subject of "Wirephoto." The meeting was transferred to the wirephoto room of the *Detroit News* where the speaker outlined the general operation of the system and then described the technical operation of each part of it. Since the equipment was in operation, the members could study its actual performance in a newspaper office. About thirty minutes are required from the time of starting to the finished eight-by-ten photograph, and covers transmission, developing, and printing. The speed of the scanning motors is governed by a temperature controlled 300-cycle tuning fork. They are of the induction type for 60-cycle operation but have an extra winding energized by power from the fork amplifier circuits. The forks are adjusted weekly by means of a standard tone sent from the telephone laboratory.

New York Meeting

The March New York meeting of the Institute was held on the 6th in the Engineering Societies Building and was presided over by President Ballantine.

A paper on "Recent Developments in Miniature Tubes" by Bernard Salzberg and D. G. Burnside of the RCA Radiotron Division of RCA Manufacturing Company was presented by Mr. Salzberg.

The development of two indirectly heated miniature tubes, a triode, and a sharp cut-off amplifier pentode especially suited for use at high frequencies, was described. The electrical and mechanical factors involved in the design and application of these tubes were discussed, and their novel structural appearance described.

Various types of developmental high-frequency receivers and auxiliary equipment which have been built up around such tubes were exhibited, and a demonstration was given of the relative amplification which can be obtained at high frequencies by conventional tubes and by the miniature tubes.

A number of the 350 members and guests in attendance participated in the discussion of these papers.

Philadelphia Section

A meeting of the Philadelphia Section was held at the Engineers Club on February 7 and was attended by 300. Fifteen attended the dinner which preceded it. The meeting was presided over by E. D. Cook, chairman.

A paper on "The Multifactor Tube and Some of Its Applications" was presented by P. T. Farnsworth, vice president in charge of research of Television Laboratories, Ltd. This tube is a cold cathode vacuum tube employing the phenomena of secondary emission. Under favorable conditions conductors which are bombarded with electrons may give off as many as ten secondary electrons for each primary electron impinging on it. By suitable electrode construction energy may be imparted to secondary electrons by applying a high-frequency field of a frequency at which the electrons within the tube are resonant. The cathodes may be two plain parallel disks with a ring-shaped anode midway between them. A magnetic focusing field guides the electrons between the cathodes and prevents their hitting the anode. Specially shaped cathodes and anodes may be made self-focusing. These tubes can be made to multiply small currents such as of photo-electric nature to as high as 10^{15} of the starting current. They are used as amplifiers and oscillators having several different modes of oscillation.

Pittsburgh Section

The February 21 meeting of the Pittsburgh Section, which was sponsored jointly with the Pittsburgh Physical Society, was held at

Carnegie Institute of Technology. R. D. Wyckoff, chairman of the Physical Society and vice chairman of the Pittsburgh Section, presided.

A paper on "Recent Progress in Television" was presented by P. T. Farnsworth, research engineer for Television Laboratories, Ltd. The system of television described differed from many others in that scanning was accomplished by shifting the image of the object about a fixed spot instead of shifting the scanning speed over the picture. A highly efficient electron gun used for receiving supplies an emission of several milliamperes.

The system was said to be commercially practical with a receiver selling at a price of $250. Several European countries, especially Germany and England, are now preparing for commercial television services.

The electron multiplier developed by the author was demonstrated as a radio-frequency generator and uses two cold cathodes on opposite sides of a centrally located spiral anode. A high-frequency voltage is applied to the cathodes in opposing phase in addition to the anode voltage to accelerate electrons leaving one cathode for the opposite cathode. Their arrival velocity and phase relation result in secondary electrons being emitted and drawn back through the first cathode to dislodge additional electrons. With a realizable electron gain of six each trip across the elements, very high gains would be possible and the tubes are claimed to have an amplification of 10^7 although in practical use it is limited to several hundred. Many members and guests of the 300 present participated in the discussion.

Rochester Section

A meeting of the Rochester Section was held on January 10 at the Sagamore Hotel with R. H. Manson presiding. The attendance totaled fifty and there were eight at the dinner held before the meeting.

"Recent Developments in Metal-Clad Mercury-Arc Rectifiers for Radio and Power Work" was the subject of a paper by W. E. Gutzwieler of the Allis-Chalmers Company. The theory of operation of mercury rectifiers utilizing a third element or grid for output control was first outlined. The steel-clad grid-controlled rectifier was then compared with the glass envelope types and its two chief advantages for broadcast and other radio transmitters outlined. First, no high speed circuit breakers are required as the grid control rectifier becomes the circuit breaker, automatically disconnecting the plate supply when a disturbance occurs and reëstablishing it upon the termination of the surge. All this takes place within a fraction of a second without interrupting the program. Second, large powers may be handled in a small space with extremely high efficiency. The performance of these rectifiers was illustrated by a series of oscillograms. Additional applications of them for such purposes as direct-current power transmission, the control of large size variable speed motors, and frequency changes were indicated. The discussion which followed was participated in by a number of those present.

The February meeting of the section was held on the 14th at the Sagamore Hotel. H. J. Klumb, chairman, presided and seventy seven members and guests attended. Ten were present at the dinner before the meeting.

A paper on "Ultra-High-Frequency Radio Transceivers and their Applications" was presented by L. C. F. Horle, consultant. The problems confronted in the use of frequencies of thirty megacycles or higher for practical two-way radio communication were outlined. Treatment of the subject was divided into two classes, the division being based upon whether the equipment was portable or nonportable. Existing installations of both types were described and a number of transceivers were available for inspection together with a superheterodyne receiver for reception of ultra-high-frequency broadcasts.

The use of these waves for broadcasting was considered and it was pointed out that transmission to the receiver is almost entirely by ground wave, the sky wave being absorbed or completely transmitted through the ionosphere. This limits the range of transmission and would permit a duplication of channels by stations spaced only 100 or 150 miles apart. Methods of reflecting and transmitting these waves in narrow beams thereby extending the distance over which very small amounts of radio energy could be received dependably were explained. It was pointed out that static was almost entirely missing in this portion of the radio spectrum.

San Francisco Section

A meeting of the San Francisco Section was held at the Bellevue Hotel on February 20 with A. H. Brolly, chairman, presiding and twenty seven present. Eight attended the dinner which preceded the meeting.

"Radio Applications in the National Park Service" was the subject of a paper by J. F. Maxwell who is a radio engineer in the National Park Service. He discussed the uses of radio in our national parks describing its great advantage in fighting forest fires and in furnishing communication in the winter after telephone lines have been broken by winter storms. Various types of radiophone equipment used were described and slides showed the widely varying conditions under which the equipment must serve.

Toronto Section

The November meeting of the Toronto Section was held on the 29th at the University of Toronto. W. F. Choat, past chairman, presided and the attendance was thirty five.

S. F. Fisher of the Special Applications Section of the Northern Electric Company presented a paper on the "Synchronization of Broadcast Stations." He opened his discussion with a review of experiments conducted over the past ten years in the synchronizing of broadcast stations. Methods employed were outlined and reasons given for the distortion at the receiver which is due to slight frequency differences between the two stations to be synchronized. Methods of eliminating several of the distorting factors were described. The talk was followed by a practical demonstration using two miniature broadcast transmitters. The frequency of one was kept constant while that of the other was varied from that frequency by from a quarter to 50 cycles. The various distortion effects were heard by the audience from a receiver connected to the two transmitters and their effects on both speech and music were demonstrated. The paper was discussed by Messrs. Bayley, Hackbusch, Hepburn, Leslie, and Pipe.

The January meeting of the section was held on the 9th at the University of Toronto with F. J. Fox, vice chairman, presiding.

Austin Armer, chief engineer of the Magnevox Company, presented a paper on "Commercial Developments of Loud Speakers." He covered the development of loud speakers from their original application before the time of radio broadcasting and showed that the first of these obviously was the result of a search for a better telephone receiver. The advances made from the first horn type speaker to the present cone type dynamic speaker were reviewed and the work of Lodge, Pridham and Jensen, and Kellogg and Rice noted.

The principles of loud speaker design were explained. Sound pressure curves were shown and the reasons for the various resonance points given together with methods for their correction. It was shown that loud speaker efficiency and acoustic output depended on flux density, length of leakage path, and actuating current. Methods of improving speaker efficiency consistent with cost were given in detail and it was shown how improvements in design and production have made possible a reduction in cost to one tenth of that of ten years ago.

The reasons for the improved characteristics of new speakers using curvilinear cones rather than those with straight sides were given. The paper was followed by a motion picture showing the development of the production of loud speakers for the Magnavox Company. Messrs. Bayley, Fox, and Hepburn of the forty eight members and guests in attendance participated in the discussion. Seventeen were at the dinner which preceded the meeting.

Washington Section

A meeting of the Washington Section was held at the Potomac Electric Power Company Auditorium on February 11. E. K. Jett, chairman, presided and seventy five were present at the meeting. Eighteen attended the dinner which preceded it.

G. W. Fyler of the Radio Transmitter Engineering Department, General Electric Company, presented a paper on "Parasitics and Instability in Radio Transmitters." Parasitics are classified as any spurious oscillations taking place in addition to the normal oscillation for which the circuit is designed. They occur as any normal oscillation does when the necessary conditions exist. In many instances troubles which are attributed to other causes actually are due to parasitics. They may cause additional carriers and side bands to be radiated, voltage flashover, instability, loss of efficiency, and short life or failure of vacuum tubes and other circuit elements. They cannot be fully anticipated in design of new transmitters and usually it is necessary to eliminate them after the transmitter has been constructed. In many cases the determination of the parasitic circuit may require considerable study and the use of cut-and-try experiments. Detuning and damping of the offending circuit to stop the oscillation is then quite simple and methods of determining and eliminating commonly encountered parasitics were described.

Personal Mention

C. A. L. Lowry has become chief engineer of Polymet of Canada at Hamilton, Ontario, having formerly been with the deForest Radio Corporation of Toronto.

K. A. Norton previously with the Bureau of Standards has joined the staff of the Federal Communications Commission of Washington, D.C., as an associate engineer.

Formerly with the deForest Radio Company, C. M. Wheeler is now on the staff of the Federal Telegraph Company of Newark, N. J.

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TECHNICAL PAPERS

THE APPLICATION OF SUPERHETERODYNE FREQUENCY CONVERSION SYSTEMS TO MULTIRANGE RECEIVERS*

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Summary—*In this paper certain problems involved in the application of the superheterodyne circuit to high frequencies are discussed. The effects of coupling between oscillator and radio-frequency circuits are considered, and it is shown that such coupling may occur as the result of space-charge variations in the converter tube. The problem of noise reduction is considered. The means recommended for improvement of high-frequency performance are the use of higher intermediate frequencies, special circuit arrangements to avoid undesired couplings, higher minimum L/C ratios in the high-frequency bands, and the use of sufficient radio-frequency amplification to minimize noise.*

Introduction

THE superheterodyne circuit for receivers is so well known that little discussion of its features is necessary. However, the effects of some of these features are important in connection with the performance of the superheterodyne circuit at high frequencies.

The selectivity of a superheterodyne receiver depends principally upon the design of the intermediate-frequency amplifier which is tuned once and for all at the time the receiver is built. The result is constant arithmetical selectivity—the band widths are practically the same at all signal frequencies. The shape of the selectivity curve can be controlled accurately, since the intermediate-frequency tuning adjustments are independent of the tuning operation performed by the listener.

Tuned radio-frequency circuits are required to supply efficient energy transfer from the antenna to the receiver and to provide suppression of image responses. These radio-frequency circuits and the oscillator circuit are the only ones which must be tuned by the operator and which must be changed for each additional band.

The frequency-conversion system is the heart of the superheterodyne receiver. Voltage from a local oscillator is impressed on an electrode of a converter tube, which may be either a tube designed es-

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pecially for this function or a tube designed primarily for some other function but adapted to this application.

Some of the features desired in converter tubes and circuits for multirange receivers are:

1. Transconductance of the oscillator tube or the oscillator section of the converter tube high enough to insure sufficient amplitude of oscillations with practical circuits.
2. Minimum reaction between oscillator and radio-frequency circuits.
3. A high degree of frequency stability in the oscillator.
4. Selectivity in all bands as good as that required in the broadcast band.
5. High conversion gain, approximately uniform in all bands.
6. Minimum switching requirements.
7. Low tube noise.

Some of the requirements are difficult to achieve in the high-frequency band. While limitations of the tubes employed contribute to these difficulties, certain design practices are also responsible to an even greater degree.

The use of a 465-kilocycle or lower intermediate frequency is responsible for the most important defects in multirange receivers. In the high-frequency bands, 465 kilocycles is only a few per cent of the signal frequency, and the radio-frequency circuits will be almost in resonance with the oscillator circuit. Very small coupling between the oscillator circuit and the radio-frequency circuit connected to the converter control grid will result in an oscillator-frequency voltage of considerable magnitude at this grid. Conversion gain is substantially reduced by this condition. Alignment of oscillator and radio-frequency circuits becomes difficult because of the reaction of one upon the other. Image response ratios are very poor.

The practice of using the same tuning-capacitor range in all bands, while desirable to minimize the number of bands, results in unfavorable L/C ratios at the low-frequency ends of the high-frequency bands. When the L/C ratio is low, the signal voltage obtained from the antenna is also relatively low, and the noise-to-signal ratio is high. The design of oscillator circuits becomes difficult, and the oscillator voltages at the low ends of the high-frequency bands are often inadequate for complete modulation of the converter current. This results in relatively high converter noise.

When the selectivity requirement is met, very small fluctuations in oscillator frequency result in large changes in output. The motor boating and acoustic feed-back difficulties so frequently encountered in multirange receivers are generally chargeable to this cause. A sharply peaked selectivity curve will tend to aggravate these difficulties. Since oscillator circuits must generally be designed for maximum coupling to make them operate at all in the high-frequency bands, there is little latitude for the improvement of the inherent stability of these circuits. Heavy filtering of voltage supplies helps to prevent motor boating, and careful attention to the mountings of the speaker and the tuning units will minimize acoustic feed-back.

Converter Circuits

Converter circuits used at high frequencies are almost identical with those used at broadcast frequencies. Modifications are sometimes necessary to facilitate switching, and other modifications are sometimes made with the object of improving oscillator performance. An example is the use of a separate oscillator tube coupled to a pentagrid converter.

Pentode Circuits

The triode pentode (type 6F7) is not usable for the higher frequency bands on account of the relatively low transconductance of the triode section. The small cathode area available for the triode is responsible for the low transconductance.

Single-tube pentode-converter circuits, such as were popular at broadcast frequencies before the introduction of the pentagrid converter, are not available for use at high frequencies because of the low usable oscillator transconductance and the reaction between oscillator and radio-frequency circuits. A separate oscillator may be used with a pentode in several ways. Oscillator voltage may be applied inductively to the signal circuit, in series with the cathode, to the screen, or to the suppressor. All of these methods have been used at broadcast frequencies.

Application of oscillator voltage in the cathode circuit is probably the most popular method. The oscillator voltage required for complete modulation of the plate current is obtained with least difficulty by this connection. Coupling between oscillator and radio-frequency circuits is due principally to the grid-to-cathode capacitance of the tube, but the variation of space charge at oscillator frequency plays some part in determining performance.

Inductive coupling between oscillator and radio-frequency coils is difficult to apply at high frequencies, because of the wide range of variation of the impedance of the radio-frequency circuit to the oscillator frequencies. It can be used in combination with other methods to provide a partial correction for the effect of undesired couplings. Capacity coupling may also be used for this purpose in some cases.

The voltage and power requirements for modulation in the screen-grid circuit are great enough to discourage the use of this method at high frequencies. Coupling occurs through the screen-to-control-grid capacitance and through variation of the space charge.

Modulation in the suppressor-grid circuit of a pentode does not result in appreciable direct capacitance coupling between oscillator and radio-frequency circuits. Space-charge coupling is small, but some exists as a result of the electrons, turned back at the suppressor, entering the space between screen and control grid. The oscillator voltage required for complete modulation is fairly high but the power required is negligible. Plate impedance is rather low with suppressor modulation. At high frequencies, direct grid current may be observed in the signal-grid circuit even when this grid is several volts negative; this effect occurs when the potential of the suppressor changes appreciably in the time required for a returning electron to pass between suppressor and screen, and results from the additional acceleration of the returning electrons under such conditions.

Pentagrid-Converter Circuits

The pentagrid converters (type 6A7, etc.) are the most popular converter tubes for multirange receivers at the present time. Among their advantages are high conversion transconductance, minimum switching requirements, and, of course, the elimination of a separate oscillator tube. Electrostatic screening in these tubes makes coupling due to interelectrode capacitances negligible under most conditions. However, coupling between oscillator and radio-frequency circuits does occur to a serious degree in the pentagrid converters, principally because of the variation in the space charge near the control grid at oscillator frequency. This variation in space charge causes a displacement current to flow through the control-grid circuit and results in a voltage of oscillator frequency across the radio-frequency circuit.

The transconductance between control grid and oscillator anode results in variations in the oscillation amplitude when control-grid bias is varied. Some frequency shift may occur along with the amplitude variation. Also, if the impressed radio-frequency signal is large, a pull-in effect may be observed. These effects may be corrected by using a separate oscillator with the converter, but the space-charge coupling effect is not eliminated by use of a separate oscillator.

There seems to be no general agreement as to the desirability of a separate oscillator with the pentagrid converter. Some manufacturers use the pentagrid alone for frequencies up to twenty-two megacycles, apparently with satisfactory results; others use a separate oscillator for excitation.

Effect of Coupling between Oscillator and Radio-Frequency Circuits

The function of the oscillator in a converter circuit is to produce an alternating component of converter current which may be modulated by the incoming signal. This can be accomplished by application of a voltage of oscillator frequency to any electrode of the converter which has an effect on the plate current. Simultaneous application of oscillator-frequency voltages to two or more points in the circuit will produce an alternating current which is the resultant of the effects of these voltages.

When coupling exists between oscillator and signal-frequency circuits, a current of oscillator frequency will flow through the signal circuit and cause a voltage of oscillator frequency across this circuit. The combined effect of this voltage and the applied oscillator voltage will result in a different magnitude of oscillator-frequency current than will be obtained if the signal circuit is shorted. Phase relations between these voltages may operate either to decrease or to increase the oscillator-frequency plate current.

The conversion transconductance of a converter has roughly the same relation to the alternating plate current as the direct transconductance has to the direct plate current. Consequently, if coupling exists between oscillator and radio-frequency circuits in a converter, the conversion transconductance will change when the impedance of the radio-frequency input circuit is changed.

In particular, the conversion gain of a converter system measured with a signal generator connected to the converter grid will be different than the gain obtained under actual operating conditions, since the signal generator shorts the radio-frequency circuit.

A typical series of receiver measurements will illustrate this. The radio-frequency gain of a receiver is measured by connecting a signal generator to the antenna terminals, by substituting a tube voltmeter for the converter, and by finding the ratio of output-to-input voltage. The over-all sensitivity is measured by reading the input voltage giving fifty milliwatts of audio output. Another sensitivity measurement is made with the signal generator connected to the converter grid. The ratio of the third to the second input reading will not be equal to the radio-frequency gain, since the conversion gain is not the same in the two cases. A rather natural error is to attribute the discrepancy to degeneration or regeneration in the radio-frequency circuit, but selectivity measurements will show that the effects of degeneration and regeneration are small. In one case, the radio-frequency gain was four times as great as the ratio of sensitivities. The observer claimed that the converter reduced the radio-frequency gain by this ratio through degeneration, but at the same time he noted that the image ratio was slightly better than separate selectivity measurements on the radio-frequency circuit indicated. The latter observation indicates regeneration in the input circuit; so the first observations must be interpreted as indicating that the conversion gain with the input circuit shorted is four times as great as the gain with a tuned circuit in the input.

The voltage induced in the control-grid circuit of a converter will not be of sufficient magnitude to have an appreciable effect on the conversion gain unless the radio-frequency circuit is nearly in resonance with the oscillator.

The voltage developed across the input circuit will frequently necessitate an increase in bias to prevent grid current. The increase in bias will partially cancel the advantage which might otherwise be obtained with a favorable phase relation between the oscillator voltage and the voltage developed across the input circuit.

In the pentagrid converter, space-charge coupling results in a voltage on the input circuit out of phase with the oscillator voltage when the oscillator frequency is higher than the signal frequency. No increase in bias is generally required in this case, since the signal grid becomes positive only when the cathode current is cut off by the oscillator grid. Conversion transconductance, and consequently conversion gain, is substantially reduced by the effect of the coupling at high frequency.

When the oscillator frequency is made lower than the signal frequency, the voltage developed across the signal circuit is in phase with the oscillator voltage. Both control grids are positive at the same time, so that an increase of bias is necessary to avoid grid current. But since the induced voltage is large only at the high-frequency end of the high-frequency band, the increased bias results in a substantial reduction of gain at other frequencies.

In the case of a pentode with oscillator voltage applied in series with the cathode, the voltage effective in modulating the plate current is the voltage across the grid-to-cathode capacitance. When the oscillator frequency is higher than the signal frequency, the signal circuit offers capacity reactance to the oscillator frequency, and the voltage divides between input circuit and grid-to-cathode capacitance. When the oscillator frequency is lower, the voltage between grid and cathode will be greater than the cathode voltage by the voltage developed across the input circuit, but additional bias will be required and will have the same effect as in the case of the pentagrid converter.

Space-Charge Coupling

The motion of a charge in the vicinity of a conductor will cause current to flow in the conductor. In a vacuum tube, a change in the number of electrons in the vicinity of a grid will be accompanied by a flow of current through a circuit connected to that grid, although the grid may be at a potential which prevents electrons from reaching it. The current is a charging current or displacement current, and is proportional to the time rate of change of the potential gradient at the grid.

In a converter tube, if the space charge in the vicinity of the control grid varies at the oscillator frequency, a current of oscillator frequency will flow through the control-grid circuit. If the control-grid circuit offers an appreciable impedance to the oscillator frequency, a voltage of oscillator frequency will appear across this circuit. The effect of such a voltage was discussed in the preceding section.

Another argument demonstrating the existence of space-charge coupling may be given. It is known that the space charge in a tube affects the capacitances between its electrodes. If a direct voltage is impressed across a capacitor and its capacitance is changed, a current will flow. Consequently, a change in the space charge between two elements at different direct-current potentials in a tube will cause a current to flow between these elements, and if the space charge varies at the oscillator frequency, the current will be of oscillator frequency.

In the pentagrid converter, as the oscillator grid becomes more positive, the negative space charge in the vicinity of the control grid increases. This increase in negative space charge tends to drive electrons out through the control-grid circuit. The capacitance between oscillator grid and control grid has the opposite effect; that is, a positive increment of voltage on the oscillator grid will tend to induce a negative charge on the control grid, drawing electrons in through the control-grid circuit. If a circuit offering capacity reactance to the oscillator frequency is connected between control grid and ground, the capacitance between control grid and oscillator grid will act to produce a voltage in phase with the oscillator-grid voltage across this circuit, and the variation in space charge will tend to produce a voltage out of phase with the oscillator voltage.

Fig. 1 shows a circuit used for measuring the space-charge coupling in a pentagrid converter. With this circuit, advantage is taken of the fact that the direct current in the tube changes when a voltage of oscillator frequency is developed at the control grid. The impedance of the control-grid circuit is varied by changing the capacitance of the circuit.

As the control-grid circuit is tuned through the oscillator frequency, the capacitance being increased from minimum towards maximum, the plate current first increases, then passes through a maximum, decreases rapidly to a minimum, and returns slowly to its initial value. The maximum and minimum indicate maximum in-phase and out-of-phase components of oscillator-frequency voltage at the control grid. When capacitance is introduced between oscillator grid and control

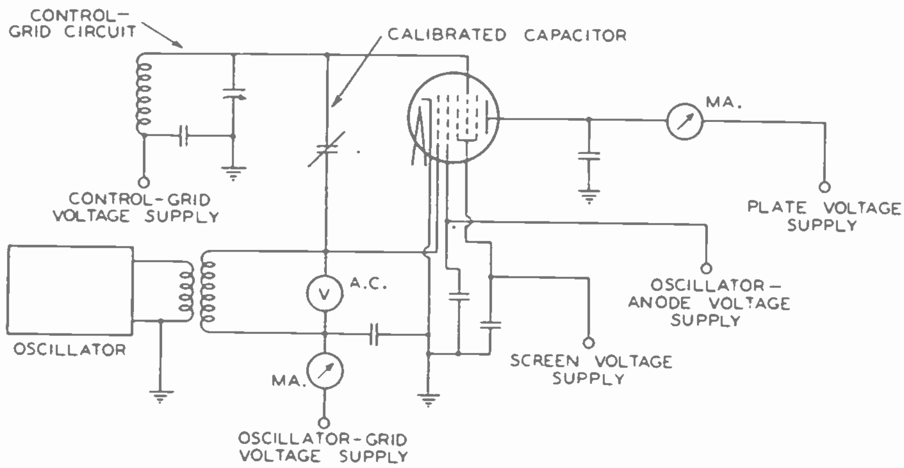

Fig. 1—Circuit for measurement of space-charge coupling.

grid, the magnitudes of the plate-current variations are reduced and may be practically smoothed out by a critical adjustment of this capacitance. A further increase in capacitance will cause fluctuations in plate current in the opposite order—first a minimum, then a maximum. This is sufficient demonstration that the space-charge coupling is greater in magnitude than the capacity coupling between oscillator grid and control grid.

The charging current through the direct capacitance will be directly proportional to the oscillator voltage. The displacement current resulting from space-charge coupling will not increase so rapidly, because the space-charge density will approach a limiting value. It is, therefore, possible to find conditions of operation under which the space-charge coupling is approximately balanced by the direct-capacitance coupling. Measurements in a circuit of the type of Fig. 1 show that this condition is realized under recommended broadcast-frequency operating conditions. Under typical high-frequency operating conditions, additional capacitance of 0.5 to 1 micromicrofarad is required to balance the space-charge coupling.

The situation is more complicated when the oscillator anode grid is used. The voltage on this electrode is out of phase with the oscillator control-grid voltage; consequently, capacitance between anode grid and signal control grid results in coupling in phase with the space-charge coupling. A regenerative effect may also be present, since the voltage developed on the signal control grid has some effect on the anode-grid current.

1. OSCILLATOR FREQUENCY HIGHER THAN SIGNAL FREQUENCY

Circuit		Capacitance Coupling		Variation of Space Charge	
		Between:	Phase°°	Between:	Phase°°
Pentode with Triode Oscillator Voltage Applied to:					
	Cathode	G_1 and Cathode	−	G_1 and Cathode G_1 and G_2	− −
	Screen	G_1 and G_2	+	G_1 and Cathode G_1 and G_2	− −
	Suppressor	G_1 and G_3	+*	G_1 and Cathode G_1 and G_2	+*
Pentagrid Converter		G_1 and G_4 G_2 and G_4	+* −*	G_3 and G_4 G_5 and G_4	− −

*Small or negligible.

°°Phase (+) indicates induced voltage in phase with oscillator voltage. Phase (−) indicates induced voltage out of phase with oscillator voltage.

2. OSCILLATOR FREQUENCY LOWER THAN SIGNAL FREQUENCY

The above table applies if all phase relations are reversed.

Fig. 2—Phase relations between oscillator voltage and voltage induced on signal grid.

Other converter circuits exhibit space-charge coupling effects, but in most of these circuits coupling through direct capacitance is more important. The phase relations between oscillator voltage and voltage induced on the control grid for several circuits are shown in Fig. 2. There is an important difference between the pentagrid type of converter and the type represented by a pentode with suppressor modulation. In a pentagrid converter, variation at oscillator frequency of the space charge near the control grid is essential to the operation of the tube. Consequently, space-charge coupling is inherently present in this type of converter. In the second type, variation of space charge in the vicinity of the control grid is not essential to the operation of the tube. If the electrons turned back by the suppressor can be prevented from entering the space between screen and control grid, space-charge coupling can be avoided.

Space-charge coupling is not exclusively a high-frequency effect. It will be observed at broadcast frequencies if intermediate frequencies of less than 100 kilocycles are used.

Converter Noise

At broadcast frequencies, the principle source of noise in a receiver using a radio-frequency stage is the thermal-agitation voltage developed in the circuit connected to the grid of the first tube. When the converter is the first tube, the converter noise may equal or slightly exceed the thermal noise. In either case, the total noise output will vary only slightly with variations in tube noise, and there can be no large difference between noise outputs with different types of tubes. This statement has been verified whenever precise measurements of noise-to-signal ratios in a receiver have been made. The maximum variation in noise-to-signal voltage ratios observed with different types of tubes in receivers operating in the broadcast band is approximately two to one.

In the high-frequency bands, the impedance of the tuned circuits is frequently so low that thermal-agitation voltages are negligible in comparison with tube-noise voltages. In such cases, the variations in noise-to-signal ratio correspond to the variations in tube noise.

A tube used as a converter produces approximately twice as much noise as the same tube used as an amplifier under the most favorable conditions. When the oscillator voltage is insufficient to modulate completely the converter current, the noise will be still higher. Moreover, the tubes designed especially for converter operation are inherently somewhat noisier than other tubes. The use of a radio-frequency amplifier stage in the high-frequency bands is, consequently, essential for quiet operation. If the gain of such a stage exceeds the ratio of noise voltages in the two tubes, the converter noise may be neglected. When the gain is less than this ratio, the noise will be approximately the converter noise divided by the gain. If converter noise is extremely high or radio-frequency gain attainable in a single stage is low, more than one radio-frequency stage may be required to reduce noise to a minimum.

An estimate of the noise in a receiver may be made by assuming that the following noise voltages appear at the grids of the tubes:

Pentodes, as radio-frequency amplifiers	1 microvolt
Pentodes, as converters (complete modulation)	2 microvolts
Pentagrid converters (complete modulation)	4 microvolts
Thermal agitation, broadcast band	3 microvolts

Thermal agitation in other bands is inversely proportional to the square root of the frequency. In a band starting at 5500 kilocycles, thermal-agitation noise will be approximately one microvolt.

The noise in a converter with incomplete modulation is inversely proportional to the gain realized with untuned input. If the gain in a pentagrid converter is one fourth of the gain realized at broadcast frequencies, the noise voltage is sixteen microvolts.

In a remote cut-off tube, the noise increases as the bias increases, and is approximately inversely proportional to the square root of the amplification. In a sharp cut-off tube, the noise is not changed materially when the bias is changed.

To obtain the equivalent noise voltage at the last radio-frequency grid, divide converter noise by radio-frequency gain. The same method applies in obtaining noise at the antenna, and at the first radio-frequency grid when two stages are used.

When all noise voltages have been referred to the same point, their resultant is obtained by extracting the square root of the sum of the squares of the noise voltages.

Correction of L/C Ratio

The best method of improving the unfavorable L/C ratios generally found at the low-frequency ends of the high-frequency bands is the use of separate capacity sections for the high-frequency bands, and the switching of capacitance along with inductance.

A second method is the use of fixed capacitors in series with the tuning capacitors in the high-frequency bands. The power factors of the series capacitors must be low.

A purely mechanical method of obtaining lower maximum capacitance may be used. This consists in limiting the motion of the tuning capacitor when the switch is in a high-frequency position. Of course, best design will require a mechanical linkage which will cause the tuning indicator to traverse the entire scale while the capacitor is rotated through the reduced arc.

An additional switch position and set of coils will probably be required to cover the desired range of frequencies when any of these methods are used, but this complication is justified by the improvements in sensitivity, noise-to-signal ratios, and stability when high standards of performance are required.

Careful attention to wiring and other factors affecting minimum capacitance will assist in limiting the reduction in tuning range produced by any of these methods.

Intermediate-Frequency Amplifier

A proper choice of the intermediate frequency for a multirange receiver may be arrived at by analogy with broadcast-frequency experience. A widely used intermediate frequency for broadcast receivers has been 175 kilocycles. This intermediate frequency requires, preferably, three tuned radio-frequency circuits for adequate image ratios, although two circuits may be used with fair results. Lower intermediate frequencies offer better selectivity, but the maintenance of satisfactory image ratios becomes difficult. Space-charge and capacitance coupling effects are just measurable under normal conditions with 175 kilocycles, but become more serious when lower frequencies are used. At the high-frequency end of the broadcast band, the difference between oscillator frequency and signal frequency is 11.7 per cent.

If this difference is to be maintained at twenty megacycles, an intermediate frequency of 2300 kilocycles is indicated. However, a frequency just beyond the high end of the broadcast band, possibly 1600 kilocycles, will give very satisfactory results.

If arithmetical selectivity is to be maintained at broadcast-frequency standards, from six to ten 1600-kilocycle circuits may be necessary. Preferred design requires the use of high-Q circuits critically coupled to give a flat-topped selectivity curve. Such a curve tends to minimize instability caused by frequency fluctuations in the oscillator.

The use of 1600-kilocycle or higher intermediate frequencies will practically eliminate the effects of coupling between oscillator and radio-frequency circuits in the high-frequency bands.

When lower intermediate frequencies are used, considerable improvement in stability may be expected if the circuits are designed for a flat-topped selectivity curve. Either staggered tuning or combinations of overcoupled and undercoupled circuits can be used to obtain such curves, with approximately equivalent results. The fact that the flat-topped selectivity curve is also necessary in the design of high fidelity receivers will probably lead to its more extensive use.

Parallel Push-Pull Circuits

Certain balanced converter systems may be used to minimize the coupling between oscillator and radio-frequency circuits. Examples of such systems are shown in Figs. 3 to 6.

In Fig. 3, two pentodes are employed as converters and a triode is used as the oscillator. Signal voltage is applied to the two pentodes in parallel, and oscillator voltage is applied in push-pull. The intermediate-frequency output is obtained from a push-pull circuit. This is necessary because reversal of phase of the oscillator voltage in one of the converters results in reversal of phase of the intermediate-frequency output from that tube. Voltage of oscillator frequency induced on the control grid in one tube is balanced by the corresponding voltage of opposite phase induced on the control grid in the second

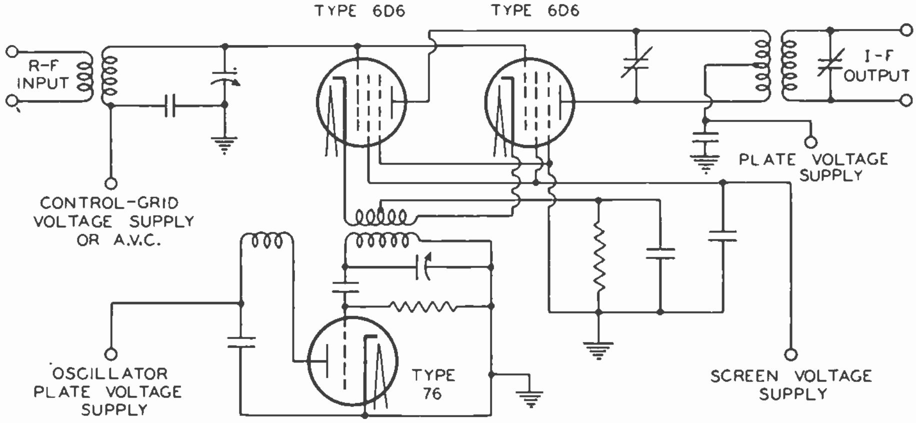

Fig. 3—Parallel push-pull converter circuit with two pentodes and triode oscillator.

tube. Consequently, with reasonably good balance between the tubes and symmetry of circuits, coupling between oscillator and control circuit will be negligible.

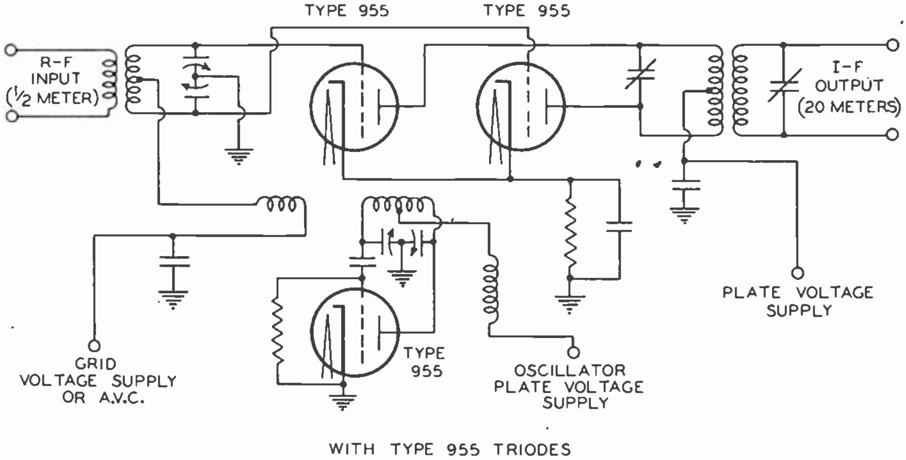

Fig. 4—Parallel push-pull high-frequency converter circuit with type 955 triodes.

Fig. 4 illustrates the application of the parallel push-pull system to very high frequencies. In this case, inputs are in push-pull, since symmetrical antenna systems are generally used. The oscillator voltage is applied to two 955's in parallel, and the output is push-pull because of the reversal of phase of the input to one of the tubes.

Fig. 5 shows a circuit similar in principle, using a type 85 tube. The triode section is used as the oscillator, and the diodes are used as converters. Conversion gain of approximately unity may be obtained with this circuit.

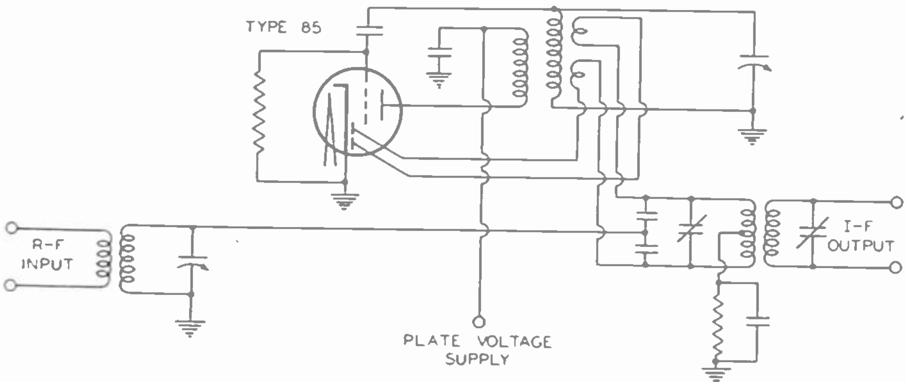

Fig. 5—Parallel push-pull converter circuit with duplex-diode triode, triode as oscillator, diodes as converters.

Fig. 6 shows two pentagrid converters, self-excited, connected with inputs in parallel, and oscillators and outputs in push-pull. This circuit will oscillate at frequencies as high as forty megacycles; conventional single tube circuits are limited to approximately twenty-five mega-

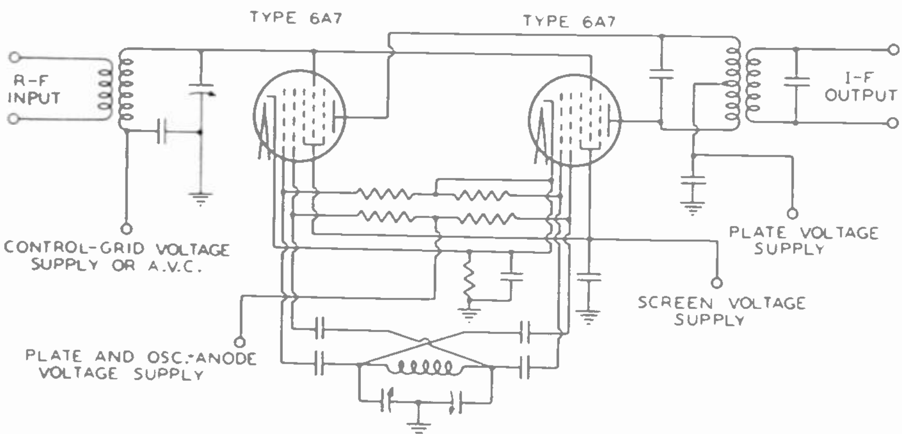

Fig. 6—Parallel push-pull converter circuit with pentagrid converters.

cycles. No more switching is required than for single tube circuits. The extra tuning-capacitor section required in this arrangement is a disadvantage. However, there are receivers now on the market using split tuning-capacitor sections for the high-frequency bands. If the two sections of the oscillator condenser are equal, the circuit shown can be employed.

Taps on the oscillator coil may be preferable for optimum performance, or a tickler coil can be wound between the turns of the grid coil. Capacity between windings will not be serious, since in this circuit adjacent turns can be at the same radio-frequency potential. These and other oscillator arrangements are shown in Fig. 7. The use of two converter tubes will result in a theoretical reduction of 30 per cent in the converter-noise-to-signal ratio.

When inputs are in parallel and output circuits are in push-pull, a signal at the intermediate frequency impressed on the input grids will produce little or no response. If oscillator voltages are applied in

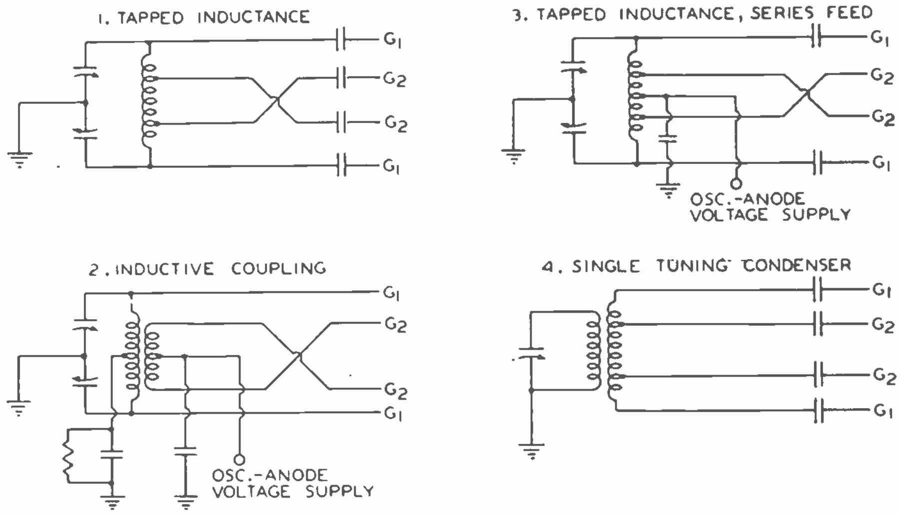

Fig. 7—Parallel push-pull pentagrid-converter circuit alternative oscillator arrangements.

push-pull and the oscillator frequency is made equal to twice the intermediate frequency, output at the intermediate frequency will be obtained. Consequently, this circuit may be used for reception in a band including the intermediate frequency. Of course, very close balancing is required to avoid a "birdie" at the intermediate frequency, but signals differing by ten kilocycles from the intermediate frequency can be received without the severe regenerative or degenerative effects encountered when this is attempted with a single tube.

Conclusion

The requirements listed for good performance in the first part of this paper can be fulfilled for the most part by application of the principles discussed. The restriction of the tuning-capacitor range in the high-frequency bands assists in the attainment of sufficient oscillator voltage, frequency stability, uniform gain in radio-frequency amplifier and converter, and minimum noise-to-signal ratios. Reaction between oscillator and radio-frequency circuits may be avoided either by the use of a high intermediate frequency or by application of the parallel push-pull circuits described. The use of a high intermediate frequency is the better plan since good image-response ratios can be obtained only in this way, but the large number of circuits required for adequate selectivity is a disadvantage. The noise can be limited approximately to that coming from the first tube by the use of sufficient radio-frequency amplification. The use of coils of reasonably large diameter is necessary to obtain maximum signal input to the first tube; such coils will also assist in obtaining sufficient radio-frequency gain and sufficient oscillator voltage. The cost of these improvements is probably not so great as to prevent their inclusion in better grade receivers with real benefit to high-frequency performance.

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A STUDY OF TELEVISION IMAGE CHARACTERISTICS*

PART TWO[1]

Determination of Frame Frequency for Television in Terms of Flicker Characteristics

By

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Summary—*During 1931 and 1932 an investigation was carried out to obtain quantitative information on the several characteristics of television images. Part One of this paper covered those characteristics relating to image detail. This paper, Part Two, covers a determination of frame frequency in terms of flicker characteristics. The analysis is based on a number of simple tests largely in terms of equivalents so that a wide range of conditions might be studied. Conclusions are reached regarding several means for minimizing flicker.*

Television images consist of rapidly superimposed, individual frames much the same as motion pictures. In the case of motion pictures a group of time related stills is projected at a uniform rate, rapid enough to form a continuous picture through persistence of vision. By present methods each frame of a television image is built up element by element in some definite order, and these time related frames are reproduced at a rapid rate.

In motion pictures the taking or camera frame frequency determines how well the system will reproduce objects in motion. This has been standardized at 24 frames per second. In television it is assumed that we shall use a frame frequency of 24 per second or greater. Since this is satisfactory for motion pictures, it is also satisfactory for television and this characteristic of frame frequency will, therefore, not be considered further.

In the reproduced image there is another effect of frame frequency which has been the subject of investigation. This is the effect of frame frequency on flicker. Motion picture projectors commonly used are of the intermittent type. The usual cycle of such a projector is that at the end of each projection period the projection light is cut off by a "light cutter," the film is then moved and stopped so that the succeeding frame registers with the picture aperture, the light cutter then opens, starting the next projection period. This is repeated for each

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[1] E. W. Engstrom, "A study of television image characteristics, Part I," Proc. I.R.E. vol. 21, no. 12, pp. 1631–1651; December, (1933).

frame—24 per second. Since projection at 24 light stoppages per second with the illumination levels used in motion pictures causes too great a flicker effect, the light is also cut off at the middle of the projection period for each frame for a time equivalent to the period that it is cut off while the film is moved from one frame to the next. This results in projection at 24 frames per second with 48 equal and equally spaced light impulses. Such an arrangement provides a satisfactory condition as regards flicker. In television we also may have a reproduced image at 24 frames per second, but because of the manner in which the image is reconstructed, a continuous scanning process, it is not practicable further to break up the light impulses by means of a light chopper in a manner similar to that used in the projection of motion pictures. We, therefore, have for the usual systems of television a flicker frequency which corresponds with the actual frame frequency (24 per second, for example). This is satisfactory at very low levels of illumination but becomes increasingly objectionable as the illumination is increased.

It is of interest to review very briefly some of the fundamental considerations regarding time relations in vision. Full treatment of this may be found in most texts on optics dealing with the eye and the physiological aspects of the subject.

When the retina of the eye (adapted to darkness) is suddenly exposed to a field of steady brightness, the sensation rises rapidly to a maximum and then falls to a lower constant value. When the stimulus is removed, the sensation does not immediately disappear but takes a finite time to decay below the limit of perception. Thus, if the eye is exposed to a source of rapidly varying intensity, the effects of the finite rates of growth and decay of sensation (or the persistence of vision) may prevent flicker from being noticeable. This is true provided that the total cycle of variations is regular and of high enough frequency.

If the frequency of a varying source is sufficiently high so that flicker is imperceptible, the eye is able to integrate the brightness over the cycle of variation. Thus $I=(1/t)\int_0^t i\,dt$, where I is the apparent brightness, t the total period of one complete cycle of variation, and i the instantaneous brightness. The effect is as if the light for each cycle were uniformly distributed over the period of the cycle.

The highest frequency at which flicker can just be detected is called the critical frequency. It has been shown that the critical frequency is practically a linear function of the logarithm of the brightness of the field (over the range of interest for television). The sensitivity of the eye to flicker is noticeably increased when increasing the field of view from a few degrees to an image of the size and viewing distances encountered in motion picture practice. The sensitivity to flicker is also greater for averted vision when viewing large fields of varying brightness.

In television there are a number of factors that contribute to flicker effects. These in general are:

Number of frames (light impulses) per second.

Brightness of image.

Percentage of time the image is illuminated for one frame cycle.

Wave form of rise and decay of light impulse.

Size of image in terms of angle subtended at the eye of observer.

Because information of the type in which we are interested for a study of television flicker was not directly available, a number of general tests were made. The first tests were with a simple flicker disk. The set-up included the elements of a motion picture light and optical system—an incandescent lamp light source, a reflector and condenser system, a light cutter (sector disk), a picture aperture, a projection lens, and a reflecting type screen. A diagram of the light cutter is shown in Fig. 1. In order to have the light cutter mechanically balanced, two light openings were used on opposite halves of the disk. A complete cycle of 360 degrees, therefore, consists of one-half revolution of the light cutter disk. The dimensions of the light cutter and picture aperture were such that the aperture was just fully covered (illuminated) by the smallest light opening used (10 degrees) and just fully cut off by the sector for the largest light opening used (350 degrees).

In this set-up the wave form of illumination was determined by the speed of rotation, the dimensions of the light cutter and picture aperture. The variables for the tests were the angle of opening on the light cutter and the speed of rotation. The size of the image used was one foot high, four-by-three ratio of width to height, and was viewed at a distance of six feet (a picture height to viewing distance ratio of one to six). The color of the light was determined by the incandescent lamp (home movie projector type) which was used within the operating voltage range. Tests were made with the above-determined factors on the relationship of number of frames, illumination, and opening of light cutter. For these tests the stray room illumination was of the general order of one-tenth foot candle.

Tests were made for a number of light openings from 10 degrees to 350 degrees. For each light opening, observations were made for several levels of integrated illumination of the screen image from one-half to twenty foot candles. The illumination was measured at the screen with a Weston direct reading illuminometer, looking toward the projection lens and with the light cutter operating. The screen was a large sheet of very white, matte surface drawing paper having a reflection coefficient of about 75 per cent. For each value of light opening and screen image illumination a frame or light impulse frequency was chosen at which the flicker on the image could just be noticed when vision was concentrated at the center of the image. At higher frequencies the flicker was unnoticeable—at lower frequencies the flicker became increasingly noticeable. Each frequency value chosen was the average for four observers. Care was taken in making these observations but, because of the type of the effect, the results are only approximate. The data taken are shown in chart form in Fig. 1.

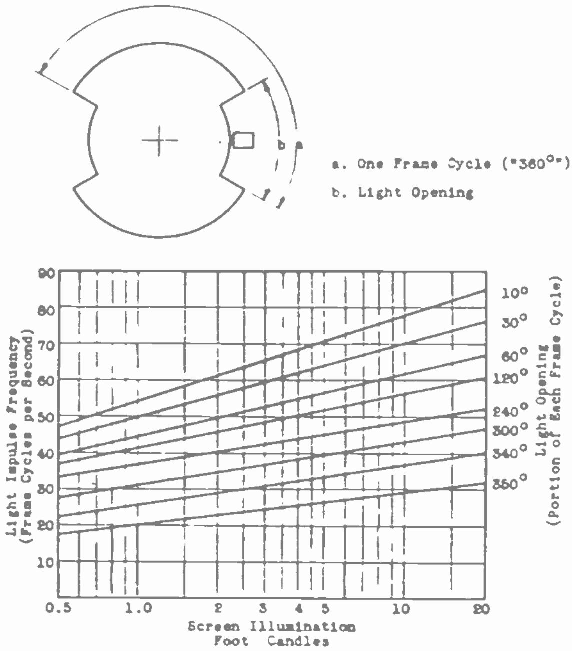

Fig. 1—Sector disk tests. Conditions for just noticeable flicker.

Similar observations were taken for another qualification of flicker. For the same conditions as used in the tests just described, frame or light impulse frequencies were chosen at which the flicker effect became disagreeably objectionable This rating was difficult to judge. Comparison of the data taken with Fig. 1 indicated that the frequency for disagreeably objectionable flicker was lower for all conditions by a fairly constant value (10 to 15 frames) than that for just noticeable flicker.

For practical conditions of viewing, flicker is most noticeable with averted vision. One of the conditions for the above observations was that vision was concentrated at the center of the screen image. If vision is concentrated at one edge of the image, then the other edge is positioned at a larger angle from the central portion of the eye than when vision is concentrated at the center of the image. In order to determine the magnitude of this effect for various image sizes in terms of viewing angles, several observations were made. It was found that with a picture height to viewing distance ratio of one-to-four, and with vision concentrated on the image edge, either left or right, flicker from the opposite edge required the choice of a frame or light impulse frequency approximately ten per cent greater than that indicated when vision was concentrated at the center of the image, for conditions of just noticeable flicker. With a picture height to viewing distance ratio of one-to-eight, a little less difference was noted between these same two conditions. With a picture height to viewing distance of one-to-twelve, very little difference was noted.

Several observations were made to determine the effect of general room illumination on flicker. The same screen material was used as in

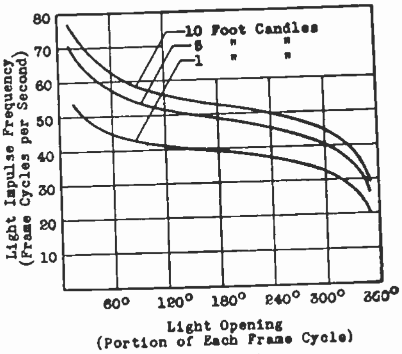

Fig. 2—Sector disk tests.

the previous tests. The screen area was several times that of the projected picture aperture. Tests were made by evenly flooding the screen and room with daylight, measuring the stray illumination at the screen, measuring the added illumination of the projection device, and making observations for the condition of just noticeable flicker. These tests indicated that if the difference between the measured illumination of the image and the measured illumination of the surrounding screen was used as the value of screen illumination, then the results obtained were the same as given in Fig. 1. These observations were carried up to the point where stray illumination equaled the illumination from the projection device.

In order to visualize more readily the effect that the time the image is illuminated for each frame cycle has on flicker, the data in Fig. 1 have been replotted in the form shown in Fig. 2. From these curves obvious conclusions may be reached regarding the portion of each frame that the image must be illuminated and the number of frames per second for the illumination level desired. The results are in terms of just noticeable flicker.

After this general study of flicker we shall proceed to consider several specific conditions more definitely related to television systems. In television systems using moving mechanical devices in combination with a light source which is varied by the incoming picture signal, each element of the image is illuminated for only that portion of each frame that the light source covers that picture element—an extremely short period of time. Persistence of vision is relied upon to carry the effect from one frame to the next.

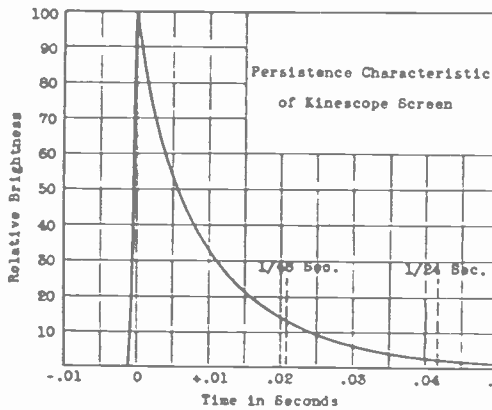

Fig. 3

In a television system using a cathode ray tube (kinescope) in the receiver, each element of the image on the luminescent screen, when excited by the electron beam, fluoresces and assumes a value of brightness corresponding with the value of excitation. Upon removing the excitation this brightness then decays (phosphoresces) in an exponential manner dependent upon the screen material. The phosphorescence or persistence of the image screen aids the persistence of the eye in viewing the reproduced image. The persistence characteristic of a kinescope screen of the type generally used (zinc orthosilicate phosphor—"willemite") is shown in Fig. 3. The spectral distribution characteristic is given in Fig. 4 (the band maxima of this figure are not related in intensity).

In order to obtain data on the flicker from a kinescope under several conditions of operation, a series of observations was made. The deflecting circuits were arranged so that the vertical speed (frames per second) could be varied. The screen height used was six inches and the viewing distance three feet—a picture height to viewing distance ratio of one to six. The stray room illumination was of the general order of one-tenth foot candle. The luminescent screen illumination was measured with a Weston direct reading illuminometer at the kinescope glass envelope looking toward the screen. This was a measure of the light directed to the observer. Sufficient number of lines of even distribution were used to fill completely the picture height of six inches.

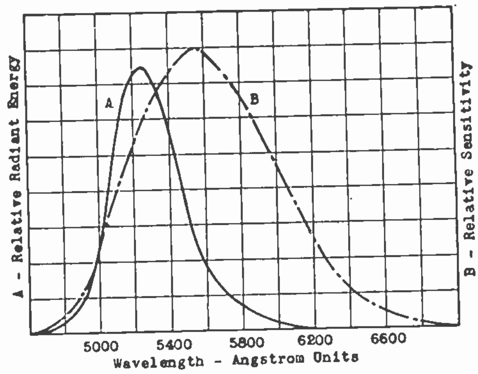

Fig. 4—A—Spectral energy characteristic of kinescope screen.
B—Sensitivity of eye.

Data were taken for three conditions of flicker:

Just noticeable; noticeable but satisfactory; disagreeably objectionable. The results of these observations are shown by Fig. 5.

These observations indicate that for even one foot-candle screen illumination 38 frames per second are required for just noticeable flicker, 35 frames per second for noticeable but satisfactory flicker, and

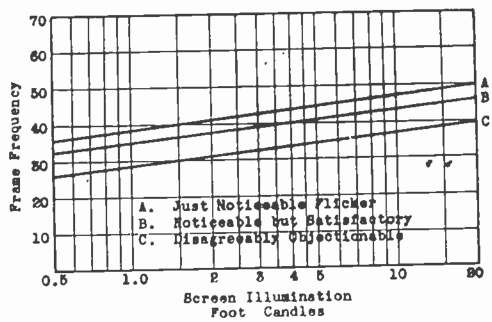

Fig. 5—Kinescope tests.

28 frames per second result in disagreeably objectionable flicker. Thus a standard of 24 frames per second cannot be justified. These data also indicate that 48 frames per second will be satisfactory from the standpoint of flicker for values of illumination likely to be encountered in television.

The light from such a kinescope is restricted to a rather narrow band in the green portion of the spectrum. In order to compare the flicker effect of this green light with the flicker effect of light from a projection type incandescent lamp, an additional set of observations was made. For these tests the same sector disk set-up was used as for the data shown on Fig. 1. The sector disk setting used for this comparison was the 240-degree light opening. Tests were repeated for just noticeable flicker with the normal light from the incandescent lamp, and then again introducing filters so that the light at the screen was about equivalent to the green light from the luminescent screen of the kinescope. The flicker results from the two colors of light of the same illumination were the same within limits of experimental error.

Because of the kinescope operating characteristics and the associated "a-c" type picture signal amplifiers of the usual type television system, the average illumination of the entire screen remains constant. Sections of the picture "ride" on this average illumination, varying above and below. For some reproduced images, sections of the picture may remain at a higher level of illumination than average and, thereby, make flicker of these sections more pronounced. However, in practice this effect is not noticeable, and observations indicate that the flicker of a plain deflection pattern of given illumination is more pronounced than when a signal of an average picture is impressed. It is probable that the television systems of the future will include methods for modifying the average illumination in accordance with the total illumination of the original scene which is being reproduced. For this condition, a frame frequency should be chosen based on the higher sustained levels of illumination in reproduction.

Reference to the persistence characteristic of the kinescope screen shown in Fig. 3 indicates that a change of frame speed affects first, the rapidity of the successive light impulses and, second, the amount of illumination existing at the end of each light impulse cycle. These operate in the same direction, that is, reducing the effects of flicker as the frame frequency is increased and increasing the effects of flicker as the frame frequency is decreased. With the usual shape of decay curve the effect of the persistence characteristic of the screen is limited in the control of flicker. Also, if the persistence is too great, a blurring effect or tail will follow bright moving objects on a reproduced image.

In order to determine the relationship between persistence of the kinescope screen and flicker, a number of tests were conducted. These tests were made through a series of equivalents using a special projection device so as to provide sufficient range for measurement. The projection device consisted of a modified motion picture projector having a constant speed sprocket in place of the intermittent sprocket so as to pull the film past the picture aperture at a uniform rate of speed. Provision was made for a wide range of operating speeds (frames per second) and for a wide range of screen illumination. A group of special films was prepared in which the light transmission characteristic of each frame from top to bottom decreased exponentially at a given rate for each film. Six films were made having attenuation characteristics in accordance with Fig. 6. The dotted lines indicate the number of frames that would be required for the transmission to be reduced to one per cent. Sample prints of one frame are shown in Fig. 7.

These films were projected in the normal manner, that is, sharply focused on the viewing screen. Since the film passed by the picture

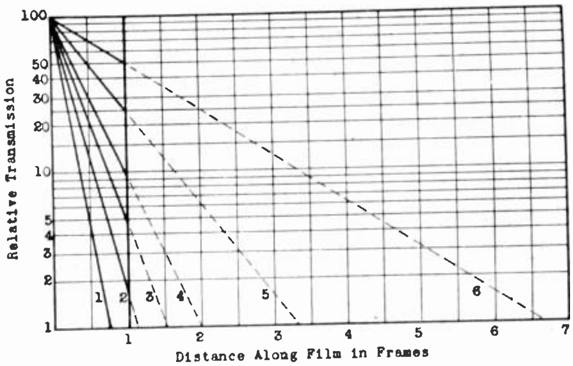

Fig. 6—Transmission characteristics of special films.

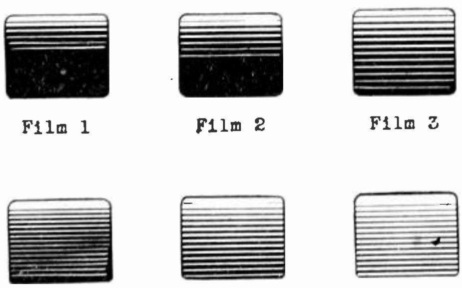

Fig. 7—Sample frames of special films for flicker tests.

aperture of the projector at a constant rate of speed, the visual effect at the screen was practically the same as if viewing a kinescope having a luminescent screen of persistence characteristics corresponding with the particular film used. Comparison with the persistence characteristic for willemite, shown in Fig. 3, indicates that at 24 frames per second film No. 2 corresponds approximately with willemite, and at 48 frames per second film No. 4 corresponds approximately with willemite for this test set-up.

Observations were made with a screen image one foot high and with a viewing distance of six feet. The screen illumination was measured for each film and for each observation at the screen with a Weston direct reading illuminometer looking toward the projection lens and with the projector in operation. For each film a series of data was obtained, for a wide range of screen illumination, on the frame frequency per second (light impulse frequency) required for just noticeable flicker. The observed values used were the average for three observers. These data are shown in chart form by the next three figures—Figs. 8, 9, and 10. This information has been presented in several forms so as to be most useful for indicating the effect of screen persistence.

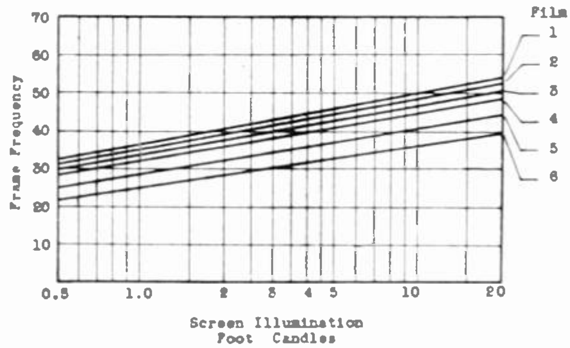

Fig. 8

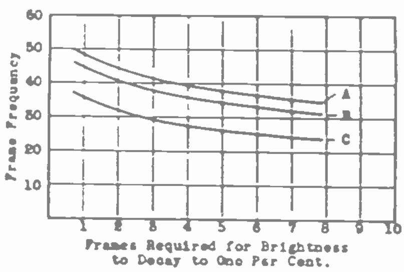

Fig. 9

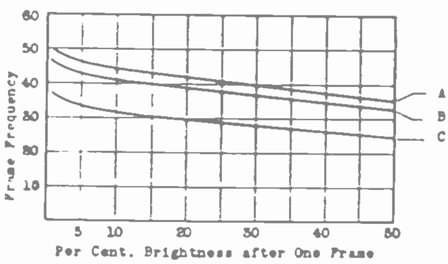

Fig. 10

A—10 foot candles.
B— 5 foot candles.
C— 1 foot candle.

Figs. 8, 9, 10—Conditions for just noticeable flicker. Special film tests.

Inspection of these charts indicates the range of control that the persistence characteristic of the luminescent screen has on flicker. During the tests made, no consideration was given to the "hang-over"—blurring effect that too great a persistence would have on bright moving objects in the reproduced image. It is probable that film No. 6 exceeds the allowable time lag from the standpoint of this characteristic. From these data it does not appear logical that the complete solution to the flicker problem can be arrived at through screens of greater persistence so long as the decay characteristic is exponential in form. It is obvious that the desired wave shape of persistence (brightness versus time) is one which would be flat-topped for the period of one frame or slightly less, and then drop rapidly to zero.

In motion picture practice the projection light is broken up at the rate of two or more times the frame frequency during projection to reduce the flicker effect. A directly similar method is not practicable for television. However, in television certain special procedures in scanning have been used to reduce the effects of flicker. In order to determine the merit of such a system a number of tests were made.

In the usual systems the object scanned is covered by equal horizontal strips or lines from top to bottom in the regular order of lines 1, 2, 3, 4, 5, 6, . . . (progressive scanning). This results in one over-all light impulse for each frame repetition in the reproduced image. If this procedure is modified so that the scanning is, for the first half of one frame period, in the order of lines 1, 3, 5, 7, 9, . . . from top to bottom of the frame, and for the second half of the frame period in the order of lines 2, 4, 6, 8, 10, . . . from top to bottom of the frame (interlaced scanning), then the flicker effect of the reproduced image is changed. Each frame period now consists of two portions with respect to time, the first of alternate lines and the second of the remaining set of alternate lines, properly staggered to form a complete interlaced pattern. This results in an over-all effect of two light impulses for each frame repetition—twice that of the usual method of scanning. However, another effect is now present. In progressive scanning each line flickers at the rate of once per frame, and neighboring lines differ in time relation only by the time required for scanning one line. There is, therefore, no noticeable interline effect. In the interlaced pattern each line also flickers at the rate of once per frame, but neighboring lines differ in time relation by one half a frame period. This results in two flicker effects—an over-all effect and an interline effect.

A test set-up was made to obtain preliminary information on an interlaced scanning pattern. Fig. 11 is a reproduction of a disk, reduced in size, which was prepared for this test. It was mounted on a mechanism and rotated at 24 revolutions per second. The disk was masked except for one aperture having an opening 30 degrees wide and high enough to include the two sections of lines. It was evenly illuminated with daylight; about 20 foot candles. The inner section of the disk corresponds to a condition where each line is illuminated for two thirds of each frame cycle and (for 24 revolutions per second) at the rate of 48 frames per second—a progressive scanning pattern. The outer section corresponds to a condition where each line is illuminated for two thirds of each frame cycle and at the rate of 24 frames per second, but so that alternate groups of lines are illuminated 180 degrees out of phase—an interlaced scanning pattern with a field frequency twice the frame frequency. The width of white lines in each section is the same.

The inner section of the disk was prepared as a check for the observations on the outer section. The results from viewing the inner section were in accordance with the data previously presented. Starting with a viewing distance considerably beyond that which allowed resolution of the individual lines, it was naturally noted that the flicker effect was not noticeable on the inner section. Approaching the disk, a definite distance was reached where the lines could just be seen as separate units. The flicker at this point was, also, unnoticeable. Ap-

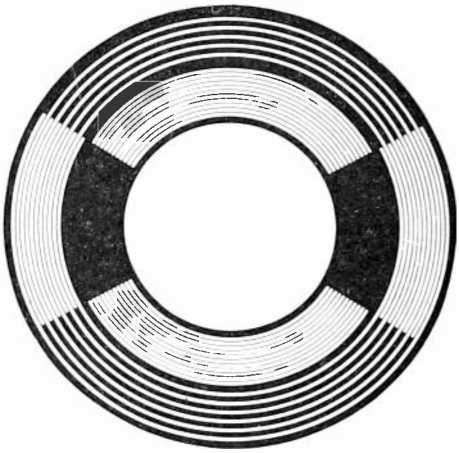

Fig. 11—Special disk for flicker tests with interlaced scanning.

proaching the disk still closer made the line structure increasingly more pronounced, but the flicker effect was still unnoticeable. The same test was repeated for the outer section (interlaced line structure). Starting with a viewing distance considerably beyond that which allowed resolution of the individual lines, a flicker effect was unnoticeable. Approaching the disk it was observed that the line structure could be resolved at the same distance as for the inner section. In addition, a peculiar interline effect was observed and this started sharply at the distance where the line structure was first defined. This effect is rather difficult to describe, but, roughly, the adjacent lines appear to interweave. Approaching the disk closer caused this action to become very pronounced and jumpy. Assuming that the line structure could just be resolved at ten feet, then this jumpy, interweaving effect would first be noticed at this distance. At nine feet the effect would be more definite, and at closer distances it would be objectionable and tiring to the eyes. It was noted that blinking, rapid movements of the eye, or jerky movements of the head, particularly movements at right angles to the lines, caused the interlacing to be momentarily destroyed. The visual effect was apparently to see only one set of alternate lines with spaces between.

From these observations some approximate generalizations may be drawn. An interlaced pattern of the type just described is a means of minimizing flicker. The reproduced image must be viewed at distances equal to or greater than that at which the line structure can be resolved, to prevent undesirable effects from interline flicker. In Part One of this paper[1] it was determined desirable to view reproduced images at distances which were also in accordance with the above findings from considerations of image detail. With progressive scanning the only effect at closer distances is the noticeable line and picture structure. With an interlaced pattern of scanning the limitation becomes more definite because of the interline flicker.

In the test described, each frame of the image was divided into two sets of alternate lines. It might have been broken down into other forms—three groups of every third line, four groups of every fourth line, etc. Since we are using as a base a frame frequency of 24 per second, and since dividing each frame into two over-all light impulses results in satisfactory flicker condition, further subdivision seems unnecessary. Also, with further subdivision the flicker and crawling effect between groups of three lines, groups of four lines, or whatever subdivision above two is chosen, will become increasingly pronounced, requiring in turn proportionately greater viewing distances.

Observations were made on a system using an interlaced scanning pattern with a television receiver using a kinescope. These observations confirmed in general the results of the tests just described with the rotating disk. The indications are that satisfactory flicker conditions are obtained for kinescopes having screen persistence characteristics as in Fig. 3, when using a frame frequency of 24 per second and a field frequency of 48 per second, resulting in an interlaced pattern having the equivalent of 48 light impulses per second. Viewing conditions are in general limited to those equal to or greater than the position where the line structure may be resolved. It is obvious that an interlaced scanning pattern of the type just described requires the same maximum frequency for transmission as an image resulting from progressive scanning with the same base frequency (24 per second). With the interlaced scanning pattern the horizontal line deflecting frequency remains the same as for a progressive scanning pattern but the vertical deflecting frequency is doubled—24 to 48.

The method of obtaining an interlaced pattern and the incidental problems to be solved are beyond the scope of this paper. However, one phase of this is of interest since it influences the required band width for transmission. This is in connection with the practical application of alternating-current power supply systems to cathode ray type receivers using interlaced scanning. The effect of ripple voltages from the power supply system appears in the reproduced image from numerous sources. In progressive scanning, adjacent lines are closely related in time and, therefore, the displacement (due to the effects of ripple) of one line from its true position with respect to adjacent lines is small. The effect of ripple is, therefore, somewhat of an over-all image effect causing to be superimposed on the scanning pattern a varying brightness or line density effect. If the frame frequency differs from the supply frequency —differs except in terms of whole number multiples or sub-multiples— then the effect of this ripple will move across the image. If the frame frequency is a submultiple of the power supply frequency—30 frames for a 60-cycle source—then the effect of this ripple is stationary on the image and much less pronounced. This moving ripple pattern is almost as disturbing as the true flicker and the visual effects are about the same. (The improvement due to the reduction of the ripple effects might make a 30-frame per second image desirable for 60-cycle supply sources for progressive scanning, even at the expense of the increased frequency band.) With the interlaced scanning pattern, adjacent lines are separated in time by one half the period of one frame—one forty-eighth of a second for a 24-frame image where the field frequency is twice the frame frequency. With an interlaced pattern where the frame frequency or field frequency differs from the power supply frequency (such as 24 frames, 48 groups of alternate lines per second for a 60-cycle supply source) the ripple effects also move across the image. Since adjacent lines are widely separated in time, the phase difference of the ripple effect will cause adjacent lines first to draw together and then to separate in correspondence with the time difference between the frame and supply frequencies. With even very well filtered alternating-current to direct-current supply systems and with good magnetic shielding, this effect is sufficiently pronounced and random to destroy the effect of interlacing. The means for reducing the level of ripple to that required to prevent such action are of a classical rather than a practical order. The visual effect is apparently to lose one half of the total number of lines. The solution is found in changing to a submultiple of the power supply frequency, for example, a 30-frame interlaced pattern for a 60-cycle source. The ripple effect is then stationary and, with reasonably well-designed supply systems, is quite unnoticeable. Tests were made which indicated that for a 30-frame per second picture, the transmitter supply frequency and the receiver supply frequency need not be synchronous; a difference up to one cycle per second in power supply frequencies may be tolerated. The choice of a 30-frame per second interlaced scanning has naturally increased the frequency band required in the ratio of 30 to 24.

Summary

We may summarize the results of the study of television image flicker as follows. Naturally the frequency band required for transmission is proportional to the frame frequency.

For a system of television reproduction where the light output of each elementary area is effective only during the scanning period for that area (equivalent of scanning disk), a very high frame frequency is required for useful levels of illumination—Fig. 1.

For a system of television reproduction using a kinescope as the translating device and which has screen luminescence characteristics of the same order as willemite, a frame frequency in excess of 40 per second is required—Fig. 5. A frame frequency of 48 per second will be satisfactory for levels of illumination likely to be encountered in television. (At 48 frames and with a 60-cycle power system the effects of ripple voltages will travel across the image; the choice of 60 frames per second provides a complete solution to the visual effects of both flicker and ripple.)

A generally ideal translating device would be one in which each elementary area consisted of a light source, the brightness of which would be adjusted once each frame. A kinescope luminescent screen is very roughly such an arrangement. However, the persistence of brightness after excitation follows an exponential decay. The choice of material and its fluorescent and phosphorsecent characteristics provides only a partial solution in the control of flicker for practical conditions—Figs. 8, 9, 10.

For a system of television using an interlaced scanning pattern and with a frame frequency of 24 per second, or more, satisfactory flicker conditions exist if each frame consists of two groups of alternate lines (equivalent to 48 frames per second). If, for interlaced scanning, a kinescope is used as the translating device, then the frame frequency must bear a definite relation to the power supply frequency; for a 60-cycle supply the frame frequency may be 30 per second and the field frequency 60 per second. The minimum viewing distance for such an image is limited simultaneously by interline flicker and resolution of line structure.

ACKNOWLEDGMENT

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GENERAL CONSIDERATIONS OF TOWER ANTENNNAS FOR BROADCAST USE*

By

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Summary—*The factors influencing the action of towers when used as radiators are considered. It is shown that the results predicted from the simple theory of sinusoidal distribution of current on the tower differ to a major extent from the actual results. A series of measurements using small models of actual antenna structures resulted in data correlating closely with the performance of the full-sized structures. These measurements showed that departures from the simple theory are due to nonsinusoidal current distribution.*

Several types of recently installed antenna towers are shown to be less effective than the simple theory prediction, particularly with regard to reduction of sky wave and fading. Means for correcting the current distribution and thereby improving the performance are pointed out.

The statement that low base capacity is essential to high antenna efficiency is shown to be a fallacy providing simple precautions are taken to reduce conduction losses.

The ground system and earth currents are considered from both a theoretical and experimental viewpoint. A simple method of measuring the earth currents is described and it is pointed out that such measurements indicate whether the antenna current is sinusoidal or not.

Appendix A gives a method of computing the radiation characteristics and the radiation resistance when the current distribution is known but is not expressed analytically. Appendix B contains the theory behind the ground current measurements. Appendix C shows the influence of the base insulator capacitance on the operating characteristics.

I. Introduction

THE antenna system of a broadcast transmitter is an important factor in determining the effectiveness of the coverage. This is especially true in the case of high-powered transmitters where the service area is limited by the signal fading due to reflections from the Heaviside layer. The ideal antenna would send most of the radiated energy out close to the earth and very little at angles above say twenty degrees. Several years ago, Ballantine[1] attacked the problem of a straight vertical antenna over a perfect earth, with a sinusoidal distribution of current existing on this antenna. He showed that, for a given radiated power, the field strength at the horizon was greatest

* Decimal classification. R120×R320. Original manuscript received by the Institute, December 27, 1934.

[1] S. Ballantine, "On the optimum transmitting wave length for a vertical antenna over perfect earth," Proc. I.R.E., vol. 12, p. 833; December, (1924).

when the antenna height was 0.64 of one wavelength. An antenna of this height would then yield a field intensity at the horizon which would be approximately forty per cent greater than that given by a quarter-wavelength long antenna which was radiating the same power. Recently Ballantine[2] has discussed the problem still further, showing that while an antenna 0.64 wavelengths long gives the greatest intensity along the horizon, the smallest ratio of reflected sky wave to

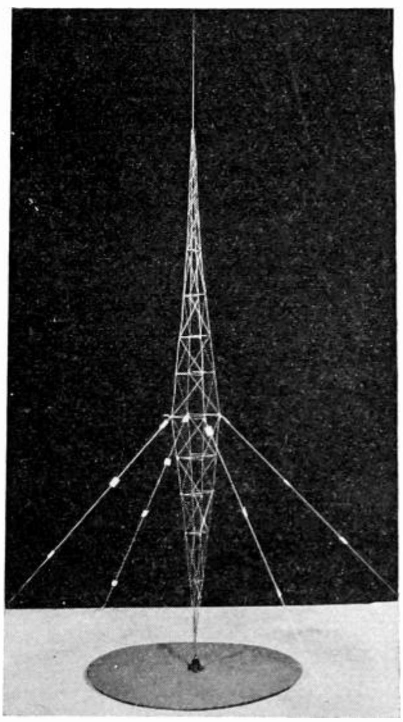

Fig. 1a—Type A antenna.

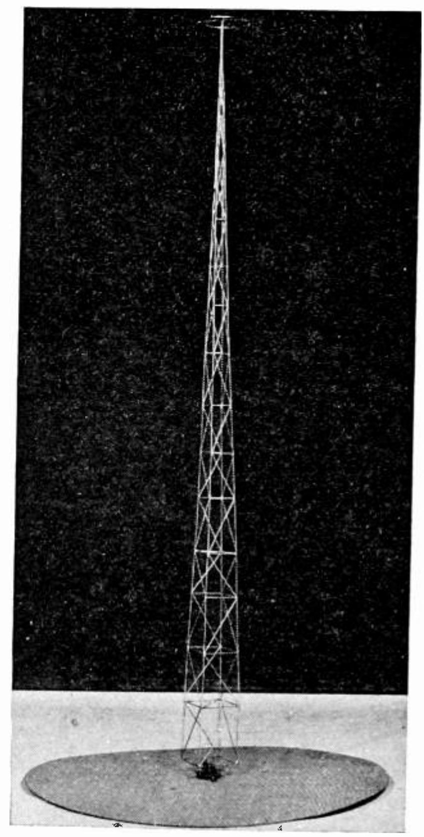

Fig. 1b—Type B antenna.

ground wave is obtained from an antenna whose length lies between 0.5 and 0.64 wavelength, depending upon the amount of attenuation suffered by the ground wave.

In the attempt to attain these optimum conditions, it has been necessary to depart from the conventional antenna construction, consisting of a wire antenna supported between two towers. The idea of using a single tower as the radiator itself was soon proposed and has been widely adopted. The guyed cantilever tower soon appeared. (Fig.

[2] S. Ballantine, "High quality radio broadcast transmission and reception," Proc. I.R.E., vol. 22, pp. 616–629; May, (1934).

1a) This consisted essentially of two conventional towers placed base to base. The lower end of the antenna rested upon a base insulator and the antenna was held in place by from four to eight guy wires. These guy wires, which were broken up at several points by strain insulators, were fastened at the largest cross section of the antenna. Additional height was obtained by a rod at the top of the tower. The conventional tower was also used for an antenna. The tower was mounted on four base insulators, and required no guy wires. (Fig. 1b.) A capacity crown was often placed at the top.

It has been found that these self-supporting antennas gave results not consistent with the theory, in regard to several points. The ground wave was not as large as would be predicted. The resistance versus frequency curve did not check the theoretical curve. Another important effect was also found. The vertical radiation pattern of an antenna whose length is somewhat greater than 0.5 wavelength and whose current distribution is sinusoidal consists of two lobes, one large one along the ground and a smaller one at high angles with a distinct minimum in the neighborhood of fifty-five degrees (curve *A*, Fig. 2). Airplane

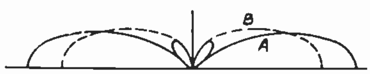

Fig. 2

measurements of the vertical radiation pattern of actually installed antennas showed a complete absence of the high angle lobe and the minimum (curve *B*, Fig. 2). Ballantine[2] has published airplane measurements on WABC, Wayne, New Jersey, that also show this absence. He gives as possible reasons, (1) the antenna current distribution is not sinusoidal, (2) the guy wires have sufficient current flowing in them to cause this skyward radiation, and (3) the ground system is inadequate.

Probably the most important factor is the nonsinusoidal distribution of current. In the type A (Fig. 1a) especially, the distributed inductance and capacitance per unit length vary rapidly along the antenna. An examination of the fundamental equations of a long transmission line shows that the current distribution is sinusoidal only if the inductance, L, and capacitance, C, per unit length are constant along the line. In general L and C vary in such a fashion that the product LC is a constant, while L/C is not. Thus the velocity on any section of line is a constant value close to the velocity of propagation of light, while the surge impedance (approximately $\sqrt{L/C}$) changes. Thus

[2] *Loc. cit.*, Fig. 41.

the surge impedance is small where the cross section of the antenna is large, and vice versa. It is evident that the frequency at which the antenna reactance becomes zero depends on the tower dimensions in a very complicated fashion. It is most certainly erroneous to use this resonant frequency to describe the mode of operation of the antenna. The ratio of operating wavelength to fundamental wavelength (λ/λ_0) is often used. This ratio is only significant when the current distribution is sinusoidal. Even when this is true, it is much better to describe the antenna in terms of its height, a, measured in wavelengths as a/λ. This fraction immediately gives a picture of what portion of a sine wave of current exists on the antenna. A still further simplicity is gained by defining a quantity $G = 2\pi a/\lambda$ radians $= 360a/\lambda$ degrees. For example

$a/\lambda = 0.125 \quad \cdot G = 45° \quad \cdot \lambda/\lambda_0 = 2$ (for sinusoidal current distribution)

$a/\lambda = 0.25 \quad \cdot G = 90° \quad \cdot \lambda/\lambda_0 = 1$

$a/\lambda = 0.50 \quad \cdot G = 180° \quad \cdot \lambda/\lambda_0 = 0.5$

$a/\lambda = 0.5975 \quad \cdot G = 215° \quad \cdot \lambda/\lambda_0 = 0.418$

$a/\lambda = 1.0 \quad \cdot G = 360° \quad \cdot \lambda/\lambda_0 = 0.25.$

When the current distribution is not sinusoidal, the length may still be expressed in electrical degrees. While it does not tell the current distribution, it does tell the ratio of antenna length to operating wavelength. Even if the current distribution were sinusoidal on the tower antennas, the value λ/λ_0 obtained experimentally would be erroneous due to the capacity currents which flow to earth through the base insulator.

II. Measurement of the Current Distribution

(a) *The use of models to determine the distribution*

There has been much speculation as to just what the current distribution is. As far as is known, no one has attempted to measure the currrent distribution on one of these large antennas. The authors soon abandoned the idea of measuring the current distribution on an actual antenna in favor of the use of models operating at short wavelengths. Two model antennas were built, one of each type. A metal plate of fifteen-inch radius was placed on the ground. A small porcelain insulator supported the antenna. The type A tower was provided with guy wires. The antennas were excited through approximately fifty feet of transmission line from the driving oscillator. A small wavemeter circuit was constructed. This consisted of a small coil (two turns), a small variable condenser, and a thermal milliammeter. This wavemeter was placed on a support which enabled it to be moved parallel to the axis of the antenna. The wavemeter was kept at a large enough distance from the antenna axis so that wavemeter readings remained constant as the antenna was rotated about its vertical axis. This constancy of wavemeter reading indicated that the flux density was constant at all points on the circumference of a circle whose radius was equal to the distance from the center of the wavemeter coil to the antenna

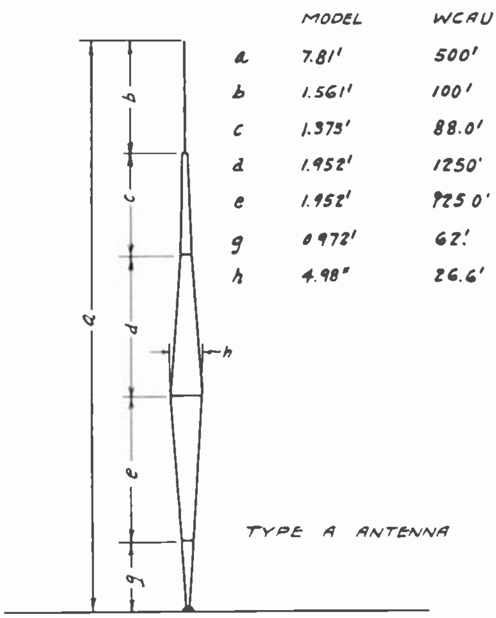

Fig. 3—Dimensions of type A model antenna.

axis. The plane of this circle is perpendicular to the vertical axis of the antenna and the vertical axis passes thru the center of the circle. Then the antenna current at the height of the wavemeter is proportional to the wavemeter reading. It is not necessary to calibrate the apparatus, since relative measurements give the desired quantity, namely, the current distribution.

(b) The current distribution on type A antennas

Since it was possible to obtain other pertinent information relating to the antenna in use at WCAU, Philadelphia, Pa., the type A model was a small scale replica of this antenna. Fig. 3 shows the antenna with the important dimensions. The WCAU tower proper is 400 feet high with an additional shaft extending 100 feet higher, making a total height of 500 feet, (152.5 meters). The operating wavelength is 256 meters. Thus $a/\lambda = 0.595$. The model was to operate at four meters when simulating the WCAU operating conditions, so that the over-all height of the model was $a_m = 0.595 \times 4 = 2.38$ meters $= 7.8$ feet. Guy wires were provided, in this case being No. 18 wire, while the actual guys are about three inches in diameter.

The first test was to determine the effect of the guy wires on the current distribution. Measurements were made with the guy wires attached to the antenna, and also with the guy wires replaced by string. The results of this test are shown in Fig. 4. A theoretical sine curve is

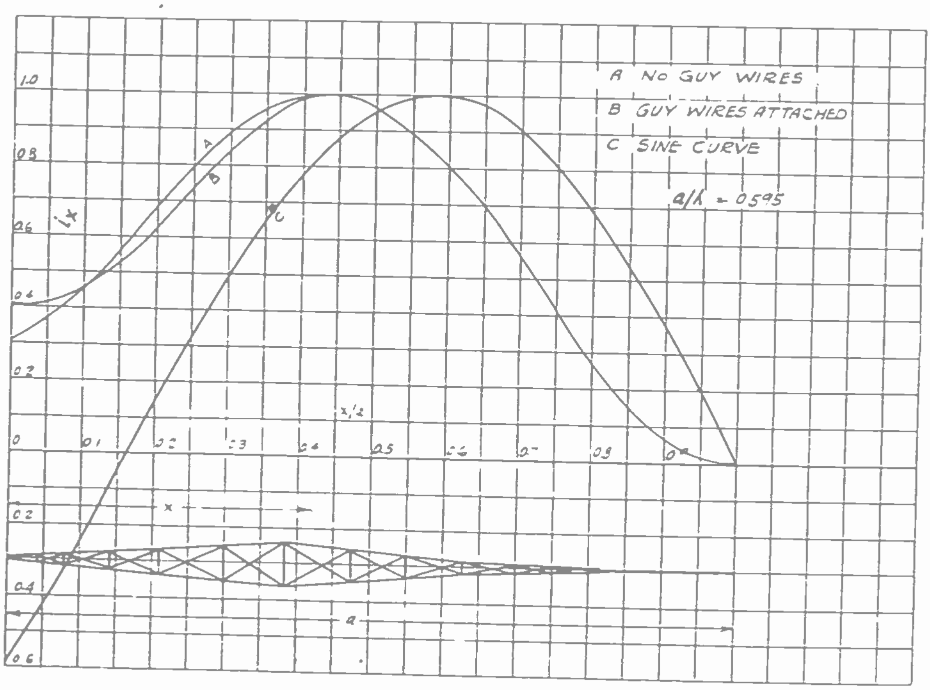

Fig. 4—Current distribution on type A antenna showing the effect of guy wires.

plotted on this same diagram to show the great departure of the current distribution from the theoretical distribution. Other points to be noted are:

(1) The guy wires affect the current distribution only slightly.

(2) The guy wires seem to have no effect on the current distribution above their point of attachment.

(3) The maximum value of current occurs at a much lower point on the antenna than the simple theory predicts.

(4) At no place on the antenna (except at the top end) does the current become zero or even approach zero.

(5) The rod which makes up the top twenty per cént of the antenna carries very little current.

The fourth remark requires amplification. On a long line with uniformly distributed constants, where the end remote from the voltage source is open circuited, the current on the line may be regarded as being made up of two components, one of which travels down the line from the generator to the open end, and the other of which travels in the reverse direction. If these two components have the same amplitude at all points along the line, which condition can occur only when the losses in the line are zero and when no radiation takes place, the result will be a standing wave of current, sinusoidally distributed along the

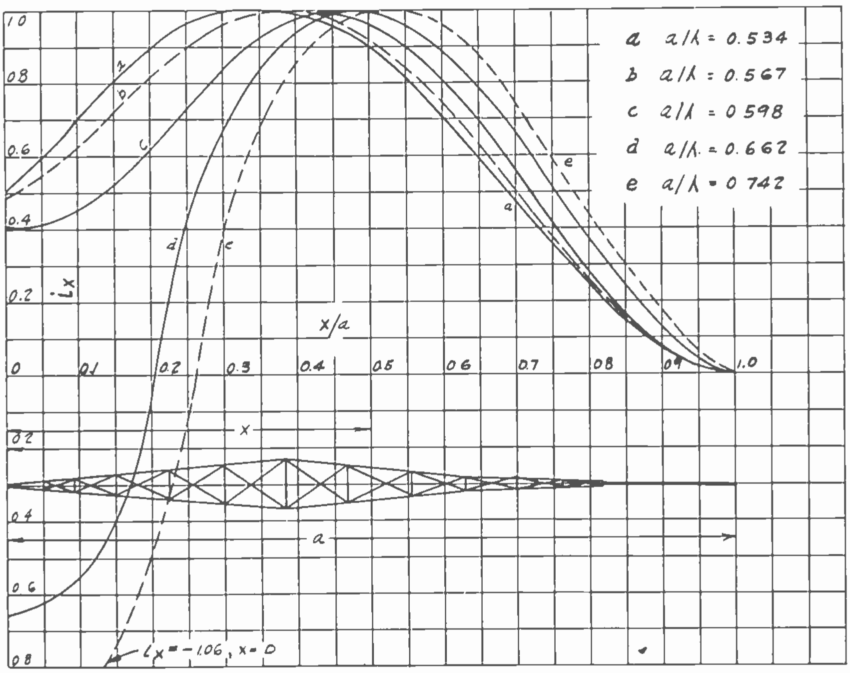

Fig. 5—Current distribution on type A antenna as a function of frequency.

line. Thus at points separated by a half wavelength, the current will go to zero, and the current in the half-wave interval on one side of a zero point will be opposite in phase to the current in the half-wave interval on the opposite side of this zero point. When losses occur in the line, the phase reversal will take place by means of a phase rotation rather than by means of the amplitude going through zero. When the losses are low, the current distribution remains very nearly sinusoidal at all points except in the region of the theoretical zero points. Here the current will be very small except when the losses are very large. On a nonuniform line no such simple picture can be made. It is very likely, however, that phase reversals take place, but not at nicely spaced intervals. In our measurements on these models, it was not possible to measure the phase of the current along the antenna, but only its amplitude. Thus our results can only be reported in terms of magnitudes, although we will plot results as negative values where we find that the current reaches small values and then gets larger. In later tests, it was found that when apparent phase reversals occurred, the current went to very small values. Because of these facts, Fig. 4 may be regarded as showing that no phase reversal occurs.

With the guy wires in place another series of measurements was made, with frequency as the parameter. These results are shown in

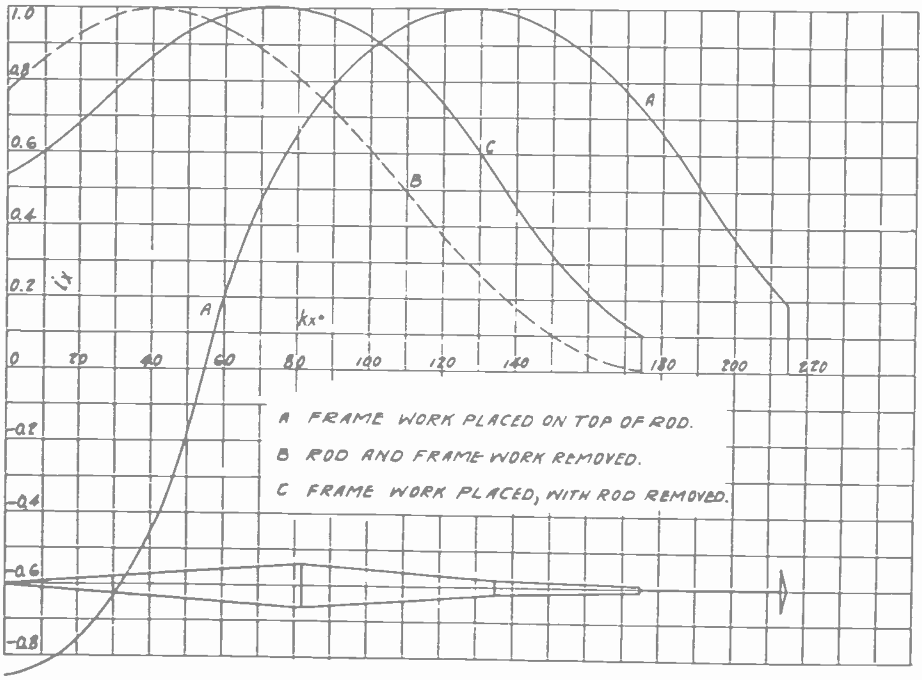

Fig. 6—The effect of some changes on the type A antenna.

Fig. 5. The frequency was varied so that measurements of current distribution are given for values of a/λ from 0.534 to 0.742. It is apparent that the current distribution is far from the simple sinusoid. These curves will be examined again when the radiation patterns are discussed.

A framework was constructed which could be placed at the top of the antenna to give added capacitance. This framework was square and of the same dimensions as the largest cross section of the antenna. The frequency was returned to seventy-five megacycles so that $a/\lambda = 0.595$, and the framework was placed on the top of the antenna. A measurement of current distribution was made. Curve A, Fig. 6 shows this effect. This has the same effect on the current distribution as lengthening the antenna. We see that the current passes through a minimum and apparently undergoes a phase reversal. The rod which formed the top twenty per cent of the antenna and the framework were next removed. Then the value of a/λ was 0.476. The resulting current distribution is shown by curve B, Fig. 6. When the framework was again placed on top of this antenna, curve C, Fig. 6, was obtained. Because the antenna length was changed, the abscissa of Fig. 6 is distance from the ground plate measured in electrical degrees.

(*c*) *The current distribution type B antennas*

The type B antenna is a conventional tower. Its dimensions are shown in Fig. 7. This antenna was a model of the antenna in use at

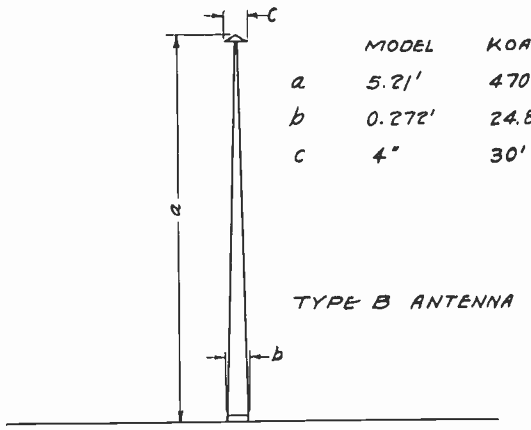

Fig. 7—Dimensions of type B model antenna.

KOA, Denver. The capacity area at the top is circular, having a full scale diameter of thirty feet. For our tests, another ring was prepared which was the equivalent to a diameter of fifty feet. The tests were made in exactly the same manner as before. The first test was made with no capacity area at the top. These results are shown in Fig. 8. The same remarks that were made regarding the distribution on the type A antenna hold true here. The type B tower used no guy wires.

Next, the small ring was placed on the top of the antenna. The results are shown in Fig. 9. Fig. 10 portrays the current distribution when the large ring is placed at the top.

These tests show that type B antenna yields current distributions more nearly sinusoidal than does the type A antenna. This is no doubt due to the fact that the type B antenna changes cross section less abruptly. A study of these diagrams shows the effect of capacity areas at the top of the antenna.

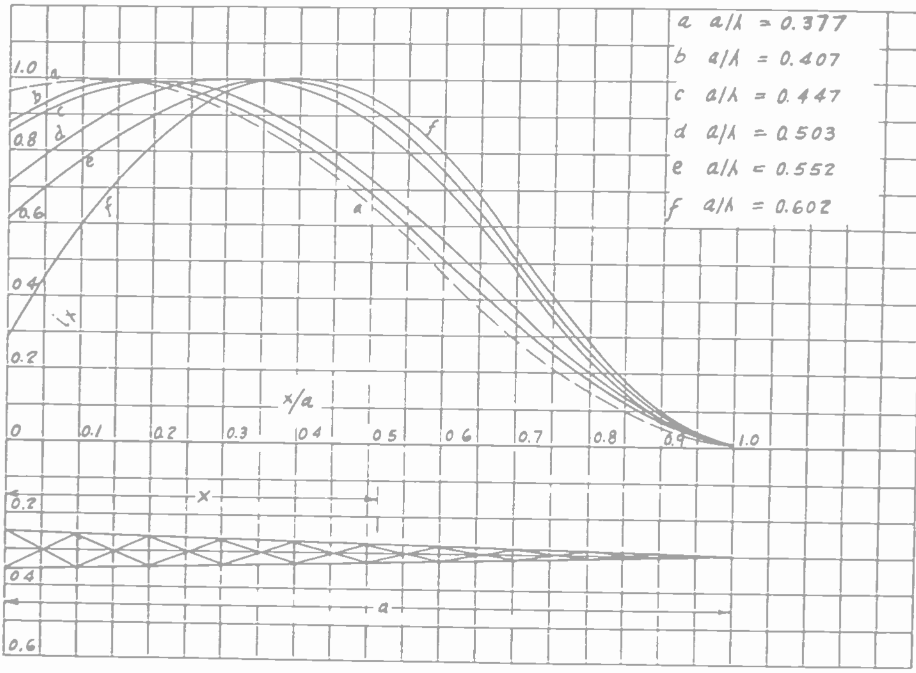

Fig. 8—Current distribution on type B antenna.

(d) *The current distribution on a vertical wire antenna*

In the above, we have attributed the departure from the sinusoidal distribution to the changing cross section of the antenna. It has been

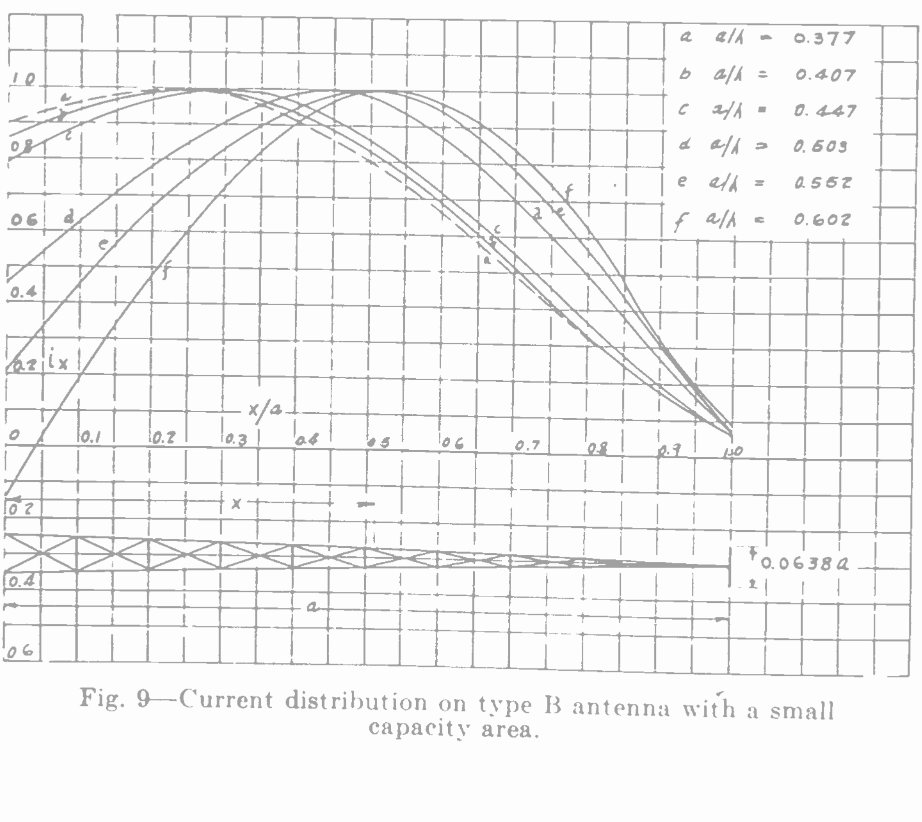

Fig. 9—Current distribution on type B antenna with a small capacity area.

assumed that an antenna of uniform cross section would yield a current distribution very nearly sinusoidal. To prove this point, the tower antenna was replaced by a single No. 14 copper wire. The wire was taped to a vertical wooden pole and the lower end of the wire was fastened to the top of the base insulator which formerly supported the tower antennas. The frequency was held constant at seventy-five megacycles (four meters) and three separate lengths of wire were used. These lengths were related to the wavelength as follows:

$$a/\lambda = 0.2265, \; G = 81.5 \text{ degrees}$$
$$a/\lambda = 0.555, \; G = 199.5 \text{ degrees}$$
$$a/\lambda = 0.6475, \; G = 233.0 \text{ degrees}$$

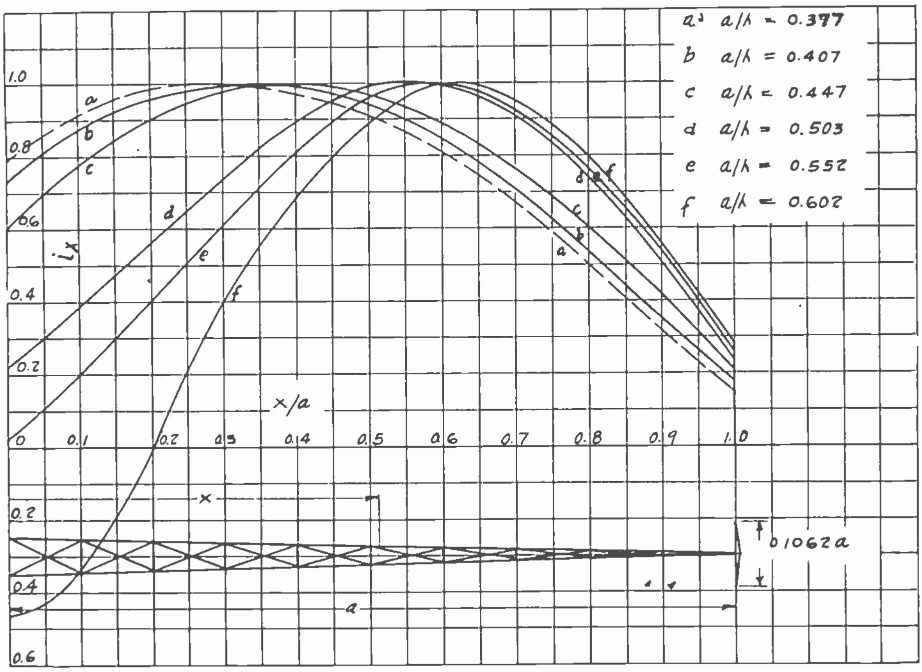

Fig. 10—Current distribution on type B antenna with a large capacity area.

Fig. 11 shows portions of sine waves G degrees long. The crosses, circles, and squares indicate experimental values. It is seen that there is a substantial agreement between the theoretical and experimental values.

III. The Vertical Radiation Patterns

When the current distribution is known, the vertical radiation pattern can be calculated. For our purposes, the earth may be considered to be a perfect conductor, since this assumption changes the high angle radiation only a little. Using the current distributions shown in Fig. 5, we have computed the vertical radiation pattern of the type A antenna for three values of a/λ. These are shown in Fig. 12. Airplane measurements of the vertical radiation from WCAU were taken and the results are shown by the circles on Fig. 12. As stated before, the

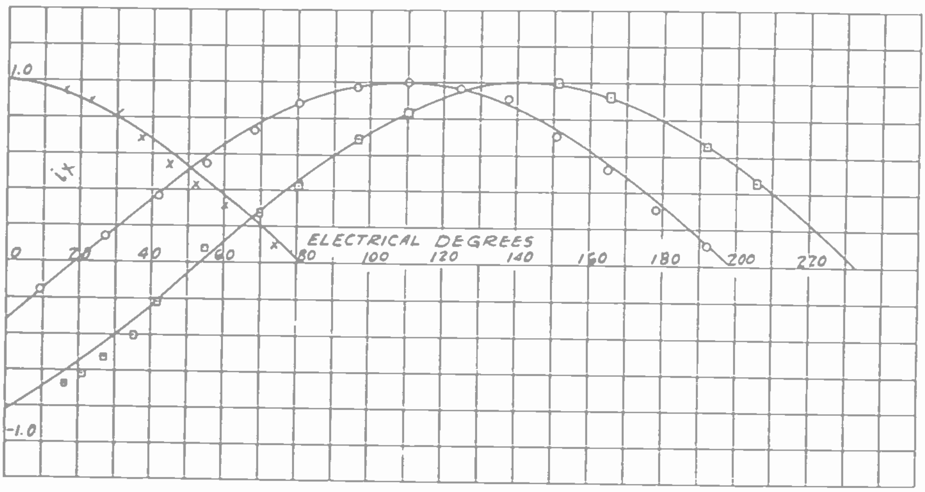

Fig. 11—Current distribution on vertical wire antenna.

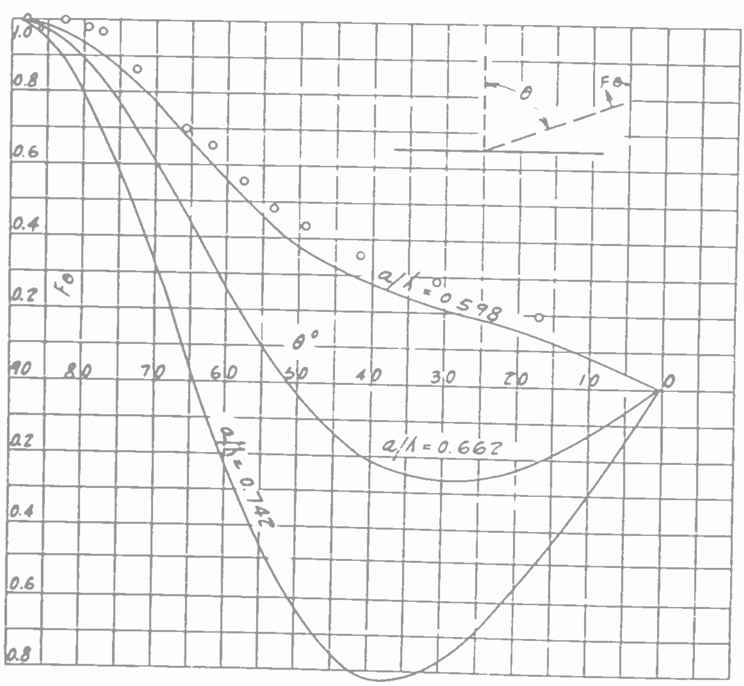

Fig. 12—Vertical radiation characteristics on type A antenna. (The circles are experimental values obtained by airplane measurements at WCAU.)

ratio, a/λ, equaled 0.595 for the WCAU antenna, so that the agreement between the results of the model experiments and the actual airplane measurements would indicate that the effect is due almost entirely to

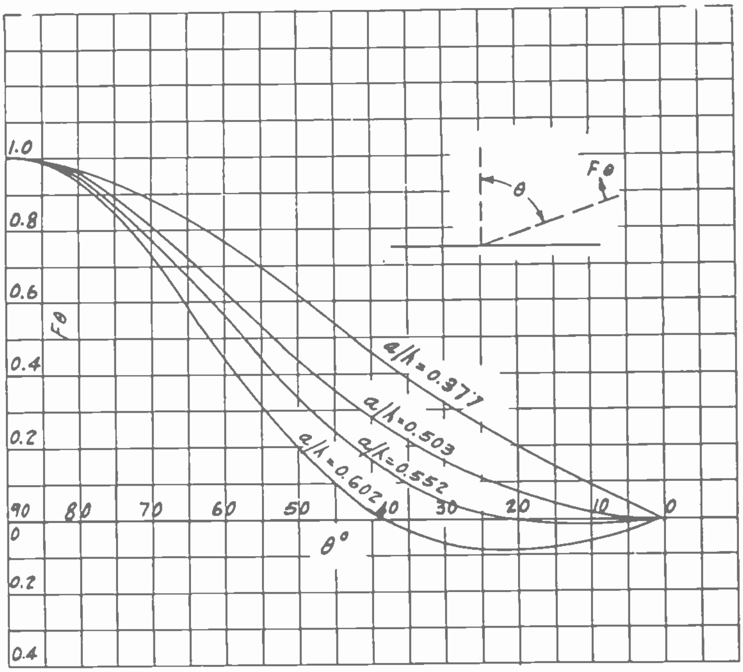

Fig. 13—Vertical radiation characteristics of type B antenna.

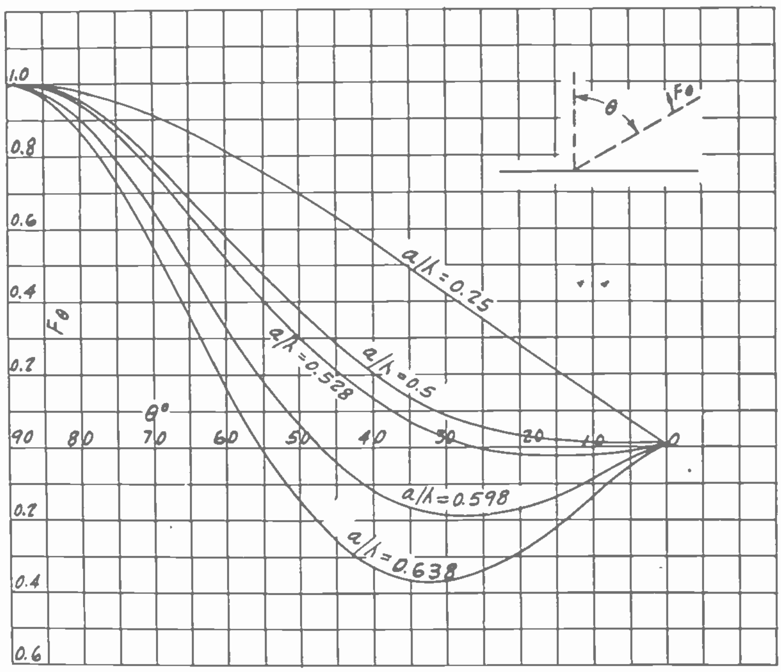

Fig. 14—Vertical radiation characteristics of straight vertical wire.

the fact that the current distribution is nonsinusoidal. Fig. 13 shows the same sort of curves for the type B antenna with the small capacity ring. As a means for comparison, we have included Fig. 14 to show the patterns that might be expected if the current distribution were sinusoidal. These diagrams show that the best height of antenna to use to reduce the sky wave to a minimum depends on whether it is a type A antenna, type B antenna, or an antenna with sinusoidal distribution.

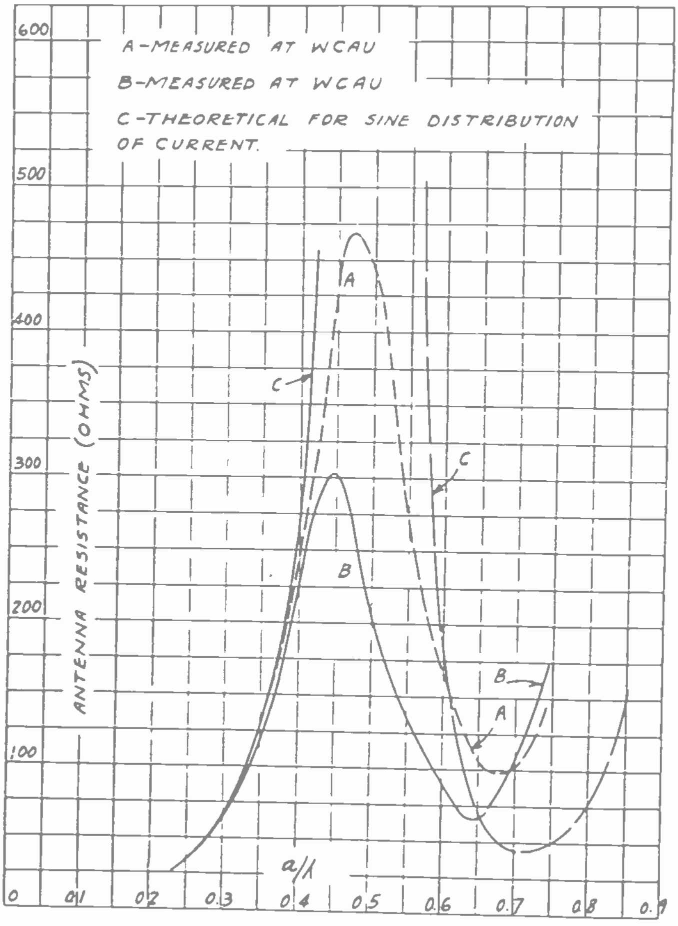

Fig. 15—Radiation resistance vs. a/λ.

IV. Information on Existing Antennas

It is desirable to compare the action of these tower antennas to the theoretical results one might expect if the distribution were sinusoidal. Resistance measurements of the WCAU antenna were made at a number of frequencies over the broadcast band. The results of these measurements are shown by Fig. 15. Curve A shows the resistance of the WCAU antenna. For this measurement, the tower stood entirely alone with no lighting equipment or static drain coils on the antenna. Later, an insulated generator and other equipment was attached to the tower, increasing the base capacity a large amount. This capacity changed the effective series resistance of the antenna system. The resistance values under these operating conditions are shown by curve *B*. Curve *C* is the corresponding theoretical curve of radiation resistance for a sinusoidal distribution of current. It is seen that there is very little agreement between the two in the region, $a/\lambda > 0.5$. It is interesting to note that the maximum value of resistance is only of the order of 400 ohms. This is due to two causes. The first is that the current at the resonant point reverses phase by a rotation rather than by passing through zero. The other is that the true antenna impedance is shunted by the capacitance of the base insulator. This shunting effect has been found to be even greater in other installations. It should be pointed out that the size of this base capacitance plays no part in determining the efficiency of the antenna, if there is no leakage resistance to ground. Since it is likely that these capacitance currents will flow through some earth before getting back to the antenna coupling coil, it is important that a conducting path be provided for these currents. This is best done by placing a screen on top of the earth around the base insulator and bonding this screen directly to the ground system.

Another test made on the WCAU antenna was the measurement of the field strength at one mile as the frequency was varied and the power was held constant. The results are given as curve *A*, Fig. 16. Curve *C* of this same figure is the theoretical curve obtained if the earth is a perfect conductor, and the antenna current is assumed to be sinusoidally distributed. The input power was a constant value, 50,000 watts. There is seen to be a great departure from the theoretical value. The crosses indicate values computed from the distribution obtained on the model antenna, assuming a perfectly conducting earth.[3] These values are in very good agreement with the actual situation and show that the departure from the simple thoery is due to current distribution. Curve *B* of Fig. 16 is a similar curve obtained on KOA, Denver.[4] This is a type B antenna. The circles are values computed from the current distributions shown in Fig. 9. We see that neither type of antenna approaches the theoretical curve through the whole range. If we refer to the vertical radiation patterns at the same time, we see that the type A antenna gives a large high angle radiation by the time it has reached the maximum value of field strength. If the antenna is shortened to reduce the high angle radiation, there is a very appreciable loss in field strength. Besides this, it seems doubtful

[3] When a/λ is large enough that the antenna current undergoes a phase rotation, one is not justified in computing the polar diagram and field intensity without knowing the phase of the current at all points. Our model measurements gave only amplitude.

[4] The authors are indebted to Mr. Raymond F. Guy of the National Broadcasting Company for the data shown by curve *B*.

whether shortening really decreases the high angle radiation in this case. It is seen that there is room for a great deal of improvement on the type A antenna. The type B antenna gives values much closer to the theoretical and seems to suppress the sky wave for values of a/λ up to 0.55. Beyond this point, the intensity curve drops off rapidly. The maximum value of intensity obtained is far short of the theoretical maximum. It is evident that there is much to be gained by using an

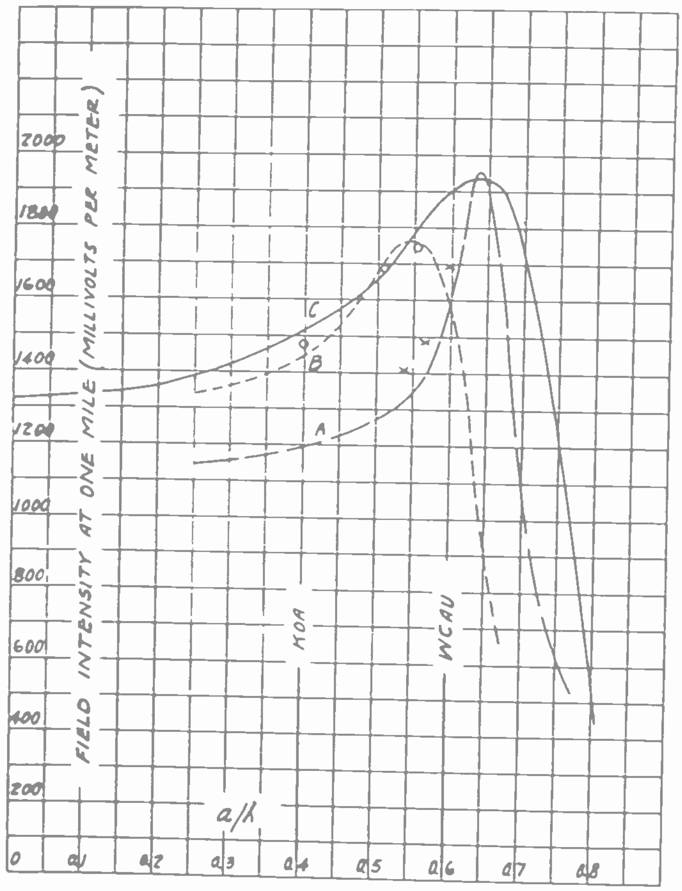

Fig. 16—Field intensity at one mile vs. a/λ. (Input power = 50,000 watts.)

antenna which will have a sinusoidal distribution of current. Such a result might be had if the tower were of uniform cross section for its entire length. This of course involves some difficulties in the design of the taller towers. A method of obtaining a sinusoidal distribution of antenna current will be discussed below.

V. A Method of Attaining a Sinusoidal Distribution of Current on the Type A and B Antennas.

As it has just been pointed out, it is desirable to make the cross section of the antenna be constant. This was accomplished in a very simple fashion. The tests were made on the models as before. The square framework was placed on top of the type A model. From each corner of this framework, a wire was dropped vertically downward to the corners of the largest cross section of the tower, and from there to the corners of a square frame placed at the base of the antenna. The arrangement is shown by the sketch on Fig. 17. The current distribution was measured and found to approximate the sinusoidal much better than when the wires were not used. (Fig. 17.) The departure

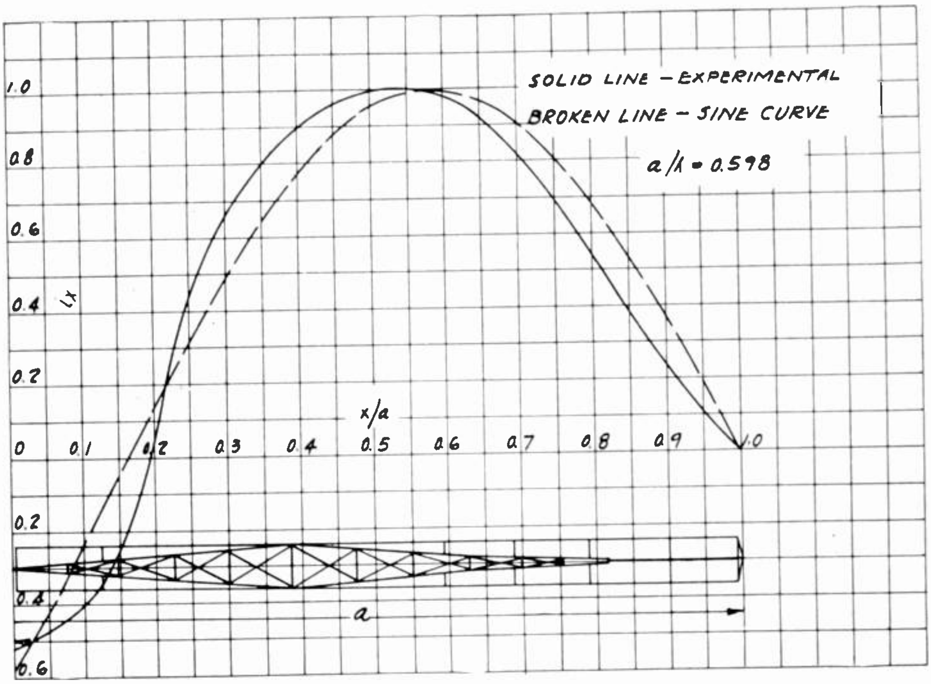

Fig. 17—Current distribution on type A antenna equipped with outrigger and vertical wires.

from the sinusoidal would no doubt be made less by the use of more wires. The minimum of four wires was used to show the great correction that was obtained with the minimum number of wires.

The same procedure was followed with the model of the type B antenna. The small circular ring was placed at the top and four wires were dropped from the diameter of this ring to join the tower a short distance above the base insulators. The current distribution was measured for $a/\lambda = 0.602$. This distribution is shown in Fig. 18. This curve agrees well with the expected sinusoidal distribution. The frequency was shifted so that $a/\lambda = 0.407$, and a series of measurements made. These results are shown by Fig. 19. This curve was so close to a true sine wave that it was not possible to plot a sine wave on Fig. 19 and have it show as a separate curve.

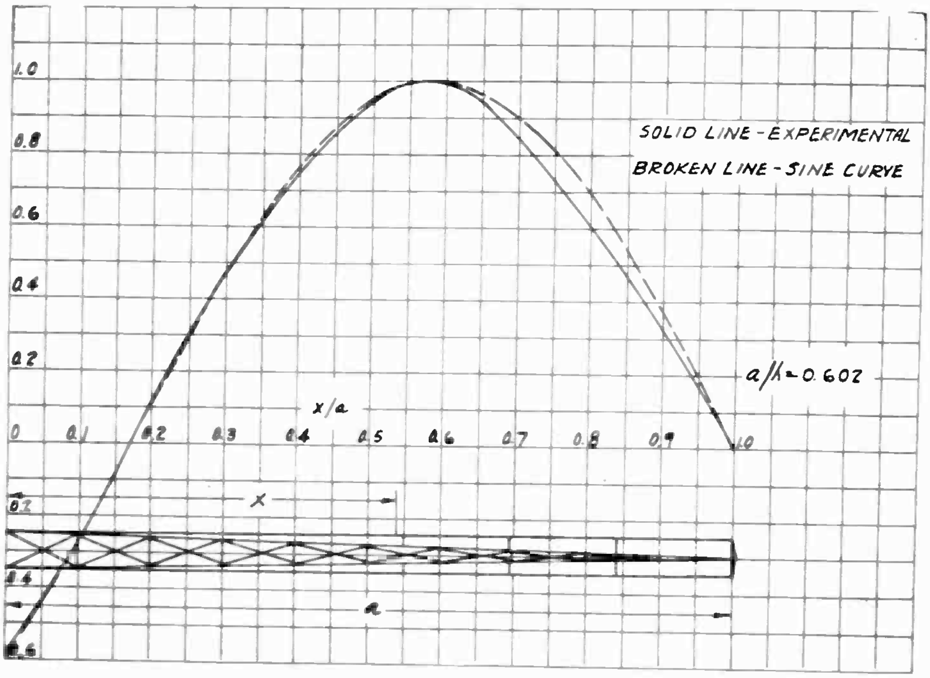

Fig. 18—Current distribution on type B antenna equipped with outrigger and vertical wires ($a/\lambda = 0.602$).

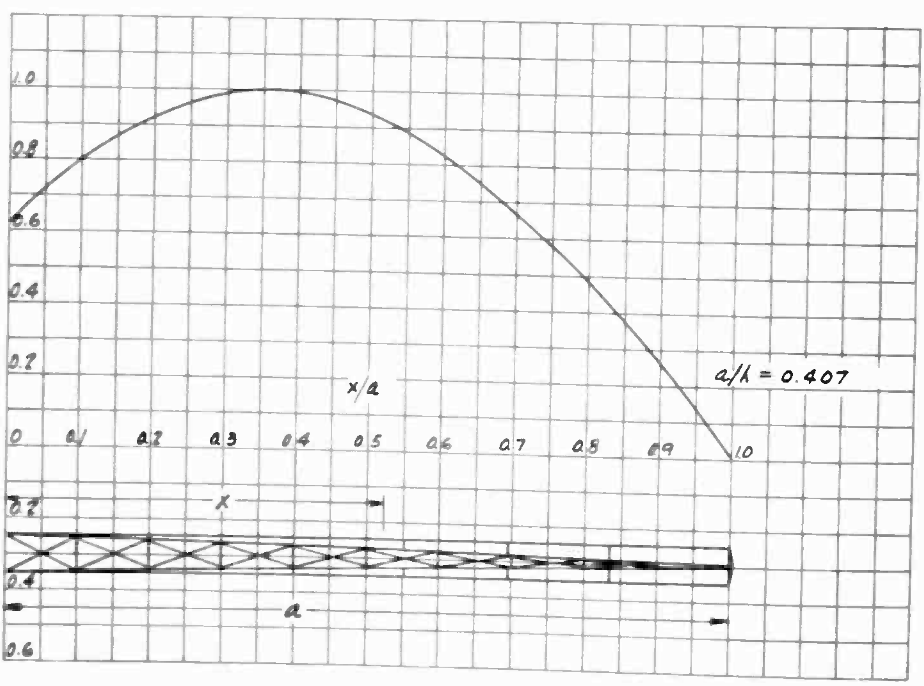

Fig. 19—Current distribution on type B antenna equipped with outrigger and vertical wires ($a/\lambda = 0.407$).

Other arrangements of the wires were tried without correcting the current distribution. For instance, on the type A antenna, wires were placed only from the top framework down to the bulge of the tower. This gave a distribution far from the sinusoidal. Another test was made with wires strung from the bulge to a point near the ground. This was also found to be unsatisfactory.

While it may be argued that there may be other distributions of current than the sinusoidal which will give equal or better results, it seems more desirable to use the sinusoidal distribution. For every particular distribution proposed, it would be necessary to go through a great deal of tedious computation to examine the theoretical results that we might expect. So much of the theory for a sinusoidal distribution has already been worked out that the antenna designer's problem is very much simplified if he can assure himself that the current distribution is sinusoidal.

VI. Earth Currents Near the Antennas

It has been pointed out by other writers that ground systems beneath antennas play a dual rôle. One function of the ground system (usually buried radial wires, with the base of the antenna as the common point) is to provide a good conducting path for the earth currents, so that these currents will not flow through a poorly conducting earth. This is necessary close to the antenna where the earth current densities become high. The other function is to act as a good reflector for waves coming from various points on the antenna, so that the vertical radiation pattern will be close to that obtained if the earth under the antenna were perfectly conducting. Actually these two functions are synonymous. If the ground system is complete and extensive enough to eliminate all power expenditures in the earth, the reflection of each little incident wave will be perfect. We will concern ourselves only with an examination of the earth currents. We will discuss ground systems which consist of buried wires running radially in all directions from the base of the antenna.

To form a picture of what is happening in the earth beneath an antenna, let us examine Fig. 20. Displacement currents leave the antenna, flow through space, and finally flow into the earth where they become conduction currents. If the earth is homogenous, the skin effect phenomena keep the current concentrated near the surface of the earth as it flows back to the antenna. Where there are ground wires present, the earth current consists of two components, part of which flows in the earth itself and the remainder of which flows in the buried wires. As the current flows in toward the antenna, it is continually added to by more displacement currents flowing into the earth. It is not necessarily true that the earth currents will increase because of this additional displacement current, since all these various components differ in phase. In Fig. 20, let C be a cylindrical surface of radius, x, where the vertical antenna and the cylindrical axis are coaxial. Then we will denote the total earth current flowing radially inward across the surface of C as $\bar{I}_x$. If buried wires are present, $\bar{I}_x = \bar{I}_w + \bar{I}_e$, where $\bar{I}_w$ is the component flowing in the wires and $\bar{I}_e$ is the part which actually flows

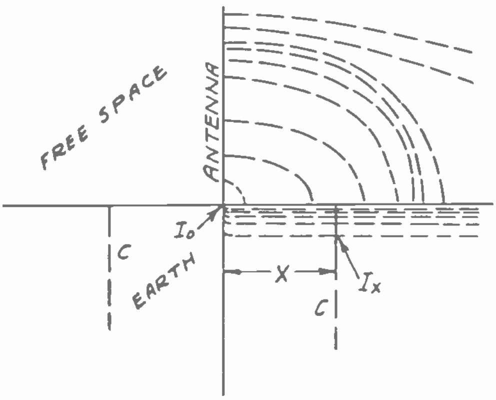

Fig. 20

in the earth. It is interesting to examine the factors which determine the proportions of these two components. An approximate relation is[5]

$$\bar{I}_e/\bar{I}_w = \gamma_e\left[\pi(c^2 - r^2)R_{ac}(1 + j) + j4\pi \cdot 10^{-9}f\left\{\pi c^2 \log\frac{c}{r} - \frac{\pi}{2}(c^2 - r^2)\right\}\right] \tag{1}$$

where,

γ_e = earth conductivity (mhos per centimeter cube)
R_{ac} = radio-frequency resistance per centimeter length of wire used in the ground system
r = radius of the wire in the ground system (centimeters)
f = frequency (cycles per second)
x = distance from the antenna (centimeters)
n = number of equally spaced radial wires
$c = \pi x/n$

The assumptions are that the earth is homogenous, the wires are buried between six and twelve inches deep, and the ends of the wires are well connected to the earth, possibly through deep rods or large plates. Because of the relative size of the quantities involved, (1) is quite accurately given as

$$I_e/I_w = j\gamma_e \cdot 4\pi^2 \cdot 10^{-9}fc^2\left\{\log\frac{c}{r} - 0.5\right\}. \tag{2}$$

[5] G. H. Brown, "A theoretical and experimental investigation of the resistances of radio transmitting antennas," Chapter V. This is an unpublished thesis filed in the University of Wisconsin Library.

Then the current in the wires is expressed in terms of the total earth current as

$$\left| \bar{I}_w/\bar{I}_x \right| = \left| \frac{1}{1 + \bar{I}_e/\bar{I}_w} \right|. \tag{3}$$

Using (2) and (3), we have computed the ratio of the current in the radial wires to the total earth current for 30, 60, and 120 radial wires,

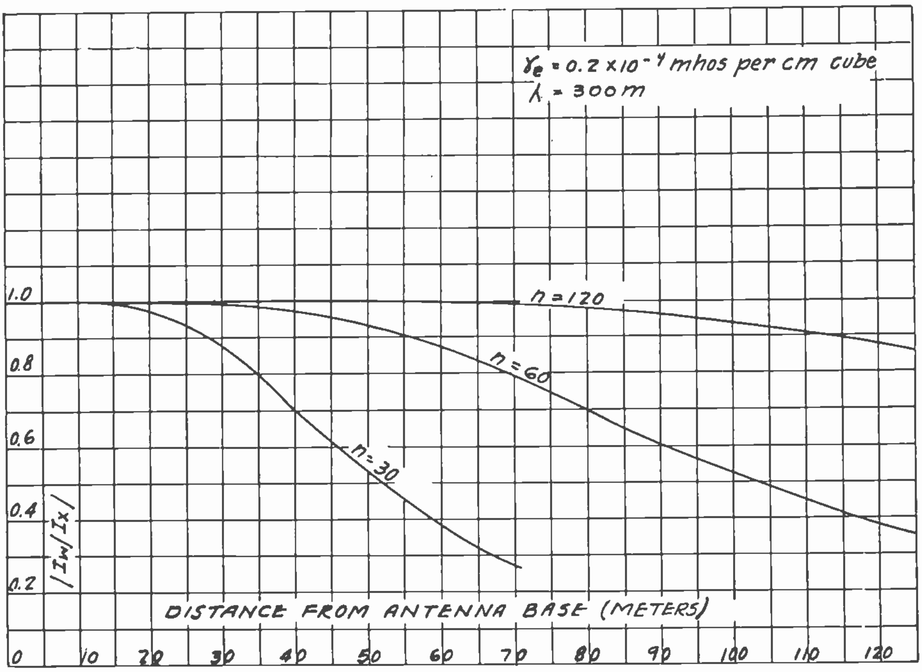

Fig. 21

(No. 8 annealed copper) and a frequency of 1000 kilocycles per second. Three different earth conductivities were chosen. These were:

$\gamma_e = 0.2 \times 10^{-4}$ mhos per cm cube $= 20.0 \times 10^{-15}$ e.m.u.
$\gamma_e = 0.5 \times 10^{-4}$ mhos per cm cube $= 50.0 \times 10^{-15}$ e.m.u.
$\gamma_e = 1.0 \times 10^{-4}$ mhos per cm cube $= 100.0 \times 10^{-15}$ e.m.u.

The results are expressed graphically in Figs. 21, 22, and 23. It is seen that as the conductivity of the earth increases, a greater part of the total current flows in the earth and less in the buried wires. The number of buried wires is also an important factor. These curves show merely the *ratio* of the current in the wires to the total earth current. We will now examine the way in which the total earth current varies.

If it is assumed that the earth below the antenna is a perfect conductor, it is a simple matter to calculate the earth current flowing across any cylinder of radius, x. This current is a function of the distance from the base of the antenna, the wavelength, and the dimensions of the antenna. When the antenna is a straight vertical wire of height,

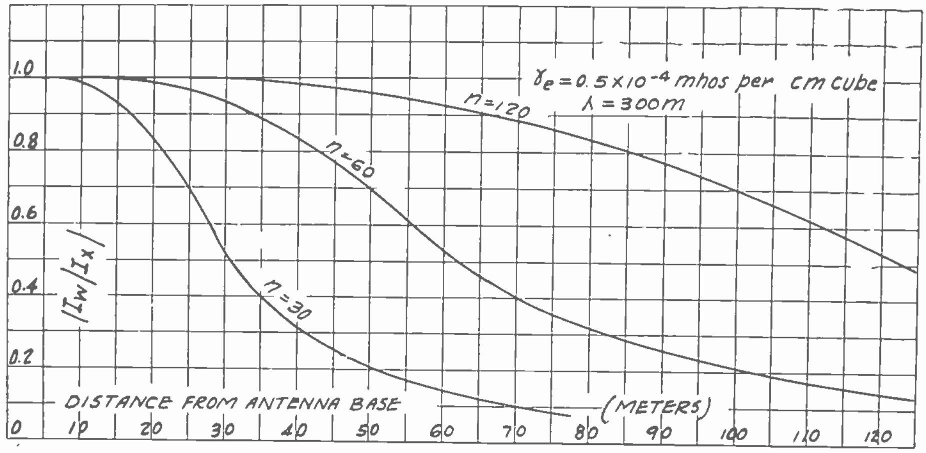

Fig. 22

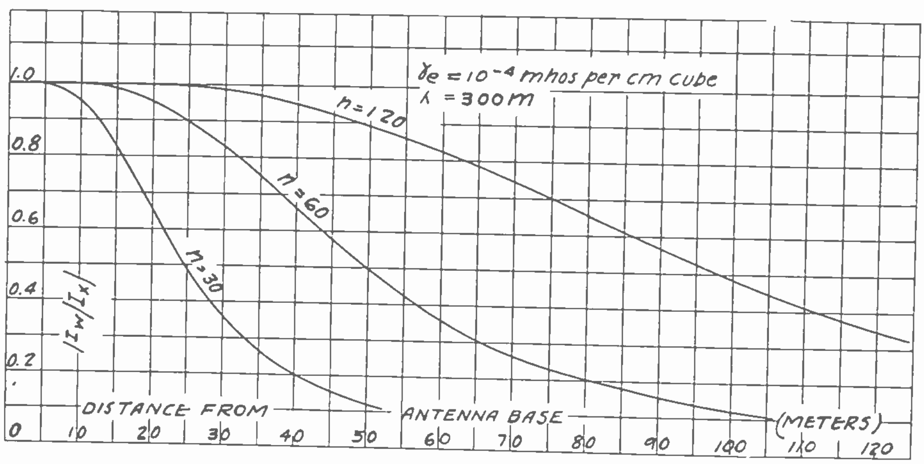

Fig. 23

a, with the current distributed sinusoidally on the antenna, the total earth current is given by[6]

$$\bar{I}_x = j\frac{\bar{I}_0}{\sin G}\left[\epsilon^{-jkr_2} - \cos G\epsilon^{-jkx}\right], \tag{4}$$

where, I_0 = current at the base of the antenna

$r_2 = \sqrt{a^2 + x^2}$

λ = wavelength

$k = 2\pi/\lambda$

$G = ka$.

[6] G. H. Brown, "The phase and magnitude of earth currents near radio transmitting antennas," Proc. I.R.E., vol. 23, no. 2, pp. 168–182; February, (1935).

The absolute value of this earth current is

$$| I_x | = \frac{I_0}{\sin G} \sqrt{1 + \cos^2 G - 2 \cos G \cos k(r_2 - x)}. \quad (5)$$

This equation has been used to compute the earth currents for several different antennas, namely, $G=60$ degrees, $G=90$ degrees, $G=120$ degrees, $G=180$ degrees, $G=215$ degrees, $G=230$ degrees, and $G=270$ degrees. The magnitudes of the earth currents are plotted in Fig. 24.

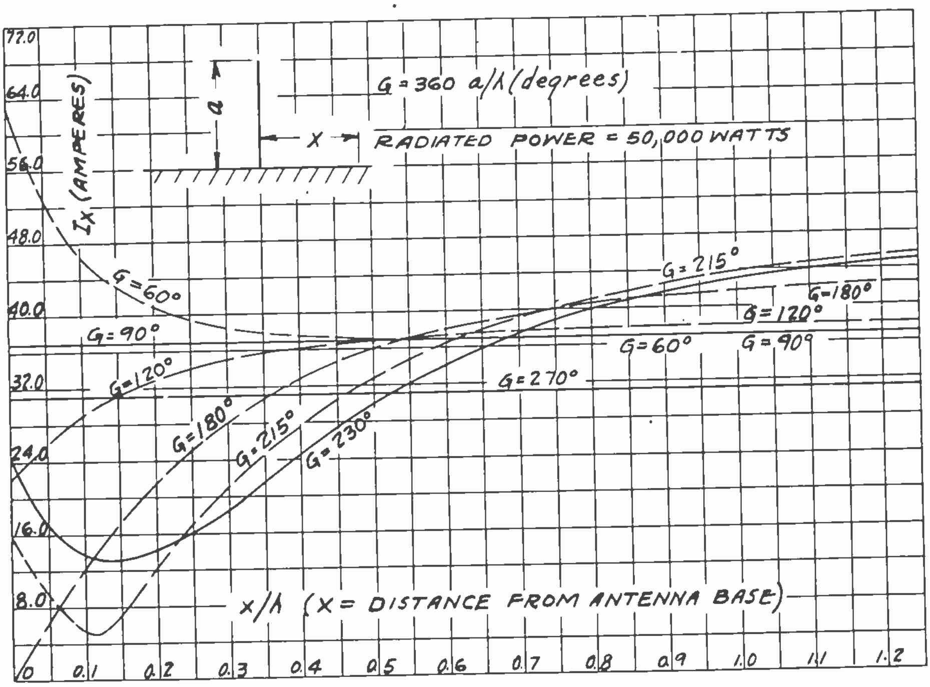

Fig. 24—Earth currents in the neighborhood of transmitting antennas.

It was assumed that 50,000 watts was radiated in each case. It is seen that the earth currents are larger at remote points than the current at the base of the antenna when the antenna is longer than a quarter wavelength. These values were computed on the assumption of a perfect earth, but it is highly probable that the same general law is followed when the earth becomes conducting, at least in the range of frequency used for broadcasting. It has previously been shown that the maximum losses in the earth beneath a half-wave antenna occur in the region about 0.35 of one wavelength from the base of the antenna.[6] A similar treatment of the curves of Fig. 24 shows that this same value, 0.35 of one wavelength, is true for the other antennas whose height is greater than a half wavelength. Thus, an examination of Figs. 21 to 24 shows the importance of using many wired radial ground systems and extending them much farther than is the present practice. We thus see that with a wavelength of 300 meters, it is important to use at least 120 radial wires, whose lengths are at least 130 meters. It is also evident that if only a small number of radial wires is used, there is little point in running them out to great distances, since they will carry little current.

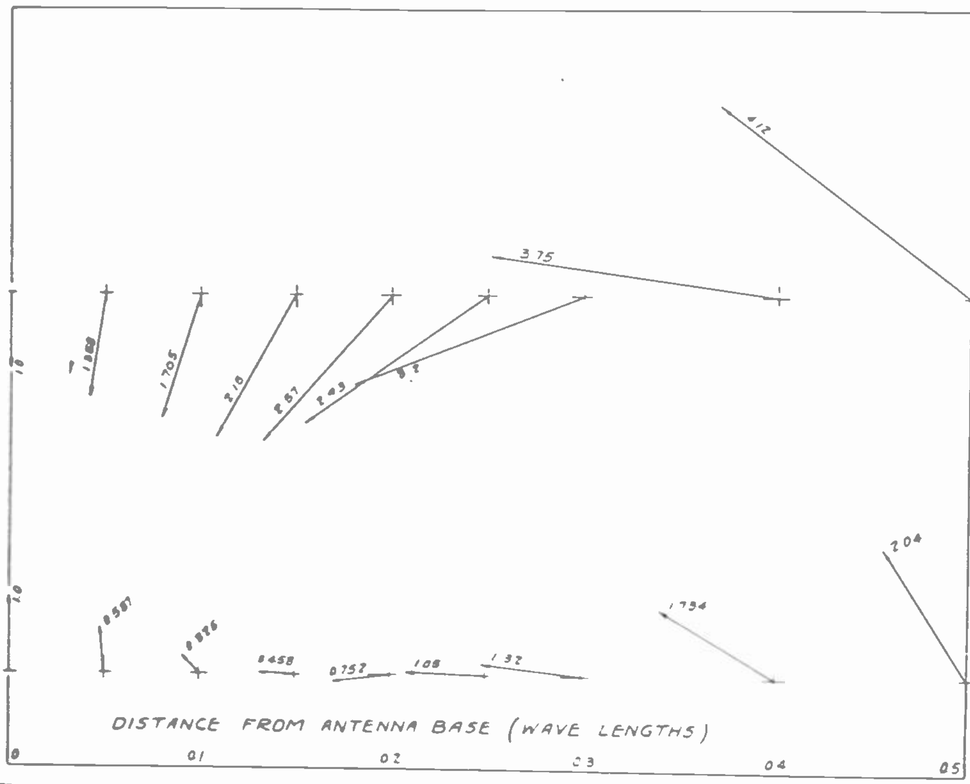

Fig. 25—Phase and magnitude of earth current in the neighborhood of a 215-degree antenna. The top row of vectors shows the earth currents computed from the current distribution of curve *B*, Fig. 4. The bottom row of vectors shows the earth currents when the antenna current is sinusoidally distributed. In each case, the vector at the extreme left is the current at the base of the antenna.

Returning to Fig. 24, we see that in the case of the group of antennas longer than one-half wavelength the earth current drops off very rapidly for a short distance from the antenna and then builds up. It is possible to make use of this peculiar characteristic to determine whether there is a reversal of phase on the antenna. Let us first examine these earth currents in more detail. In Fig. 24, we have shown the magnitude of the earth currents for a number of antennas whose current distribution is sinusoidal. Equation (4) gives not only the magnitude but the phase position of the earth current with respect to the current at the base of the antenna. These alternating-current vectors have been computed from (4) for a 215-degree antenna with a sinusoidal distribution of current on the antenna. The results are shown by the lower row of vectors shown in Fig. 25. We see that as we proceed from the base of the antenna, the earth current advances in phase and decreases in magnitude and then starts a phase lag and increases in magnitude. These vectors were computed on the basis of one ampere into the base of the antenna. This value of current was chosen so that one can easily see the relative magnitudes of current. A similar set of

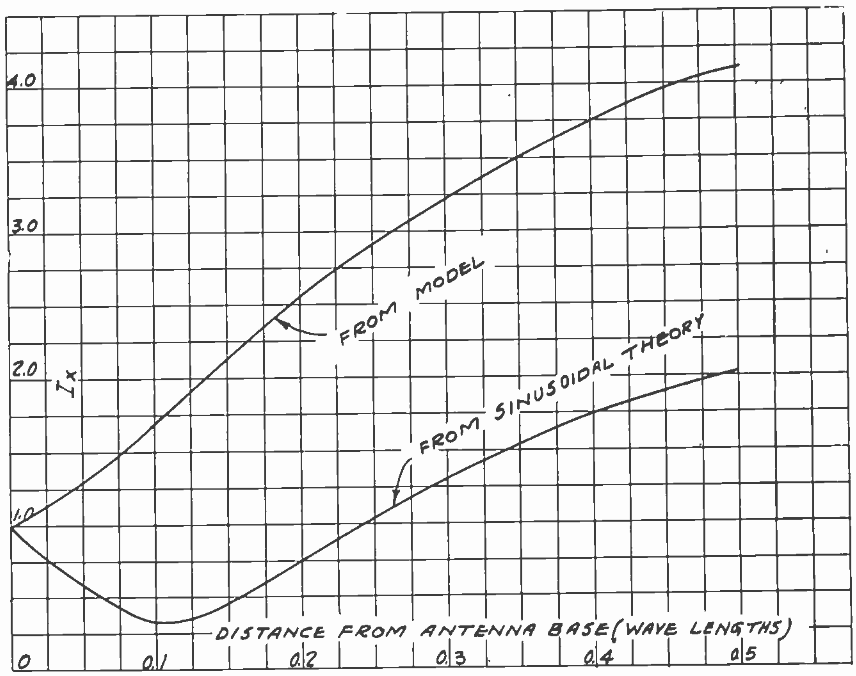

Fig. 26—The magnitude of the vectors shown in Fig. 25 vs. distance from antenna.

vectors was computed using the antenna current distribution on a type A tower which is 215 electrical degrees tall. This antenna current distribution is given by curve *B*, Fig. 4. The earth currents were necessarily computed by a graphical method. The results are shown by the upper row of vectors of Fig. 25. We see that in this case the earth current immediately begins to lag and increase in magnitude as we proceed from the antenna. The magnitude of the vectors is plotted for both cases in Fig. 26. If we could measure the earth current at various distances from the antenna, we could easily tell whether the current on the antenna had a complete phase reversal or not.

Actually we can measure this current in a rather simple fashion. It has been shown[6] that the flux density in space at the surface of the earth is related to the total earth current by the relation

$$2\pi x B = \mu I_x \tag{6}$$

where x is the radius of the cylinder shown in Fig. 20 (centimeters), I_x is the total earth current flowing through the surface of the cylinder (amperes), B is the electromagnetic flux density at a point on the surface of the earth x centimeters from the antenna base, (webers per square centimeter), and μ is the permeability of free space ($4\pi \times 10^{-9}$). It is of course assumed that the earth current is distributed uniformly around the periphery of the cylinder. The measurement of flux density can be made with a loop antenna and a calibrated vacuum tube voltmeter. The loop is untuned so that the vacuum tube voltmeter reads the induced voltage. The induced voltage is related to the electromagnetic flux density by[7]

$$e_i = 2\pi f N A B \tag{7}$$

where,

f = frequency (cycles per second)
N = number of turns on the loop
A = area of the loop (square centimeters)
B = electromagnetic flux density (webers per square centimeters)
e_i = induced loop voltage (volts).

Combining (6) and (7),

$$I_x = \frac{2\pi x e_i}{\mu 2\pi f N A} = \frac{x e_i}{60h}. \tag{8}$$

where $h = 2\pi f N A/c$ is the effective height of the loop measured in centimeters, and c is the velocity of propagation of light (3×10^{10} centimeters per second).

If x is measured in meters, (8) becomes

$$I_x = \frac{x e_i}{0.60h}. \tag{9}$$

If x is measured in feet, (8) is

$$I_x = \frac{x e_i}{1.97h} \tag{10}$$

where h is still expressed in centimeters.

Thus we have a simple method of measuring the earth currents in the vicinity of an antenna. We simply proceed along a radial line measuring the distance from the antenna and reading the vacuum tube voltmeter. Since some antenna effect may exist, it is desirable to re-

[7] Appendix B, equation (16).

verse the loop at each point and take the average of the two readings. Where the ground around the antenna is fairly flat, such a measurement should be of value in determining whether or not the current distribution on the antenna is close to sinusoidal. Such a procedure was carried out in the vicinity of the WCAU antenna, which is a type A tower of height 215 electrical degrees and is the case for which the calculations of Fig. 25 were made. Since the ground was rather rough, the average of a number of readings was taken. The results are shown in Fig. 27. The earth current rises quickly to over sixty amperes at a

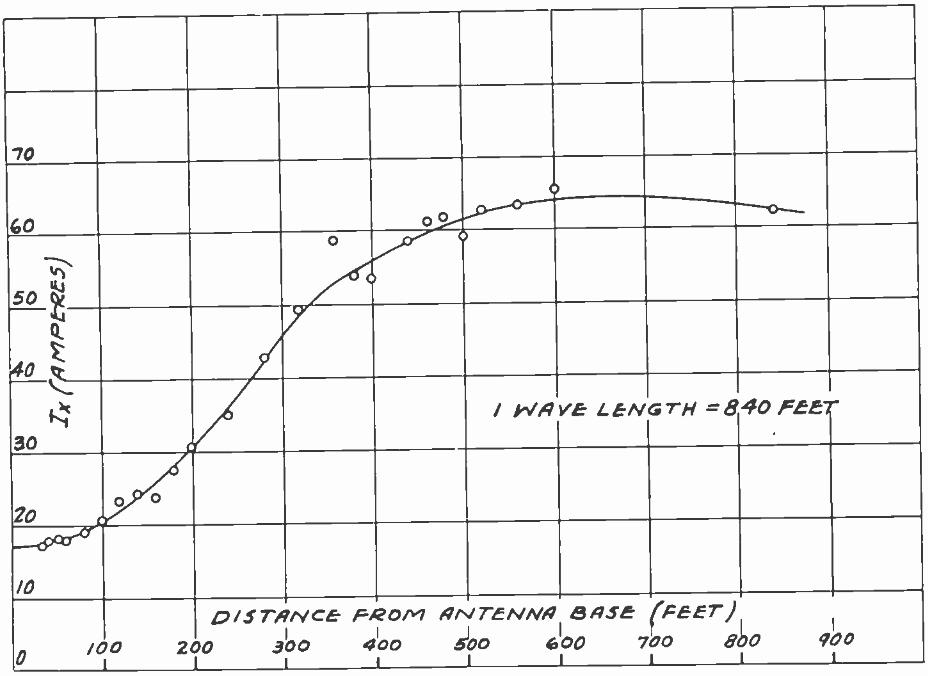

Fig. 27—Measured earth currents in the vicinity of the WCAU antenna.

little beyond one-half wavelength, (420 feet). It has dropped slightly at one wavelength distant from the antenna, (840 feet). A measurement at one mile (6.28 wavelengths) showed that attenuation of the earth had dropped the earth current to 43.2 amperes. These measurements show no tendency to have a dip as predicted for a sinusoidal distribution, and check rather well with the theoretical results arrived at from our tests with the models. It should be noticed that the curve when extrapolated to zero abscissa indicates an antenna current of 17.3 amperes at 50,000 watts. The antenna current meter actually indicated 22.7 amperes. This apparent discrepancy is easily accounted for. The measured antenna resistance with all the lighting apparatus, etc., tied on is 97.0 ohms. Then the input power is $I^2R = 22.7^2 \times 97.0 = 50{,}000$ watts. The resistance measured before any equipment was attached was 165.0 ohms. This is closer to the true antenna resistance.

The true antenna current as shown by Fig. 27 is 17.3 amperes. Then the input power to the antenna proper is $17.3^2 \times 165.0 = 49{,}300$ watts. Thus we see that the true antenna current is masked by the capacity at the base of the antenna. This effect is more completely discussed in Appendix C.

VII. Conclusion

Vertical radiators with varying cross section along the tower depart considerably in their performance from the theoretical antenna with sinusoidal current distribution. The effect on the ground wave for the type A tower is shown on Fig. 16, (compare curves *A* and *C*) and the effect on radiation at vertical angles is shown on Figs. 12 and 14.

It was possible from current distributions obtained on model antennas to calculate the values corresponding to curve *A*, Fig. 16, which are shown on the figure and are in good agreement. It was also possible by methods described to calculate the vertical pattern shown on Fig. 12. Calculated and measured values are in good agreement.

The departure of the type B antenna from theoretical values is shown by curve *B*, Fig. 16. No vertical patterns were obtained by actual airplane measurements on the type B, antenna. However, the agreement obtained in the case of the WCAU antenna (type A) substantiates the method of attack.

The present tower antennas on the whole are unsatisfactory compared to what may be obtained. The cause of the discrepancy lies in the varying cross section of these towers which causes the current distribution to be nonsinusoidal. Experiments on models from which the current distribution was obtained for type A and B towers show that sinusoidal distribution is again approached when the cross section is made uniform. In the case of a single wire, the distribution is sinusoidal as shown in Fig. 11.

The effect of nonsinusoidal distribution in the cases mentioned is highly detrimental since the horizontal radiation is less and the skyward radiation more than for sinusoidal distributions, which exactly counteracts the effect that was sought in the use of such radiators. Hence, in order to approach the theoretical values for sinusoidal distribution, it is necessary to have a sinusoidal distribution of current on the tower which, as shown by the model experiments, can be obtained only by a tower of uniform cross section. Other expedients such as capacity crowns help slightly, as shown in Figs. 9 and 10, but there are further reasons for making the distribution absolutely sinusoidal.

The chief limitation on the primary night service area for cleared channel stations having sufficient power to lay down a serviceable signal from 60 to 100 miles is fading. In order to extend the service area, the fading zone must be located further from the transmitter. This can be accomplished only by reducing the sky-wave radiation. Since it is expensive to experiment with radio antennas at broadcast frequencies, it is necessary to be able to predict what will happen before the antenna is built. Such predictions can only be accurately made when the designer is assured of sinusoidal distribution, which in turn means that the tower must be of uniform cross section.

The most important phase of antenna design is sky-wave elimination since this is apparently the most serious limitation of the night service area. The increase in ground-wave signal is important but must be sacrificed for sky-wave elimination.

The optimum electrical length of 230 degrees, which gives a theoretical 41 per cent increase in horizontal radiation as compared to a 90-degree antenna, would cause very serious fading. Hence, 190-degree antennas are now used, giving a 27 per cent increase in ground wave as compared to a 90-degree antenna. A 190-degree antenna with sinusoidal distribution will increase the distance to the fading zone considerably.

It is desirable to have some quantitative guide as to what effect a change in the output of the antenna has on the day service radius. In van der Pol's discussion of the Sommerfeld propagation theory,[8] it is pointed out that for a numerical distance of more than twenty the field strength drops off very nearly as the inverse square of the distance. Under average conditions in the broadcast band, the numerical distance is more than twenty at the edge of the service area, so that in this zone, the service radius may be said to vary as the square root of the magnitude of the field intensity measured a few wavelengths from the antenna. Hence, the use of a 190-degree antenna instead of a 215-degree antenna will give a service radius of $\sqrt{1.27/1.39}$ or 96 per cent of that obtained with a 215-degree antenna. On the other hand, the service radius of a 190-degree antenna is $\sqrt{1.27/1}=1.13$ times greater than that of a 90-degree antenna, or an increase of 13 per cent. Even the 230-degree antenna which gives the greatest intensity along the ground yields an increase in service radius as compared to a 90-degree antenna of only 19 per cent.

The above is given in an attempt to show that the increase in service radius is not so much the important factor in these antennas as the reduction in sky wave. In the broadcast band, the optimum antenna for the reduction of sky wave at distances from 60 to 120 miles is the 190-degree antenna.

[8] Balth. van der Pol, "The propagation of electromagnetic waves," *Zeit. für Hochfrequenz.*, vol. 37, April, (1931).

The present practice in the use of ground systems is to have too few radials of insufficient length. Since for antennas of 180 degrees and over, the maximum loss occurs in the vicinity of 0.35λ from the antenna, the ground system should at least extend beyond this point, preferably to 0.5λ. The effect of not using enough radials is shown on Fig. 23. If a small number of radials are used, the ratio of the current in the wires to the total ground current drops rapidly, thus increasing the losses. At broadcast frequencies, for average conditions, at least 100 to 120 radials should be used. The wires need not be buried more than six to twelve inches. The wires, however, should be in the earth and not lying on the surface.

In Appendix C is given a discussion of the effect of base capacity. It is shown that under the worst conditions in average practice, the power loss due to base insulator capacity is of the order of one to two per cent. The effect of this loss on the field intensity is the square root of this amount and the decrease in service radius is the square root of this latter figure, so that the effect of the losses in the base insulator capacitance can be said to be entirely negligible. It is true that the resistance and reactance values measured at the base of the tower will be different with and without capacity, the reason for which is obvious from a study of the results given in Appendix C. There is, however, no effect on the current distribution and hence no effect on the properties of the tower as a radiator. It is assumed that the only loss is due to the resistance component in the porcelain insulators. The capacity currents introduce no loss if they are properly conducted. This can be accomplished in practice by placing a large metal screen or mat on the surface of the earth below the antenna and tying this mat directly to the ground system.

Owing to the difficulty of making measurements on a radio tower, much of their performance has been shrouded in mystery. The performance characteristics that are of value are current distribution on the antenna, vertical radiation pattern, and field intensity at one mile versus frequency for a constant power input. Current distribution on a large radio tower is almost impossible to obtain, vertical patterns are obtained only at great difficulty by means of airplane measurements, while the field strength on the ground as a function of frequency is easily obtained. This latter measurement can be made by varying the frequency, keeping I^2R into the antenna a constant, and measuring the field strength at a fixed point about one mile distant. This can be done with very small power. Unless the amount of base capacitance is known, resistance and reactance measurements are of little value. Static capacitance measurements are a total loss as far as gauging the performance of the tower as a radiator is concerned. Measurement of the fundamental frequency also tells nothing about the performance since the resonance point of an antenna does not fall at 90 degrees. It is usually in the vicinity of 80 to 88 degrees operation, and closer to 80 degrees for an antenna of large cross section.

Definite indications can be obtained which will show whether or not the current distribution is sinusoidal. These are:

1. By making magnetic flux density measurements as described in Part VI, and calculating the total earth current for a number of points up to one wavelength. Points should be spaced about 0.025λ up to about 0.3λ, 0.05λ up to 0.5λ, and 0.1λ up to one wavelength. This current can be plotted and compared to the curves of Fig. 24. For an antenna over 180 degrees, the earth current drops off rapidly, then begins to rise, and soon approaches a steady value. For antennas between 90 and 180 degrees, the current rises to a steady value. For antennas of less than 90 degrees, the current starts at a high value and comes down to a steady value. The general shape of the measured curve and the curve in Fig. 24 should be the same for an antenna of given electrical height if the antenna current distribution is sinusoidal. The value at the base of the antenna, zero distance on the curve, will correspond to the true value of the antenna current. This may differ from the antenna current meter reading because of the effect of base insulator capacitance.

2. A field strength versus frequency curve from $a/\lambda = 0.25$ to $a/\lambda = 0.75$ as described should have the same shape as the calculated curve C shown on Fig. 16, with a peak at 230 degrees if the current distribution is sinusoidal.

3. If the equipment is available, a measurement of the vertical characteristic can be made. This should agree closely with the proper curve of Fig. 14.

Thus there are three indications from which a determination can be made as to whether the current distribution is sinusoidal. If these indications show that the current distribution is not sinusoidal, the only recourse is to build a model and measure the current distribution as described in Part II to determine the performance of the antenna.

Referring to Fig. 4, curves A and B, it can be seen that the effect of guys is not as detrimental as one might expect. Very little information is available concerning the effect of guys on the antenna characteristics. Experiments can be carried out by the method of models to determine the difference in performance, if any, between guyed and unguyed towers. It is interesting to note that in the flux density measurements close to the WCAU antenna, no important difference could be found in the readings taken along the radials running out close to the guy wires and those radials remote from the guy wires.

Appendix A

Method of Computing Field Intensity and Radiation Resistance of a Vertical Antenna Over a Perfect Earth, When the Antenna Current Distribution is Nonsinusoidal

To determine the probable action of an antenna array, it is desirable to calculate the radiation pattern and the radiation resistance. Many authors have treated the case where the current distribution on the antenna is sinusoidal. We will outline a method of determining these quantities when the current distribution on the antenna is known but is not sinusoidal.

Let us consider the element of current shown in Fig. 28. The current in the element is assumed to be sinusoidally varying with time with a frequency of f cycles per second. The length of the element is an infinitesimal quantity, dy. Then it can be shown that, at a point P several wavelengths removed from the current element, the electric field intensity lies in the plane formed by the axis of the current element and the radius vector, r, and this intensity is normal to r (Fig. 28). Thus, if θ is the angle between the current axis and r, the electric intensity will point in the direction of increasing θ. This electric intensity is given by

$$dF_\theta = +j30k\frac{i_y}{r}\epsilon^{-jkr}dy\sin\theta \qquad (1)$$

where,

i_y = r-m-s current in the element (amperes)
dy = length of element (centimeters)
λ = wavelength of radiated wave (centimeters)
r = distance from current element to remote point, P (centimeters)
$k = 2\pi/\lambda$
$\epsilon^{-jkr} = \cos(kr) - j\sin(kr) = \angle -kr$
$j = \sqrt{-1}$
θ = angle between dy and r
dF_θ = electric field intensity due to current element (volts/centimeter).

Let us now consider the situation when the current element is a part of the antenna shown in Fig. 29. The antenna is a vertical wire with its lower end adjacent to a perfectly conducting earth. The length of the antenna is designated by a. The current along the antenna is some known function of y, the distance from the ground.

From (1), the electric field at point P due to a current element y centimeters from the earth is

$$dF_{\theta,1} = +j30k\frac{i_y}{r_1}\epsilon^{-jkr_1}dy \sin\theta_1 \tag{2}$$

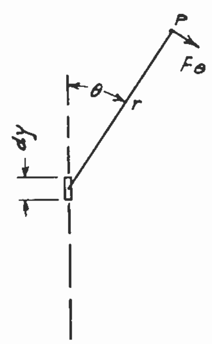

Fig. 28

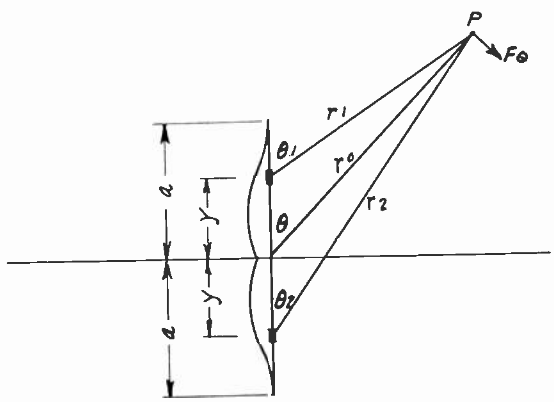

Fig. 29

and the contribution due to the image of this element is

$$dF_{\theta,2} = +j30k\frac{i_y}{r_2}\epsilon^{-jkr_2}dy \sin\theta_2. \tag{3}$$

It is assumed that P is sufficiently remote that

$$\left.\begin{aligned} \theta_1 &\doteq \theta_2 \doteq \theta \\ \frac{1}{r_1} &\doteq \frac{1}{r_2} \doteq \frac{1}{r_0} \\ r_1 &\doteq r_0 - y\cos\theta \\ r_2 &\doteq r_0 + y\cos\theta \end{aligned}\right\}. \tag{4}$$

Then the addition of (2) and (3) yields

$$dF_\theta = +j60\frac{ki_y}{r_0}\cos(ky\cos\theta)dy\sin\theta\epsilon^{-jkr_0}. \tag{5}$$

The total field due to the entire antenna is obtained by integrating over the antenna.

$$F_\theta = +j60\frac{k\epsilon^{-jkr_0}}{r_0}\sin\theta\int_{y=0}^{y=a} i_y\cos(ky\cos\theta)dy. \tag{6}$$

If the current distribution is an analytic function of y, the electric intensity is given by integrating (6). If the equation of the distribution is unknown but the distribution is given by a plotted curve (obtained by experiment in the case of the tower antennas) two procedures are possible. One is to try to find an analytic function which will fit the current distribution curve. This is generally not very fruitful. The other procedure is to plot the integrand and measure the included area. At best, the method is rather tedious.

For convenience, let us write

$$F_\theta = +j\frac{60}{r_0}Kf(\theta)\bar{I}_0\epsilon^{-jkr_0}. \tag{7}$$

where $\bar{I}_0$ = current at the base of the antenna. We will call K the *form factor* of the antenna and $f(\theta)$ the *vertical radiation characteristic.*

$$K = \left[k\sin\theta\int_{y=0}^{y=a}\frac{i_y}{I_0}\cos(ky\cos\theta)dy\right]_{\theta=90^\circ}$$

$$= \left[k\int_{y=0}^{y=a}i_y\frac{dy}{I_0}\right] \tag{8}$$

and,

$$f(\theta) = \left[k\sin\theta\int_{y=0}^{y=a}\frac{i_y}{I_0}\cos(ky\cos\theta)dy\right]/K. \tag{9}$$

It should be noted that $f(\theta)$ is unity when θ is 90 degrees.

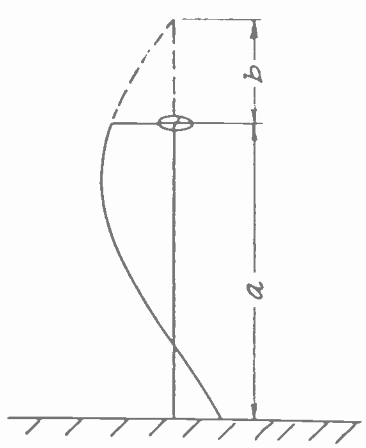

Fig. 30

It is of interest to show the form taken by K and $f(\theta)$ when the current distribution is sinusoidal. Suppose that the antenna has a non-radiating capacity area at the top so that the current distribution is as shown in Fig. 30. Then b is the length of the portion of sine wave suppressed by the capacity area.

Then we define the quantities

$$B = 360b/\lambda \text{ (degrees)}$$

$$A = 360a/\lambda \text{ (degrees)}$$

$$G = A + B.$$

The current distribution is

$$i_y = \frac{I_0 \sin (G - ky)}{\sin G}. \tag{10}$$

Then,

$$K = \frac{\cos B - \cos G}{\sin G} \tag{11}$$

and,

$$f(\theta) = \frac{\cos B \cos (A \cos \theta) - \cos \theta \sin B \sin (A \cos \theta) - \cos G}{\sin \theta [\cos B - \cos G]}. \tag{12}$$

When there is no capacity at the top of the antenna, the current at the top is zero and $B=0$, $G=A$, so

$$K = \frac{1 - \cos G}{\sin G} \tag{13}$$

and,

$$f(\theta) = \frac{\cos (G \cos \theta) - \cos G}{\sin \theta (1 - \cos G)}. \tag{14}$$

At the point on the earth ($\theta=90$ degrees) a distance r from the antenna

$$| F_{90°} | = \frac{60 I_0}{r} \frac{(\cos B - \cos G)}{\sin G} \text{ (volts/centimeter)}. \tag{15}$$

In (15), when I_0 is measured in amperes and r in centimeters, the electric intensity is given in volts per centimeter. These units are not particularly convenient. A more suitable expression is

$$| F_{90°} | = \frac{37.25 I_0}{r} \frac{(\cos B - \cos G)}{\sin G} \text{ (millivolts/meter)} \tag{16}$$

where now r is expressed in miles and the electric intensity is given in millivolts per meter.

The current at the base of the antenna depends on the input power and the antenna resistance thus

$$I_0 = \sqrt{\frac{P}{R_a}}. \tag{17}$$

The antenna resistance is made up of the loss resistance and the radiation resistance. In a well-designed antenna, the radiation resistance predominates.

The calculation of radiation resistance is a fairly simple matter when the current distribution is sinusoidal and the results are well known. When the current is sinusoidally distributed as shown in Fig. 30, the radiation resistance is[9]

$$\begin{aligned} R_r(\text{ohms}) = \frac{30}{\sin^2 G}\Bigg[& \sin^2 B\left\{\frac{\sin(2A)}{2A} - 1\right\} \\ & - \frac{\cos(2G)}{2}\{C + \log(4A) - Ci(4A)\} \\ & + \{1 + \cos(2G)\}\{C + \log(2A) - Ci(2A)\} \\ & + \sin(2G)\left\{\frac{Si(4A)}{2} - Si(2A)\right\}\Bigg] \end{aligned} \tag{18}$$

where A, B, and G are the quantities defined previously and expressed in radians. $C = 0.57721+$ is Euler's constant and $Ci(x)$ and $Si(x)$ are respectively the integral-cosine and the integral sine as defined on page 19 of the Jahnke-Emde "Funktionentafeln."

When the current distribution is nonsinusoidal, one most resort to other tactics. A simple graphical method will be outlined. It can be shown that the total power radiated through the surface of a hemisphere of radius, r_0, and whose center is at the base of the antenna is

$$P_r(\text{watts}) = \int_{\theta=0}^{\theta=\pi/2} \frac{F_\theta^2 r_0^2 \sin\theta}{60}\, d\theta. \tag{19}$$

Substituting (7) in (19)

$$\begin{aligned} P_r &= \int_{\theta=0}^{\theta=\pi/2} 60K^2 f^2(\theta) I_0^2 \sin\theta d\theta \\ &= 60K^2 I_0^2 \int_{\theta=0}^{\theta=\pi/2} f^2(\theta) \sin\theta d\theta = I_0^2 R_r. \end{aligned} \tag{20}$$

The integral of (20) may be integrated by plotting $f^2(\theta)\cdot\sin\theta$ against θ in rectangular coördinates and determining the included area. This operation requires squaring the quantity $f(\theta)$ and multiplying by $\sin\theta$.

[9] Balth. van der Pol, Jr., *Jahrbuch d. drahtl. Telegr.*, vol. 13, p. 217, (1918).

We have used a slight variation from this procedure. It first requires the preparation of a new type of graph paper. It is a slightly modified polar paper (Fig. 31). The angular coördinates are still radial lines. The magnitude coördinates are no longer circles, but are circles multiplied by $\sqrt{\sin\theta}$. We merely plot $f(\theta)$ for any antenna on this paper. The area lying under the resultant diagram is then proportional to the

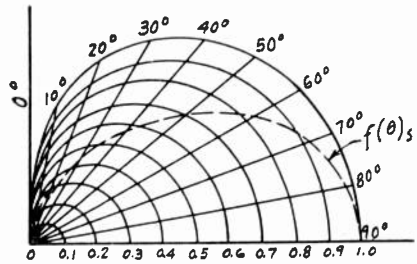

Fig. 31—Radiation polar paper.

integral, $\int_{\theta=0}^{\theta=\pi/2} f^2(\theta)\sin\theta\, d\theta$. For reference, we also plot an $f(\theta)$ curve for a quarter-wave antenna. We will use the quarter-wave antenna as a standard. Then

$$f(\theta)_s = \frac{\cos(90°\cos\theta)}{\sin\theta}. \tag{21}$$

This curve is shown on Fig. 31 and its included area is designated as A_s. A_0 is the area of the antenna in question. Then the power radiated from this particular antenna is

$$I_0^2R_r = 60K^2I_0^2A_0 \tag{22}$$

and the power radiated by the standard quarter-wave antenna is

$$I_s^2R_s = 60I_s^2A_s. \tag{23}$$

Dividing (22) by (23),

$$R_r = K^2\frac{A_0}{A_s}R_s \tag{24}$$

where $R_s = 36.6$ ohms. Thus to determine the radiation resistance of any given antenna whose form factor and vertical radiation pattern we know, we merely plot the vertical radiation pattern on the graph paper of Fig. 31, planimeter the area of this diagram and the standard area and substitute in (24).

The ratio of the intensity at the horizon to the intensity due to a quarter-wave antenna radiating the same power is given by

$$F_0/F_s = \sqrt{\frac{A_s}{A_0}}. \tag{25}$$

Appendix B

Magnetic Flux Density Measurements with a Loop Antenna

The ordinary field intensity measuring set makes use of a loop antenna. This device inherently measures the magnetic flux density of a radiated field, and really yields the electric intensity by virtue of the fact that at remote points from the source of radiation the magnitudes of the electric vector and the magnetic flux density vector are related in a constant ratio. This simple relation no longer holds when the measurement is made close to the source of radiation. In the usual calibration of a loop antenna, the induced voltage is taken as

$$e_i = Fh \tag{1}$$

where F is the vertical electric intensity, and h is the effective height of the antenna in centimeters, given by the equation

$$h = \frac{2\pi f N A}{c} \text{ (centimeters)} \tag{2}$$

where,

f = frequency in cycles per second
N = number of turns on the loop
A = area of loop (square centimeters)
$c = 3\times 10^{10}$ centimeters per second = velocity of propagation in free space.

Actually, the induced voltage is given by

$$e_i = 2\pi f NAB = chB \tag{3}$$

where B is the magnetic flux density measured in webers per square centimeter. At a point remote from the transmitting antenna,

$$|\overline{F}| = c \cdot |\overline{B}| \text{ (volts/centimeters)} \tag{4}$$

and (1) may be used to measure the magnitude of the electric intensity, F. At any point where (4) does not hold, we can only measure the magnetic flux density as given by (3), and obtain a false value of F by using (1).

To illustrate the point, let us examine the case of a small loop antenna placed in the vicinity of a quarter-wave transmitting antenna. (Fig. 32). The loop is placed at the surface of the earth, which is assumed to be perfectly conducting. The plane of the loop is placed in the plane determined by the axis of the transmitting antenna and the center of the loop. H is the actual height of the loop, while W is its width. We will find the voltage induced in the loop.

At any point, P in space (Fig. 32) the components of electric intensity are

$$\bar{F}_z = -j30\bar{I}_0\left[\frac{\epsilon^{-jkr_2}}{r_2}+\frac{\epsilon^{-jkr_1}}{r_1}\right] \tag{5}$$

and,

$$\bar{F}_x = +j30\bar{I}_0\left[\frac{\epsilon^{-jkr_2}}{r_2}\,\frac{(z-a)}{x}+\frac{(a+z)}{x}\,\frac{\epsilon^{-jkr_1}}{r_1}\right] \tag{6}$$

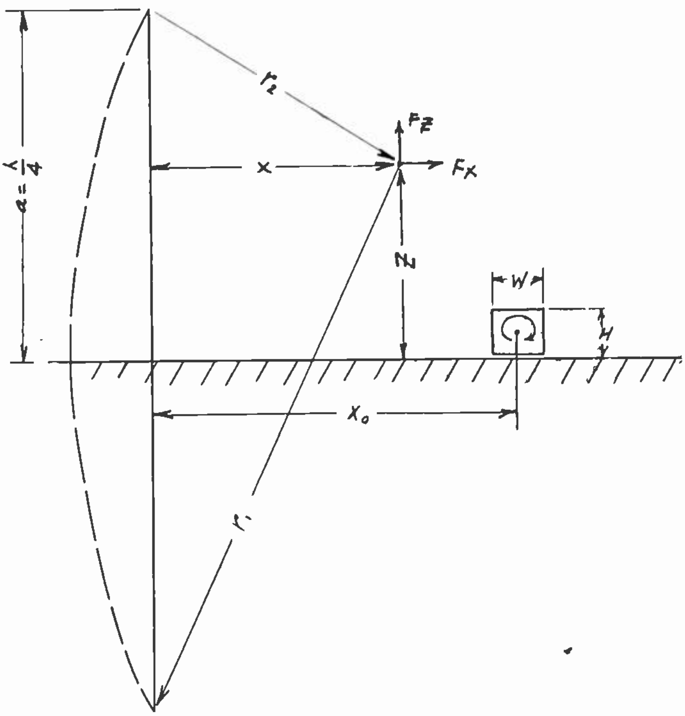

Fig. 32

where, $a=\lambda/4$, the height of the transmitting antenna,

$r_2=\sqrt{(a-z)^2+x^2}$

$r_1=\sqrt{(a+z)^2+x^2}$

$\bar{I}_0=$ current at the base of the transmitting antenna

$k=2\pi/\lambda$.

Then the voltage induced in the vertical side of the loop nearest the transmitting antenna is

$$e_1 = -j60\bar{I}_0NH\left[\frac{\epsilon^{-jkr_0+j(kWx_0/2r_0)}}{r_0-\dfrac{Wx_0}{2r_0}}\right] \tag{7}$$

where $r_0=\sqrt{a^2+x_0^2}$

and, if the dimensions of the loop are small compared to a, the following approximation holds true,

$$r_2 = r_1 = \sqrt{a^2 + \left(x_0 - \frac{W}{2}\right)^2} = \sqrt{a^2 + x_0^2 - x_0 W + \frac{W^2}{4}}$$

$$\doteq r_0 \sqrt{1 - \frac{W x_0}{r_0^2}} \doteq r_0 - \frac{W x_0}{2 r_0}.$$

Under a similar approximation, the voltage induced in the opposite vertical side is

$$e_2 = + j60\bar{I}_0 NH \left[\frac{\epsilon^{-jkr_0 - j(kWx_0/2r_0)}}{r_0 + \frac{Wx_0}{2r_0}} \right]. \tag{8}$$

In (8), the sign has been reversed so that e_2 has the same direction around the loop as e_1. Adding (7) and (8), we find

$$e_1 + e_2 = - j60\bar{I}_0 \frac{HWN}{r_0^2} \epsilon^{-jkr_0} \left[\frac{x_0}{r_0} + jkx_0 \right] \tag{9}$$

where it has been observed that

$$\cos\left(\frac{kWx_0}{2r_0}\right) \doteq 1$$

and,

$$\sin\left(\frac{kWx_0}{2r_0}\right) \doteq \frac{kWx_0}{2r_0}.$$

The voltage induced in the horizontal section of the loop most remote from the ground is found from (6) by placing $z = H$, and noting that

$$r_2 = \sqrt{(a - H)^2 + x_0^2} \doteq r_0 - \frac{aH}{r_0}$$

and,

$$r_1 = \sqrt{(a + H)^2 + x_0^2} \doteq r_0 + \frac{aH}{r_0}.$$

This voltage is

$$e_3 = + j30\bar{I}_0 WN \left[\frac{(H - a)}{x_0} \frac{\epsilon^{-jkr_0 + j(kaH/r_0)}}{\left(r_0 - \frac{aH}{r_0}\right)} + \frac{(a + H)}{x_0} \frac{\epsilon^{-jkr_0 - j(kaH/r_0)}}{\left(r_0 + \frac{aH}{r_0}\right)} \right]. \tag{10}$$

If,

$$\cos\left(\frac{kaH}{r_0}\right) \doteq 1$$

and,

$$\sin\left(\frac{kaH}{r_0}\right) \doteq \frac{kaH}{r_0},$$

(10) becomes

$$e_3 = +j60\bar{I}_0WHN\frac{\epsilon^{-jkr_0}}{r_0^2}\left[\frac{x_0}{r_0} - j\frac{ka^2}{x_0}\right]. \tag{11}$$

Since the bottom horizontal section of the loop is at the surface of the earth, the voltage induced in this section is zero. This may be seen by setting $z=0$ in (6).

Then the total induced voltage is obtained by adding (9) and (11),

$$\begin{aligned} e_i &= +60\bar{I}_0HWN\frac{\epsilon^{-jkr_0}}{r_0^2}\left[kx_0 + \frac{ka^2}{x_0}\right] \\ &= +\frac{60I_0}{x_0}\epsilon^{-jkr_0}KHWN = +\frac{60I_0}{x_0}h\epsilon^{-jkr_0}. \end{aligned} \tag{12}$$

It is seen from (5) that the actual vertical electric intensity vector at this point is

$$\bar{F} = -j60\frac{\bar{I}_0}{r_0}\epsilon^{-jkr_0} \tag{13}$$

so that, if (1) is applied, we would obtain an erroneous value for the electric intensity.

It can be shown that the flux density at the center of the loop is

$$\bar{B} = +j\frac{\mu}{2\pi}\frac{\bar{I}_0}{x_0}\epsilon^{-jkr_0} \tag{14}$$

where $\mu = 4\pi\cdot 10^{-9}$ = permeability of free space. The total flux linking the loop is

$$\phi = BNWH. \tag{15}$$

The induced voltage is

$$\begin{aligned} e_i &= -\frac{d\phi}{dt} = -j2\pi f\phi = -j2\pi fNWHB = -jhcB \\ &= +\frac{\mu c}{2\pi}\frac{\bar{I}_0}{x_0}\epsilon^{-jkr_0}KNWH = +\frac{60\bar{I}_0}{x_0}h\epsilon^{-jkr_0} \end{aligned} \tag{16}$$

which result checks with (12). Thus we see that while it is not valid to use (1) to determine F, it is true that (3) may be used to determine the electromagnetic flux density. This is the procedure that was followed in determining experimentally the distribution of earth currents in the neighborhood of the WCAU antenna.

It is interesting to observe the amount of error one makes in using equation (1) to measure the vertical electric intensity. If the transmitting antenna is a straight vertical wire of height, a, with a sinus-

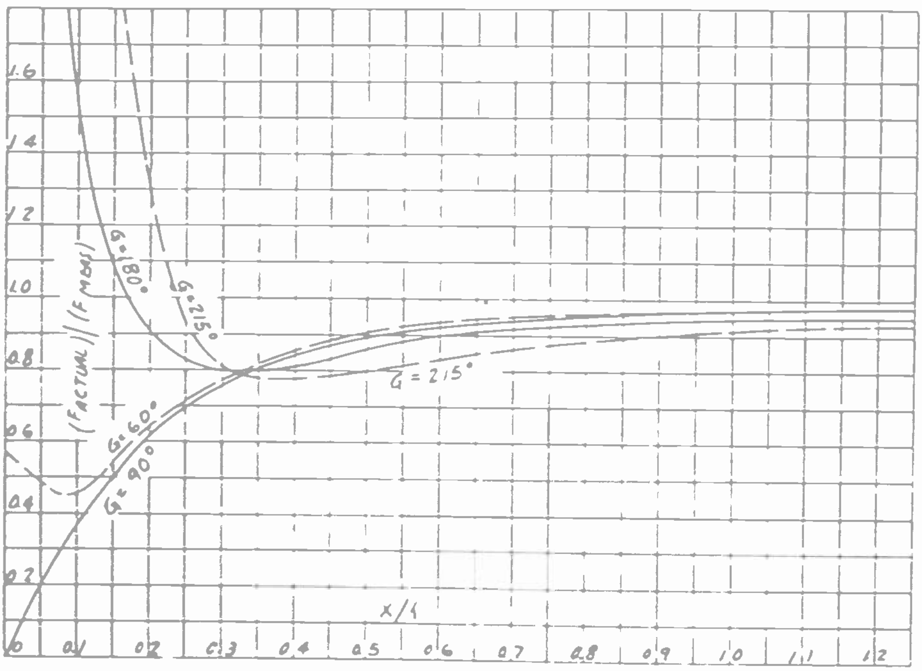

Fig. 33—Ratio of the actual vertical electrical intensity to that measured by a loop antenna in the neighborhood of a transmitting antenna.

oidal distribution of current, the true electric intensity at a point on the earth a distance, x, from the base of the antenna is given by

$$|\bar{F}_{\text{actual}}| = \left| \frac{60 I_0}{\sin G}\left[\frac{\epsilon^{-jkr_0}}{r_0} - \frac{\cos G}{x}\epsilon^{-jkx}\right]\right| \tag{17}$$

where $r_0 = \sqrt{a^2 + x^2}$ and $G = 2\pi a/\lambda$, and the apparent measured electric intensity as given by (1) is

$$|\bar{F}_{\text{meas.}}| = \left| \frac{60 I_0}{\sin G}\left[\frac{\epsilon^{-jkr_0}}{x} - \frac{\cos G}{x}\epsilon^{-jkx}\right]\right|. \tag{18}$$

Fig. 33 shows the ratio of $|\bar{F}_{\text{actual}}| / |\bar{F}_{\text{meas.}}|$ as a function of the distance from the base of the antenna for a number of transmitting antenna heights. We see that all the curves approach unity as we go to distances greater than one wavelength. It is also seen that for distances from the antenna less than one-half wavelength, the error in measuring the electric intensity becomes very large. Thus we see that a loop antenna may be employed to measure the magnetic flux density in the neighborhood of transmitting antennas; some other arrangement, such as two parallel disks, must be used as an antenna to measure the electric intensity in the same region.

Appendix C

The Effect of the Base Insulator

Since the tower type antennas necessarily rest on one or more base insulators, the capacitance of these insulators is shunted across the antenna circuit itself, thus altering the effective impedance offered to the driving voltage. We will examine this effect quantitatively. The equivalent circuit is shown in Fig. 34. In this figure,

R_A = antenna resistance
X_A = antenna reactance (inductive or capacitive, depending on the antenna length)
C_B = capacitance of base insulators
R_B = equivalent series resistance of the base insulator capacitance. (This series resistance may be due to losses occurring in the base insulators or to losses occurring in the earth if the capacitance current flows through a layer of earth in returning to the ground connection.)
$\bar{I}_A$ = true antenna current
$\bar{I}_B$ = current flowing to earth through the base insulators
$\bar{I}_0$ = total current supplied to the antenna system (the vector sum of I_A and I_B.)
$\bar{E}_0$ = driving voltage at the base of the antenna
$\bar{Z}_A = R_A + jX_A$
$Z_B = R_B + jX_B$ $(X_B = -1/2\pi f C_B)$
$Z_0 = R_0 + jX_0$ = equivalent impedance of the combined circuits.

We may arrive at the equivalent impedance of the combined circuits by the relation,

$$Z_0 = R_0 + jX_0 = \frac{\bar{Z}_A \bar{Z}_B}{\bar{Z}_A + \bar{Z}_B}. \tag{1}$$

From the equation,

$$\bar{E}_0 = \bar{I}_0 \bar{Z}_0 = \bar{I}_A \bar{Z}_A = \bar{I}_B \bar{Z}_B \tag{2}$$

we find,

$$\bar{I}_A/\bar{I}_0 = \bar{Z}_0/\bar{Z}_A, \quad \bar{I}_B/\bar{I}_0 = \bar{Z}_0/\bar{Z}_B. \tag{3}$$

The power lost in the base insulator circuit is,

$$P_B = I_B{}^2R_B = I_0{}^2R_BZ_0{}^2/Z_B{}^2 \tag{4}$$

while the total power supplied to the combined circuits is

$$P_0 = I_0{}^2R_0. \tag{5}$$

Thus, the power wasted in the base circuit expressed in per cent of the total input power is

$$100P_B/P_0 = \frac{100R_B}{R_0}\frac{Z_0{}^2}{Z_B{}^2} = \frac{(\text{p.f.})_BZ_0{}^2 \times 100}{R_0Z_B}, \left[(\text{p.f.})_B = \frac{R_B}{Z_B}\right]. \tag{6}$$

We thus have available enough expressions to tell us the complete story. We see from (6), if R_B is made zero, that the total input power is spent in the antenna itself. This is an obvious conclusion from inspecting Fig. 34. While it is impossible to make R_B precisely equal

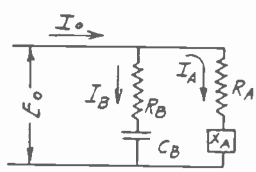

Fig. 34

to zero, it will in general be small compared to the reactance of C_B. For instance, if the total displacement current flowed entirely through the porcelain, we would assign a power factor of approximately 0.7 of one per cent. Since some of the path is through air the power factor will be lowered. This is of course on the assumption that due precautions have been taken to prevent these displacement currents from flowing through any high resistance layers of earth. The various quantities expressed by (1), (3), and (6) have been computed as a function of C_B for the following conditions:

Antenna height = 0.5975 wavelength = 215 degrees
Effective radius of antenna = 12.5 feet
$R_A = 191.0$ ohms
$X_A = -97.0$ ohms
Frequency = 10^6 cycles per second.
$(p.\ f.)_B = R_B/Z_B = 0.01.$

The results are shown in Fig. 35. We see that the effective resistance decreases while the reactance increases with an increase in the base capacitance. The power loss in the base circuit increases, but has only attained a value of 0.6 of one per cent of the total input power when the base capacitance is 300.0 micromicrofarads.

Fig. 36 shows similar results for the following conditions:

Antenna height = 0.4166 wavelength = 150 degrees
Radius of antenna = 12.5 feet

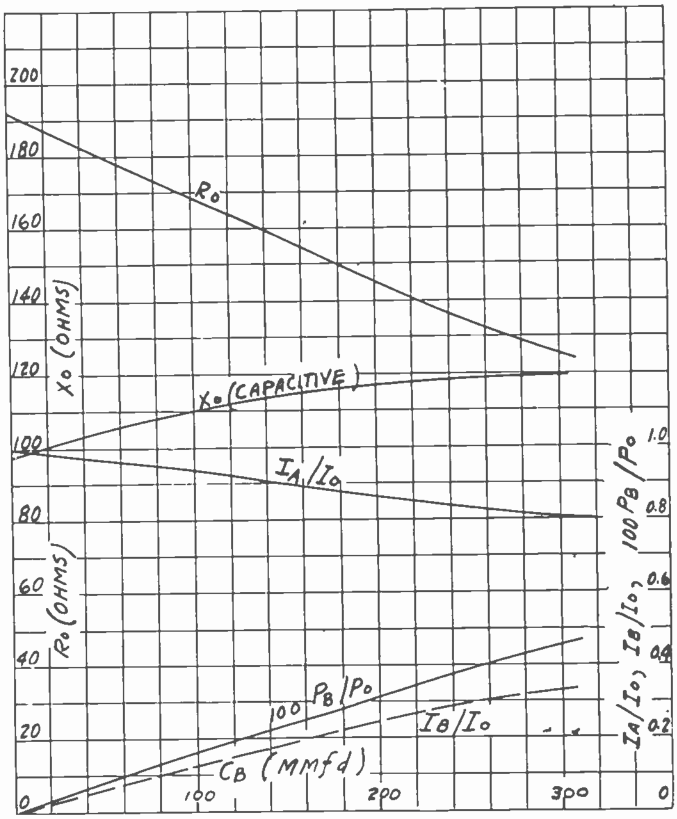

Fig. 35—Antenna height = 0.5975λ = 215 degrees. Effective radius of antenna = 12.5 feet. $R_A = 191.0$ ohms, $x_a = -97.0$ ohms. Frequency = 10^6 cycles per second. $(\text{p.f.})_B = 0.01$.

Frequency = 10^6 cycles per second
$R_A = 400.0$ ohms
$X_A = 472.0$ ohms
$(p.f.)_B = 0.01$.

In this case, we see that parallel resonance may be obtained for an antenna much less than a half wavelength long. We also see that the power wasted may be of the order of 1.5 per cent of the total input power. When it is observed that the field intensity is proportional to the square root of the power radiated, one sees that a loss of power of 1.5 per cent lowers the field intensity about 0.75 of one per cent.

A similar calculation for an antenna whose height is close to one-quarter wavelength shows that the power wasted is less than 0.1 of one per cent.

These calculations show that a large amount of base capacitance is harmless as long as the equivalent resistance of this circuit is kept

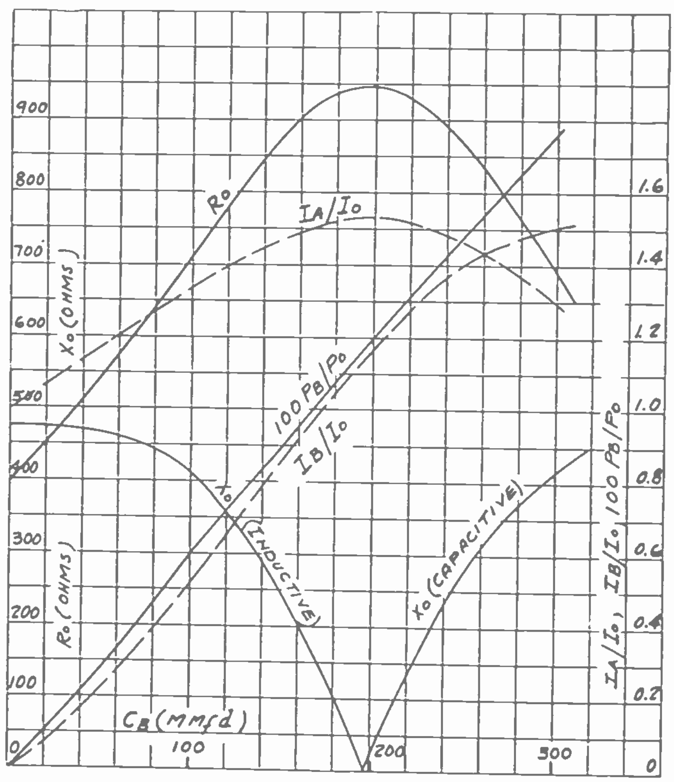

Fig. 36—Antenna height $= 0.4166\lambda = 150$ degrees. Effective radius of antenna $= 12.5$ feet. $R_A = 400.0$ ohms, $x_a = +472.0$ ohms. $(\text{p.f.})_B = 0.01$.

small. In fact, an added capacitance is often placed across the antenna to match the resultant circuit to the transmission line which feeds the antenna. It has been found very desirable to place metallic plates or mats beneath the base insulator to receive any displacement currents which might otherwise flow through a layer of earth to reach the ground system. These mats are of course bonded directly to the ground system. In one particular case, the use of such mats served to increase the field strength 11.0 per cent.

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EXPERIMENTS WITH DIRECTIVITY STEERING FOR FADING REDUCTION*

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Summary—Short-wave fading is largely due to phase interference between multiple path signals of varying path length. Fortunately, stable angular differences usually exist between these paths at the point of reception. It is therefore desirable to employ antenna directivity which is "steerable" and sufficiently sharp to accept only one of the several paths in order to reduce this fading.

This paper describes experiments made with a "steerable" directive antenna during reception of transoceanic short-wave signals. The results demonstrate that sharp angular discrimination is a basically sound method of combating fading which is due to phase interference.

INTRODUCTION

RAPID fading in radio communication has been recognized for some time as being due to the interaction of distinct components having different transmission times. The possibility that these components might arrive from slightly different directions was suggested by various observed facts, among which was the behavior of sharply directive antennas.

It has been noticed in the past that fading was affected by the directivity of the receiving antenna. An example is given in the oscillograph records of Fig. 1 showing observations made by the authors some years ago at Cliffwood, New Jersey. These illustrate a condition of less fading on a large "inverted vee"[1] antenna than on a small nondirectional antenna, using telegraph signals received from station GBK in England. Beating the signal with a local oscillator provided the audio frequency which was recorded. The directive antenna output was recorded on the upper trace while the lower strip recorded the output of the substantially nondirective, comparison antenna.

Such observations as these suggest the possibility of controlling and reducing fading by a systematic use of sharp directivity. The present paper reports some experiments in which changes in fading are correlated with changes in the directive pattern of a rhombic antenna[1] made by mechanically changing its shape.

* Decimal classification: R113.1. Original manuscript received by the Institute, November 23, 1934.

[1] E. Bruce, "Developments in short-wave directive antennas," PROC. I.R.E., vol. 19, pp. 1406–1433; August, (1931).

It may be reasoned that, where the total differences in the path lengths are small, variations can result only in the carrier and side

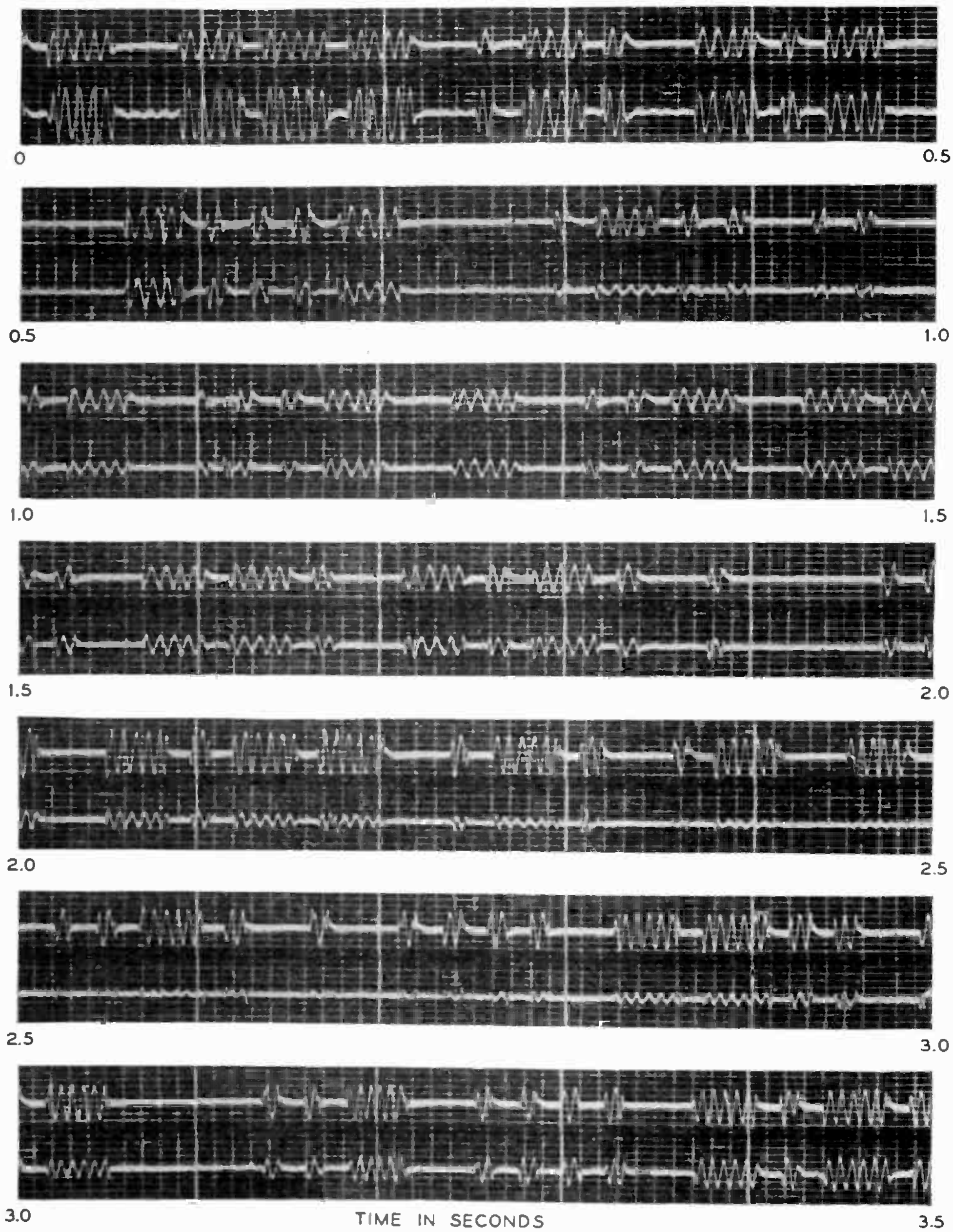

Fig. 1—Oscillographic record of carrier fading reduction. The upper trace is proportional to the output of a large "inverted vee" antenna, and the lower trace to the output of a half-wave vertical antenna when receiving station GBK on 16.6 meters. Taken at Cliffwood, N.J., November 16, 1927, 4:00 P.M., E.S.T.

bands fading in and out together or in other words "general" fading. In such cases, there either may or may not be appreciable angular separation between the multiple waves at the point of reception. However, there is little question that, where multiple waves cause a "selective" fade over a speech band which is, of course, a very small percentage of the carrier frequency, a material path length difference must exist. Where this is the case, it is difficult to conceive of wave routes which do not possess appreciable angular separation between them at the place of reception. The truth of this latter point is of vital importance in this discussion.

The hope of success in fading reduction through directivity rests on the possibility of a continuous, stable angular separation between the interfering waves during times when fading is really troublesome. Fortunately this possibility is reasonably existent; therefore it should be possible to reject all but one of the interfering paths, by means of sharp directivity, with a consequent reduction in selective fading.

Description of Equipment

Tests have shown[2] that a greater degree of angular spread between the multiple waves exists in the incident vertical plane than in the horizontal plane. It might be expected, then, that such a scheme as that illustrated in Fig. 2 would be worth trying. Here the steep edge

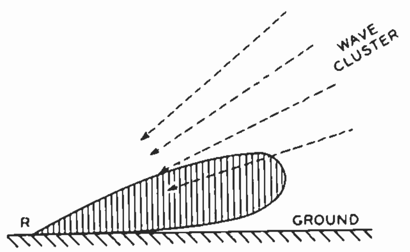

Fig. 2—Edge system for achieving fading reduction with moderate antenna directivity.

of a moderately sharp directional characteristic is moved just far enough into the wave cluster, assumed directively stable, to accept the first wave. Obviously it is possible to approach the wave cluster from the bottom as illustrated or we may approach the cluster from above. A primary essential in this scheme is that no minor ears of the directive diagram be of appreciable size.

Using this scheme, it is not necessary to discriminate completely against the adjacent waves for practical benefit. A discrimination of ten decibels between two adjacent waves of equal amplitude will

[2] H. T. Friis, C. B. Feldman, and W. M. Sharpless, "The determination of the direction of arrival of short radio waves," Proc. I.R.E., vol. 22, pp. 47–78; January, (1934).

make improbable a fade deeper than 5.7 decibels from their sum. Fading of this depth would be relatively unimportant for ordinary speech transmission.

An edge wave may at times be much smaller in amplitude than the adjacent waves. The scheme under discussion may be usefully operative even in this situation since the very smallness of the edge wave means that it cannot be seriously harmful. When signals are weak, the edge of the directive diagram should be advanced until a large amplitude wave is encountered. Some fading of small depth would then exist.

It was stated above that the antenna system used should have no minor ears of appreciable size. At the same time, the edge position of the major loop must be continuously adjustable. A simple method of meeting these requirements is that of mechanically moving the elements of a "long-wire" antenna in space so as to alter the manner of its exposure to the space waves.

Fig. 3 is a rectangular plot of the incident plane directive diagram of a large horizontal rhombic antenna when used for GBW on 20.78 meters. The essential antenna dimensions are indicated on that figure as well as the equation for the directive diagram.

Each bracketed quantity in the directive equation of Fig. 3 is separately plotted on that figure together with the final resulting product. Factor 3, known as the "phasing" factor, exerts the greatest influence on the shape of the major lobe. This factor contains only the variables of length l and the angle ϕ, defined as half of the side interior angle. The length cannot be made easily variable but the angle ϕ can be readily adjusted. When an adjustment in ϕ is available for this antenna, Fig. 4 gives the directions of the major lobe maxima, and the first nulls, above the horizontal, for a series of wavelengths. It is evident that a useful degree of steering is provided without limiting the desirable variable wavelength features of the antenna. In all cases, the minor ears remain small.

In Fig. 5 is shown a remote controlled power-winch system for altering the interior angles. This experimental system in slightly modified form was in operation at Holmdel, N. J., for some time, without any antenna breakages. This was primarily possible because the angles of flexing were very small and copper-clad steel wire was employed in the antenna. The power-winch was equipped with automatic safety stops at the extreme positions, also with a potentiometer which was coupled to the winch to permit the use of a voltmeter as an antenna position indicator. This position indicator was located at the operator's position. By using counterweights, the required size of the winch motor is reduced.

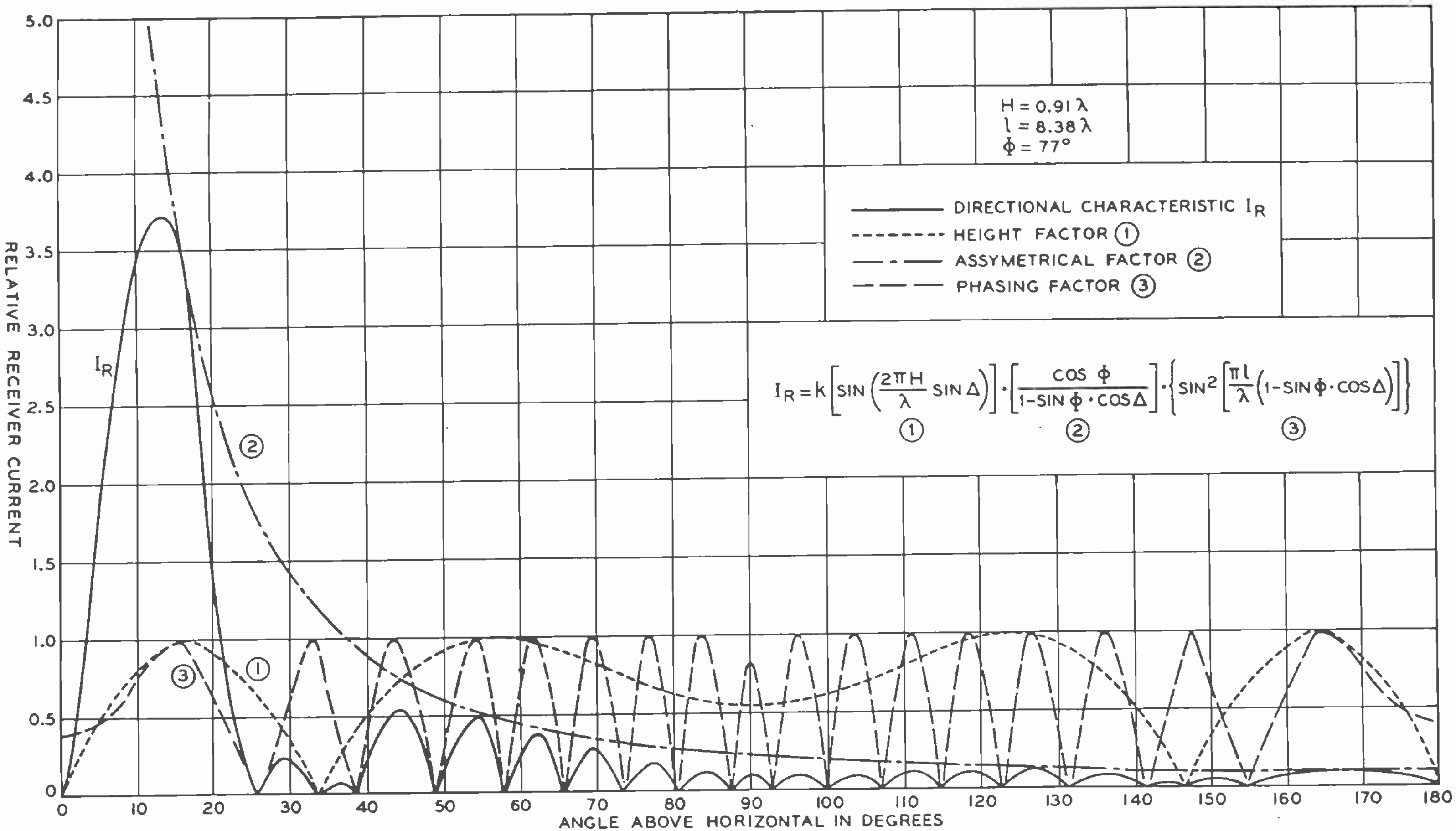

Fig. 3—Receiver current diagram of incident plane directivity for one adjustment of a steerable rhombic antenna. The factors in the equation are plotted separately.

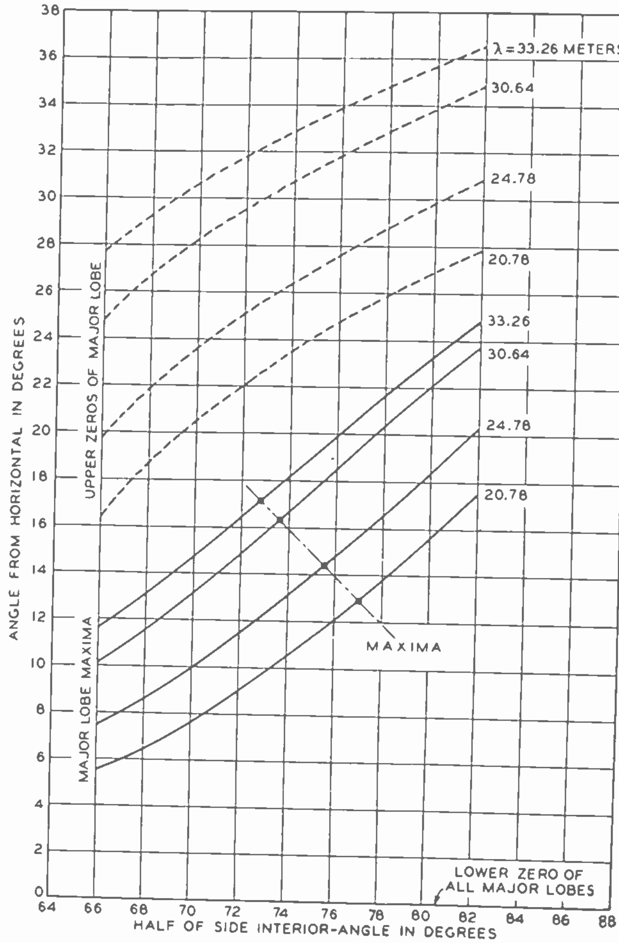

Fig. 4—Steerability, at several wavelengths, of the horizontal rhombic antenna used for the fading reduction studies. The antenna element lengths were 184 meters and their height 19 meters.

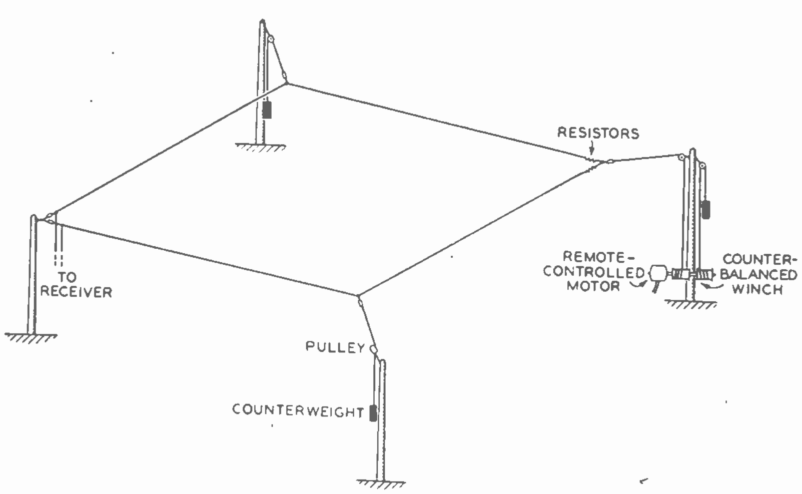

Fig. 5—Mechanical layout of the steerable horizontal rhombic antenna.

The adopted system for observing selective fading required a frequency-wobbled carrier from the transmitting station. By beating this frequency-wobbled carrier with a local fixed frequency, a wobbled audio note was obtained after detection. This audio output was impressed on the horizontal plates of a cathode ray tube, after being

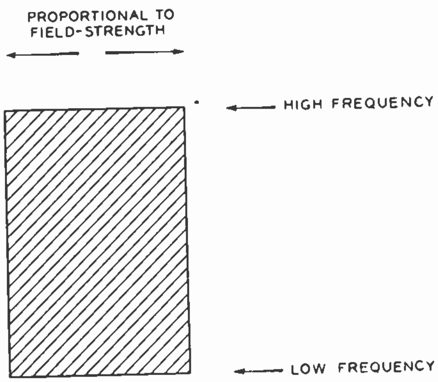

Fig. 6—Cathode ray oscillograph figure for no selective fading when observed with wobbled carrier.

amplified by an audio amplifier. This produced a horizontal spot deflection on the tube screen which was directly proportional to the field strength of the signal. The vertical plates had a locally adjusted sweep circuit voltage impressed on them to produce vertical spot deflections. The sweep frequency was synchronized with the wobble rate

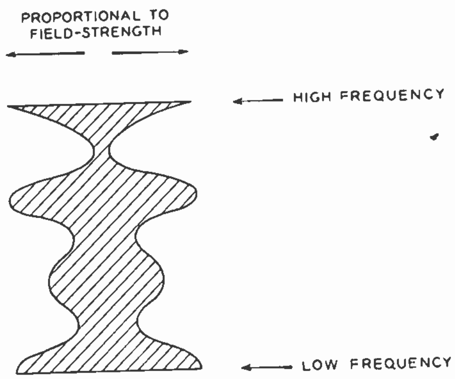

Fig. 7—Cathode ray oscillograph figure for severe selective fading when observed with wobbled carrier.

so that the extreme upper and lower deflections occurred at the same instant as the respective upper and lower frequencies of the wobble. Fig. 6 indicates the cathode ray picture of a signal without selective fading while that of Fig. 7 shows a severe case of selective fading. It is apparent that general fading was revealed by the horizontal collapse of the rectangle of Fig. 6.

It is an interesting fact that upon the first appearance of the cathode ray figure, with the wobble rates employed, it is a horizontal line moving up and down, but after a few seconds, the traced solid figure stands out clearly, due to the persistence of vision.

One of the surprising results of experience with this system was that, at times of severe fading, eight or ten depressions were occasionally seen within a sweep of a few hundred cycles.

For comparison purposes, there were two complete outfits, as described, with their cathode ray tubes mounted side by side. One outfit operated on a simple antenna system, as a standard of comparison, while the second was connected to the adjustable directive antenna. Fig. 10 is a photograph of this apparatus.

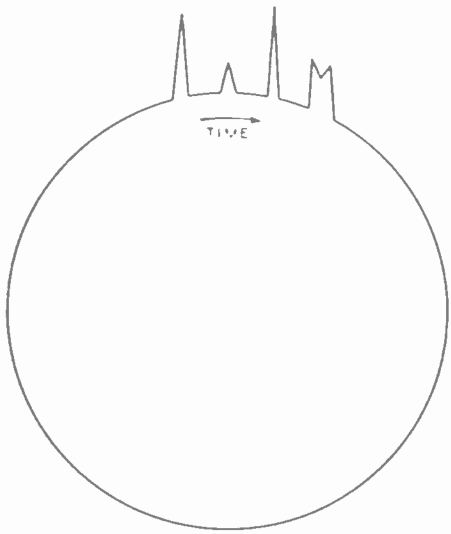

Fig. 8—Cathode ray oscillograph pulse figures when using the circular sweep circuit. The circumference is traversed by the spot in twenty milliseconds.

Other tests also going on at Holmdel, N. J., were concerned with the measurement of the comparative delay times and the respective angles of the various paths of the waves.[2] To permit this, the British Post Office transmitter sent pulses of very short duration. At the receiving point, a single transmitted pulse frequently appeared as several spaced pulses when a sweep circuit was employed. The spacing enabled the measurement of the relative time delays. It was found to be the apparently invariable fact that the earlier arriving pulses are the lower in angle with the horizontal and are relatively stable in direction. These tests suggested that a somewhat similar scheme of observations would be useful to the present work since, if pulses were similarly employed, one would actually see the effect on each individual path of steering the antenna.

Accordingly, cathode ray equipment was constructed employing a circular sweep system, in place of the usual linear sweep, thus making the entire time interval always in view. Fig. 8 illustrates how the pulses sometimes appeared during this sweep. Since the pulses were always vertical, their definition was lost if permitted to slide down into the "3 o'clock" or "9 o'clock" positions of the circle. This possibility was considerably reduced by employing the ellipse in Fig. 9 instead of the circle. For general observation purposes the ellipse was used but for more accurate time delay measurements the circle was employed.

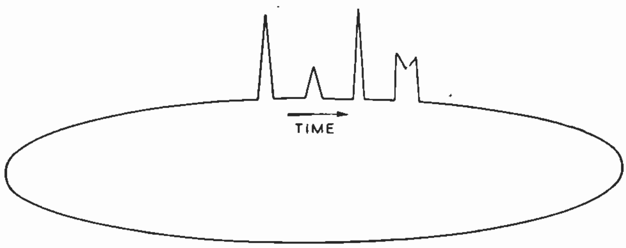

Fig. 9—Cathode ray oscillograph pulse figures when using the elliptical sweep circuit.

The British Post Office station transmitted pulses at intervals of 0.02 second. In order to synchronize with them, an oscillator variable about 50 cycles was used to keep the pulse position stationary. A split-phase circuit feeding the four cathode ray plates produced the circular or elliptical sweep. This equipment is also shown in Fig. 10.

Some studies of general carrier fading were made with a pair of magnetic counters actuated by trigger gas tubes. These fading counters were operated together with automatic recorders so as to maintain the same integrated average signal output. Since, in the recorder integration, ten-second intervals elapsed between gain readjustments, the fading counters operated to record all quick fades, during these intervals, which fell below the average output level by any prescribed amount. A photograph of this equipment is shown in Fig. 11.

Results

Cathode ray tube observations of selective and also general fading were made on the British Post Office stations GBW and GBU using wobbled carrier. Whenever possible, these observations were made at half-hourly intervals. For record purposes, arbitrary numbers ranging from 0 to 4 were adopted. Zero meant very little fading (five per minute or less) and the most severe cases were represented by 4. These figures were recorded separately for the standard antenna and for the rhombus. The difference between the numbers assigned to each antenna gave an indication of the fading reduction accomplished.

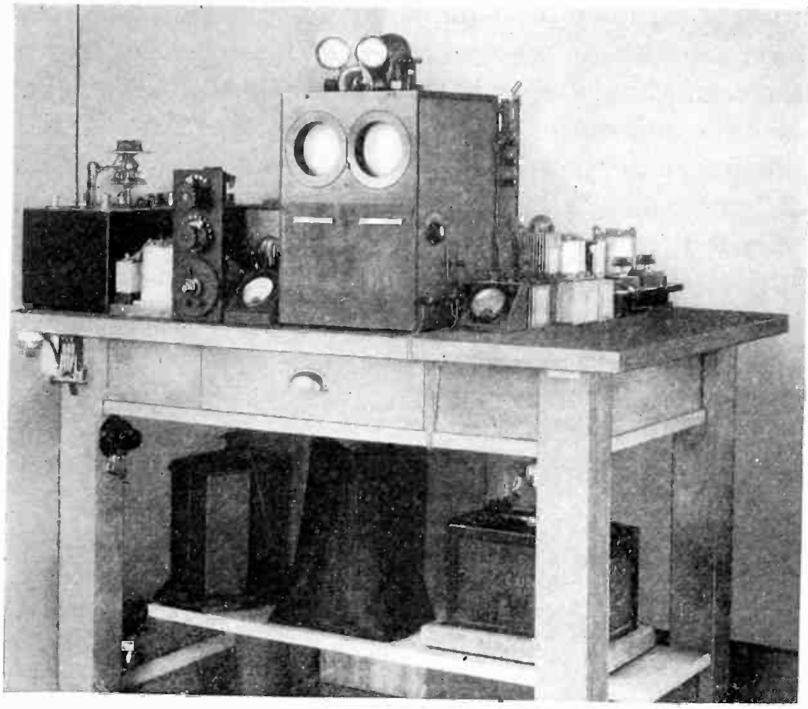

Fig. 10—Cathode ray oscillographs, their amplifiers, and the sweep circuit installation. The meter in the center of the table is the antenna position indicator.

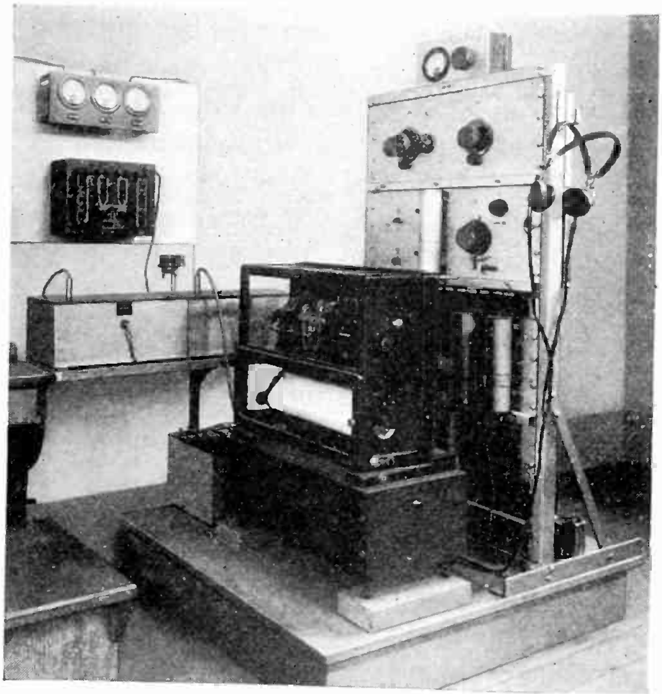

Fig. 11—Field strength recorder and fading counters used for fading reduction studies.

Fig. 12 is a summary of results of these half-hourly observations made during the working hours of March and April, 1933. Disregarding the fact that portions of that figure are shaded, the total lengths of the vertical bars represent percentage of the total number of observations plotted against the degree of the selective type of fading, observed on the comparison antenna, as indicated on the abscissas.

During each of the above observation intervals, the rhombus was steered over its available range to determine the best position for reduction in selective fading. Each of the vertical bars in Fig. 12 is subdivided by shading into the various degrees of fading reduction obtainable at the best position of the adjustable rhombus. The solid sections

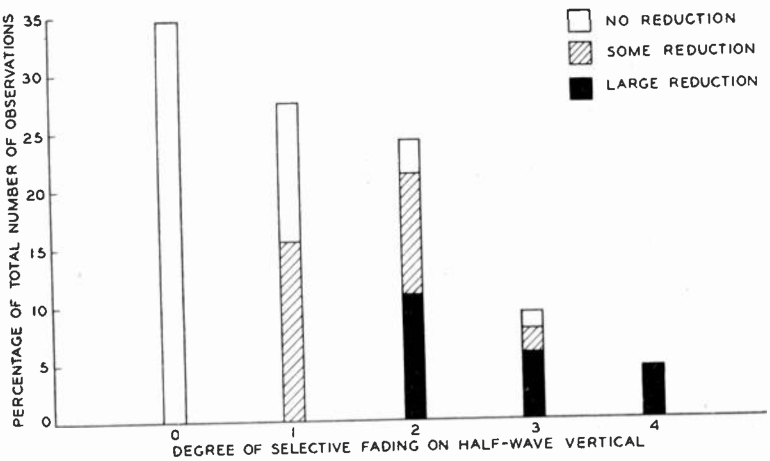

Fig. 12—Selective fading severity and its reduction at the best positions of the rhombic antenna. Stations GBU and GBW, March and April, 1933.

represent large selective fading reductions, the crosshatched sections are fair reductions, while the unshaded portions indicate that the reductions were not of appreciable magnitude.

Analyzing Fig. 12, the results show that 51 per cent of the readings gave no reduction in selective fading. However, for 35 per cent of the readings there was practically no selective fading to be reduced. On the other hand, if one disregards the rather mild and therefore relatively harmless fading cases, graded 0, 1, and 2, rhombic fading reductions were possible 89 per cent of the remaining time, so that when selective fading on the comparison antenna was really troublesome, it is important to note that an appreciable rhombic selective fading reduction was nearly always accomplished. By deliberately steering the rhombus to a disadvantageous angle, it was possible four per cent of the time to make the selective fading worse on the rhombic antenna than on the comparison antenna, but no case has been observed where, at an ordinary rhombic antenna setting, the selective fading was not at least equal to or less than that on the comparison antenna.

While the cathode ray tube figures indicated some degree of general fading, where all frequencies fade together, it was evident that this type of fading is of far less importance than the selective type of fading, in fact it was rarely noticeable except when the selective fading was almost absent.

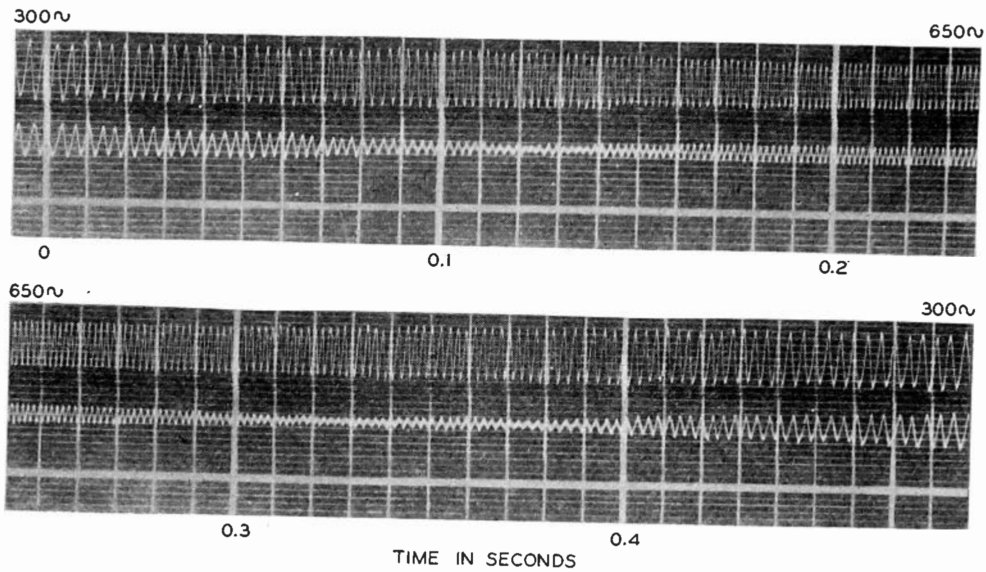

Fig. 13—Oscillographic record of selective fading reduction. The upper trace is proportional to the output of the rhombic antenna, when the angle ϕ equalled 69 degrees, and the center trace is proportional to the output of the half-wave vertical antenna. The lower string was idle. Wobbled carrier from station GBU, April 19, 1933, 4:00 P.M., E.S.T.

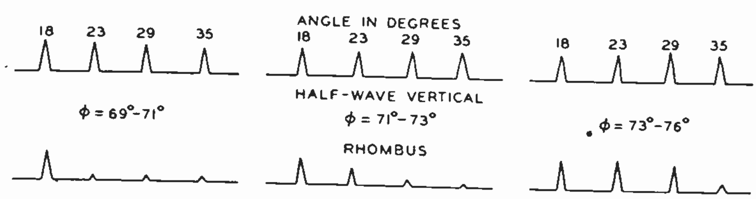

Fig. 14—Pulse pattern changes with steering, March 8 and 9, 1933. Station GCS on 33.26 meters.

Fig. 13 is a photograph of permanent wobble records of selective fading as recorded by the string oscillograph previously mentioned. The center string was actuated by the signals from the half-wave vertical comparison antenna while the rhombus signal was fed to the upper string. The third string was not utilized. The frequency wobble can be seen on close examination and as each small timing division is 0.01 second, the audio frequency is recorded. The record has been marked at the wobbled frequency extremities.

Figs. 14, 15, and 16 are sketches of three interesting series of pulse patterns observed on the rhombic and comparison antennas. The three groups reading from left to right show the effects on the individual pulses of the steering of the rhombus, as indicated by the angle ϕ. The steering achieved at these angles can be seen by referring again to Fig. 4. Marked over the individual pulses are the arrival angles above the horizontal, measured through the coöperation of coworkers.

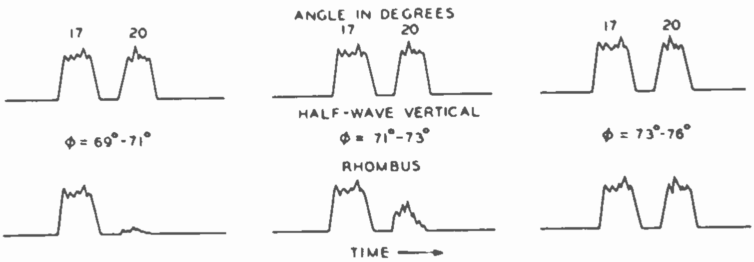

Fig. 15—Pulse pattern changes with steering, April 8, 1933, 2:00 P.M., E.S.T. Station GBW on 20.78 meters.

Fig. 14 is of a test, at thirty-three meters, during a period when a wide angular spread of the cluster prevailed. Four narrow pulses of similar magnitude appear on the half-wave antenna. The progressive effect of suppressing the higher angle waves by steering the rhombus is shown. Very appreciable selective fading reductions are possible under such conditions.

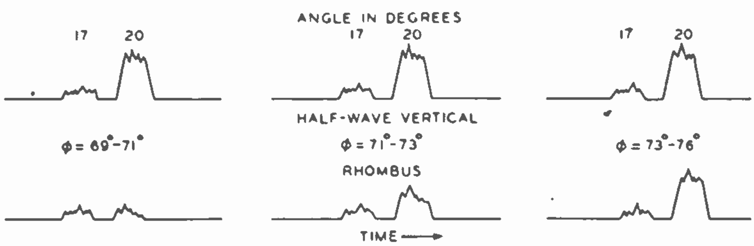

Fig. 16—Pulse pattern changes with steering, April 8. 1933. 2:40 P.M., E.S.T. Station GBW on 20.78 meters.

Fig. 15 is a sketch of twenty-meter observations during a period when selective fading reductions were achieved at the lower angle antenna settings. The broad, flat tops of the pulses are incidentally an interesting contrast to those in Fig. 14. These are possibly due to an increased horizontal spread of wave angles.

Fig. 16 is of a case where it was possible deliberately to make the fading on the rhombus worse than that on the comparison antenna. Since the later pulse had a higher amplitude than the earlier one, rhombic steering by equalizing the relative amplitudes, as shown in the left-hand figure, made the selective fading very bad indeed. The opportunities for producing a result of this nature are rather rare, in fact in our previously mentioned wobble studies it was possible to make the selective fading worse only four per cent of the total time of observations.

Occasionally, and in particular on twenty meters, only slight selective fading was observed. When pulse transmissions were available during these times, only one major pulse could be seen. Really bad fading invariably occurs when multiple pulses, which are widely spaced in time, are observed.

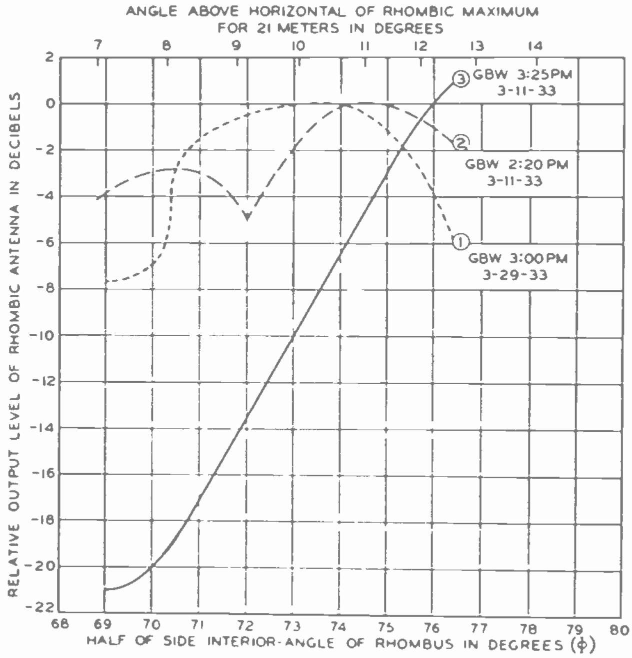

Fig. 17—Horizontal rhombic antenna output changes with steering as shown by automatic field strength recorders. Corrections for changes in signal level with time, as obtained from a half-wave vertical antenna, have been applied.

It may be evident, from the previous discussions in this paper, that the change in antenna output, with steering, is closely related to the number and spread of the waves arriving and to the selective fading improvement obtainable. Fig. 17 shows three cases of results secured by reading relative gain changes, as shown by automatic recorders.

Case 1 is typical of a closely spaced wave cluster arriving at an average angle of about ten degrees above the horizontal. Case 2 can be explained as due to a narrow wave cluster at eleven degrees plus another of less amplitude at eight degrees. We would ordinarily expect annoying selective fading in such an event. Should we deal with many closely spaced waves having a large angular spread, very little gain change would be evident while steering the rhombic antenna, but selective fading improvements over the comparison antenna might still be possible.

Curve 3 is of considerable interest in that it served as one of the experimental checks of the theoretical directive pattern calculations. The change in gain with steering is so well defined that probably only one wave direction existed. This belief was supported by an absence of noticeable fading. Independent measurements, made by an average angle measuring installation[2] consisting of two horizontal dipoles at different heights which determines the average angle by the ratio of the respective outputs, gave the arrival angle at from eighteen to nineteen degrees above the horizontal. Fig. 4 indicates that a ϕ angle of about 68 degrees would place a null at this angle. While the range of steering of the rhombic antenna in use did not permit an adjustment to less than about sixty-nine degrees, the trend of the curve leaves little doubt as to the correctness of our null point calculation.

As might have been expected, the previously described fading counters for studying general carrier fading showed that reductions were usually obtained at the directivity positions which also gave the least selective fading. This type of apparatus is incapable of determining whether general fading or selective fading conditions are affecting the amplitude of the fixed carrier frequency.

Conclusion

It is believed that the results, discussed in this paper, demonstrate that sharp angular discrimination is a basically sound method of combating selective fading.

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STEEL-CYLINDER GRID-CONTROLLED MERCURY-ARC RECTIFIERS IN RADIO SERVICE*

By

S. R. Durand

(Allis-Chalmers Manufacturing Company, Milwaukee, Wis.)

Summary—*This article describes a steel-cylinder grid-controlled mercury-arc rectifier as applied to radio transmitting service.*

The various parts of the rectifier cylinder are pointed out with reference to a picture of the rectifier unit. The use of mercury seals is described, and the operation of various auxiliaries such as vacuum pumps, vacuum meter, and ignition-excitation equipment is dealt with.

Particular reference is made to the use of grid control for deionizing the anodes after carrying current, regulation of the direct-current output voltage, ultra-high speed electronic circuit breaker protection of the rectifier and radio transmitter equipment, and inversion of electrical energy stored in the filter system during the interruption of short circuits. The operation of the grid-control protective circuit as compared to that of an alternating-current oil circuit breaker in interrupting direct short circuits is illustrated by means of oscillograph records.

THE development of steel-cylinder mercury-arc rectifiers was started about 1905 and the first commercial installations were made both in the United States and in Europe more than twenty years ago. At the present time there are about three thousand rectifier units installed throughout the world. The most extensive field of application has been in electric railway power systems, but this type of rectifier has also been applied to electrochemical, electric lighting, industrial, and radio service.

Even though a steel-cylinder mercury-arc rectifier unit delivering power at the relatively high voltage of 12,000 volts, direct current, was installed in an electrochemical plant in Germany in 1926, it was not until 1929 that this type of high voltage rectifier was first applied to radio service. Since that time almost every one of the European broadcast stations operating at high power output has been equipped with a steel-cylinder mercury-arc rectifier, and within the last year the first high voltage unit of this type manufactured in the United States was placed in service in a radiotelegraph transmitting station at Palo Alto, California.

Description

A steel-cylinder rectifier is operated on the principle that alternating-current power can be converted into direct-current power by the

* Decimal classification: R353.3. Original manuscript received by the Institute, October 29, 1934.

characteristic unidirectional valve action to the current flow of a mercury arc in vacuum. In order to maintain a vacuum within the cylinder, it is a general practice to employ a mercury-vapor diffusion or condensation vacuum pump operated in conjunction with a rotary oil vacuum pump. The steel cylinder is divided into an arc chamber and a condensing chamber. In an insulated cup at the base of the arc chamber is located a pool of mercury forming the cathode of the rectifier. Mounted on the cover plate of this chamber by insulator bushings are the six

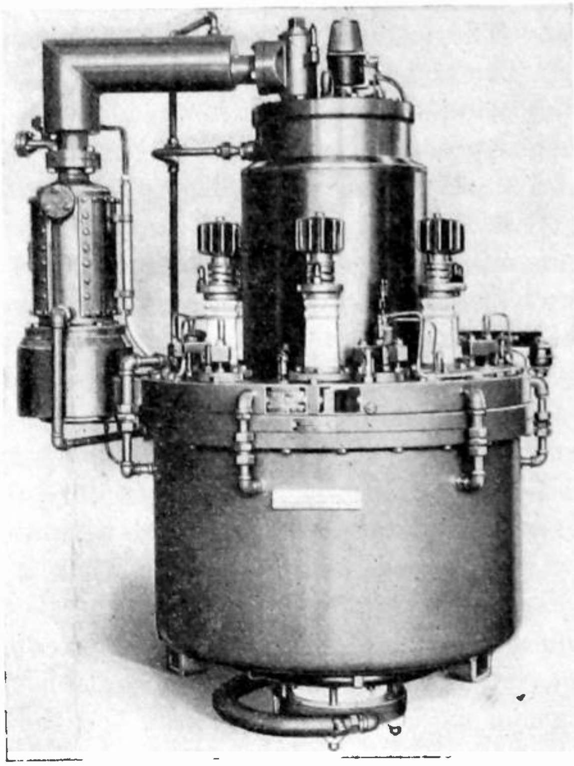

Fig. 1—High voltage grid-controlled mercury-arc rectifier tank with built-on vacuum pump.

main anodes, grid elements associated with each main anode, and two small excitation anodes. The condensing chamber is located above the arc chamber, and is closed at the top by a cover plate upon which is mounted a vacuum valve in the vacuum pump pipe line. Mercury vaporized by the arcs is condensed in both the arc chamber and condensing chamber, and due to the force of gravity automatically returns to the cathode.

Fig. 1 is a picture of a typical high voltage radio rectifier unit with attached mercury-vapor vacuum pump. The cylinder is approximately three feet in diameter by five feet in height, and weighs more than one ton. It has a maximum rating of 800 kilowatts, 25,000 volts, direct current, and is employed in commercial service at a minimum rating of about 30 kilowatts, 10,000 volts, direct current. At lower ratings a smaller size rectifier unit can be used, and at higher ratings a larger unit is employed, or two or more cylinders are connected in series or parallel.

Vacuum tight sealing of the entrance bushings to the rectifier cylinder is accomplished by the use of mercury seals. Each insulator is set in a machined recess in the anode plate and mercury is poured into the surrounding space. The seal is closed at the top by a rubber ring held down by a flange. The seals are equipped with mercury gauges to provide a visible indication of their condition at all times.

An exclusive feature of a mercury seal is that the slightest leak can be discovered and quickly repaired. A leakage is indicated by the sinking of the mercury level in the gauge, and can be remedied by tightening the press bolts on the flange. Even though a leak is not noticed for several weeks, no harm can be done to the rectifier unit since only mercury from the seal gains access to the arc chamber.

In most steel-cylinder rectifier installations, the high vacuum mercury vapor pump is maintained in service continuously, whereas the rotary oil vacuum pump is normally operated only once or twice a week for a period of a few minutes each time under automatic control of a vacuum meter. The heater plate of the high vacuum pump requires about 700 watts of electrical energy. The function of the rotary oil pump is to discharge into the atmosphere any gases extracted by the high vacuum pump from the rectifier cylinder. It is equipped with two automatically operated valves, one of which controls the discharge of gases into the atmosphere during operation, and the other of which seals the exhaust system against atmospheric pressure when the operation of the rotary pump is stopped.

An electrodynamic type of vacuum meter is employed directly to record the degree of vacuum within the cylinder in microns of mercury pressure. The rectifier is normally operated at a pressure of the order of one micron (0.001 millimeter of mercury). The vacuum meter is connected to a Wheatstone resistance bridge, two branches of which are exposed to the vacuum and two to the atmosphere. Due to the difference in radiation of heat from the resistors in the branches when current is passed through the bridge circuit, a potential difference is created across the coils of the meter which is a function of the degree of vacuum within the rectifier. Contacts on the vacuum meter control the starting and stopping of the motor of the rotary vacuum pump within required limits, and additional contacts automatically protect the rectifier plant by causing an alarm to be sounded, and later tripping the main circuit breaker, in case the vacuum should ever become too low for safe operation.

In starting up a mercury-pool type of rectifier, it is a general practice first to establish a low voltage ignition arc. An ignition rod within the rectifier is caused momentarily to dip into the mercury pool under magnetic force, and the arc struck as it is withdrawn under force of a return spring is transferred to the two low voltage excitation anodes which maintain it during operation of the unit. The ignition-excitation circuit is designed for complete automatic operation, and like all other auxiliaries is standard equipment used on more than two thousand steel-cylinder rectifier units.

Immediately after establishing a low voltage excitation arc, the main circuit breaker can be closed to apply power to the rectifier transformer. In many of the installations in radio service in Europe, a double three-phase transformer unit with interphase transformer is employed. However, it is generally considered to be more advantageous to use a transformer with delta connected primary and six-phase fork connected secondary. Whereas the rating of this transformer must be slightly higher, the voltage across each winding is considerably less, and the voltage regulation of the rectifier plant between complete no-load and full-load is much lower. The use of a fork connected transformer is particularly advisable when grid-control voltage regulation is employed for adjustment of the direct-current output voltage of the rectifier plant.

Grid Control

Perhaps the most interesting and valuable feature incorporated in steel-cylinder radio rectifiers is the use of grid control. The important advantages derived by placing grid elements in the arc path of each one of the main anodes and connecting them to control circuits are:

(1) Rapid deionization of the anodes after carrying current.

(2) Regulation of the output direct-current voltage.

(3) Ultra-high speed electronic circuit breaker protection of the rectifier and radio transmitter equipment.

(4) Inversion of electrical energy stored in the filter system back into the alternating-current network during the interruption of short circuits.

Each anode of a mercury-arc radio rectifier normally carries current for one sixth of a cycle during the time it is at highest positive alternating-current potential in relation to the other anodes. The relative instant at which the arc is transferred between one anode and the next can be controlled, however, by means of the grid elements. As long as a negative potential in relation to the cathode is maintained on a grid, it will prevent the corresponding anode of the rectifier from firing; the instant this negative potential is released and a positive potential is applied, the anode will fire and carry current.

In reapplying a negative potential to a grid immediately after an anode has ceased firing, the grid tends to collect a negative current (positive ion current) and thereby causes the space about the anode to be quickly deionized. This deionizing action of the grid is considered to be a valuable feature in preventing the occurrence of backfires within the cylinder.

By controlling the firing of each anode, the direct-current output voltage of the rectifier plant can be reduced to any desired value. A very simple method of accomplishing grid-control regulation is to apply alternating-current potentials to each grid which bear a constant phase relationship to the alternating-current potentials applied to the anodes. By adjusting a negative direct-current biasing voltage also impressed on the grids, the relative instant at which each one is permitted to become positive (and thereby allow the main anodes to fire) is controlled, and the effective mean value of the direct-current output voltage of the rectifier plant is regulated.

Since flashovers within radio transmitting tubes are most likely to occur each time a transmitter is placed in service too quickly at full voltage, grid-control voltage regulation is particularly valuable in starting up radio transmitters at very low voltage and gradually bringing them up to the normal operating value. For continuous operation, it is generally satisfactory to design the filtering system of the rectifier plant for the required ripple tolerance when the direct-current output voltage is regulated about fifteen per cent by grid control. Within this range, it is only necessary to increase the capacity of the filter by a very small amount in order to maintain the same filtering efficiency as at full voltage without any grid-control regulation. The control power required by the grid circuit is of the order of only one/ten-thousandths (0.0001) part of the output power. The grid-control apparatus is relatively smaller and much less expensive than induction type voltage regulators or transformer tap changing switches for accomplishing the same regulation of a rectifier plant.

The most important feature obtained by the use of grid control is ultra-high speed electronic protection of the rectifier unit and radio transmitter. On the occurrence of an overload or direct short circuit, power is instantly cut off by means of a negative direct-current blocking bias applied to the rectifier grids. Energy stored in the filter reactor is automatically inverted back into the alternating-current network, and the short-circuit current is thus completely interrupted in the time of about one cycle (1/60 second).

Since most short circuits in radio transmitters are caused by flashovers within the power tubes, it is often possible automatically to restore voltage to the transmitter shortly after interrupting the flashover arc current. The grid-control apparatus is generally adjusted to restore power at low voltage after two or three cycles, and automatically to increase it gradually to full voltage. In this way the likelihood of a tube flashing over again upon being replaced in service is materially reduced. In case of the occurrence of a permanent short circuit, however, wherein power cannot be automatically restored after one, two, three or any desired number of successive attempts, the grid-control circuit operates to trip the main oil circuit breaker and cause an alarm to be sounded.

Test demonstrations of direct short circuits of a high voltage mercury-arc rectifier plant made with a wire about as fine as a human hair proved that even though the short-circuit current attained values of several hundred amperes, the wire did not have time to fuse and break because of the rapidity with which the current was interrupted by means of grid control. Experiments made without grid-control protection but depending upon a high speed alternating-current oil circuit breaker to interrupt the power supply circuit resulted in the occurrence of an explosive discharge and complete fusing of the wire by a destructive arc before the breaker operated to interrupt the power supply system. These experiments demonstrated very effectively the value of grid control in protecting radio transmitting tubes from being damaged on the occurrence of flashovers within them.

The upper half of Fig. 2 shows an oscillograph record of a direct short circuit applied to a 250-kilowatt, 14,000-volt direct-current, grid-controlled radio rectifier plant. The first oscillograph element recorded the direct-current voltage at the terminals of the rectifier. The center element recorded the direct-current load current, and the third element recorded the alternating current in one phase of the three-phase sixty-cycle primary supply. It is observed that at point A when the short circuit was applied, the direct current increased rapidly, and at point B reached a peak value of nearly sixteen times normal current. Operation of the grid-control protective circuit caused the direct-current voltage to become negative at point C, so that energy stored in the filter reactor was inverted back and dissipated in the alternating-current system. Inversion of this energy can be completed in half a cycle, so that the flow of short-circuit current was fully interrupted at point D.

During a subsequent interval of eight and seven-tenths cycles, the grids were maintained negative to prevent the restoration of power to the rectifier and radio equipment. At the end of this interval at point E, normal service was instantly resumed. If desired, full voltage could have been restored slightly after point D, or after an interruption of only two cycles (1/30 second). However, it is generally advisable to restore power gradually during an interval of about ten cycles (1/6 second).

The lower half of Fig. 2 shows a similar oscillograph record of a direct short circuit applied to the same rectifier plant, but with the

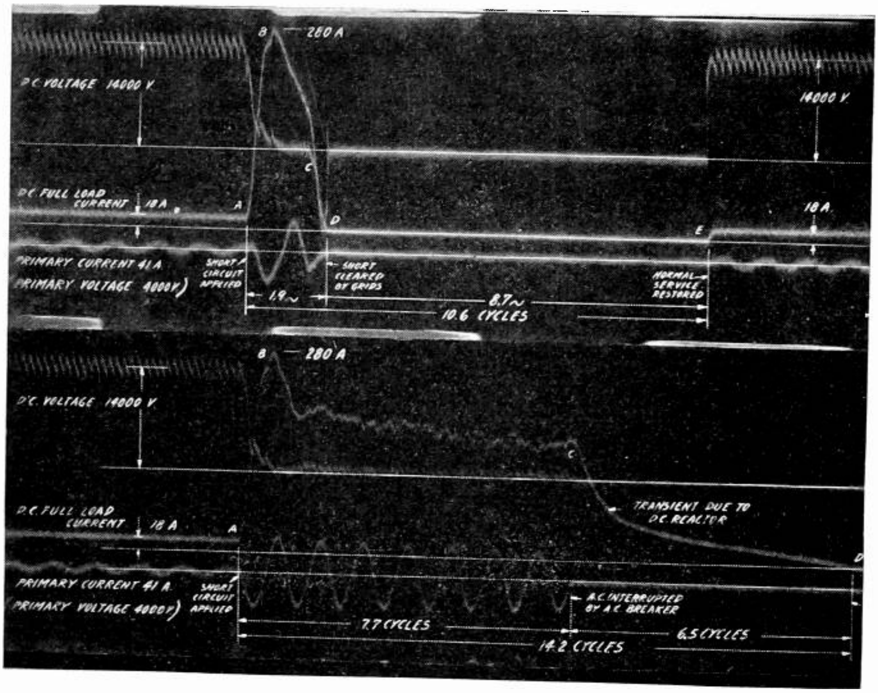

Fig. 2

grid-control protective circuit blocked and interruption being accomplished by means of automatic tripping of a high speed alternating-current oil circuit breaker. It is apparent that the short-circuit current attained about the same peak value of 280 amperes between points A and B as before. However, it required an interval of 7.7 cycles for the breaker to open completely the alternating-current power supply circuit, so that during this entire interval from point B to C, power from the alternating-current system was fed into the short-circuit path. After the breaker had opened the connections to the alternating-current network, it was impossible to invert energy stored in the filter reactor back into the alternating-current system, so that this energy had to be dissipated in the short-circuit arc. An additional interval of six and one-half cycles was required before this transient current ceased flowing at point *D*, and the short-circuit current was completely interrupted.

It is apparent that it took over seven times longer to clear a direct short circuit by means of a high speed alternating-current oil circuit breaker than by means of grid-control protection of a rectifier plant. It is also apparent that the total power dissipated in the short-circuit path is much greater and the resultant damage could, therefore, be very much worse when only an alternating-current oil circuit breaker is relied upon for protection than when grid control is also employed for ultra-high speed interruption and inversion of the short-circuit energy.

Steel-cylinder mercury-arc rectifiers operated at high voltage in radio transmitting service have proved during five years' time that they are not exceeded in efficiency by any other type of converter of similar rating, that they are absolutely reliable, and that their length of life is indefinite and their maintenance cost is negligible. This type of rectifier has already established for itself an important place in the radio field, and is employed today in several dozen broadcast, telegraph, television, and facsimile transmitting stations throughout the world.

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THE QUADRATURE OSCILLOGRAPH: AN ELECTROMECHANICAL DEVICE HAVING TWO DEGREES OF FREEDOM*

By

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Summary—*An electromechanical oscillograph galvanometer has been developed which is capable of deflection along two mutually perpendicular sets of coördinate axes. For power and low audio frequencies, the instrument presents all of the possibilities inherent in a deflecting system of two degrees of freedom.*

I. Introduction

DURING the course of the design and construction of mechanical oscillograph galvanometers, a survey of the literature disclosed the lack of a suitable galvanometer capable of two deflections in space quadrature. Such an instrument would afford the possibility of constructing an electromechanical oscillograph for power frequency work, in which the rotating or oscillating mirror is replaced by a self-synchronizing sweep circuit applied to the horizontal deflector. The conventional mechanical oscillograph with variable speed mirror is difficult to use for observation of phenomena not of constant frequency, and the use of a synchronous motor to drive the mirror limits decidedly the frequency range of the apparatus. A mechanical galvanometer giving quadrature deflections overcomes these limitations and provides as well the means for producing Lissajous figures, hysteresis and other loss loops, and other two-coördinate figures. This is in effect a low-frequency substitute, of indefinite life, for the cathode ray tube.

This paper outlines such a galvanometer, and describes one instrument built for use on sixty cycles.

II. Galvanometer Principle

The direct projection method of the Einthoven galvanometer is employed, using longitudinally bored poles and optical magnification by means of a projecting microscope. However, instead of projecting the shadow image of a vibrating string, a slitted mask mounted on parallel strings permits restricted passage of light. One such mask when stationary will produce on a viewing screen a straight line of light on a dark field. Two slitted masks, one on each vibrator, with

* Decimal classification 621.374.7. Original manuscript received by the Institute, October 9, 1934; revised manuscript received by the Institute, December 15, 1934.

their planes parallel to each other and with their slits at right angles, will produce on the screen a square spot of light resulting from the intersection of the two slits. This is indicated in Fig. 1.

If either vibrator alone moves under the application of an alternating electromotive force, a bright line will be traced on the screen in either the horizontal or the vertical direction. The movement of both vibrators will result in the production of Lisajous figures precisely as in other cases of quadrature vibrations.

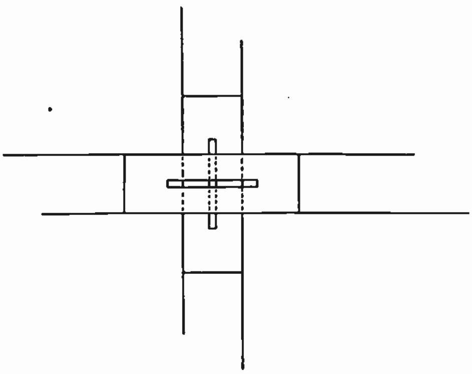

Fig. 1—Arrangement of masks.

III. Vibrator

A rugged vibrator used in a galvanometer intended for sixty cycles has a suspension of two parallel tungsten wires 0.05 millimeter in diameter and seven centimeters long, separated one millimeter. The mask is of aluminum foil, three millimeters long, with width equal to the separation of the wires. The mask is bent around the wires and cemented to them. A slit approximately 0.0065 millimeter wide (measured by miscroscopic examination) allows the passage of light as already mentioned.

Fig. 2 shows the complete suspension holder. It is removable as a unit from the field structure. Adjustments are provided for lateral movement of both horizontal and vertical suspensions for the purpose of centering the slits. Tension adjustments are likewise provided for both suspensions.

The horizontal and vertical suspensions are so spaced that the masks have a clearance of approximately 0.8 millimeter. This spacing is sufficiently small to allow focusing of the microscope on both slits without excessive "fuzziness" of the spot, and at the same time provides an adequate clearance of the suspensions.

As used on sixty cycles, the suspensions are given a free period of about 1000 cycles. By the use of a calibrating voltage to which both vibrators may be connected, the tension and hence the period and sensitivity of both vibrators can be made very nearly alike. A free period of 1000 cycles is sufficient to include the effect of the seventh harmonic with reasonable accuracy.[1] In order to avoid exaggeration of any particular harmonic it is necessary to avoid harmonic points in tuning the suspensions; this, because of the sharpness of resonance of an air-damped vibrator,[2] is not difficult.

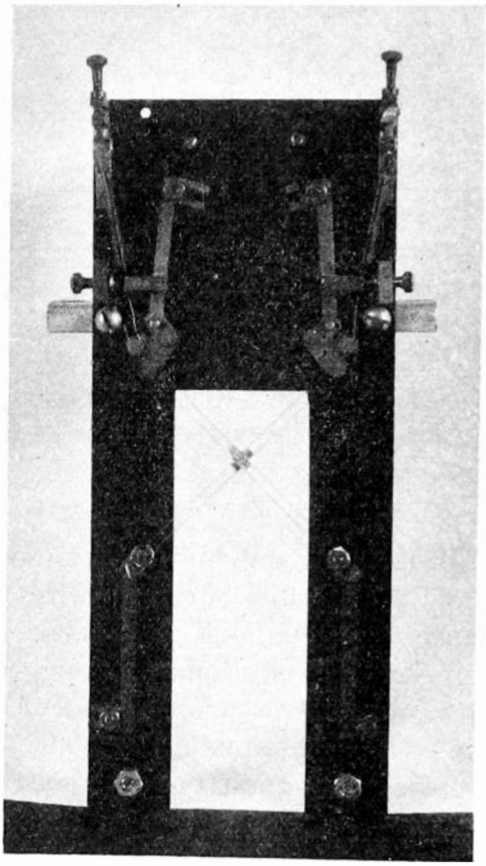

Fig 2—Suspension holder.

IV. Optical System, Etc.

Fig. 3 shows the optical system used. By mounting the prism on a universal joint operable from the panel, small movements of the spot on the screen can be made optically without the necessity of actually moving the suspensions. Focusing of the microscope is accomplished by a rack and pinion likewise controlled from the panel.

[1] For quantitative effect of the galvanometer period, see Laws, "Electrical Measurements," pp. 639–640.
[2] Kennelly, "Electrical Vibration Instruments," p. 217.

With over-all optical magnification of 250, the spot is approximately 1.6 millimeters square. The practical limit to magnification is fixed by the minimum feasible width of the slit in the mask. Using a small 50-watt projection lamp, the spot is readily visible in a moderately lighted room. No attempt has been made to incorporate recording equipment; the problem however does not differ from that of other oscillographic apparatus.

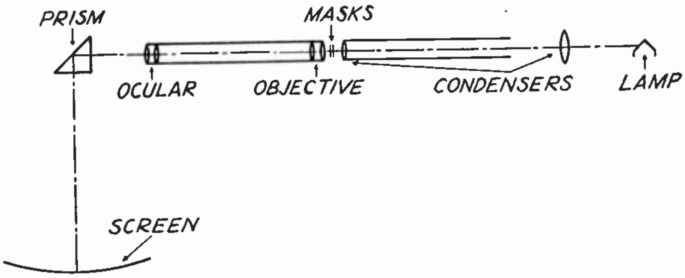

Fig. 3—Optical system.

With a gap density of approximately 30,000 lines per square centimeter, the current sensitivity at sixty cycles with suspensions tuned to 1000 cycles is 0.017 ampere per millimeter half amplitude. This represents a power sensitivity of 6×10^{-4} watt per millimeter, the resistance of the suspension being two ohms.

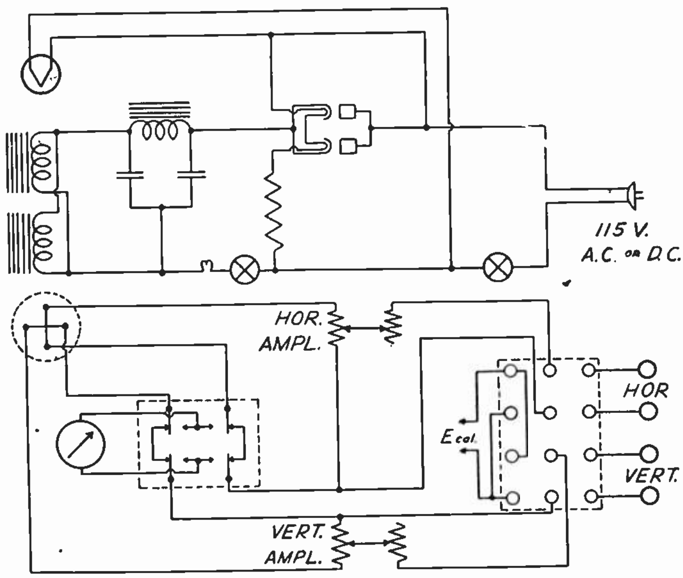

Fig. 4—Schematic wiring diagram circuit.

Fig. 4 shows the complete circuit diagram. A "universal" power unit is provided for the galvanometer field so that the apparatus may be connected to either an alternating- or direct-current source. Adjustments are provided for amplitude of both horizontal and vertical deflections.

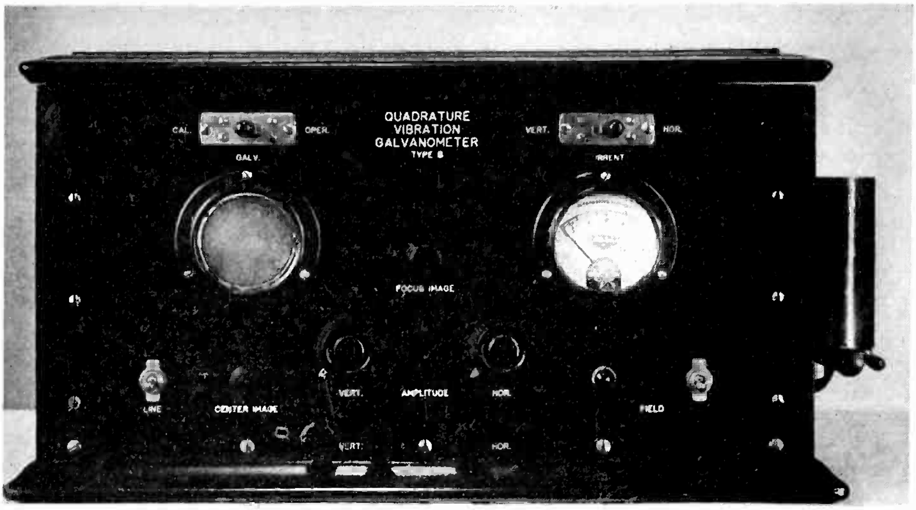

Fig. 5—Front view of apparatus.

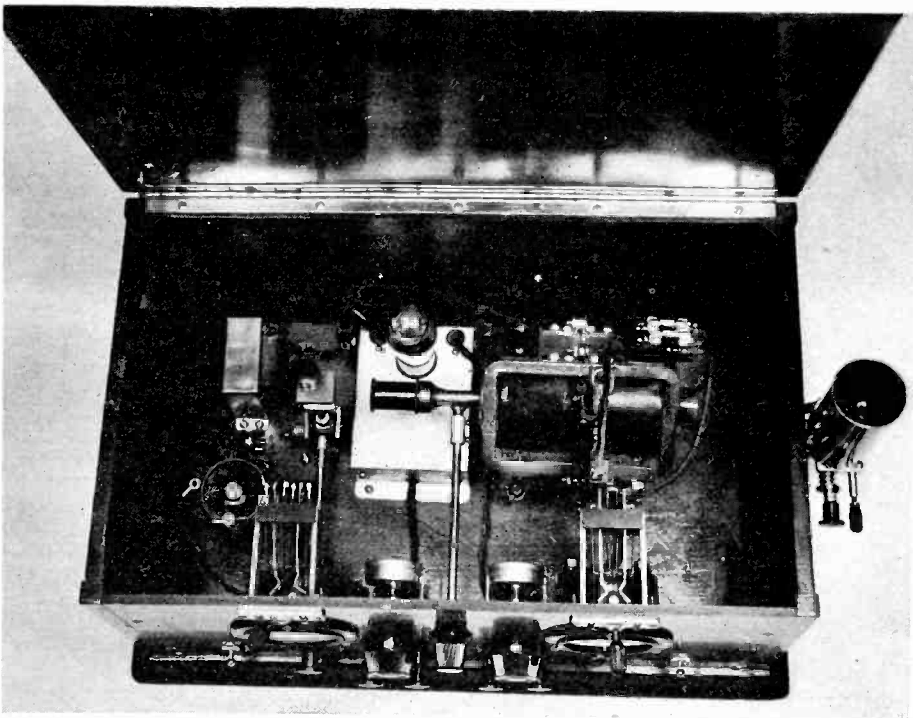

Fig. 6—Inside view of apparatus.

A single switch permits both vibrators to be connected either to a calibrating voltage or to the sources under investigation. An ammeter may be thrown into either vibrator circuit by means of another switch.

Figs. 5 and 6 show the complete apparatus with projection lamp mounted on the side of the cabinet. Referring to Fig. 6, the galvanometer unit may be seen with one condensing lens at its right and the projecting microscope with focusing mechanism at its left. To the left of the microscope is the prism and on the panel in front of the prism is the ground-glass screen. The field supply with rectifier tube is at the left of the galvanometer unit. The complete apparatus measures approximately $19'' \times 10'' \times 10''$.

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A DETERMINATION OF SOME OF THE PROPERTIES OF THE PIEZO-ELECTRIC QUARTZ RESONATOR*

By

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Summary—Entirely consistent determinations of the logarithmic decrement of X-cut quartz rods whose natural frequencies are of the order of seventy kilocycles have been obtained during the last few years. The decrement is obtained from measurements on oscillographic records of the decay of the free vibrations of the resonator. The most extreme (lowest) values of decrement have been checked by an entirely independent method; namely, by tracing the resonance curve, the agreement between the two methods being to within a few per cent. Extended studies by the former method have shown the logarithmic decrement to be a linear function of the pressure of the surrounding gas, and the energy losses in the two gases tried, air and hydrogen, to be in the ratio of their radiation resistances. This assumes previous elimination of mounting losses, which for ordinary mountings are even larger than radiation losses. When radiation losses have been eliminated, the effect of the abrasive used in making the usual ground finish of the surfaces of the quartz becomes obvious, the lowest decrements being obtained when the surfaces have been etched. Representative decrements measured under various conditions are as follows, the corresponding reactance-resistance-ratio Q being inserted in parentheses following each logarithmic decrement.

Ground surfaces, in air at atmospheric pressure,

126×10^{-6} *(25,000)*

Ground surfaces, in hydrogen at atmospheric pressure,

31×10^{-6} *(101,000)*

Ground surfaces, in vacuum,

$11–17 \times 10^{-6}$ *(290,000–180,000)*

Etched surfaces, in vacuum,

6.4×10^{-6} *(490,000)*

Etched surfaces, polished, in vacuum,

5.4×10^{-6} *(580,000)*

Decrements quoted are for short-circuited resonators with thin silver electrodes chemically plated on to the crystal, the short-circuit values being obtained by extrapolation from determinations at various resistance loads. The method yields also the resistance and inductance elements of the equivalent network,[1] and further by the equations for the network a value 5.29×10^4 c.g.s.e. for the piezo-electric coefficient which is effective in driving Y waves in X-cut rods. Temperatures are of the order of 27 degrees centigrade throughout. Resonator dimensions are $0.5 \times 4.0 \times 0.5$ centimeters along X, Y, and Z, respectively.

* Decimal classification: 537.65×R214. Original manuscript received by the Institute, December 4, 1934. For a preliminary report cf. *Phys. Rev.*, vol. 42, p. 587, (1932), Abstract No. 2. This paper was distributed in practically its present form as Document A. G. 1934, No. 24, Comm. I, of the general assembly, Union Radio-Scientifique International, London, July 16, 1934. Much of the present paper was also presented at the Williamstown meeting of the American Physical Society. Cf. K. S. Van Dyke and J. P. Hagen, *Phys. Rev.*, vol. 46, p. 938, (1934), Abstract No. 8.

[1] K. S. Van Dyke, "Piezo-electric resonator and its equivalent network," Proc. I.R.E., vol. 16, pp. 742–765; June, (1928).

CONSISTENT determinations of the logarithmic decrement of an X-cut quartz rod ($N25$, dimensions approximately $4.0\times 0.5 \times 0.5$ centimeters) for lengthwise resonance at 64.5 kilocycles have been obtained by the two following methods: (1) from the persistence of the decay phenomena after excitation as a resonator, and (2) from the sharpness of the resonance curve of the rod used as a resonator. These logarithmic decrements were obtained when the resonator, after etching and then silvering on the YZ faces, was suspended as described below in vacuum (0.1 millimeter or less). The decay determination was made in the midst of a series of three independent resonance curve determinations and the resonator conditions were identical except that for the decay experiments the resonator had to be carried into another room, mounted in a different container, and reëvacuated. The single decay curve determination itself involves some sixteen decay experiments with from six to twelve measurements of each decay event, so that in itself it is a fairly good mean. The values obtained by the resonance curve method at 23 degrees centigrade were 8.6, 8.4, and 10.0×10^{-6}, and by the decay method at 26 degrees centigrade, 9.0×10^{-6}. These values are in agreement to within the limits of precision of the methods, and are considerably smaller than are known to have been reported for any resonator. The resonator in these tests was known to be not entirely free from grease. A previous determination by the decay method on this same resonator when it had been carefully cleaned was 6.2×10^{-6}.

Because of the very low values of decrement which we had found by the decay method, it seemed important to develop a second method to the same precision and then to compare their results. The resonance curve method was selected, and its complete agreement with the decay method for this one resonator $N25$ is taken as a check on the latter method with which an extensive study of the properties of the quartz resonator has been in progress for several years. It must be added, however, that the decay method itself offers a number of checks on its own validity in addition to the consistency of the results which it yields. All of the other determinations which are here being reported are made by the decay method.

For the rest of the report most of the experimental work has to do with a different resonator than the $N25$ above. It is numbered $N19$ and is of similar size, shape, and mounting, but of different orientation in the quartz and hence of somewhat different frequency. $N19$ is cut with its length parallel to the Y-axis of the quartz while $N25$ is at 75 degrees to the optic axis instead of 90 degrees. The 75-degree cut crystal has its length very nearly in the direction of minimum compressional wave velocity in the quartz. Both are X-cut crystals. The decrement found for the resonator X19 when the etched crystal was polished with rouge before silvering was 5.4×10^{-6} at 67.5 kilocycles.

In the investigation of the dependence of decrement upon various factors it was found that ordinary mounting losses are enormous and quite erratic. We have not been able to secure reproducible values of decrement when the resonator is in contact with rigid electrodes except when the crystal is firmly clamped, and then the decrements are very large. The mounting which was finally adopted for the present work uses a fine suspension thread or wire fastened at the center of the rod, which is a node of motion, with a touch of shellac (or of Wood's metal when conduction to electrodes plated on the quartz is desired). In some of the work to be reported the use of electrodes directly in contact with the surface of the crystal is essential to the theory of the method, and silver plating is used in these cases. Light silver plating of electrodes on to the crystal is not found to make an appreciable increase in the decrement. In all cases the effect of moisture is very serious, and precautions are taken to keep the crystal quite dry.

Evacuation of the crystal container reduces the decrement to but five per cent of its dry atmospheric value. The internal viscous losses in the quartz are thus seen to be very small in comparison to the losses to the air and to the usual commercial mounting. The energy losses to the gaseous medium are believed to be largely in acoustic radiation, for the decrement varies linearly with the pressure, as does also the radiation resistance of the gas (product of gas density and velocity of sound). Furthermore, a comparison of the dependence of decrement on pressure in air and in hydrogen on another resonator $B5$ of approximately the same frequency (67.5 kilocycles), but of quite different shape, showed losses to the two gases approximately in the ratio of their radiation resistances. The experimental ratio of the gas losses in air to those in hydrogen is 4.6 while the accepted values of the velocity of sound give a ratio of the computed radiation resistances of 3.9. These velocity of sound determinations are unfortunately not available at the frequencies used here so the values used are those for acoustic frequencies.

Etching of the surface of the crystal apparently removes chips of quartz which have been broken loose by the abrasive and left locked by peaks on the surface, or perhaps widens up the cracks left by the abrasive so that in the vibration process the opposing walls of a crack no longer rub against each other. A ground crystal has a vacuum decrement two or three times that of the etched crystal, while polishing with rouge after etching reduces the decrement still further.

The following list shows values of logarithmic decrement for various conditions of the resonator N19, though in the two cases noted these conditions were not directly studied for this crystal, but the values were estimated by comparison of equivalent conditions with other crystals.

Ground surfaces in air at atmospheric pressure	126. $\times 10^{-6}$
Ground surfaces in hydrogen at atmospheric pressure (estimated)	31.
Ground surfaces in vacuum	11–17.
Etched surfaces in vacuum (estimated)	6.4
Etched surfaces polished in vacuum	5.4

Thus far we have limited our discussion to the decrement of the resonator. The present investigation yields also the elements of the resonator's equivalent network. From these latter and their theoretical expressions derived in a former paper[1] the piezo-electric coefficient and the internal friction constant of quartz are obtainable. These two mechanical and electromechanical properties of quartz are thus determined by purely electrical measurements except for density and linear dimensions of the quartz.

The parallel capacitance in the equivalent network represents the dielectric capacitance between the electrodes of the resonator. In considering the discharge of energy stored in the series chain of the network this parallel capacitance branch may be neglected if the electrodes in contact with the quartz are directly connected to an external load whose admittance is large in comparison with that of the parallel capacitance. With relatively small resistances connected to the crystal electrodes the measured decrements of crystal and load as this external load resistance is varied should obviously be a linear function of this load, which is found to be the case experimentally. The intercepts of the measured decrement vs. external resistance load curve are respectively the decrement of the resonator on short circuit; i.e., the true crystal decrement when there is no external dissipation,—and the negative of the resistance in the series chain of the equivalent network. This latter intercept represents the negative resistance which the load should have to yield a decrement of zero by complete negation of the crystal's internal viscous losses. The short-circuit decrement here defined is that for which values have been quoted throughout the early part of this report. The inductance of the network is obtained from the short-circuit decrement, together with the value of the resistance element and the known frequency of resonance by the usual expression for the logarithmic decrement of a simple series circuit $R/2fL$. It turns out that the inductance is inversely proportional to the slope, tan θ, of the decrement load curve and in practice is computed from the relation $L=1/2f \tan \theta$. For the resonator $N19$, whose decrement has been reported above as 5.4×10^{-6}, $R=320$ ohms, and $L=420$ henrys.

The internal resistance of the resonator $N25$, whose logarithmic decrement has already been quoted for the two methods, was similarly obtained as 650 ohms, while the substitution of resistance for the resonator to give a current equal to that found for the peak of the resonance curve gives a value 660 ohms. Thus the two methods agree well in their determinations of resistance as well as decrement, and hence also, as will later be evident, in the values of inductance and piezo-electric coefficient computed therefrom.

From equations derived for the elements of the network in the former paper, the piezo-electric coefficient ϵ_{12} is obtained by substitution of the values of the inductance and of the mechanical constants of the resonator. These latter, using the notation of the former paper, are for $N19$ in its etched and polished state, $l=4.05$, $b=0.47$, $e=0.49$ centimeters, and $f=67.5$ kilocycles. Hence, $\epsilon_{12}=5.4\times10^4$ in cgse units. This value is slightly higher than had been found for this crystal before etching. Both $N19$ and $N20$, which were cut parallel to each other from the same piece of quartz and of identical dimensions, had yielded on several different occasions the value $\epsilon_{12}=5.2\times10^4$ when both crystals had ground surfaces. So far[2] there has been made but the single determination quoted for the etched crystal, but its precision is believed to be somewhat higher than the others. These values are to be compared with 5.1×10^4 preferred by Sosman,[3] the value selected by Cady for inclusion in the International Critical Tables, the only existing determinations being by static methods.

The above determinations of the decrement indicate that the internal friction constant of quartz which is effective in damping the Y-wave vibrations of an X-cut rod is 0.007×10^{-15}. This constant, ξ, is defined by Kimball as the ratio of the logarithmic decrement to Young's modulus. Here the value chosen for Young's modulus is the elastic constant which, used with the length and density of the rod, will yield the frequency of resonance. Quartz is apparently in a class by itself for low internal losses. None of the other materials listed in Kimball and Lovell's table[4] compare with it, the nearest being phosphor-bronze with a value 0.3×10^{-15}.

A brief statement should be included in the present report regard-

[2] Note added November 26, 1934: Further refinements in the method of reducing the data and the use of least squares solutions yield 5.29×10^4 as the best value of the piezo-electric coefficient effective in driving Y waves in these X-cut rods.

[3] Properties of Silica.

[4] Kimball and Lovell, *Phys. Rev.*, vol. 30, p. 956; December, (1927).

ing the resonance curve determination of the decrement which was quoted above. The apparatus and method were developed and preliminary determinations of decrement made by John Walstrom in this laboratory. John Hagen, who has assisted the author in much of the present investigation and who has obtained most of the experimental values listed in this report, has more recently, using Walstrom's apparatus, secured the resonance curves which are quoted above as checking the decay curve determinations. The principal difficulty in the resonance curve method is in obtaining a reading of the current at the true peak of the very sharp resonance curve. Though currents used are of the order of one milliampere, their heating effect in the crystal causes serious shifts in the resonance frequency of the crystal. Hence, due to the temperature variation, the frequency of resonance varies with the proximity of the driving oscillator setting to the frequency of resonance at the instant. A final setting on the peak of the curve is thus one of unstable equilibrium. A small change in the frequency of the driver from the peak of the curve would involve a diminution in current and thus in the very heating effect which is holding the temperature in equilibrium during the reading. If the change is in the right direction, without there being any change made in the setting of the driver, the crystal resonance curve appears to slide right away from the frequency of the driver, which, a few seconds before, had produced peak current. It has been found possible, however, with a considerable development in technique, to obtain very close approximations to the peak current and to reproduce the readings. Following this the width of the curve at the half-energy points is determined in cycles. It is to be noted that this width is less than 0.2 cycle at the frequency of $N25$ (64.5 kilocycles).

In the method used by Walstrom and Hagen in tracing the upper half of the resonance curve, the resonator to be studied is driven by a crystal oscillator and the frequency varied through the resonance range by the use of a condenser in parallel with the oscillator crystal. A third crystal is used as a standard oscillator and kept at a constant frequency to serve as a reference standard. The difference in frequency between the driver and the standard is obtained by counting the slow fluctuations of a meter in a detector which monitored these two oscillators. Great care was taken in the design of the oscillators and the intermediate buffer amplifier stages to insure freedom from coupling phenomena and from any reaction of the resonator on the driver. Sensitive thermocouples are used to read both resonator current and voltage.

In the decay method the resonator is first driven by an oscillator and then switched to the oscillograph for the decay period. The large voltages required by the oscillograph and the necessity of keeping crystal amplitudes and voltages very small so as to avoid electrical discharge through the gas require the use of an amplifier between the resonator and the oscillograph. A three stage screen-grid amplifier with tuned stages is used giving an over-all amplification of the order of 10^4. Every precaution was taken to avoid regeneration phenomena, both within the amplifier itself, and in combination with the resonator being studied, and to make certain that the amplification obtained is always independent of the input voltage at the instant. Tests of the system have given frequent satisfactory assurance on all of these points. The decay phenomena persist from one-half to ten seconds, during which time from six to twelve photographs of the amplitude at regularly timed intervals are made by the oscillograph, which is of the cold cathode type.

Preliminary studies of the variation of decrement with temperature indicate that the decrement doubles in value as the temperature is changed from ten to forty degrees centigrade, while the piezo-electric coefficient decreases slightly, perhaps five per cent, during the same change.

The selectivity corresponding to the decrement may also be described in terms of the reactance-resistance ratio Q of either the mechanical resonator or the equivalent network. By the definitions of Q and δ, $Q=\pi/\delta$, hence the largest Q here reported is 580,000 at 67.5 kilocycles.

Acknowledgment

Acknowledgment is made of Grants-in-Aid by the National Research Council for assistance in the experimental work, of the coöperation of the General Electric Company in making possible the securing of their cathode ray oscillograph, and also of the coöperation of the Bell Telephone Laboratories in the design and construction of the amplifier used in connection with the oscillograph.

Publication of the complete paper on the investigation described in this abbreviated report is planned for an early date.

Proceedings of the Institute of Radio Engineers
Volume 23, Number 4 *April, 1935*

BOOK REVIEWS

Bessel Functions for Engineers, by N. W. McLachlan. Published by the Oxford University Press. xii+192 pages. Price $5.00.

This book forms a very direct practical and teachable approach to this important branch of applied mathematics. Its distinguishing features are the emphasis placed on the applications, the considerable number of problems worked through in the text, and the large number of problems (some six hundred all told) with answers and many with hints for the solution which are placed at the end of each chapter. Other features are a summarized list of all of the important formulas, tables of the principal functions which while not as extensive as those available elsewhere nevertheless appear to be adequate for the solution of the problems given, an up-to-date bibliography and a convenient subject index. The typography is good and misprints and errata appear to be rather unusually rare for a work of this nature.

Matters which are primarily of more interest to the professional mathematician are subordinated to those of importance to persons mainly concerned with the use of the functions as a tool for the solution of problems of physical or engineering significance. Thus questions of rigor are conspicuous by their absence, while the concrete numerical solution of problems of a type frequently met with in practice is stressed. The problems discussed cover pretty well the whole range of those for which the functions have been found to be useful. While the author's well-known interest in and contributions to the theory of loud speakers has naturally resulted in the inclusion of considerable material of primarily acoustical interest, nevertheless the other important fields in which Bessel Functions play a conspicuous part, such as the theory of the "skin effect," submarine cables, and transmission lines, are by no means slighted.

In the way of adverse criticism there is little to be said. Mathematicians may be pained at the omission in the bibliography of any reference to the historically important works of Heine and Graf and Gubler, but to the engineer for whom the book was written there is probably ample compensation in the fact that the recent literature is quite thoroughly covered and that the great majority of the references are to work in the English language. The reviewer cannot agree with the author that the method of approach to the subject adopted is more suitable for engineers than the more usual one via Laplace's equation, and mildly regrets that the classical method is not sketched even briefly as an alternative. But these are minor matters and do not affect appreciably the usefulness of the book.

This is the latest volume in the Oxford Engineering Science Series and is fully up to all that that implies. It is a good book and should prove very useful both to engineering teachers and to designing engineers.

*L. P. WHEELER

* Washington, D. C.

Gasentladungs-Tabellen, by M. Knoll, F. Ollendorf, and R. Rompe. Published in German by Julius Springer, Berlin. x+171 pages. Price, bound, 29 Reichmarks.

This is a handbook of tables, formulas, and curves which summarizes the available data on the physics (including technical applications) of electrons and ions. The first section includes the data on single or isolated particles—atoms, molecules, electrons, ions, and photons. The second covers the data on the phenomena due to statistical aggregates of such particles. Together these two sections give a résumé of our knowledge of what may be termed the theory of space conduction. The remainder of the book is devoted to information as to the more technical aspects of the subject. The third section covers the data on conduction in electron and gas tubes (including arcs); the fourth and fifth give information on the materials used in making tubes and on vacuum technique; the sixth gives the AEF definitions and classification of discharge phenomena. Finally, there are two sections on aids to computations in this field—one on units and general constants, and the other devoted to tables and graphs of mathematical functions useful in probability and statistical work. An adequate index is provided.

To the reviewer it seems that the authors have covered the field quite comprehensively, and arranged the great mass of material in a logical and easily accessible manner. The handbook should be of value to all interested in the field, and especially to vacuum tube designers.

*L. P. Wheeler

Proceedings of the Institute of Radio Engineers
Volume 23, Number 4 *April, 1935*

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed on this page may be obtained without charge by addressing the publishers.

The Ward-Leonard Electric Company of Mt. Vernon, N. Y., describe in Bulletin 22 their vitrohm plaque resistors, direct-current magnetic contactors in Bulletin 1901, and alternating-current magnetic contactors in Bulletin 4401.

The RCA Radiotron Division of the RCA Manufacturing Company, 415 S. 5th Street, Harrison, N. J., has issued Application Note No. 45 on the use of the 57 or 6C6 to obtain negative transconductance and negative resistance.

The RCA Victor Division of the RCA Manufacturing Company at Camden, N. J., has issued a leaflet on its portable alternating-current-operated cathode ray oscillograph.

Gen-Ral coils and circuit arrangements in which they may be used are described in the 1934–1935 catalog issued by the General Manufacturing Company of 116 Broad Street, New York City.

Sensitive Research Instrument Corporation of 4545 Bronx Boulevard, New York City, has issued a leaflet on high-frequency measurements.

Allan B. DuMont Laboratories, Upper Montclair, N. J., have issued a leaflet on their electronic switch, a device permitting simultaneous observation of two phenomena on the cathode ray oscillograph. Their Type 145 cathode ray oscillograph is described in another leaflet as are their cathode ray tubes.

DeJur Amsco Corporation, 95 Morton Street, New York City, has issued Bulletin No. 35 on tuning control devices.

Bliley Quartz Crystals is the name of a booklet issued by Bliley Electric Company of Erie, Pa.

The 7th edition of the Technical Data leaflet issued by Raytheon Production Corporation of 30 East 42nd Street, New York City, gives data on several dozen tubes.

Type 250B radio broadcast transmitter is described in a bulletin issued by Doolittle and Faulkner of 1306 W. 74th Street, Chicago, Ill. A second leaflet describes concentric cable transmission line and antenna coupling units. Their Type FD1 visual frequency indicator is described in another leaflet.

Catalog No. 11 of the Ohmite Manufacturing Company, 636 North Albany Avenue, Chicago, Ill., describes rheostats and resistance units.

The Miller All-Wave superheterodyne coil kit, test oscillator and inductances, are covered in leaflets issued by J. W. Miller Company, 5917 S. Main Street; Los Angeles, Calif.

CONTRIBUTORS TO THIS ISSUE

Beck, Alfred C.: See PROCEEDINGS for January, 1935.

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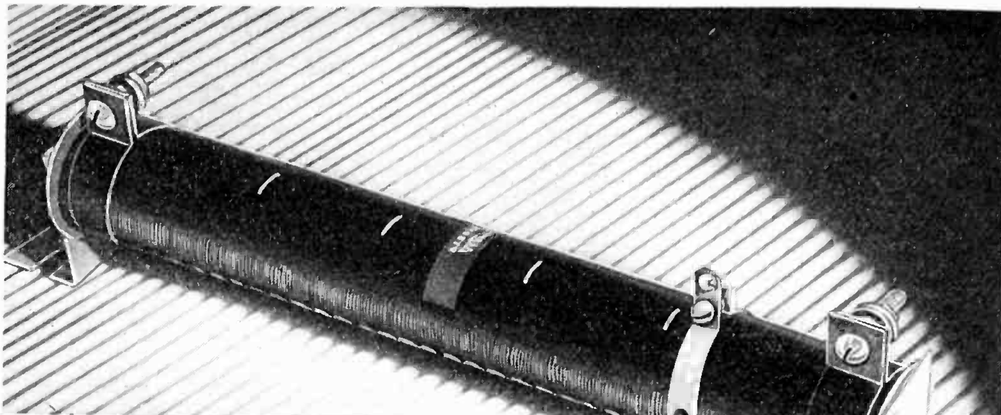
Sherman, Jesse B.: See PROCEEDINGS for January, 1935.

Van Dyke, Karl S.: Born December 8, 1892, at Brooklyn, New York. Received B.S. degree, 1916; M.S. degree, 1917, Wesleyan University; Ph.D. degree, 1921, University of Chicago. Assistant in physics, Wesleyan University, 1916–1917; general engineering department, American Telephone and Telegraph Company, 1917–1919; assistant in physics, University of Chicago, 1919–1921; assistant professor of physics, Wesleyan University, 1921–1925; associate professor, 1925–1928; professor, 1928 to date. Fellow, American Association for the Advancement of Science; Member, American Physical Society, Acoustical Society of America, and Sigma Xi. Associate member, Institute of Radio Engineers, 1915; Member, 1926.



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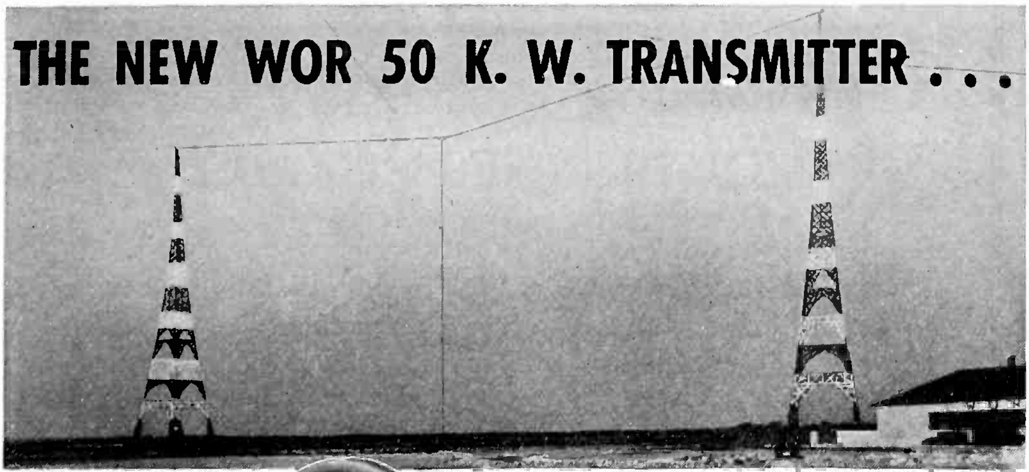

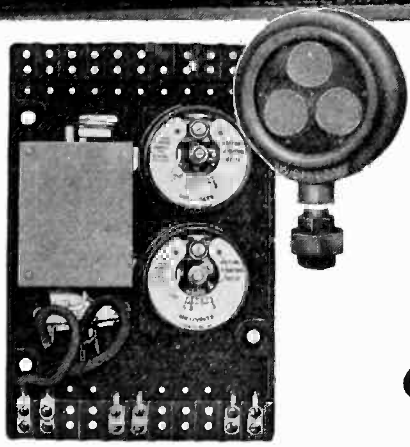

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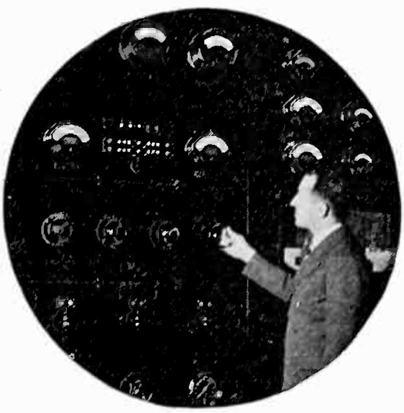

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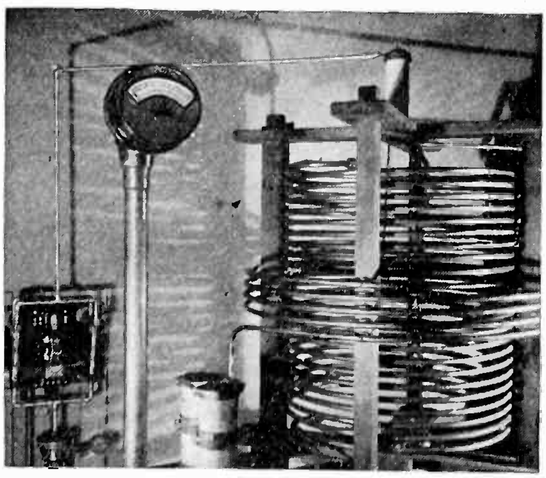

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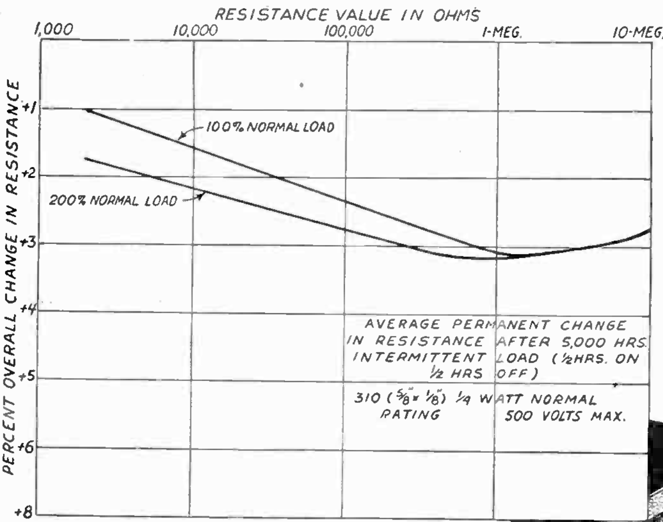
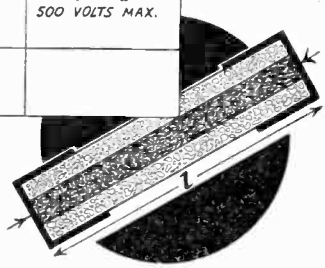


$$R = \frac{\rho l}{a}$$



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$$R = \frac{\rho\, l}{a}$$

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ρ	Specific Resistance
l	Effective Length
a	Cross sectional area

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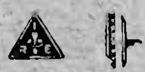
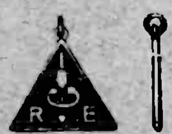


The Lapel Button is of 14k gold, with background enameled in the membership color, and with the lettering in gold. The screw-back on the back of this button fastens it securely to your coat.

$2.75 pospaid—any grade.

The Pin is also of 14k gold. It is provided with a safety catch and is appropriately colored for the various grades of membership.

$3.00 postpaid—any grade.



The Watch Charm handsomely finished on both sides, is also of 14k gold. This charm is equipped with a suspension ring for attaching to a watch fob or chain.

$5.00 postpaid—any grade.

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