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VOLUME [illegible] DECEMBER, 1935 NUMBER 12

PROCEEDINGS
of
The Institute of Radio Engineers

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Institute of Radio Engineers
Forthcoming Meetings

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December 13, 1935

CLEVELAND SECTION
December 19, 1935

CINCINNATI SECTION
December 17, 1935

CONNECTICUT VALLEY SECTION
December 19, 1935

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NEW YORK MEETING
December 4, 1935
January 8, 1936

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December 9, 1935

PROCEEDINGS OF

The Institute of Radio Engineers

VOLUME 23 | December, 1935 | NUMBER 12

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The Institute of Radio Engineers

GENERAL INFORMATION

INSTITUTE. The Institute of Radio Engineers was formed in 1912 through the amalgamation of the Society of Wireless Telegraph Engineers 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.

RESPONSIBILITY. It is understood that the statements and opinions given in the PROCEEDINGS are views of the individual members to whom they are credited, and are not binding on the membership of the Institute as a whole. Papers submitted to the Institute for publication shall be regarded as no longer confidential.

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BUSINESS, EDITORIAL, AND ADVERTISING OFFICES

Harold P. Westman, *Secretary*

330 West 42nd Street, New York, N. Y.

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BOSTON—Chairman, E. L. Bowles; Secretary, Roland G. Porter, Northeastern University, Boston, Mass.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED NOVEMBER 6, 1935

Transferred to the Fellow Grade

New Jersey	Boonton, Boonton Radio Corp.	Loughlin, W. D.
New York	Great Neck, L.I., 18 Melbourne Rd.	Wheeler, H. A.

Elected to the Fellow Grade

Belgium	Brussels, 23 Avenue de Sumatra, Uccle 1	Braillard, R.
England	Crowthorne, Berks., "Overdale"	Barton, F. S.
Germany	Berlin Lichterfelde Ost., 6 Parallel St.	Vrany, C. L.

Transferred to the Member Grade

New Jersey	Ampere, Wired Radio, Inc.	Walter, J. C.
New York	New York, c/o Hazeltine Service Corp., 333 W. 52nd St.	Harnett, D. E.
Argentina	Buenos Aires, International Tel. and Tel. Corp., Defensa 143	Braggio, J. C.
Australia	East Sydney, N.S.W., 55–57 Dowling St.	Thom, F. W. P.
Hungary	Budapest, University of Technical Sciences, Muegyetem	Babits, V. A.
New Zealand	Christchurch, c/o New Zealand Broadcasting Board, P.O. Box 219	Harrison, W. L.

Elected to the Member Grade

New Jersey	Boonton, Boonton Radio Corp.	Franks, C. J.
	Camden, RCA Manufacturing Co., Inc.	Baker, I. R.
France	Paris 15, 52 Avenue de Saxe	de Mare, J.

Elected to the Associate Grade

Arizona	Holbrook, Box E	Dargie, D. A.
California	Palo Alto, 856 Middlefield Rd.	Fontaine, G. K.
	San Pedro, Radio Division, U.S.S. *Maryland*	Seidl, F. A.
District of Columbia	Washington, c/o Capitol Radio Engineering Inst., 1413 Park Rd. N.W.	Martin, W. M.
Georgia	Atlanta, 871 Moreland Ave. S.E.	Alexander, M. S.
Kentucky	Ashland, 1314 Maryland Ct.	Leachman, E. G.
Massachusetts	Boston, 694 Washington St.	Cifre, J. S.
New Hampshire	Manchester, 440 Belmont St.	Rattigan, J. H.
New York	Brooklyn, 55 Hanson Pl.	Stromwall, C. W.
	Buffalo, 91 North Dr.	Seeman, W. F.
	New York, c/o Columbia Broadcasting System, 485 Madison Ave.	Busch, W. R.
	Utica, 1020 Smith Pl.	Wolcott, W. E.
Ohio	Columbus, 1042 Highland St.	Beard, C. E.
Texas	El Paso, 3008 Altura Blvd.	Gemoets, E. L.
Canada	Ottawa, Ont., 249 Lisgar St.	Cahoon, R. D.
China	Canton, Canton Municipal Telephone Administration	Deane, T. Y.
England	London W. 3, 99 Western Ave.	Oxley, R. F.
	Patricroft, Man., H. Clarke & Co. Ltd., Atlas Works, George St.	Whitehead, E. D.
	Portsmouth, Hants., H.M. Signal School	Gregory, D. E.
India	Bangalore, Dept. of Elec. Tech., Indian Institute of Science	Kaul, J.

Elected to the Junior Grade

Iowa	Waterloo, 126 Reber Ave.	Cade, P. J.

Elected to the Student Grade

Scotland	Airdrie, Lanarkshire, "Deanston"	Macfarlane, G. G.

Proceedings of the Institute of Radio Engineers
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APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Admissions Committee. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before December 31, 1935. These applications will be considered by the Board of Directors at its meeting on January 8, 1936.

For Election to the Associate Grade

California	Berkeley, 2510 Bancroft Way	Jones, C. B.
	San Diego, 1060 "C" St.	Meloan, J. R.
District of Columbia	Bellevue, Naval Research Lab.	Keesey, E. R.
	Washington, 7218 Blair Rd.	Cottrell, A.
	Washington, Radio Material School, Naval Research Lab.	Nelson, J. D.
	Washington, 3342-18th St. N.W.	Takase, R. H.
Illinois	Urbana, 606 W. Illinois St.	Eddy, W. L.
Kentucky	Covington, Radio Station WCKY	Glatstein, H. B.
Michigan	Detroit, 16221 Sorrento	Shepley, L.
Missouri	Kansas City, 839 W. 59th St.	Stevens, S. S.
	St. Louis, 4252 Maryland Ave.	Boskamp, J. B.
New Jersey	Harrison, RCA Radiotron Division, RCA Manufacturing Co., Inc.	Johnson, H. W.
New York	Hempstead, L.I., Cottage Pl.	Pell, W. F., Jr.
	New York, Electrical Engineering Dept., Columbia University	Russell, J. B., Jr.
North Carolina	Durham, Radio Station, WDNC	Dalton, R. A.
North Dakota	Grand Forks, Box 102, University Station	Johnston, W. L.
Ohio	Cincinnati, 1885 Taft Rd.	Gillette, A. D.
	Cleveland, Case School of Applied Science	Seletzky, A. C.
Oklahoma	Ardmore, Radio Station KVSO	Holaday, D. F.
Pennsylvania	Upper Darby, 5 E. Wilmot Ave., Llanerch	Kirk, W. C.
Virginia	Quantico, First Signal Co., Marine Barracks	Taylor, W. T.
Australia	Croydon, N. S. W., "Rockleigh," Lang St.	Slade, C. W.
	Sydney, E. S. and A. Bank Bldg., Broadway and City Rd.	Graham, L. B.
China	Shanghai, c/o Leede Engineering Corp., 395 Bubbling Well Rd.	Chien, T. P.
England	Hayes, Middlesex, 16 Waltham Ave.	Ferguson, R. A.
	Kingsbridge, Devon, Cliftonville, Belle Cross Rd.	Pressler, W.C.
	Swanley, Kent, "Wayside," 11 Crockenhill Rd.	Thwaites, H. J.
New Zealand	Pahiatua, Main St.	Glassey, R. B.
South Africa	Johannesburg, 7 Kerry Rd., Parkview	Walker, G. D.

For Election to the Junior Grade

Michigan	Three Rivers, R.F.D. 3	Perkins, M. L.

For Election to the Student Grade

California	Pasadena, Dabney House, Calif. Inst. of Tech.	Rector, E. M.
Georgia	Atlanta, 139 North Ave., N.W.	Brown, J. G.
	Atlanta, 652 W. Peachtree St.	Owens, W. B.
Indiana	Fort Wayne, Y.M.C.A.	Conley, W. H.
	Fort Wayne, 1016 W. Washington Blvd.	Jackson, W. C.
	Fort Wayne, 914 W. Jefferson St.	Rising, V. K.
	Kendallville, 124 W. Grove St.	Trowbridge, B.
	West Lafayette, 137 Cary Hall, E.	Bernstein, J. R.
	West Lafayette, 227 E. Cary Hall	Spicer, F. D.
Massachusetts	Boston, 326 Bay State Rd.	Angevine, O. L., Jr.
	Boston, 526 Beacon St.	Rulon, M. C.
	Cambridge, Hayden 413, M.I.T. Dormitories	Laforge, L. H., Jr.
	Cambridge, 41 Linnaean St.	Lewis, G. A.
	West Medford, 22 Vernon St.	Ayer, J., Jr.
Michigan	Detroit, 13130 Lawton	Katz, L.
	Detroit, 2972 Garland Ave.	Roggenbuck, R. A.
New York	Brooklyn, 281 Fountain Ave.	Wolk, N. I.
	Troy, 2151-12th St.	Adamson, A. A.
South Dakota	Brookings, Box 15	Rawlins, R. E., Jr.
Washington	Seattle, 4318 Brooklyn Ave.	Larson, H. K.
China	Shanghai, Chiao-Tung University	Chao-Han, C.
	Shanghai, Chiao-Tung University	Moo, C. Z.
	Shanghai, Chiao-Tung University	Ting, F.
	Shanghai, Chiao-Tung University	Tse, C. S.
	Shanghai, Chiao-Tung University	Wang, W. Y.
	Shanghai, Chiao-Tung University	Ye, C. Z.
England	Bristol 6, 25 Hampton Rd., Redland	Tozer, B. L.

OFFICERS AND BOARD OF DIRECTORS

Underwood and Underwood

With deep regret we record the death of

Alfred Henry Grebe

Director 1924–1926, 1928

Alfred Henry Grebe was born on April 4, 1893 at Richmond Hill, New York. At fifteen years of age, he became a licensed radio operator and spent three years in the marine service. From 1912 to 1914 he was employed by the Telefunken Company. In 1915 he started manufacturing experimental radio apparatus and from this developed the company which bore his name and was active in the field for a number of years.

His death in New York City on October 24, 1935, followed an operation.

INSTITUTE NEWS AND RADIO NOTES

Monthly Meetings of the Board of Directors

The October meeting of the Board of Directors was held on the 23rd in the Institute office and those present were Stuart Ballantine, president; Melville Eastham, treasurer; Arthur Batcheller, O. H. Caldwell, Virgil M. Graham, R. A. Heising, F. A. Kolster, Haraden Pratt, A. F. Van Dyck, L. E. Whittemore, and H. P. Westman, secretary.

The secretary reported that during July, August, September, and October, 169 Associates, 8 Juniors, and 49 Students were admitted to membership.

A revision of the standard on "Power-Operated Radio Receiving Appliances" proposed by the Underwriters' Laboratories for adoption by the American Standards Association was approved.

An invitation from the Cleveland Section for the Institute to hold its eleventh annual convention in that city was tabled pending the obtaining of additional information on convention facilities.

It was considered desirable that the 1937 Convention be held in New York City as this will be the twenty-fifth anniversary of the founding of the Institute. The San Francisco Section has requested that consideration be given to designating that city as the place for the 1938 convention in view of the preparations now being made for the holding of a World's Fair there at that time.

Upon recommendation of the Sections Committee, the secretary was authorized to withold all monies due sections until financial statements are received for the preceding years.

The president was authorized to appoint a committee to investigate the matter of membership grades, dues, entrance fees, and transfer fees.

The Committee on the Registration of Engineers submitted a report on its operation during the past year and some material on this subject is being prepared for inclusion in the PROCEEDINGS.

The Board reaffirmed its appointment of H. M. Turner as its representative on the Sectional Committee on Dry Cells and Batteries, the personnel of which is being revised.

The Emergency Employment Service placed thirty-four men during the period from June 1 to October 1. The total registration is 781 of whom 561 are members.

The November 6 meeting of the Board of Directors was held in the Institute office. Those in attendance were Stuart Ballantine, president; Melville Eastham, treasurer; Arthur Batcheller, Alfred N. Goldsmith, Virgil M. Graham. R. A. Heising, C. M. Jansky, Jr., F. A. Kolster, George Lewis, E. L. Nelson, Haraden Pratt, H. M. Turner, A. F. Van Dyck, L. E. Whittemore, William Wilson, and H. P. Westman, secretary.

W. D. Loughlin and H. A. Wheeler were transferred to the grade of Fellow and F. S. Barton, Raymond Braillard, and C. L. Vrany were admitted to that grade. V. A. Babits, J. C. Braggio, D. E. Harnett, W. L. Harrison, F. W. P. Thom, and J. C. Walter were transferred to Member, and I. R. Baker, J. DeMare, and C. J. Franks admitted to the grade of Member. Twenty Associates, one Junior, and one Student were elected to membership.

A report of the Tellers Committee on the count of ballots cast in the election of officers was accepted and L. A. Hazeltine declared elected president and Valdemar Poulsen elected vice president for 1936. E. H. Armstrong, H. H. Beverage, and Virgil M. Graham were declared elected directors for 1936–1938.

A request for a list of Institute members residing in the State of Connecticut by the Connecticut State Board of Registration for Professional Engineers and Land Surveyors was approved. The State Board will mail to these members information on a recently enacted law requiring the registration of engineers.

An invitation to be represented on a newly formed Sectional Committee on Radio-Electrical Coördination was accepted and C. M. Jansky, Jr. was designated to serve thereon.

During the month of October the Emergency Employment Service placed nine registrants and received registration data from seven new men.

Committee Work

Membership Committee

A meeting of the Membership Committee was held in the Institute office on November 6 and was attended by I. S. Coggeshall, chairman; J. M. Clayton, F. W. Cunningham, H. C. Humphrey, E. W. Shafer, C. E. Scholz, and H. P. Westman, secretary. Plans were made for the securing of additional members of Student grade and for the annual canvass of the field through the coöperation of the general membership.

Tellers Committee

A meeting of the Tellers Committee was held in the Institute office on November 1. Malcolm Ferris, acting chairman; A. G. Campbell, and H. P. Westman, secretary, attended. The ballots cast in the election of officers were counted and the report prepared for the Board of Directors.

STANDARDIZATION

Technical Committee on Electronics—IRE

Subcommittee on Large Vacuum Tubes

On October 11 a meeting of the Subcommittee on Large High Vacuum Tubes operating under the Technical Committee on Electronics was held in the Institute office and was attended by L. W. Larson, chairman; R. D. Hall, E. S. Spitzer, and H. P. Westman, secretary. The section on "Standard Methods of Testing Vacuum Tubes" in the 1933 Standards Report was reviewed and a number of changes recommended. Some additional material on operation tests of large high vacuum diodes and oscillation tests for multielectrode transmitting tubes were prepared.

Correlating Subcommittee

At a meeting of its Correlating Subcommittee consisting of B. J. Thompson and H. P. Westman, secretary, reports on definitions of the five subcommittees operating under the Technical Committee on Electronics were correlated into a single document for presentation to the Technical Committee.

Technical Committee on Electronics—IRE

The Technical Committee on Electronics met in the Institute office on October 6 and those present were: B. E. Shackelford, chairman, E. N. Dingley (representing W. B. Goggins), F. R. Lack, R. W. Larson, G. F. Metcalf, I. E. Mouromtseff (representing Dayton Ulrey), H. W. Parker, O. W. Pike, B. J. Thompson, P. T. Weeks, and H. P. Westman, secretary. The committee reviewed in detail the complete report on definitions submitted by the five subcommittees and made such changes as it felt were necessary to put the report in final form.

Technical Committee on Radio Receivers—IRE

Subcommittee on Definitions

The Subcommittee on Definitions operating under the Technical Committee on Radio Receivers met in the Institute office on October 24 and those present were F. X. Rettenmeyer, chairman; A. V. Loughren, and H. A. Wheeler. The definitions in the 1933 Standards Report which are within the scope of the committee were reviewed and several changes recommended. Some new definitions were proposed. Several modifications were recommended in the definitions which appear in the Section on Testing of Receivers.

Subcommittee on Testing Apparatus

A. E. Thiessen, chairman; J. F. Dreyer, Jr., Malcolm Ferris, C. J. Franks, H. A. Wheeler, and H. P. Westman, secretary, attended a meeting of the Subcommittee on Testing Apparatus which was held in the Institute office on October 15. The material given in the Sections on "Standard Tests of Broadcast Radio Receivers" and "Standard Tests for High-Frequency Receivers" in the 1933 Standards Report was reviewed and criticized. Two subsubcommittees were appointed to prepare preliminary drafts of the proposed changes for consideration at the next meeting.

Subcommittee on Test Procedures

The Subcommittee on Test Procedures met in the Institute office on November 6. Those present were D. E. Foster, chairman; L. F. Curtis, E. T. Dickey, Sarkes Tarzian, H. A. Wheeler, and H. P. Westman, secretary. The committee reviewed the existing material on the test procedures for radio receivers appearing in the 1933 report and prepared a number of recommendations concerning its modification.

Institute Meetings

Atlanta Section

A meeting of the Atlanta Section was held at the Atlanta Athletic Club on August 22. P. C. Bangs, secretary, presided and nine members and guests were present. Five were at the informal dinner which preceded the meeting.

A paper on "The Lead-Acid Storage Battery" was the subject of a paper by N. B. Fowler, technical employee of the American Telephone and Telegraph Company of Atlanta. In it he presented a brief history of these batteries and discussed in detail the actions involved in the charge and discharge cycles. The uses to which storage batteries are applied in the maintenance of telephone services were described. The paper was discussed by Messrs. Akerman and Bangs.

The September meeting of the Atlanta Section was held on the 26th at the Atlanta Athletic Club. I. H. Gerks, chairman, presided and nine were present.

H. L. Penn, a technical employee of the American Telephone and Telegraph Company, presented a paper on "Program Transmission."

A brief historical sketch of the origin and development of chain broadcasting was given and the part played by the telephone system in providing program supply networks outlined. A discussion was then presented of the transmission requirements of program circuits and the effects of attenuation, phase distortion, and transmission levels. It included transmission over open-wire lines and cable circuits and appropriate means for equalizing the line and equipment to compensate for distortion. Characteristics of the lines and associated equipment were given and operation and maintenance practices covered. The paper was discussed by Messrs. Daugherty, Fowler, Gerks, and Owen.

Boston Section

The Boston Section met at Harvard University on October 25. E. L. Bowles, chairman, presided and seventy-five members and guests were present. The informal dinner was attended by twelve.

E. A. Guillemin of the communication division of the Department of Electrical Engineering of Massachusetts Institute of Technology presented a paper on "An Analytical Method of Predetermining Behavior of the Vacuum Tube Oscillator, Which Yields Regions of Oscillation and Nonoscillation, the Amplitude of Oscillation, and Wave Form."

In it, the behavior of the vacuum tube oscillator circuit was determined in terms of an analytic approximation to the tube characteristic consisting of three confluent straight lines, two of which are horizontal and simulate the characteristic below cutoff and above saturation (or within the grid-current region) while the third assumes the average slope of the intervening portion. The period of oscillation is thus broken up into regions within which the circuit behavior is linear and hence determined by linear circuit analysis. The steady-state cycle is then represented by a series of confluent transients which repeat periodically. Convenient expressions are obtained for the amplitude of oscillation as a function of the tube and circuit parameters and the bias voltages from which the regions of oscillation and nonoscillation may readily be explored as well as all currents and voltages obtained explicitly. A Fourier analysis of the response yields expressions for the ratio of harmonics to the fundamental, by means of which the various regions of oscillation may readily be compared from the standpoint of quality. Within the regions of operation having practical significance, the approximation is sufficiently good to make the result useful.

The paper was discussed by Drs. Barrow and Chaffee, Professor Bowles, and Mr. Field.

Buffalo-Niagara Section

The October 24 meeting of the Buffalo-Niagara Section was devoted to a two-hour trip through the Charles R. Huntley Steam Electric Generating Plant at Tonawanda, N.Y. This station consists of nine General Electric Curtiss Turbo-Generators having a total capacity of 480,000 kilowatts. The two largest and most modern units are each 80,000 kilowatts and develop 12,000 volts at 60 cycles when driven at 1800 revolutions per minute. They are fed by four Babcock and Wilcox boilers burning pulverized coal and supplying steam at 450 pounds per square inch and 750 degrees Farhenheit. The frequency is controlled so that electric clocks in the load line give accurate time. The output is stepped up to 110 kilovolts for delivery.

Ninety-four members and guests went on the trip.

Chicago Section

A meeting of the Chicago Section was held on October 4 in the auditorium of the Western Society of Engineers. Alfred Crossley, chairman, presided and 175 members and guests were present.

This meeting was devoted to a series of papers on the communication system of United Air Lines. H. M. Hucke presented the first paper which covered the description of the complete system and covered telephone, telegraph, and radio facilities. He described the work of maintaining these facilities, the problem of operating numerous radio transmitters on a few channels, the high cost of equipment development due to the small number of units produced, and the problems of providing the necessary laboratory equipment.

The second paper by C. A. Petry covered the design considerations involved in radio equipment. These were studied from the viewpoints of reliability, simplicity, weight, and serviceability. He related these problems to the difficulties encountered with vibration, temperature variations, and the need for rapid servicing and exchanging of units. A general review of the frequency range, fidelity, automatic volume control characteristics, sensitivity, and noise reduction was then made. A short résumé of constructional details of receiving units illustrating the sealing of tuning elements, the care and arrangements of leads and cables, and similar matter concluded the paper.

C. P. Sandretto then discussed the problem of the future power supply for such auxiliary equipment as transmitters and receivers on air liners. He presented a survey of power requirements, speed of planes, passenger miles flown, and freight carried and based his conclusions on these factors. The proposed system involves the use of an auxiliary gasoline motor operating a high-frequency alternating-current generator. The limiting factors of frequency and voltage were discussed and the use of condenser type motors described. The paper was closed with a discussion of the economical and weight-saving factors involved in this type of installation.

The meeting was then adjourned to permit the inspection of a large amount of equipment which was on display.

The October meeting of the Chicago Section was held on the 25th in the auditorium of the RCA Institute. Chairman Crossley presided and 100 were present. Twenty attended the dinner which preceded the meeting.

A paper on "Radio as an Aid in Police Work" was presented by F. A. Rafferty, radio engineer for the Chicago Police Department. In introducing his subject, he outlined briefly the major difficulties faced by police installations at the present time. The objective of police radio is to increase the speed of response which is limited by psychological reactions of the personnel and inadequate apparatus. The necessity for permanency and reliability in equipment was stressed. Difficulties in securing enthusiastic coöperation from departmental officials, squad car operators, and other municipalities were pointed out. A serious situation was shown to exist with regard to the lack of technical advances exhibited by most larger manufacturers and the lack of financial responsibility and organization stability of the smaller manufacturers. An appeal was made that the patent situation and a desire for high profits be not allowed to interfere with the efficient execution of the law. A short illustrated example was presented of the increasing difficulties faced by services on the same frequency channels and the steps which are being taken to minimize it.

The second paper of the evening on "A Low Level Modulation System" was presented by W. E. Phillips of the University of Illinois. In introducing it he briefly discussed the considerations involved in amplitude modulation and the normal methods of obtaining it. An analysis of the powers involved in side bands and carriers was given indicating that the power wasted in the carrier amplification in the usual systems can be conserved to a considerable extent if class C amplification is employed for carrier only and side bands are supplied through a separate channel utilizing class B amplification.

Cincinnati Section

A meeting of the Cincinnati Section was held on September 15 at the University of Cincinnati and was attended by twenty-five. A. F. Knoblaugh, chairman, presided.

"Metal Tube Problems" was the subject of a paper by C. H. Mahoney, chief metallurgical engineer of the Ken Rad Corporation. In it, he discussed the problems of manufacturing metal tubes from the metallurgical standpoint. The multiplicity of seals in this type of tube make necessary rigid manufacturing processes if leaks are to be avoided. The large quantity of metal used results in a considerable quantity of gas being occluded, such gas being released during operation if flashing and evacuating processes are not accomplished at a suitable temperature. He demonstrated his remarks with many sample components and partially fabricated tubes. He stated that when manufacturing processes have been perfected it will be possible to build a far superior tube in metal than was ever possible in glass because of greatly lowered input shunt admittance, better shielding, and smaller elements.

Cleveland Section

The Cleveland Section met on September 26 at Case School of Applied Science. J. S. Hill, secretary, presided and thirty-eight were present.

A. W. L. Williams, President of the Brush Development Company, presented a paper on "Recent Improvements in Applied Piezoelectricity." He explained briefly how Rochelle-salt crystals are grown and cut into slabs. The crystals are impervious to ordinary ranges of temperature and humidity when properly treated. Bimorph elements made of two slabs with the C axis of one parallel to the B axis of the other will generate a potential between adjacent corners when the opposing sides are twisted in opposite directions. Phonograph pickups incorporate these elements in the shape of trapezoids. The long side is anchored while the needle is held on the short parallel side. This type of unit readily adapts itself to light weight of moving parts and is made in needle weights from two to six ounces to keep record wear low. High fidelity units are made for both lateral and vertical recordings with a half-ounce needle weight. Vertical recordings made with magnetic equipment require filtering to compensate for high frequencies when reproduced with the crystal pickup. Various types of crystal microphones were then described. The diaphragm type has no advantages over others of that variety. The crystal diaphragm type has many advantages in frequency response and application. Its construction was described. By the combination of several cells, a microphone may be made to be unidirectional, bidirectional, or nondirectional through a switching arrangement. Frequency response is controlled by the thickness and area of the crystal and the axis on which it is cut. Linear response must be sacrificed for high output. The meeting was concluded with a demonstration of various microphones and pick-ups with the new crystal headphone. The headphone features light weight, improved frequency response, high sensitivity, and a cone diaphragm.

New York Meeting

The October New York meeting of the Institute was held on the 23rd in the Engineering Societies Building. President Ballantine presided and 500 members and guests were present.

A paper on "The Secondary Emission Multiplier—A New Electronic Device" was presented by V. K. Zworykin, G. A. Morton, and L. Malter of the RCA Manufacturing Company.

In it the authors described the construction, theory, and performance of various types of fixed-field secondary emission multipliers. Detailed consideration was given of multiplier photocells employing crossed electrostatic and magnetic fields and of electron multipliers using electrostatic focusing alone, to serve as coupling and amplifying units for cathode-ray tubes such as the "iconoscope."

It was shown that while the power required for the operation of the secondary multiplier is about the same as that for the conventional amplifier, it is superior to the latter from the standpoint of noise. In the case of the multiplier photocell the signal-to-noise ratio is essentially determined by the shot noise of the photoemission, and is therefore sixty to one hundred times greater than that for a thermionic amplifier and photocell under conditions of low light intensity. Multiplier photocells have been built with an amplification factor of several millions and serve to replace the conventional photocell and accompanying amplifier system. Their low "noise" level, together with their excellent frequency response and extreme simplicity, make these electron multipliers a very satisfactory form of amplifier. A demonstration of the multiplier photocell was given after presentation of the paper.

The November 6 meeting of the Institute was held in the Engineering Societies Building in New York City and was presided over by President Ballantine. Seven hundred members and guests were present and a paper on "A Method of Reducing Disturbances in Radio Signaling by a System of Frequency Modulation" was presented by E. H. Armstrong, Professor of Electrical Engineering at Columbia University.

In it, Professor Armstrong presented a new method of reducing the effects of all kinds of disturbances. The transmitting and receiving arrangements of the system, which makes use of frequency modulation, were shown in detail. The theory of the process by which the noise reduction is obtained was discussed and an account was given of the practical realization of it in transmissions during the past year between the National Broadcasting Company's experimental station in the Empire State Building in New York City and Westhampton, Long Island, and Haddonfield, New Jersey. Finally, methods of multiplexing and the results obtained in these tests were reported.

A sound film record comparing reception of a forty-megacycle transmitter of the new type with reception from a regular broadcast station during a local thunderstorm over a transmission path of approximately ninety miles was given. An additional demonstration of the system was given by transmission from Yonkers, New York, to the Engineering Societies Building using the new system and a carrier frequency of 120 megacycles.

Philadelphia Section

On October 3 a meeting of the Philadelphia Section was held in the Engineers Club with Knox McIlwain, chairman, presiding. Two hundred and fifty members and guests attended and twenty-five were present at the informal dinner which preceded the meeting.

R. M. Wise, chief radio engineer of the Hygrade Sylvania Corporation, presented a paper on "Production of Metal Radio Tubes." In it he discussed the problem of manufacturing metal tubes and not their merits in relation to glass tubes. Their plant started producing metal tubes in May and by July the output was 30,000. This was raised to 600,000 for the month of October. The plant had to be practically re-equipped with new machinery and tools; large thyratron welders were installed for welding the metal shields to the headers. These welders use currents as high as 70,000 amperes and make 2000 welds per hour. Copper is used around glass beads to improve the weld of the lead-in wires and prevent leakage. All metal parts are fired in hydrogen to remove occluded gasses. A number of degassing problems remain to be answered and the best type of "getter" has not been found. The most recently used is a combination of barium, strontium, and magnesium in a copper envelope and placed to flash up along the sides of the steel shell. It requires from six to eight minutes to exhaust a tube. They cannot be heated above 1000 degrees as can glass. Full-wave rectifier tubes are hard to manufacture and the design of one type had to be changed. At present twenty-five per cent of the metal tubes manufactured are condemned on inspection while for glass the loss runs from five to ten per cent. The cost for labor and materials is at present about double that for glass tubes.

At the close of the paper, T. M. C. Lance of the Baird Television Company of London, gave a short interesting talk on the tube situation in England and he was followed by C. M. Spahr of San Francisco who discussed the situation on the Pacific Coast. S. Tarzian of the Atwater Kent Company exhibited and explained the operation of an Atwater Kent receiving set employing ten metal tubes.

Pittsburgh Section

On October 15 a meeting of the Pittsburgh Section was held at the Fort Pitt Hotel. Twenty-eight were present and R. D. Wyckoff, chairman, presided.

"The Use of Vacuum Tubes as Class C Amplifiers" was the subject of a paper by I. E. Mouromtseff, research engineer of the Westinghouse Electric and Manufacturing Company. A new and novel method of graphic representation of vacuum-tube operating characteristics was presented and explained. Comparative curves on the 848, 863, and 207 type tubes were shown and it was pointed out that the tubes with higher amplification factors were preferable as class C amplifiers from the standpoint of distortion. The paper was discussed by Messrs. Peters, Place, Sannergren, and Wyckoff.

San Francisco Section

"Ultra Short-Wave Practice" was the subject of a paper by F. C. Jones, consultant on radio design. This was presented at the October 16 meeting of the San Francisco Section, which was held in the Bellevue Hotel and R. D. Kirkland, chairman, presided. Thirty-one attended the meeting and nine were present at the informal dinner which preceded it.

Mr. Jones presented a brief history of ultra-short waves, an outline of refinement to date, probable future refinements, and present-day uses and advantages. He spoke in considerable detail on circuits, tubes, effective range at certain power outputs, stabilizing methods, antennas, and propagation characteristics. The paper was discussed by Messrs. Cone, Freiermuth, Harbridge, Kirkland, Sandfort, and Sharp.

Incorrect Addresses

A list of members whose incorrect addresses are unknown to the Institute is given on pages vii and viii of the advertising section. Any one who has more recent information as to their whereabouts will assist these members and the Institute by sending this information to the Secretary.

Personal Mention

J. C. Mevius formerly with WHAT has become manager of WEMP at Milwaukee, Wis.

S. C. Milbourne has left the Wholesale Radio Company to join the engineering staff of Supreme Instruments Corporation of Greenwood, Miss.

R. N. Palmer of Hygrade Sylvania Corporation has been transferred from Clifton, N. J. to Emporium, Pa.

H. L. Pitts, Lieutenant Commander, U.S.N., has been transferred from Washington, D. C., to the U.S.S. *Wright* with base at San Diego, Calif.

D. P. Poteet, Lieutenant, U.S.A., has received his captaincy and has been transferred to Fort Bragg, North Carolina.

E. H. Schoenfeld formerly with RCA Communications has joined the engineering staff of Heintz and Kaufman of South San Francisco, Calif.

R. A. Swan, Jr., of Hygrade Sylvania Corporation has been transferred from Emporium to Salem, Mass.

A. R. Taylor, Lieutenant, U.S.N., has been transferred from the U.S.S. *Augusta* to the U.S.S. *Smith Thompson*, basing at Seattle, Wash.

P. B. Taylor previously with the Federal Communications Commission has joined the staff of the Signal Corp of the Aircraft Radio Laboratory, Wright Field, Dayton, Ohio.

Ansel Challenner formerly with Bell Telephone Laboratories is now on the faculty of the college of engineering of the University of Oklahoma at Norman, Okla.

A. L. Green formerly of the University of Sydney has joined the Research Laboratories of Amalgamated Wireless (Australasia) at Sydney, Australia.

V. D. Hauck previously with DeForest Radio Company is now an engineer in the transmitter section of RCA Victor, Camden, N. J.

Formerly with the United American Bosch Company, K. L. Henderson has joined the staff of Stromberg-Carlson Telephone Manufacturing Company, Rochester, N. Y.

V. N. Janes formerly with WLB has joined the engineering staff of RCA Manufacturing Company in Camden.

W. C. Lent previously with General Communications Laboratories has joined the staff of the National Broadcasting Company of New York City.

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TECHNICAL PAPERS

AUTOMATIC SELECTIVITY CONTROL*

By

G. L. Beers

(RCA Victor Division, RCA Manufacturing Company, Camden, N. J.)

Summary—*A receiving system is described in which the selectivity automatically varies with the sensitivity. The variation in selectivity is obtained through the use of several triodes whose plate-to-cathode impedance is shunted across a number of the receiver tuned circuits. The circuits with which these selectivity control tubes are associated are described. Circuit diagrams illustrate the method by which the control-grid potentials of the automatic selectivity control tubes are caused to vary with the strength of received signals. A manual control is described which permits the user to increase the selectivity over that determined by the strength of the received signal. Curves are given showing the change in over-all selectivity with sensitivity, the relation between fidelity and sensitivity, and distortion as a function of power output.*

Introduction

The recent trend towards increasing the modulation frequency band which a broadcast receiver will accept has presented an additional problem in the design of such receivers. In order that a receiver which under favorable conditions will give satisfactory reproduction of modulation frequencies up to 8000 cycles will also be capable of giving equally satisfactory reception of weak signals, it is essential that the receiver be equipped with some means for varying its selectivity. Unless the selectivity can be increased when weak signals are received, excessive high-frequency noise and cross talk will be obtained. The high-frequency noise can be minimized by restricting the frequency response range of the audio-frequency portion of the receiver. This expedient, however, does not reduce the cross talk which may result from the relatively wide band of frequencies which is permitted to reach the detector. Both high-frequency noise and cross talk can be minimized by increasing the receiver selectivity. It is therefore desirable that a receiver be equipped with some means whereby its selectivity can be increased as its sensitivity is increased if the maximum fidelity consistent with reasonable freedom from cross talk and high-frequency noise is to be obtained.

This interdependence of sensitivity and useful selectivity in a

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broadcast receiver suggests the possibility of automatic selectivity control. An automatic selectivity control (A.S.C.) has two distinct advantages over the manual arrangement for varying selectivity. The first advantage is that the user of a receiver equipped with it is prevented from obtaining excessive noise when receiving weak signals. The second advantage is that if an automatic selectivity control receiver is used to receive signals which are subject to periods of fading, a considerably better signal-to-noise ratio will be maintained than if the receiver is equipped with only automatic volume control.

During the past several years a number of automatic selectivity control systems have been investigated. It is the purpose of this paper to describe one of the systems which has been incorporated in an experimental receiver.

The Automatic Selectivity Control Circuits

The automatic selectivity control system employed in this receiver obtained its control of selectivity by varying the plate impedance of several triodes in whose plate circuits were included a like number of

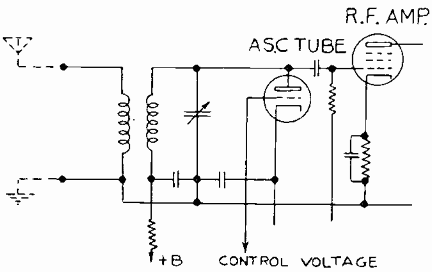


Fig. 1—Antenna transformer.

tuned circuits. These tuned circuits were used in the antenna coupling transformer and the first and second intermediate-frequency transformers. Since the control circuits employed with these transformers differed considerably, a brief description of each circuit will be given.

Fig. 1 shows the circuit used to control the selectivity of the antenna coupling transformer. The single tuned circuit employed in this transformer was tuned throughout the broadcast frequency band in the conventional manner. As indicated in the figure, a triode was so connected that its plate impedance shunted the tuned circuit. The negative potential on the control grid of this triode was varied to cause its plate impedance to change from 10,000 ohms to one megohm. The effect of this change in impedance on the selectivity characteristic of the transformer is shown by curves *a* and *b* in Fig. 2. The voltage step-up from the antenna to the grid of the first tube decreased from 4.4 to

0.75 with this reduction in selectivity. The gain and selectivity provided by this transformer increased with an increase in the plate impedance of the automatic selectivity control tube.

The first intermediate-frequency transformer employed two coupled tuned circuits. An automatic selectivity control tube was associated with each tuned circuit as shown by Fig. 3. Litz coils were used in both tuned circuits to provide the desired selectivity. The coupling between

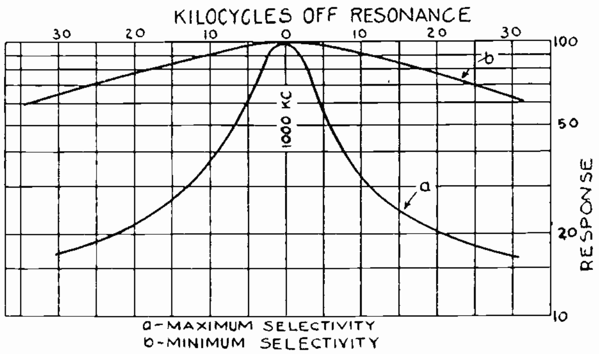


Fig. 2—Selectivity of antenna transformer.

the two circuits was adjusted to be just below the critical value when the plate impedance of the automatic selectivity control tubes was maximum. The selectivity characteristic of the transformer under this

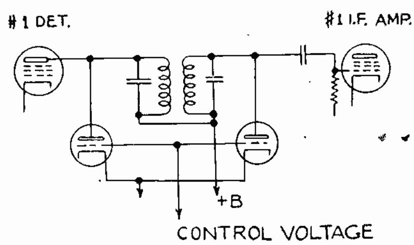


Fig. 3—No. 1 intermediate-frequency transformer.

condition is given by curve *a* in Fig. 4. Curve *b* in this figure shows the change in selectivity obtained by decreasing the plate impedance of the automatic selectivity control tubes to 10,000 ohms. The voltage gain provided by this transformer and a 58 amplifier tube decreased from 50 to 0.33 with this reduction in selectivity.

The circuit employed to vary the selectivity of the second intermediate-frequency transformer is illustrated by Fig. 5. This transformer likewise employed two coupled tuned circuits. In this case, however, an automatic selectivity control tube was used with only the secondary. The transformer primary was wound with Litz wire to minimize the primary tuned circuit losses. The secondary inductance was kept as high as consistent with satisfactory frequency stability. The coupling between the primary and secondary of this transformer was adjusted so that it passed the frequency band desired for maximum

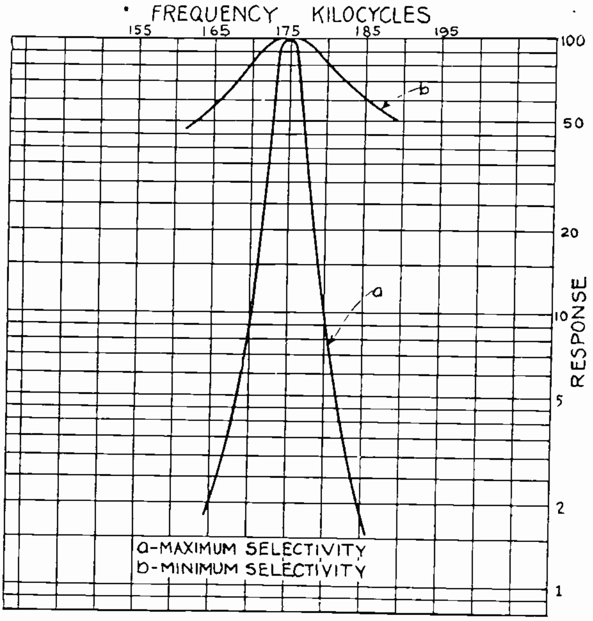


Fig. 4—Selectivity No. 1 intermediate-frequency transformer.

fidelity when the plate impedance of the automatic selectivity control tube was a maximum. A resistor was shunted across the transformer secondary to give the top of the selectivity characteristic the desired flatness. When the plate impedance of the automatic selectivity con-

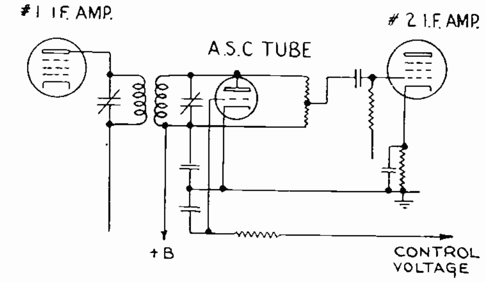


Fig. 5—No. 2 intermediate-frequency transformer.

trol tube used with this transformer was decreased, the effective primary resistance introduced by the tuned secondary decreased and the selectivity characteristic of the transformer became sharper and approached the selectivity characteristic of the low loss tuned circuit primary when the automatic selectivity control tube plate impedance was reduced to 10,000 ohms.

This change in selectivity is illustrated by the curves in Fig. 6. Curve *a* shows the maximum selectivity provided by the transformer. This curve was taken when the plate impedance of the automatic selectivity control tube was 10,000 ohms. The selectivity characteristic obtained when the plate impedance of the automatic selectivity control tube was increased to its maximum value is shown by curve *b*. The decrease in selectivity illustrated by these curves was accompanied by a 5-to-1 increase in the voltage gain provided by the amplifier stage in which this transformer was employed.

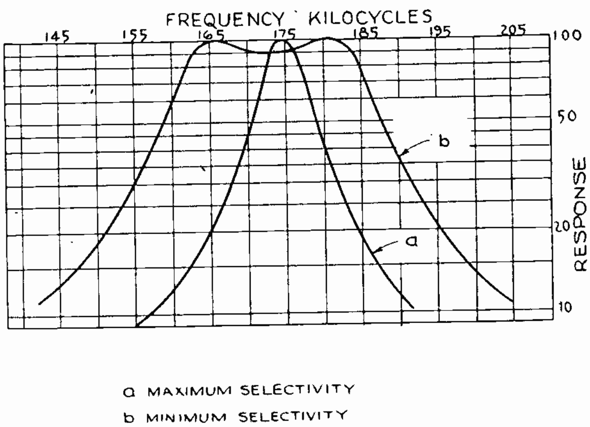


Fig. 6—Selectivity No. 2 intermediate-frequency transformer.

Decreasing the plate impedance of the automatic selectivity control tubes associated with the antenna coupling and first intermediate-frequency transformers caused a reduction in both the voltage gain and selectivity provided by these transformers. Decreasing the plate impedance of the automatic selectivity control tube used with the second intermediate-frequency transformer, however, caused a reduction in its voltage step-up but increased its selectivity. In order to cause a simultaneous reduction in the selectivity of the three transformers it was necessary that the plate impedance of the radio-frequency and first intermediate-frequency automatic selectivity control tubes be decreased while that of the second intermediate-frequency automatic selectivity control tube was increased. The reason for using transformers with different gain vs. selectivity characteristics will be discussed later.

Fixed Selectivity Circuits

In addition to the automatic selectivity control circuits which have been described the receiver contained three transformers whose selectivity characteristics were not varied. A coupled tuned circuit tuned radio-frequency transformer designed to pass a frequency band of sixteen kilocycles was used between the radio-frequency amplifier tube and the first detector. Two intermediate-frequency transformers which were also designed to pass a sixteen-kilocycle frequency band were used to provide the interstage coupling between the second and third intermediate-frequency amplifier tubes and between the third amplifier tube and the second detector. These transformers were employed so that the receiver would have adequate selectivity when strong signals were received and the automatic selectivity control circuits were adjusted to accept a wide modulation frequency band.

The amplifier stages in which these transformers were employed were designed to give a relatively high voltage gain. The gain in the amplifier stages with which the automatic selectivity control tubes were associated was kept quite low, in order to limit the receiver sensitivity to a practical value and to keep the signal potentials applied to the plates of the automatic selectivity control tubes sufficiently low to avoid the possibility of distortion.

Sources of Control Grid Potential for the Automatic Selectivity Control Tubes

The automatic selectivity control tubes were operated with a fixed plate potential and the plate impedance of these tubes was altered by varying their control grid bias. In order to cause each of the transformers with which the automatic selectivity control tubes were associated to reduce simultaneously the receiver selectivity, it was necessary to decrease the negative control grid bias of three of the automatic selectivity control tubes, while that of the fourth was increased. To fulfill this requirement a master control tube and an auxiliary control tube were used. The master control tube supplied a negative potential which increased with an increase in the strength of a received signal while the negative potential derived from the auxiliary control tube decreased with an increase in signal strength. The circuit in which these control tubes were employed is illustrated by Fig. 7. Referring to this diagram the control grid of the master control tube (A) was directly connected to the cathode of the diode (D). The diode output resistor R_1 and the cathode of the master control tube were connected to the B supply system at points which were sufficiently different in potential to bias the master control tube beyond the point of plate current cutoff when no signal was being received. The control grids of both the auxiliary control tube B and the second intermediate-frequency automatic selectivity control tube C were connected to the anode end of the master control tube output resistor R_2. The cathodes of these two tubes were connected to the *B* supply system at the same point as the anode of the master control tube. The plate potential for these tubes was adjusted so that when no signal was being received the plate current of the auxiliary control tube was four milliamperes and the plate impedance of the second intermediate-frequency automatic selectivity control tube was 10,000 ohms. As previously explained, the selectivity of the second intermediate-frequency transformer was a maximum when the plate impedance of the associated automatic selectivity control tube was 10,000 ohms.

The control grids of the radio-frequency and first intermediate-frequency automatic selectivity control tubes *E*, *F*, and *G* were connected to the anode of the auxiliary control tube which in turn was connected to the *B* supply system through a 6000-ohm resistor R_3. The cathodes of these automatic selectivity control tubes were connected to the *B* supply end of this resistor. The four-milliampere current flowing through the auxiliary control tube plate resistor impressed a negative bias of twenty-four volts on the grids of the radio-frequency and first intermediate-frequency automatic selectivity control tubes. The plate impedance of these tubes due to this negative control grid potential was sufficiently high to cause no reduction in the selectivity of the circuits with which these tubes were associated. Thus the selectivity of all the automatic selectivity control transformers was a maximum when no signal was being received.

The signal voltage developed across the primary of the last intermediate-frequency transformer was impressed on the diode *D* through a coupling condenser. Since the control grid of the master control tube was connected to the cathode end of the diode output resistor, a signal applied to the diode caused the grid of the master control tube to become less negative. When the signal level at the diode was sufficient to cause the control-grid potential of the master control tube to become positive with respect to the point of plate current cutoff, current flowed through resistor R_2. This current varied in accordance with the strength of the signal applied to the diode *D*. The flow of current through resistor R_2 caused an increase in the negative bias on the grids of the auxiliary control tube and the second intermediate-frequency automatic selectivity control tube.

The increase in negative bias on the grid of the second intermediate-frequency automatic selectivity control tube caused the selectivity of the second intermediate-frequency transformer to decrease. Increasing the negative potential on the grid of the auxiliary control tube caused a reduction in its plate current and thus decreased the negative bias on the radio-frequency and first intermediate-frequency automatic selectivity control tubes. The selectivity of the circuits with which these tubes were used was thus reduced. Since the current through the master control tube plate resistor varied with the strength of received signals, the antenna coupling transformer and the first and second intermediate-frequency transformers were caused to decrease simultaneously the receiver selectivity as the strength of the received signal was increased.

Control of Relation Between Field Strength and Fidelity

It has previously been mentioned that as the selectivity of the automatic selectivity control transformers was reduced the voltage gain in the radio-frequency and first intermediate-frequency transformers decreased while that of the second intermediate-frequency transformer increased. If the gain of all the transformers had decreased as the selectivity was reduced, the signal strength required to cause the receiver to give maximum fidelity would have been much greater than desired. The use of transformers having these opposing gain versus selectivity characteristics made it possible to cause the receiver fidelity to be a maximum for all signal strengths in excess of ten millivolts, without increasing the maximum sensitivity of the receiver beyond a reasonable value. By utilizing the potential developed across the master control-tube output resistor as automatic volume control bias for one or more of the amplifier tube control grids, it was possible to vary the relation between receiver sensitivity and fidelity through fairly wide limits.

Delayed Automatic Selectivity Control

Hiss measurements made on a receiver employing the circuit illustrated by Fig. 7 showed that when the sensitivity was high the signal-to-hiss ratio obtained did not compare favorably with that derived from a standard receiver. This excess hiss was caused by the automatic selectivity control tube shunting the first tuned circuit. Since the plate current of this tube flowed through the first tuned radio-frequency circuit, a shot effect voltage was impressed on the grid of the first radio-frequency amplifier tube. This difficulty was eliminated by delaying the action of the radio-frequency automatic selectivity control tube until the receiver's sensitivity was reduced to 500 microvolts. The manner in which this was accomplished is shown in Fig. 8. The resistance in the plate circuit of the auxiliary control tube was increased to 9000 ohms and the resistor was provided with a tap at 6000 ohms. The plate potential for the tube was increased to maintain the four-milliampere plate current when no signal was being received. The voltage drop across the entire resistor was applied to the grid of the radio-frequency automatic selectivity control tube while the drop

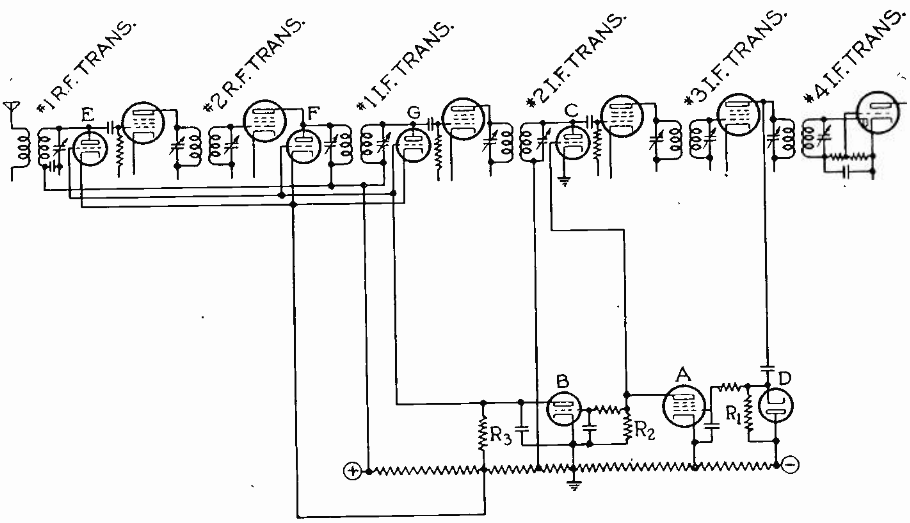


Fig. 7

across the 6000-ohm section was impressed on the grids of the first intermediate-frequency automatic selectivity control tubes. The potential on the grid of the radio-frequency automatic selectivity control

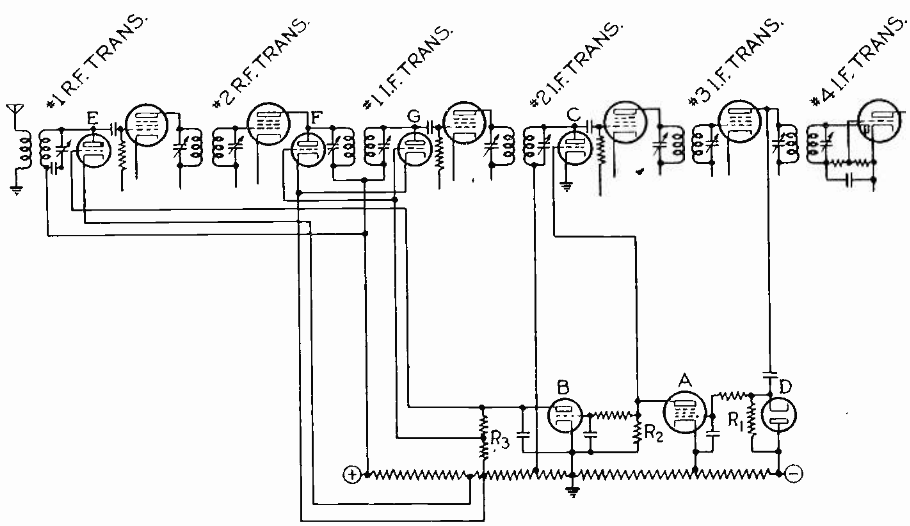


Fig. 8

tube with this arrangement was thirty-six volts with no signal applied to the receiver. The control-grid potential of the first automatic selectivity control tubes was still twenty-four volts as was employed in the first system. Neither the radio-frequency nor the first intermediate-frequency automatic selectivity control tubes produced any decrease in the receiver's selectivity until the control-grid bias for these tubes was reduced to less than twenty-four volts. When a received signal caused the negative bias on the radio-frequency automatic selectivity control tube to decrease to twenty-four volts the control-grid potential for the first intermediate-frequency automatic selectivity control tubes had been reduced to sixteen volts and a considerable reduction in the receiver's sensitivity had occurred. A further increase in the strength of a received signal caused all the automatic selectivity control tubes to decrease simultaneously the receiver's selectivity.

It was also found desirable to limit the maximum effect of the radio-frequency selectivity control tube by providing it with a fixed bias of four volts, otherwise the gain between the antenna and the grid of the first tube would have been reduced considerably below unity and the selectivity would have been decreased more than necessary for the desired maximum fidelity.

Manual Adjustment of Selectivity

Tests on a receiver equipped with the automatic selectivity control arrangement, which has been described, indicated the need for some means for increasing the selectivity of the receiver under certain receiving conditions. In case it was desired to receive signals from a station operating on a channel adjacent to that of a strong local transmitter and the desired signal was relatively strong the receiver might decrease its selectivity sufficiently to permit considerable interference by the local transmitter. Abnormal interference conditions were occasionally encountered where increasing the selectivity of the receiver over that determined by the strength of the received signal would reduce the interference to the point where satisfactory reception could be obtained. Several arrangements which permitted the user of the receiver to increase manually its selectivity were considered before the one illustrated in Fig. 9 was adopted. This manual control permitted the user of the receiver to increase its selectivity up to the maximum value regardless of the strength of the received signal. The control, however, could not be used to decrease the receiver selectivity beyond that automatically determined by the signal strength. Automatic control of the receiver output was obtained regardless of the selectivity control setting. As shown by Fig. 9 the plate of the master control tube was connected to a potentiometer JKL through the two resistors R_5 and R_6. These resistors each had a resistance of 100,000 ohms while the potentiometer element JL had a resistance of 500,000 ohms. The end of the potentiometer designated L was connected to the grids of the intermediate-frequency amplifier tubes while the end marked J was connected to the control grids of the second intermediate-frequency selectivity control tube and the auxiliary control tube. The grid of the radio-frequency amplifier tube was directly connected to the anode of the master control tube. When the movable contact K was adjusted to the position L, the potential developed by the master control tube appeared across the resistor combination LJ and R_5 in parallel with R_6. The resistance of LJ was five times that of R_5, therefore over eighty per cent of the voltage was developed across LJ. Since the grids of the

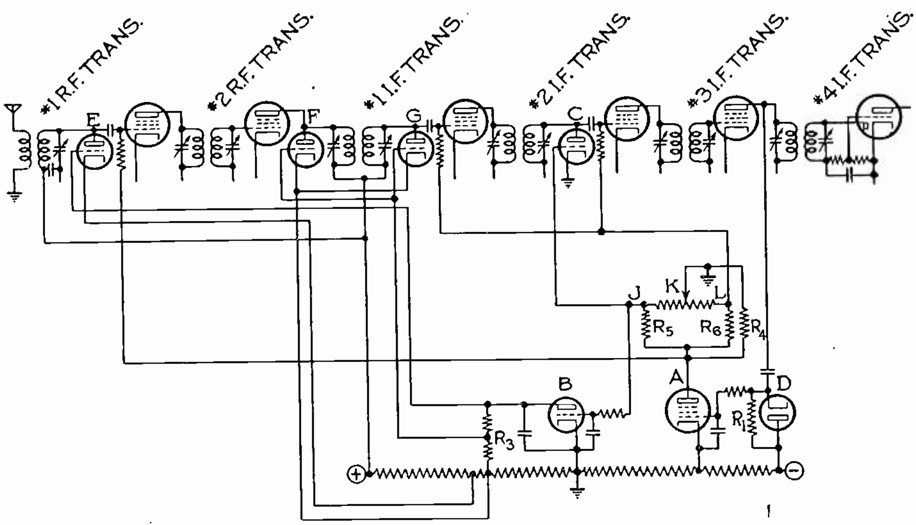


Fig. 9

automatic selectivity control tubes were connected to the point J, the receiver selectivity was varied in accordance with the strength of a received signal. No control potential was applied to the grids of the intermediate-frequency amplifier tubes, since the contact member K connected the point L to ground. By moving the contact element K to the end of the potentiometer designated J, the control potential developed by the master control tube was applied to the grids of the amplifier tubes and the volume control action was then obtained in the conventional manner. Under this condition the grids of the second intermediate-frequency automatic selectivity control tube and the auxiliary control tube received no control potential from the master control tube, since their grids were connected to ground by the contact device K. When the contact element K was set at intermediate positions between J and L, the relative control potentials applied to the grids of the two groups of tubes were determined by the ratio of JK to LK. Thus as the movable contact was varied from J to L the effect of the automatic selectivity control tubes on the receiver selectivity gradually diminished and the control of the receiver output through the variation of the amplifier tubes' control-grid potential correspondingly increased. The resistor R_4 was shunted from the plate of the master control tube to ground to keep the total resistance in the plate circuit of the master control tube reasonably constant when the contact element K was moved.

This system provided, through the use of a single potentiometer, a manual control for increasing the receiver selectivity without detracting from the performance of the automatic selectivity control system.

Audio-Frequency System

The audio-frequency amplifier employed in the receiver made use of a low-pass filter which provided an attenuation in excess of forty decibels for frequencies above 8000 cycles. The effectiveness of such a filter in minimizing high-frequency noise and "monkey-chatter" has already been discussed by Ballantine.[1]

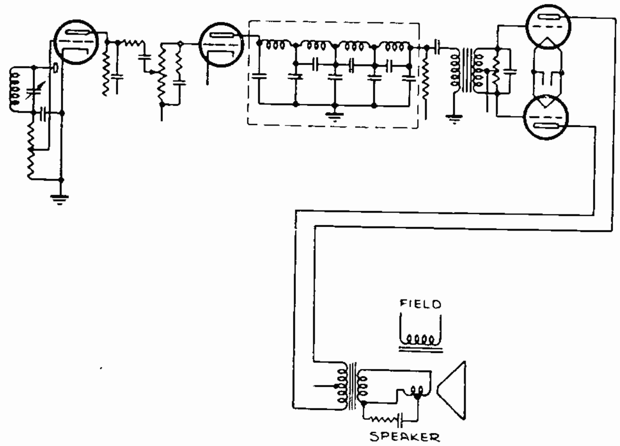


Fig. 10—Audio system.

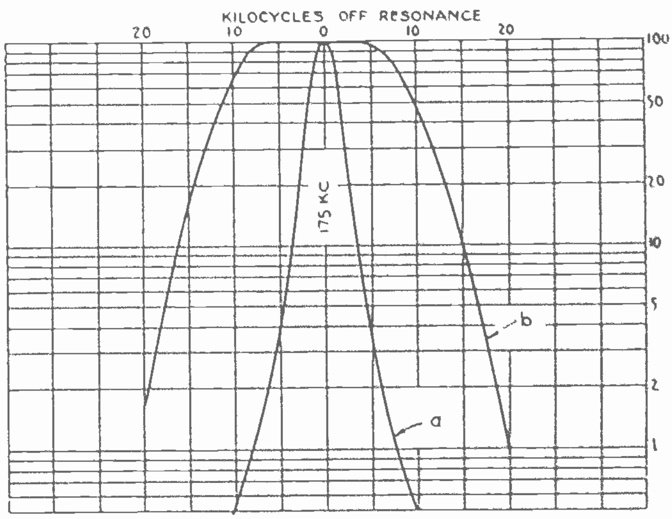
A schematic diagram of the audio-frequency system is shown in Fig. 10. The push-pull class A output stage employed two 2A3 tubes which were operated with semifixed bias.

Performance Characteristics

The effectiveness of the automatic selectivity control system may best be illustrated by several curves which show the various characteristics of the experimental receiver in which it was incorporated. The two curves in Fig. 11 give the effect of the automatic selectivity control system on the selectivity of the intermediate-frequency ampli-

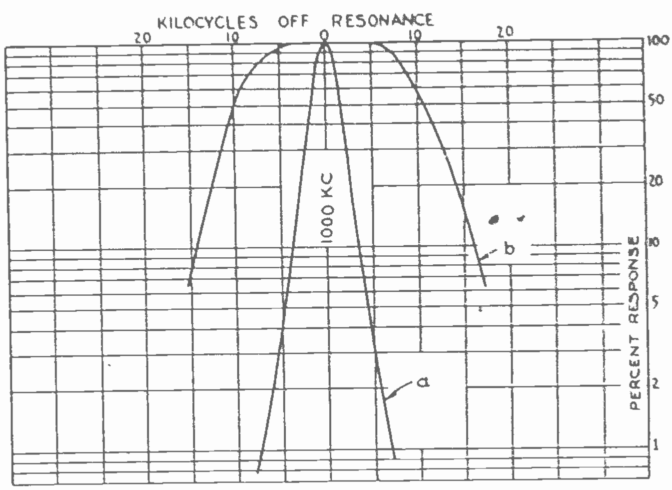
[1] Stuart Ballantine, "High quality radio broadcast transmission and reception," Proc. I.R.E., vol. 22, pp. 564–629; May, (1934).

fier. Fig. 12 shows the corresponding selectivity curves of the complete receiver when tuned to a frequency of 1000 kilocycles. These curves not only show the change in over-all selectivity which was produced



a MAXIMUM SELECTIVITY
b- MINIMUM SELECTIVITY

Fig. 11—Over-all intermediate-frequency selectivity.



a- MAXIMUM SELECTIVITY
b- MINIMUM SELECTIVITY

Fig. 12—Over-all selectivity.

by the automatic selectivity control system as the receiver's sensitivity was varied, but also indicate the limits through which the selectivity could be adjusted with the manual control when a strong signal was being received.

The receiver's over-all frequency response characteristic was likewise dependent on its sensitivity and this relationship is illustrated by

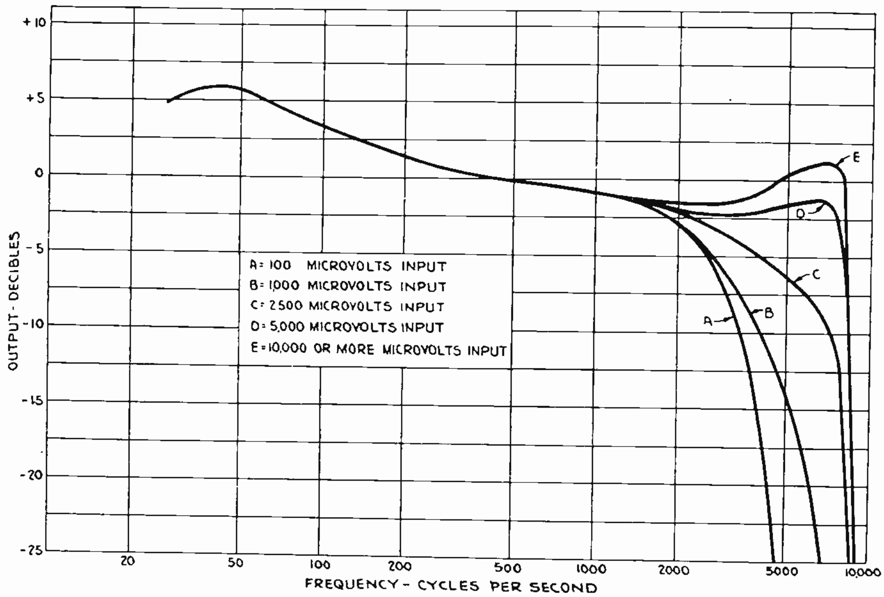


Fig. 13—Over-all fidelity.

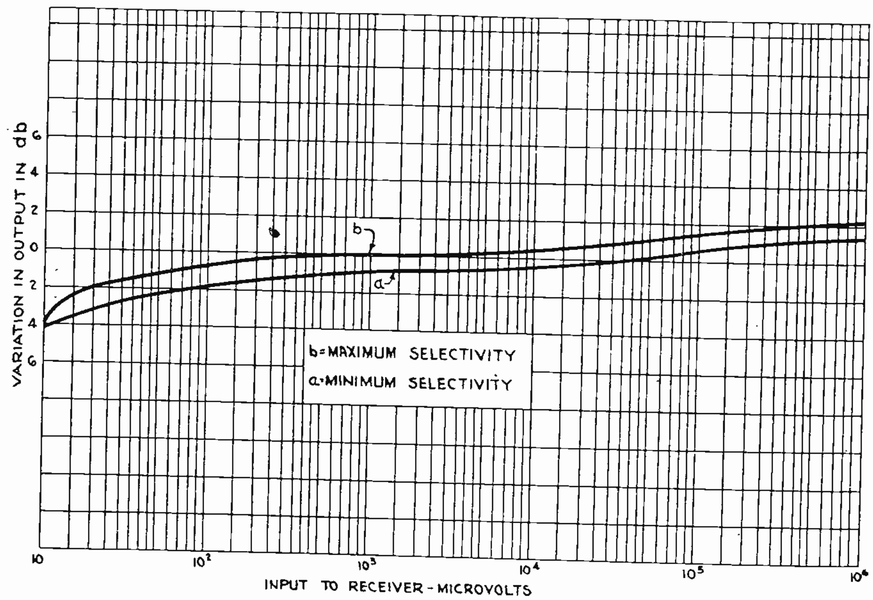


Fig. 14—Automatic volume control characteristic.

the curves in Fig. 13. It will be noted from these curves that the receiver was designed to give its maximum fidelity for all signal inputs in excess of ten millivolts. The factors which can be varied in the design of the automatic selectivity control system to alter the sensitivity-fidelity relationship have already been discussed.

Curve *a* in Fig. 14 illustrates the ability of the automatic selectivity control system to maintain a constant signal level at the second detector as the strength of the signal applied to the antenna terminals was varied. Curve *b* was obtained with the manual selectivity control adjusted for maximum selectivity. Under this condition the system functions as a conventional automatic volume control since the signal

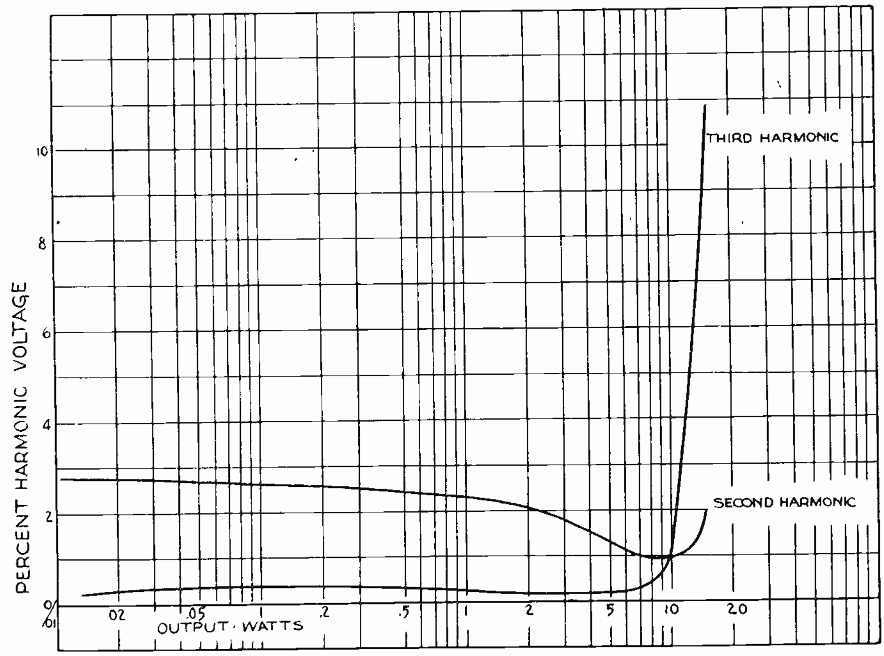


Fig. 15—Over-all distortion.

controlled bias potential was applied only to the grids of the amplifier tubes.

An indication of the distortion introduced by the receiver is given by the curves in Fig. 15. These curves were obtained by applying a ten-millivolt signal modulated 97 per cent, with 300 cycles to the antenna terminals of the receiver. The distortion in the signal generator was less than 0.5 per cent. The distortion was measured at the loud speaker terminals with a harmonic analyzer. Measurements made with signal inputs up to one volt gave substantially the same results as shown in this figure.

Conclusions

The effectiveness of the automatic selectivity control system in preventing excessive noise when receiving weak signals was demonstrated by numerous tests in which the signal-to-noise ratio obtained from the experimental receiver was compared to that derived from a standard receiver having a limited over-all frequency response characteristic. The signal-noise ratio obtained from the automatic selectivity control receiver during these tests was at least as good and sometimes better than that derived from the standard receiver. When strong signals were received from stations broadcasting high fidelity programs, the superior over-all response characteristic of the experimental receiver was quite evident.

Receiving conditions were occasionally encountered at night, which made it necessary to use the manual selectivity control to eliminate "monkey-chatter" interference. The effectiveness of an automatic selectivity control receiver in minimizing such interference may be increased by restricting the frequency response of the audio-frequency portion of the receiver simultaneously as the selectivity of the radio-frequency and intermediate-frequency amplifiers is increased. This result may be accomplished either by using automatic tone control in conjunction with the automatic selectivity control or through the use of a manual tone control which is so arranged that it is actuated simultaneously with the manual selectivity control of the automatic selectivity control system.

Acknowledgment

It is desired to acknowledge the valuable assistance of Messrs. W. R. Koch, G. L. Grundmann, and F. E. Cone in the development of this system.

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PHOTORADIO APPARATUS AND OPERATING TECHNIQUE IMPROVEMENTS*

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Summary—*A brief review of photoradio inception and progress up to 1928 is included in the introductory part of the paper.*

Improvements to terminal equipment which make possible greater fidelity of half-tone transmission over long-distance radio circuits are described. Radio circuit distortion is discussed and compensation methods suggested. A mathematical analysis of the photoradio keying is appended.

Introduction

The present photoradio system of RCA Communications, Inc., had its inception in 1923. R. H. Ranger has outlined in three papers[1,2,3] given before the Institute, the activity of early and contemporary workers in the facsimile field as well as covering the activities of the Radio Corporation of America up to and including 1928. It is the intent of the authors of this paper to list and describe the improvements to photoradio apparatus and in operating technique which have taken place since 1928.

I. General Description of the Modern Photoradio System

The photoradio equipment employed by RCA Communications, Inc., makes use of a rotating drum and a lead screw upon which is mounted either a photocell scanner or a recording device. The subject drum and its associated lead screw are driven, through suitable reduction gears, by a thermionic brake system. The driving unit of the thermionic brake system consists of an alternator and a direct-current motor mounted on a common shaft. The alternator supplies anode voltage and current to a pair of triodes connected in push-pull. The excitation for the push-pull connected triodes is furnished by a tuning fork standard having a high order of accuracy.

The transmitting head or scanner consists of a balanced modulator circuit, the phototube portion of which is energized by light reflected

* Decimal classification: R581. Original manuscript received by the Institute, May 15, 1935. Presented before New York Meeting, November 7, 1934.

[1] R. H. Ranger, "Transmission and reception of photoradiograms," Proc. I.R.E., vol. 14, pp. 161–180; April, (1926).

[2] R. H. Ranger, "Photoradio developments," Proc. I.R.E., vol. 17, pp. 966–984; June, (1929).

[3] R. H. Ranger, "Mechanical developments of facsimile equipment," Proc. I.R.E., vol. 17, pp. 1564–1575; September, (1929).

from the subject drum. The scanner output is a carrier, six decibels below reference value of six milliwatts, amplitude modulated in direct proportion to the subject light densities reflected into the phototube.

Short-wave radio circuit distortion[4] necessitates the conversion of the amplitude modulation of the scanner output into a series of varying length dots which can be used as control for a class C operated radio

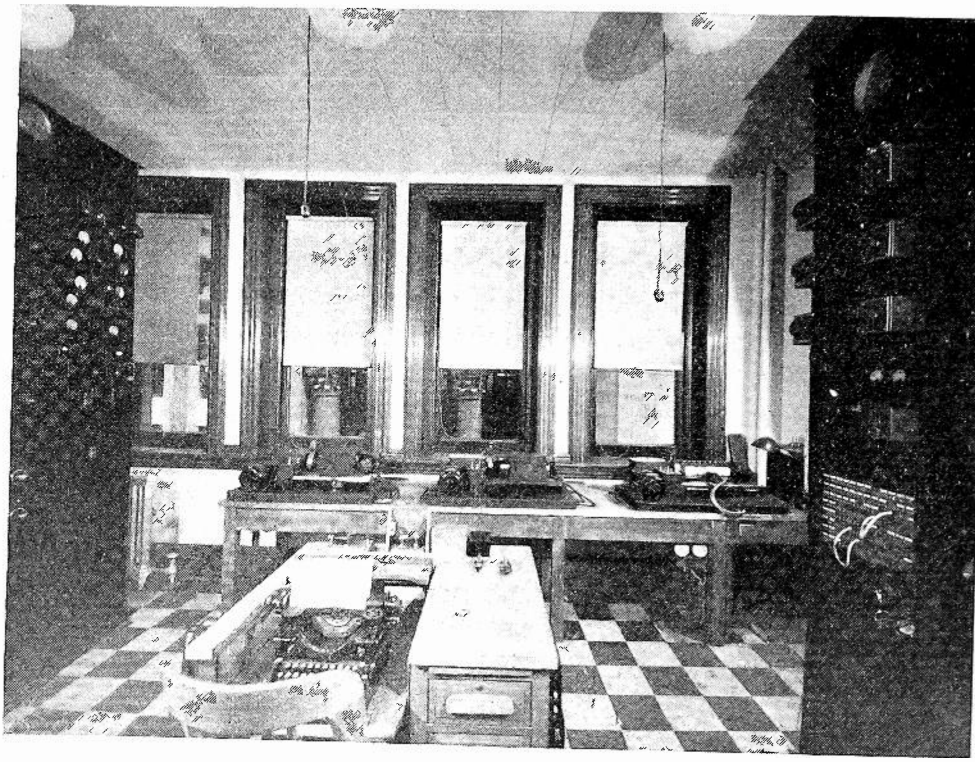

Fig. 1—Photoradio installation, 66 Broad St., New York City.

transmitter. The dot converter supplies to the radio-frequency transmitter control line an on-off keyed carrier, the keying rate and amplitude of which are held constant with weight or length of the mark intervals varying in direct proportion to the scanner output. This form of keying is called "Constant-Frequency Variable-Dot" or "CFVD" keying.

Photoradio service makes use of the standard radio transmitting and receiving station facilities normally used for radiotelegraph communication. These arrangements have been reported on by Beverage, Hansell and Peterson.[5]

[4] T. L. Eckersley, "Multiple signals in short-wave transmission," Proc. I.R.E., vol. 18, pp. 106–122; January, (1930).

[5] Beverage, Peterson, and Hansell "Radio plant of R.C.A. Communications, Inc.," *Proc. A.I.E.E.*, vol. 51, no. 8, p. 586; August, (1932).

The radio receiving station delivers to the Central Office a "CFVD" keyed carrier. The carrier is then rectified and supplied to either or both visual and photographic types of recorders. In the visual recording system the rectified receiving station signal controls a shutter which regulates the flow of ink vapor directed toward the record sheet. In the case of photographic recording the rectified signal is caused to flash

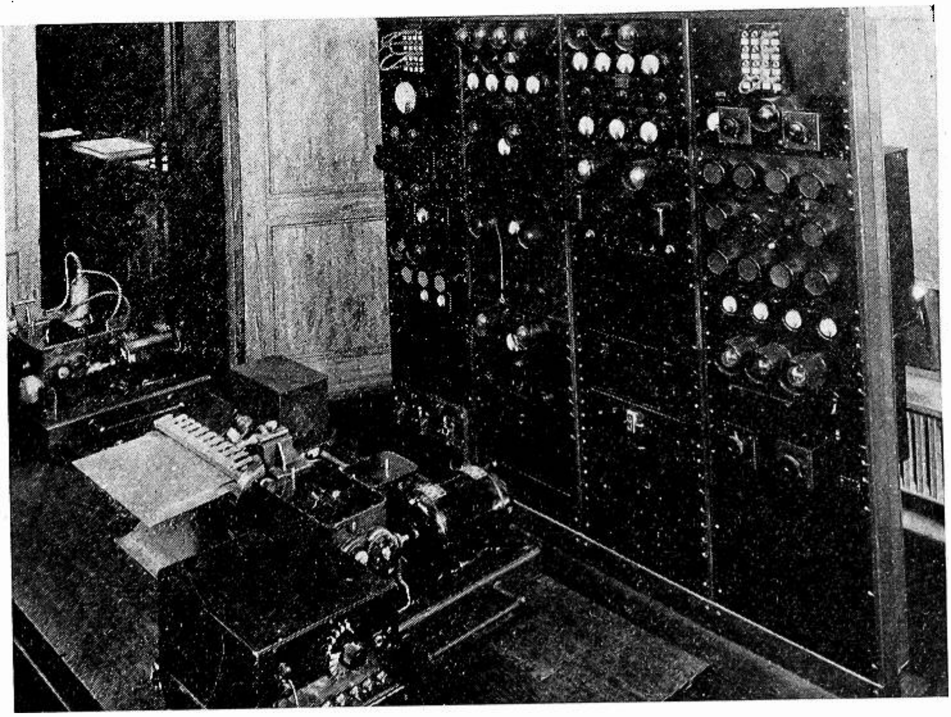

Fig. 2—Photoradio installation, 28 Geary St., San Francisco.

a glow tube, the flashes being directed by a suitable optical system, toward a record surface of either photographic film or bromide paper.

The apparatus associated with the above briefly described photoradio system is illustrated by several pictures. They are identified as Figs. 1 to 5 inclusive.

Fig. 1 is a photograph of the RCA Communications, Inc., photoradio apparatus room at 66 Broad Street, New York City. The machine bench, the fork standard equipment rack on the left, and the transmitting and receiving vacuum tube equipment rack on the right can be noted.

Fig. 2 is a photograph of the RCA Communications, Inc., photoradio apparatus installed at 28 Geary Street, San Francisco. The machine bench and the apparatus rack as well as the door leading into the dark room can be noted.

Fig. 3 is a photograph of the Cables and Wireless, Ltd., photoradio apparatus room at London. The machine equipment is shown in-

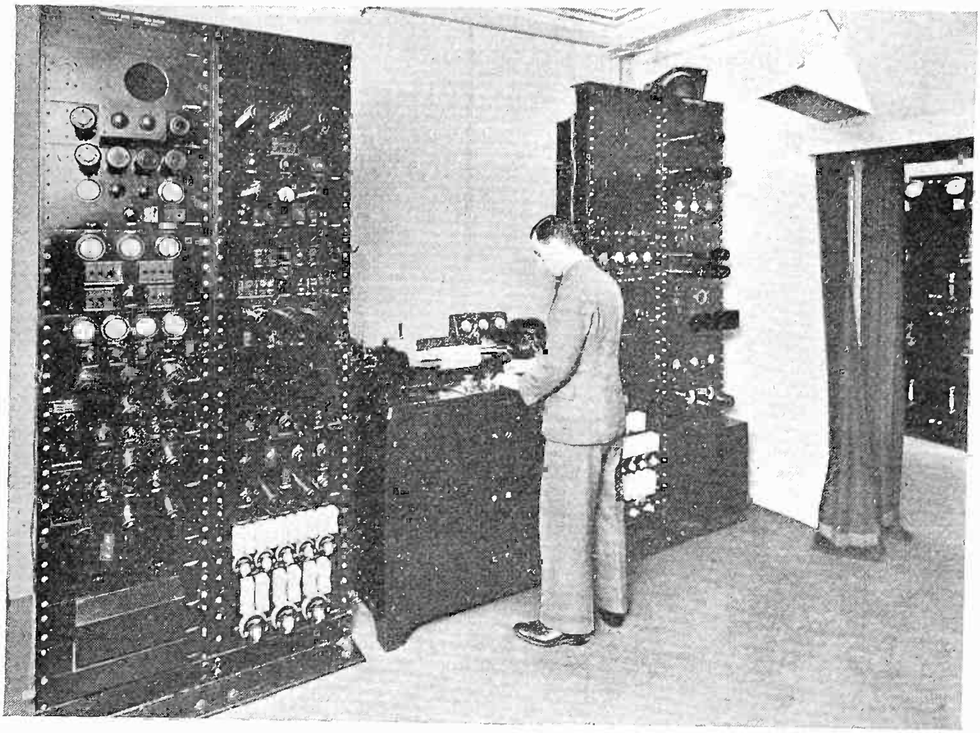

Fig. 3—Cables and Wireless, Ltd., London photoradio installation.

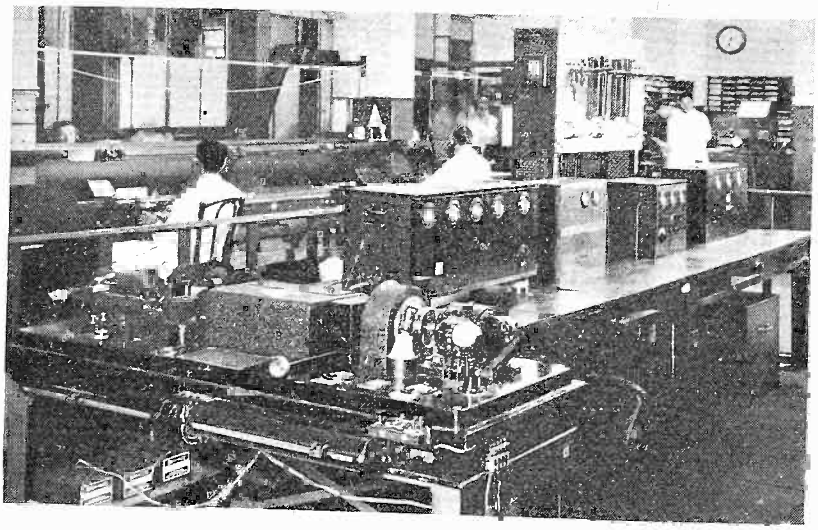

Fig. 4—Transradio Buenos Aires photoradio installation.

stalled between two apparatus racks, which hold the fork standard and the transmitting and receiving vacuum tube apparatus.

Fig. 4 is a photograph of the Transradio photoradio installation at Buenos Aires. The machine apparatus is shown in the foreground with the vacuum tube apparatus mounted on a table just to the rear.

Fig. 5 is a photograph of the Reichpost photoradio and telephoto apparatus room at Berlin. Apparatus used for continental wire-line and radio service is also shown.

A more detailed description of the component parts of the system follows.

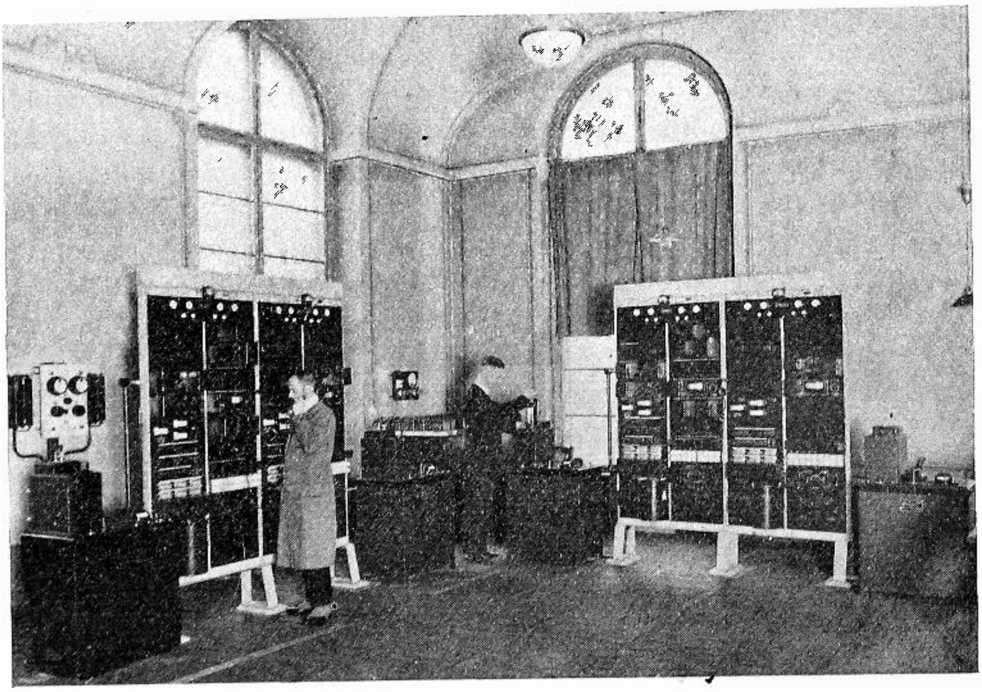

Fig. 5—Reichspost Berlin photoradio installation.

II. International Equipment and Operating Standards

Coöperative operation of the present international photoradio circuits by RCA Communications, Inc., and the several foreign administrations, makes necessary equipment and operating standards. It is impossible to expect all interested administrations to agree absolutely on design and manufacturing details. Agreement on basic design points with the various equipments arranged to meet an "Index of Coöperation," will make possible commercial results.

There are four basic design points upon which agreement is necessary if successful international photoradio operation is to be experienced. International agreement has been reached on these four points as indicated below:

1. A drive accuracy of at least one part in 100,000.
2. A drive source correction of seven parts in 100,000.
3. A positive means of pretransmission phasing. This can either be automatic or manual.
4. Agreement on at least two drum speeds is provided. These speeds are 30 and 60 revolutions per minute.

In addition to agreement on the above design points, it is necessary in order to allow reasonable freedom of choice to the individual companies in the matter of drum diameter and line advance pitch, to agree on an "Index of Coöperation." A standard Index of Coöperation permits two stations having different drum diameters, for example, to work with one another without introducing distortion, since only an enlargement or reduction of the transmitted material will take place. The *basic* Index of Coöperation can be defined as the product of the stroke length or drum circumference times the line advance, the units of both being constant. The *international* Index of Coöperation is 352, which is obtained by multiplying the drum diameter by the line advance which is equivalent to the *basic* index divided by π. The Index figure must be met within one per cent in order to hold the distortion within commercial limits. Table I illustrates the importance of the Index of Coöperation principle.

TABLE I

	RCAC	*LONDON*	*BERLIN*
Diameter of cylinder	74.54 mm	88 mm	66 mm
Line advance (Lines per mm)	4.72, 5.90, 9.45	4, 5, 8	5-1/3
Index of Coöperation	352, 440, 703	352, 440, 704	352

It will be observed from this table that although three different drum diameters are used, nevertheless, because of a common Index of Coöperation of 352 between RCAC, London and Berlin, these three stations can work with one another. In addition, RCAC can work with London using two other values of line advances since again they have a common Index of Coöperation.

International coöperation on photoradio standards has been through the medium of the CCIT (International Consulting Committee on Telegraph Communications), Berne. The suggestions on standards advanced by interested parties can be found in the CCIT Committee reports.[6,7]

[6] CCIT Third Conference, Berne, May, 1931.
[7] CCIT Fourth Conference, Prague, August, 1934.

III. Synchronizing System

1. *Brake Control*

The arrangement employed to furnish drive power to the subject drum and line advance gear boxes consists of a special driving unit and thermionic brake.

The driving unit consists of a shunt-wound direct-current motor and an inductor alternator, the combination mounted on a common shaft. With the common shaft turning at its operating speed of 1800

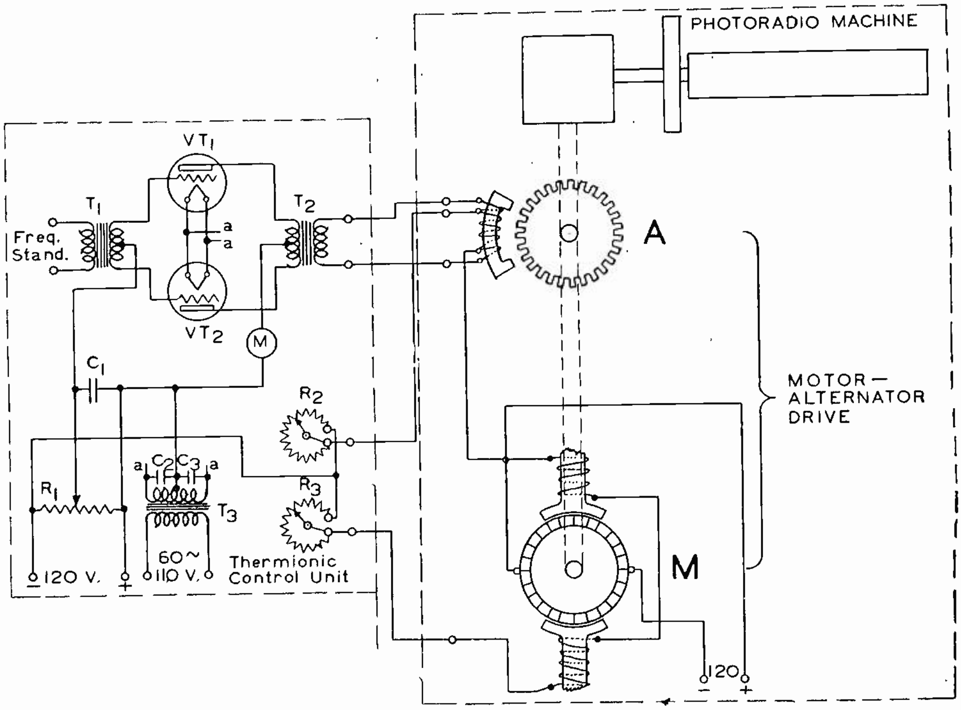

Fig. 6—Schematic layout of thermionic brake.

revolutions per minute, the alternator will generate a frequency of 810 cycles per second, which is the same as that delivered by the frequency standard.

The thermionic brake consists of a pair of vacuum tubes connected in push-pull with transformer coupling between the anodes and the alternator, and between the frequency standard and the control grids.

Fig. 6 illustrates the drive arrangement schematically. The control grids of VT_1 and VT_2 are normally biased to cutoff. When tone excitation is supplied through input transformer T_1, the control grids of VT_1 and VT_2 are alternately blocked and unblocked. The anode voltage for VT_1 and VT_2 is supplied through transformer T_2 from the armature windings of the alternator A. Motor M of the driving unit

A-M, is adjusted so it will run slightly faster than the synchronous speed, which is 1800 revolutions per minute. With constant 810-cycle excitation of a high order of accuracy supplied to the control grids of VT_1 and VT_2, the action of the anode circuit on the alternator will furnish the desired braking by varying the flux through the pole pieces shown in conjunction with *M*.

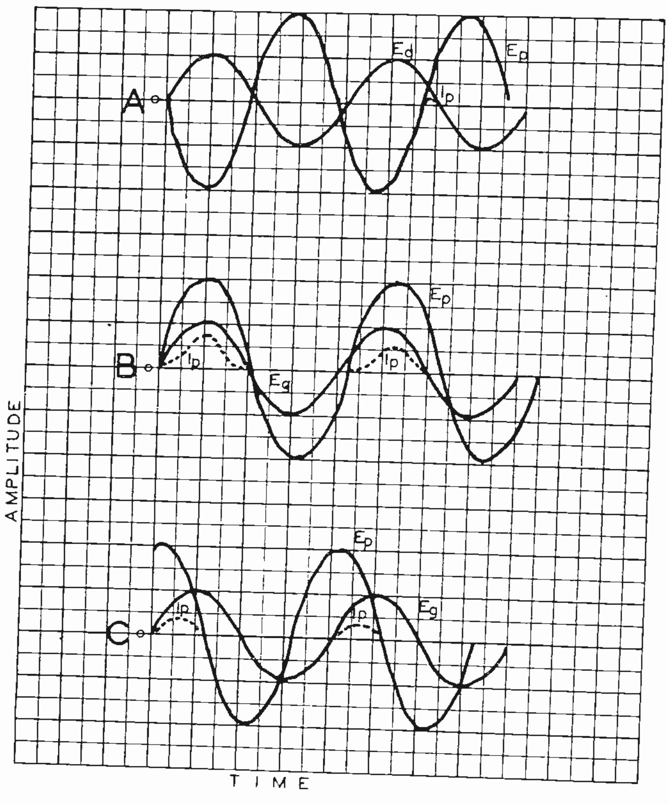

Fig. 7—Characteristic curves for thermionic brake.

Fig. 7 illustrates in graphic form the action of the brake circuit. Curve *A* is a possible condition immediately after the system has been switched on. The anode voltage E_p in this case has started exactly 180 degrees out of phase with the standard frequency grid control voltage E_g and is of course slightly higher in frequency. The anode current I_p is thus practically negligible at the start, therefore there is very little load on the alternator and the driving unit is free to run ahead of synchronous speed. As this condition continues, however, E_p draws more and more nearly into phase with E_g and I_p and thus the alternator load increases to a value sufficient to slow the motor below synchronous speed, producing a condition illustrated by curve *B*. Since the alternator output E_p is now of lower frequency than E_g, the two voltages will once more slip out of phase, and the driving unit, being more lightly loaded, will again speed up. This condition results in mechanical oscillation or "hunting." The constant of the system is such that the oscillations are highly damped, and the system soon reaches the condition of equilibrium illustrated by curve C. In this state the speed of the driving unit is exactly synchronous, so that a

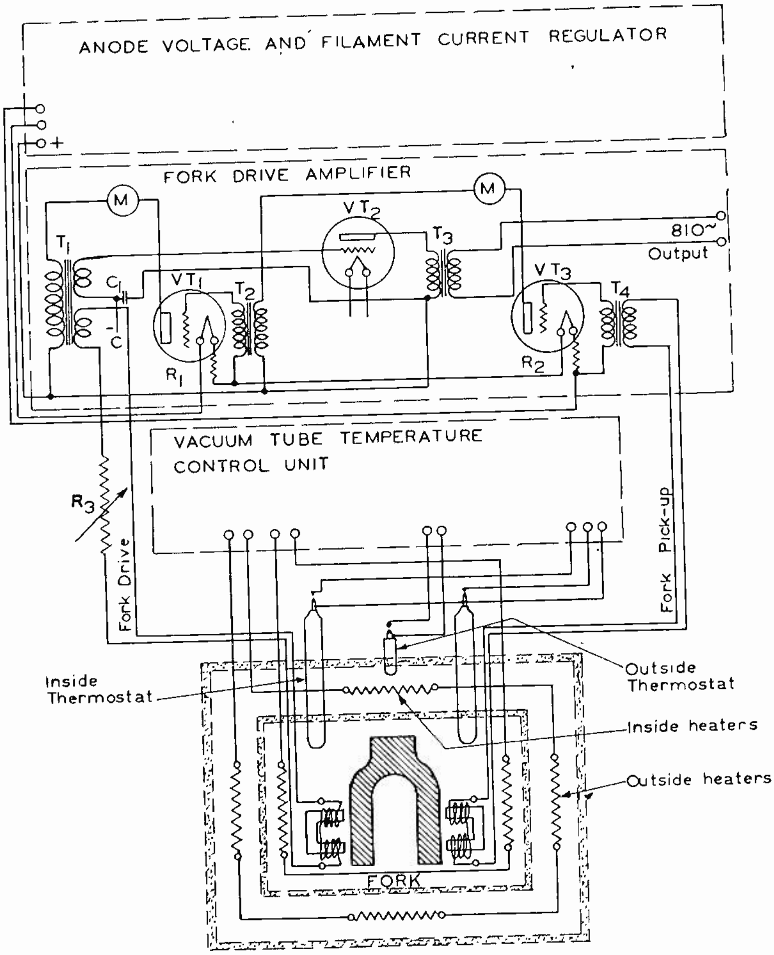

Fig. 8—Schematic layout of frequency standard.

steady load I_p is supplied by the alternator, its value depending on the phase angle between E_p and E_g. It follows that at equilibrium the phase angle between the synchronizing tone and the alternator electromotive force must supply a load current of such value as to maintain the equilibrium phase angle.

2. *Frequency Standard*

In order to maintain the drive accuracy specified by international standards, a temperature controlled tuning fork with vacuum tube drive is used as a primary standard.

Maintenance requirements dictated the devision of the standard into several specific units. Fig. 8 illustrates diagrammatically the several units. In Fig. 1, on the left-hand rack, there are two of the standards in operation at the New York central office. The outside bays house the standards, the fork units being at the bottom with their associated units mounted above.

The fork unit proper is a double temperature controlled cabinet. The thermostats are of the sealed contact mercury type. The inner and outer compartments are held to approximately 45 degrees centigrade. The fork is made from cold-rolled steel and cut to vibrate very close to 810 cycles per second. Temperature control and a variable resistance in the fork drive circuit are used for adjustment to the desired frequency.

The thermostat contacts in the fork unit control the excitation for a pair of vacuum tubes in the control unit, which, in turn, actuate relays, thereby causing voltage to be thrown on and off the fork box heater units. Control is held to 0.01 degree centigrade.

The drive amplifier is a straightforward transformer coupled arrangement. VT_1 and VT_3 are the drive and pickup tubes with VT_2 supplying the necessary power amplification for the output bus circuit. The filament, bias and anode supply for VT_1 and VT_3, and the anode supply for VT_2 are regulated in order to minimize the effect of supply voltage variations. R_3, in the fork drive circuit, provides the seven parts in 100,000 matching control required by international standards.

IV. Universal Machines

The photoradio machine is arranged to serve either as a transmitting scanner or a recorder, the unit mounted on the lead screw carriage determining the function to be performed.

The component parts are mounted on an aluminum base casting thirty-three inches long by nineteen inches wide by two inches high. On the base are mounted the motor alternator unit, subject drum gear box, line advance gear box, subject drum assembly, and the line advance lead screw with carriage track assembly. The complete assembly is illustrated by Fig. 9. The machine shown is arranged for visual recording.

The important operating features consist of a wide range of drum speeds and line advances, service flexibility, and subject-drum loading. They will be described briefly as follows:

1. Drum Speeds and Line Advance

The subject drum speed can be varied by means of a simple four-step gear shift from 20 to 60 revolutions per minute. By changing the worm and worm-wheel combination inside the gear box from a "single entry" to a "double entry," two more steps are provided, namely, 80 and 120 revolutions per minute. The line advance may be varied in twelve steps between 40 and 300 lines per inch by means of a simple gear shift.

2. Service Flexibility

As mentioned above, either a transmitting scanner or recording unit can be readily fitted to the line advance carriage. This is a necessary feature judged from an initial expense and operating standpoint.

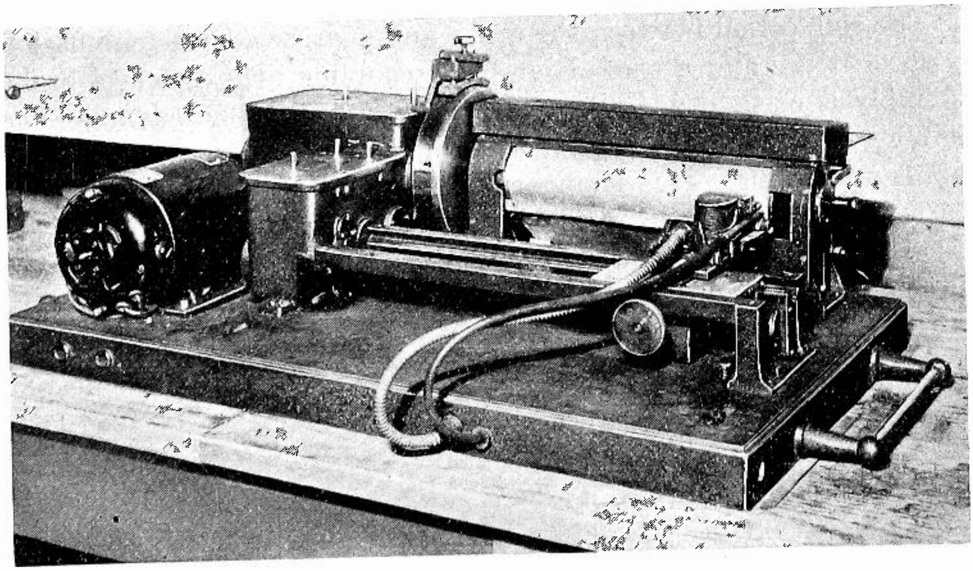

Fig. 9—Universal photoradio machine.

3. Subject-Drum, Loading, etc.

The subject drum assembly consists of the subject drum proper, loading arrangement, and gear box coupling.

The subject drum is made of aluminum alloy. It is carefully balanced and its end shafts connected to rigid supports through ball bearings. A portion of the drum periphery is cut away in order to provide space for the gripper fingers, which engage the leading and trailing edges of the subject or recording surface.

To load, the recording surface is placed on a platform positioned directly to the rear and tangent to the top of the drum. A set of feed rollers and stops are mounted directly over the drum, the point of fastening for this assembly being the drum supports. The stops at all times just clear the drum surface, whereas the feed rollers are let down on to the drum surface during the loading operation by means of a lever. Bringing the feed rollers into play, clears the subject or record surface from the stops, forcing it against the drum surface. When the cutaway portion of the drum periphery is below the rollers, the surface in question is forced downward, thereby causing the grippers to engage first its leading edge and then when the drum revolution has been completed, its trailing edge. To lift the feed rollers after the loading operation has been completed, the control lever is thrown back.

The cam motion required for the unloading operation is started by pressing a push button which is mounted on the right-hand drum shaft support. This operation clears the grippers, thereby allowing the subject or record surface to be removed by the motion of the drum.

The connection between the drum and gear box is a combination flywheel brake and a special jaw type coupling. The flywheel brake assembly is used to minimize gear ripple and drum unbalances. The coupling is of the general jaw type but is arranged to work against rubber and springs. This arrangement absorbs the loading shock, eliminating the sudden load shift on the drive source. Table II lists detailed information on the photoradio machine used by RCA Communications, Inc.

TABLE II

Diameter of cylinder	74.54 mm	2.93″
Circumference of cylinder	234.19 mm	9.22″
Gripping loss	19.05 mm	0.75″
Exposed circumference	215.15 mm	8.47″
Max. skew-index 1100*	3.36 mm	0.132″
Total gripper and skew loss	22.41 mm	0.86″
Useful subject width	211.6 mm	8.34″
Total length of cylinder	304.8 mm	12.″
Useful length of cylinder	304.8 mm	12.″
Scanning lines per mm or inch	mm	*Inch*
	1.58 4.72	40 120
	1.97 5.90	50 150
	2.36 6.30	60 160
	2.95 7.87	75 200
	3.15 9.45	80 240
	3.94 11.81	100 300
Scanning Strokes or Drum Speeds per min.	20, 30, 40, 60, 80, 120	
Index of Coöperation	117, 146, 176, 220, 234, 293, 352, 440, 469, 586, 703, 879	

* Max. skew shown $=1100\times\frac{1}{100{,}000}\times$ total length of cylinder

V. Transmitting Scanner

1. *Requirements*

The transmitting scanner assembly provides the means for converting variations in picture density to audio carrier modulation. If the carrier modulation is a linear function of the picture density variations, and various forms of distortion are held under control, an accurate conversion of picture density changes will result.

The points which must be given attention if an efficient transmitting scanner is desired are as follows:

(a) Scanning lamp
(b) Optical system
(c) Scanner output fidelity
(d) Black-white contrast
(e) Black output level
(f) Setting drift
(g) Maintenance

(a) The scanning lamp should be of the single unit type with a power requirement of not more than 75 watts.

(b) The optical system as a whole should be as simple as possible consistent with reasonable efficiency. The diaphragm should have height and width adjustments and should be calibrated. The design should be arranged for reflected light, using positive subject copy. Ordinary room lighting should offer little or no interference.

(c) The scanner output should be linear as a function of density. The wave form should be sinusoidal.

(d) The ratio between the black and white output levels should be at least three to one.

(e) Noise pickup dictates some output on black in place of the theoretical zero. A minimum of −25 decibels below zero level (1.92 volts across 600 ohms) allows for satisfactory operation.

(f) Contrast setting drift must be held to zero at least during the time required to transmit a picture.

(g) Although size suitable for mounting on the line advance carriage is essential, the layout of parts should be consistent with practical maintenance.

2. Scanner Optical System

The reflected light principle of scanning is used. That is, a light source is projected on to the positive subject copy and the result reflected into the phototube.

The optical system assembly is mounted on the subject drum side of the scanner chassis box which, in turn, is fastened to the line advance carriage. The scanner chassis box is positioned five inches from the subject drum and the side facing the drum is six and one-half by six inches. Therefore, the space available for the optical system assembly is eight and one-half by six by five inches. Of this space the assembly makes use of approximately five by four by five inches. Fig. 10 illustrates the physical arrangement.

A standard 75-watt automobile headlight lamp is used for the scanning lamp. It is operated at approximately three fourths of its rated value, thereby increasing the useful life. The housing provided completely encloses the lamp but louvres in the sides and bottom supply sufficient ventilation. Horizontal and vertical adjustment of the lamp base is furnished in order that the lamp may be properly centered.

The lens barrel between the lamp and the subject drum contains two lenses and the diaphragm assembly. The lens directly after the

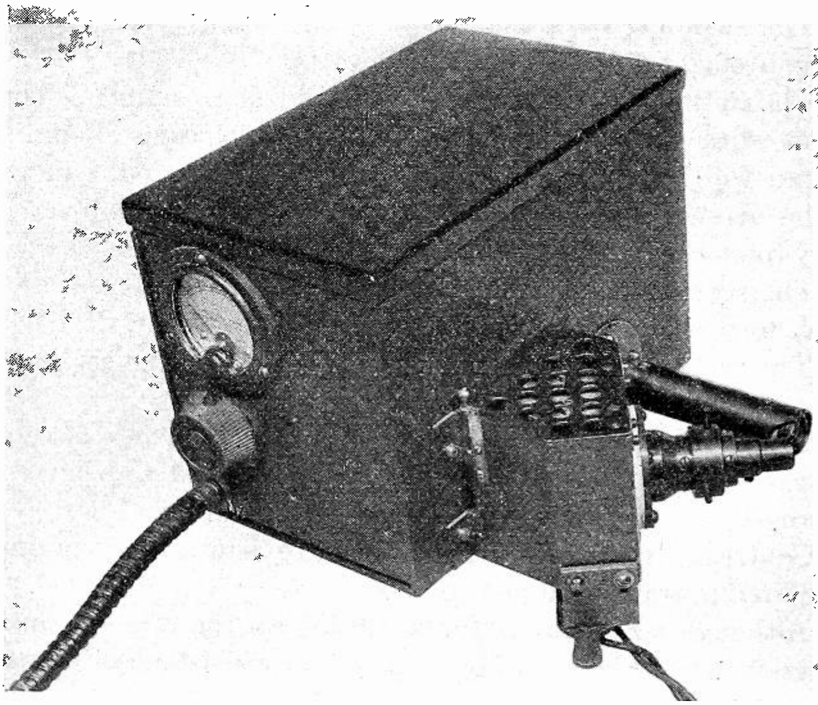

Fig. 10—Transmitting scanner.

lamp is supplied for condensing purposes and the lens facing the subject drum is supplied to focus the diaphragm image on the subject drum. The diaphragm adjustments are such that a rectangular image will result with the height and width under control. These adjustments are calibrated, the width in lines per inch corresponding to the line advance and the height in thousandths of an inch. The width adjustment is necessary to avoid scanning line overlapping. The height adjustment is necessary to minimize aperture distortion, which has been reported on in detail by Mertz and Gray of the Bell Telephone Laboratories.[8]

[8] Mertz and Gray "Theory of scanning," *Bell Sys. Tech. Jour.*, vol. 13, pp. 464–515; July, (1934).

The lens barrel between the subject drum and the phototube contains one lens, a pickup lens, and is mounted at the subject drum end of the barrel. This lens picks up the reflection of the diaphragm image and directs it, slightly enlarged, on to the phototube. The system is only concerned with the reflection of the diaphragm image and therefore ordinary room lighting does not seriously affect the phototube.

3. Scanner Circuit Arrangement

There are possibly several circuit arrangements available which would come close to or meet the scanner specifications listed in Section V, 1. A straight audio-frequency amplifier following the phototube with mechanical interruption of the light beam, is possibly the most favored arrangement. If the modulated carrier output of the scanner could be used for direct control of a wire system or a radio transmitter, then the mechanically produced carrier would be entirely satisfactory. Where it is necessary to convert the modulated carrier output of the scanner into varying length dots, it is important to remove all chance of spurious modulation. The spurious modulation, if present, will cause serious beat interference in the recorded dot pattern. A slight misalignment of bearings, dirty mirror faces, etc., are typical causes of spurious modulation.

In order to avoid the troubles just mentioned, a circuit arrangement has been adopted which is entirely electrical in operation, employing vacuum tube technique throughout. The arrangement is in the form of a balanced grid modulator, the carrier being obtained by frequency multiplication of the 810 cycles supplied by the frequency standard, and the unbalancing effect by the phototube current changes. The output of the modulator is then amplified by a single push-pull stage in order to avoid the possibility of distorting the desired sinusoidal wave form.

Fig. 11 illustrates diagramatically the complete transmitting scanner. *LP* is the scanning lamp, L_1 the diaphragm condensing lens, *DF* the double diaphragm assembly, and L_2 the lens which performs the function of focusing the diaphragm image on the subject drum. The line *S*-*S* is the subject drum surface. The reflected diaphragm image is picked up by lens L_3, a slight enlargement being impinged on the phototube *PC*. The phototube employed consists of two electrodes contained in a small gas-filled glass bulb. One electrode, the cathode, is a semicylindrical sheet of metal which has a surface sensitized with a film of caesium. The other electrode, the anode, consists of a small wire placed in the axis of the cathode surface. VT_1 and VT_2 are the modulator tubes. The carrier tone is fed to VT_1 and VT_2 through transformer T_1, the connection being push-push. R_1 and R_3 are used to set the initial balance and R_3 alone for contrast change. The phototube current changes through R_{11} and R_{12}, and the voltage divider resistance assembly, furnish the modulating potential. The anode circuit unbalance between VT_1 and VT_2 caused by the modulating potential, allows VT_3 to be excited by a modulated carrier, the modulation being a linear function of the voltage appearing in the input cir-

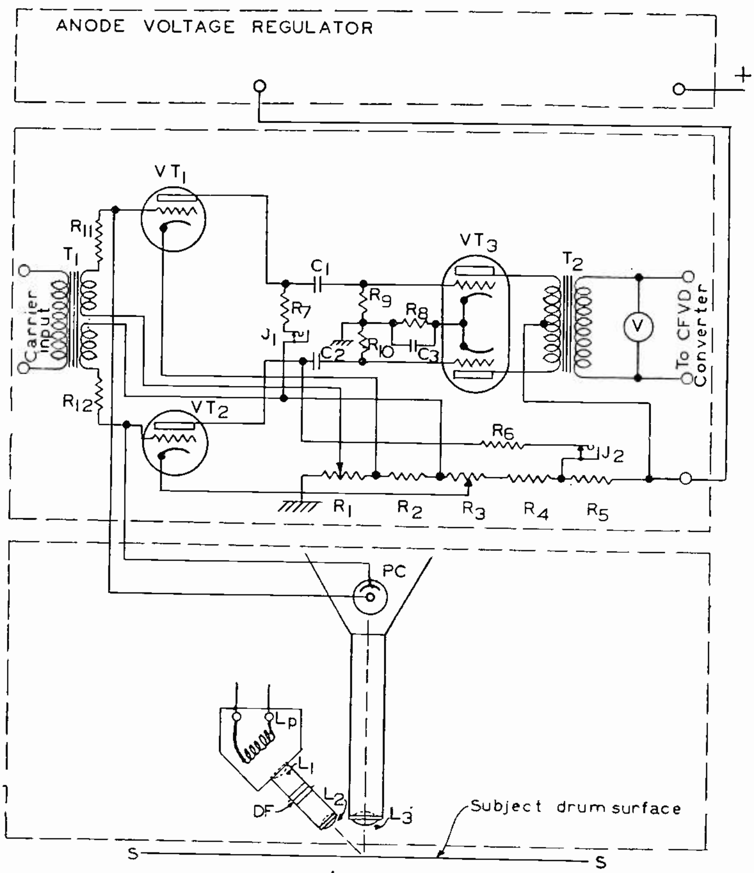

Fig. 11—Schematic layout of transmitting scanner.

cuit. VT_3 is a double triode. The meter V connected across the output terminals is a rectifier type voltmeter and is useful for checking contrast settings. Low voltage direct current is supplied to the heaters of the vacuum tubes from a common battery supply. The anode requirements are furnished from the common battery supply through a glow tube regulator in order to minimize supply voltage variations.

Fig. 12 shows graphically the type of response obtained from the balanced grid modulator type of transmitting scanner. The response is essentially a linear function from 0.4 to 1.4 volts (600-ohm load).

This provides a black-white ratio of at least three to one and a black level high enough to eliminate noise pickup troubles.

The arrangement illustrated by Figs. 10 and 11 is mounted in compact chassis form, the chassis being mounted inside the scanner box. As has been mentioned previously, the box is fastened to the line advance carriage. The right-hand end of the scanner box is a portion of the chassis and forms the mounting panel for the contrast control, output meter, balance checking jacks, and the supply and control cable terminations. The outside dimensions of the transmitting scanner box are eight and one-half inches long, six inches high, and five inches wide.

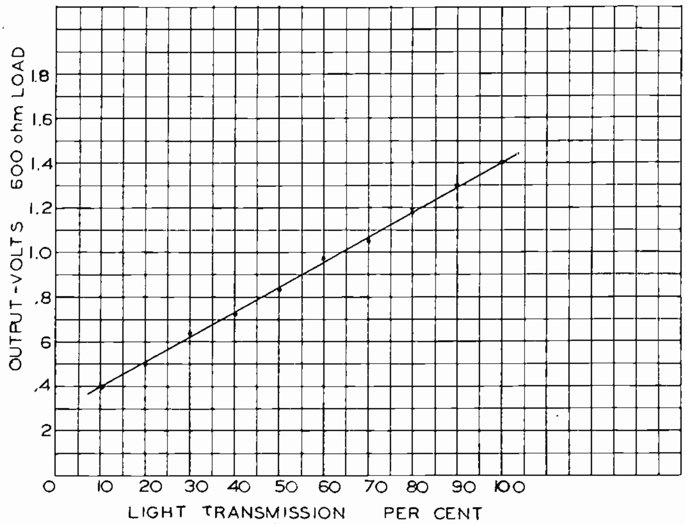

Fig. 12—Light response characteristic of RCAC BGM scanner.

4. Operating Features

The optical arrangement described in Section V, 2, has made possible an improvement in efficiency of three to one over the arrangement replaced, in each case starting with approximately the same amount of illumination. A single standard low cost lamp is now used in place of two special prefocused lamps. The use of only a small area of reflected light has reduced the effect of room lighting. Definite calibrated adjustments make possible accurate duplication when required.

The balanced grid modulator type scanner circuit eliminates the lack of fidelity experienced heretofore on the black end of the density scale. It provides an output free of spurious modulation which is essential for dot conversion. The picture contrast adjustments are simple, and stable operation allows accurate holding of a given contrast setting and permits a return to a given setting if desired.

VI. Transmitting Converter

1. *Radio Circuit Transmission Characteristics*

The use of the band from 3000 to 30,000 kilocycles, as at present, for long-distance communication, has brought about a marked improvement in over-all efficiency, but at the same time has introduced circuit distortion which is a serious handicap to picture handling if normal technique is employed. The circuit distortion experienced takes the form of round-the-world echoes, multipath echoes, and complete or selective fading. The effect of these various forms of circuit distortion have been studied carefully by our receiving engineering group and also by other engineers, notably T. L. Eckersley.[4] The circuit distortion phenomena are subject to wide variations from minimum to maximum, both slow and rapid. This being the case, direct control of the radio transmitter by the scanner modulation will produce a picture modulated radio-frequency carrier, which, when intercepted at the radio receiving station, is subject to distortion due to selective fading, resulting in a "streaky" picture under adverse fading conditions.

Radiotelegraph research teaches that if the radio-frequency carrier can be keyed definitely on and off, using a time basis for signaling, the troublesome circuit distortion can be compensated for, to a commercial degree, at the receiving station. This means of minimizing the circuit distortion at the radio receiving station when telegraph time basis keying is used to control the radio-frequency carrier is called "Diversity Receiving Technique." The diversity receiving system employed by RCA Communication, Inc., has been reported on by H. H. Beverage and H. O. Peterson.[9] Either this system or methods which will produce similar results are used on all successfully operated long-distance short-wave circuits.

2. *Photoradio Application of Keyed Radio-Frequency Carrier*

With stable high speed radiotelegraph circuits available, means were found to convert the scanner modulation into dot keying. With the method used prior to 1930, both the dot keying rate and weight were varied in something that closely approached a linear function of the half-tone picture controlled scanner modulation. A double modulated multivibrator circuit was employed to accomplish this conversion. It has been described in papers given before the Institute by R. H. Ranger.[1,2,3] This arrangement proved to be a marked improvement over direct modulation control of the radio transmitter but still left much to be desired, both from a circuit handling and half-tone reproduction standpoint.

[9] Beverage and Peterson "Diversity receiving system of R.C.A. Communications, Inc.," Proc. I.R.E., vol. 19, pp. 531–584; April, (1931).

Since 1930, improvements in the photoradio dot keying method have been made which make possible the linear response and half-tone detail desired, as well as a keying arrangement which allows for maximum radio circuit efficiency. The new method holds the dot keying rate and amplitude constant, varying only the weight as a linear function of the picture half-tone controlled scanner modulation. As mentioned in Section I, the new keying method is called "CFVD" or "Constant-Frequency Variable-Dot Keying." A comparison of a recording made by the two dot keying methods is shown in Fig. 13.

Fig. 13—Multivibrator—CFVD recording comparison.

The right-hand portion was recorded using the multivibrator dot converter and the left-hand portion by using the CFVD dot converter. An improvement in the amount and smoothness of half-tone detail can be noted. The multivibrator dot keying rate used was of the order of 300 cycles per second and the CFVD dot keying of the order of 175 cycles per second, the same drum and line advance being used in each case.

A description of the CFVD system follows.

3. CFVD Converter

(a) General Theory

A broad picture of the CFVD arrangement is essential before the associated parts are examined in detail. Reference to Fig. 14, which is in block diagram form, will assist in viewing this broad picture.

The component parts directly associated within the CFVD converter, are shown within the dotted line. They are a single stage audio amplifier, a full wave rectifier, a mixing tube, a square wave amplifier, and a line carrier frequency keyer. The synchronized screen frequency and line carrier frequency units although mounted outside, are essential parts of the converter. The operating sequence and the functions performed by the various parts of the arrangement are as follows:

The transmitting scanner, which has been described under Section V, supplies the modulation which results from scanning the subject

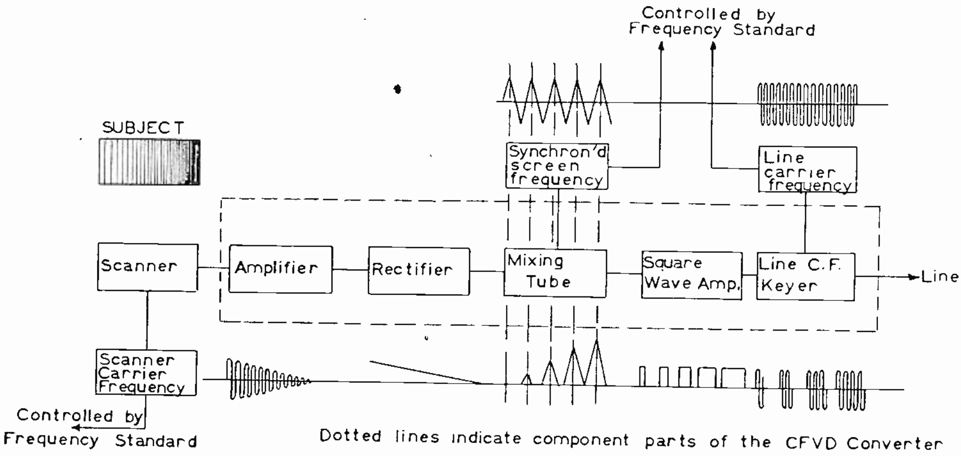

Fig. 14—Block diagram of constant-frequency variable-dot photoradio system.

illustrated. The scanner output is then amplified and rectified. The resulting wave form is indicated below each unit. The mixing tube, in the form of a triode, has a portion of its grid bias under the control of the rectifier output. The fixed portion of the mixing tube grid bias is adjusted so the tube is at cut-off when the rectifier output is minimum. With this arrangement, saw-tooth wave form voltage from the synchronized screen frequency unit is supplied in series with the mixing tube grid. The saw-tooth wave form voltage and the rectifier output voltage combine to produce the varying anode current pulses illustrated below the rectangle marked "Mixing Tube." The next step is to provide a circuit arrangement which will deliver constant amplitude varying weight dots, the weight and recurring rate governed by the mixing tube anode circuit pulses. This function is performed by the square wave amplifier. With the CFVD dots available in direct-current form, the next step is to use this direct-current form as keying bias on a tone keying stage, the output of which supplies keyed carrier to the radio transmitter control line. These steps will be described in more detail later. It is important to note that all audio-frequency requirements are controlled by the same source, namely, the 810-cycle frequency standard. This is necessary in order to eliminate recorded dot pattern beat interference.

(*b*) Circuit Arrangements

Low voltage direct current is supplied from a common bus to the heaters of all vacuum tubes. Individual unit glow tube regulation of the common source of high voltage supply is provided.

(*1*). *Converter*

Fig. 15 illustrates schematically the circuit arrangement of the component parts directly associated inside the converter unit. The

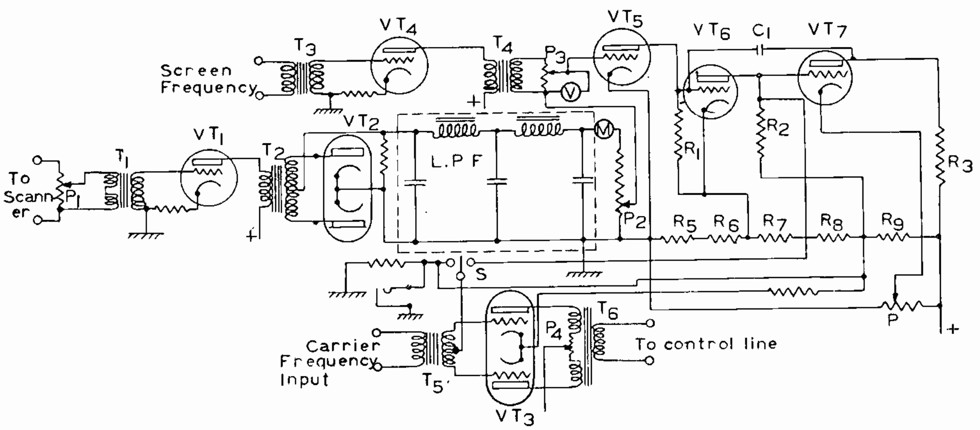

Fig. 15—Schematic layout of constant-frequency variable-dot converter.

scanner modulated carrier is connected to the converter through potentiometer P_1. Its route through to the mixing tube can be traced via transformer, T_1, vacuum tube VT_1, transformer T_2, full wave rectifier VT_2, and the low-pass keying filter *L.P.F.*, with the rectified modulation voltage appearing across potentiometer P_2.

The screen frequency is supplied to the converter through T_3. It reaches the mixing tube through vacuum tube VT_4, transformer T_4, and potentiometer P_3.

The mixing tube VT_5, has potentiometer P_2 for adjustment of rectifier output voltage and potentiometer P_3 for screen frequency level adjustment.

The square wave amplifier is composed of vacuum tubes VT_6 and VT_7, resistances R_2 and R_3, potentiometer P and condenser C_1. The voltage divider R_5 to R_9 inclusive supplies the necessary anode and bias voltages. The square wave amplifier is essentially a resistance coupled amplifier trigger circuit, the trigger regeneration being furnished by C_1, and the duration of maximum amplitude being determined by the amplitude of the pulses in the mixing tube anode circuit.

The push-pull tone keying stage consists of the double triode vacuum tube VT_3, and transformers T_5 and T_6. The keying bias is obtained from the anode circuit of either VT_6 or VT_7, depending on the bias polarity desired. Switch S furnishes means for hand keying if service requirements so dictate.

(2). Synchronizing Screen Frequency Unit

The screen frequency unit is shown diagrammatically in Fig. 16. The arrangement is essentially a controlled multivibrator, the wave form of the frequency generated being converted to a saw-tooth envelope. The saw-tooth wave produced is then amplified by a high quality stage, the result then being suitable for the CFVD converter screen frequency requirements.

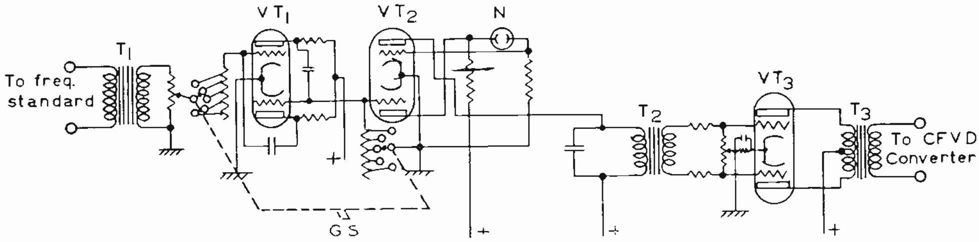

Fig. 16—Schematic diagram of synchronized screen-frequency unit.

The 810-cycle standard frequency is used for control and is connected to the multivibrator vacuum tube VT_1, which is a double triode, through transformer T_1 and its associated potentiometer. A gang switch *GS* varies tapped resistors connected to the input and output of the multivibrator divider, thereby producing the five desired frequencies. The double triode VT_2 serves the dual purpose of rendering symmetrical the distorted output of the multivibrator through the agency of the neon tube, and coupling it to the saw-tooth shaping circuit of the primary of T_2 and its associated condenser. The double triode VT_3 and its associated transformers T_2 and T_3, must be capable of passing a wide range of frequencies without discrimination in order that a symmetrical saw-tooth wave form may be passed on to the converter. Slight departures from symmetry will cause misalignment of the recorded dot pattern which will destroy half-tone detail.

The five screen frequencies provided are 90, 135, 162, 202.5 and 270 cycles. These frequencies were chosen as the result of radio circuit study extending over a period of several years and have been found best suited for photoradio operation on the international circuits.

(3). Line and Scanner Carrier Frequency Unit

Fig. 17 illustrates schematically the circuit arrangement used to produce the required synchronized carriers for the CFVD converter tone line keying stage, and for the transmitting scanner. A controlled dynatron stage with its output smoothed and amplified is provided for each service. The circuit arrangements are identical, three frequencies being produced at will in each case, namely 1620, 2430, and 3240 cycles.

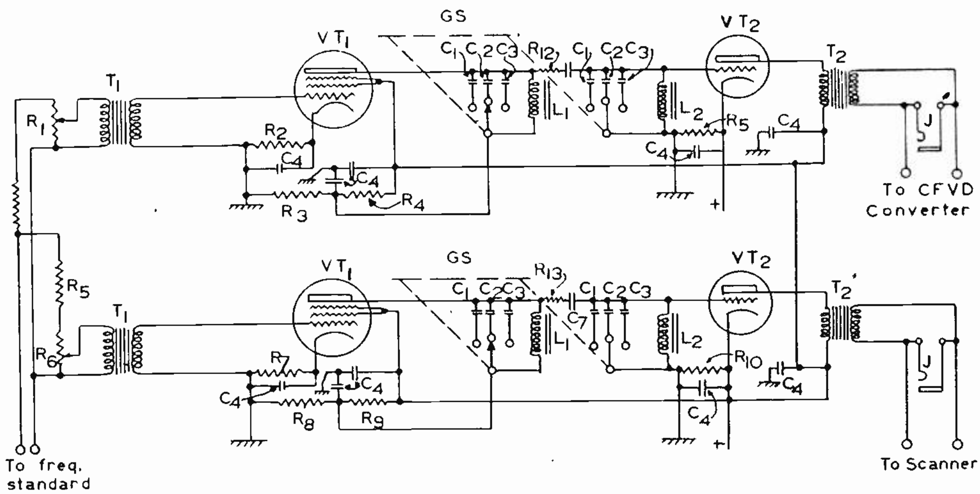

Fig. 17—Schematic layout of line and scanner carrier-frequency unit.

The 810-cycle control is connected to the individual dynatron vacuum tubes VT_1 through their respective input potentiometers and transformers. Dynatron tank circuit tuning is provided in each case. The coupling arrangement to the respective amplifying vacuum tube stages VT_2, supply the necessary smoothing. The outputs are connected respectively to the CFVD converter and the transmitting scanner.

(4). Phase Shift

The printing art has demonstrated the value of "screening" a subject in order to lay down half-tone values. The CFVD photoradio system provides an electrical screening which may be varied as the subject drum speed and screen frequency dictate. Radio circuit conditions dictate the drum speeds and screen frequencies which may be used at any given time. In order to provide the proper screen mesh when screen frequencies and drum are changed, a variation in the mesh control is required.

The electrical screening is provided by reversing the polarity of the screen frequency voltage fed to the CFVD converter at the desired rate. This is accomplished by means of a cam and contact assembly mounted on the universal photoradio machine drum shaft and a relay mounted adjacent to the CFVD converter.

Fig. 18 illustrates in schematic form the reversing arrangement and the resulting recorded dot patterns. Experience indicates that a so-called double and single phase shift will satisfy the majority of cases. If a wider range of drum and screen frequencies were normally used, a somewhat different phase shift rate possibly would be required.

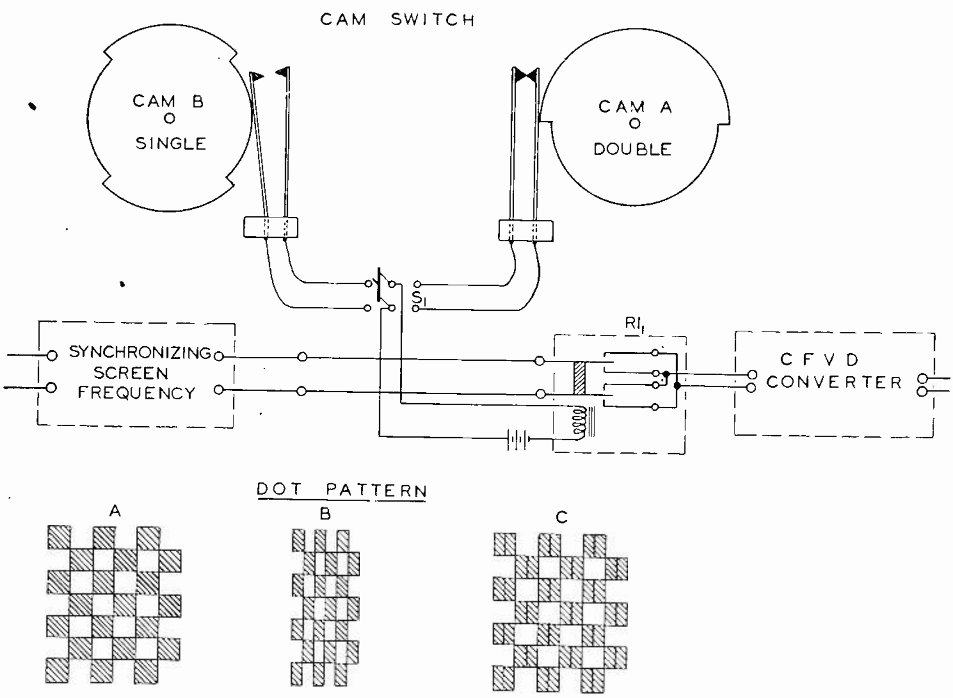

Fig. 18—Schematic layout of screen-frequency phase shift.

The dot pattern plans illustrated indicate the result of phase shift. *A* illustrates a recording at high drum speed and screen frequency using a single phase shift; *B*, a single phase shift but at lower drum speed and screen frequency; *C*, a double phase shift with the drum speed and screen frequency the same as *B*.

4. Half-Tone vs. Black-and-White Transmission

Half-tone transmission using the screen frequency converter has been described under Section VI. 3. Naturally, half-tone material represents only a part of the photoradio business handled. Therefore, means must be made available for sending black-and-white material as well as half-tone. Only a simple change is required to make possible the transmission of black-and-white material. This change is to open the synchronized screen frequency input to the CFVD converter. The rate and weight of the pulses passed on to the square wave amplifier portion of the converter will then be governed by the type of black-and-white material being scanned.

5. Keying Speed Analysis

An analysis of the CFVD keying is supplied as Appendix A. This analysis illustrates in mathematical and diagrammatic form the effect of circuit distortion when using CFVD keying and, also, suggests radio transmitting and receiving station equipment requirements if minimum distortion is desired.

VII. Recording Arrangements

The ideal recording method is, of course, visual, with the ability to view the picture during the process of recording. It is also important that a visual system be linear, laying down the proper density proportion. The ability to produce multiple copies quickly is essential. If it is not possible to supply a visual system which will furnish the type of service required, the alternative is some form of photographic recording. At the present time photographic recording either on film or bromide paper is the accepted form of commercial photoradio recording.

Although accepting photographic recording for commercial work at the present time, the idea of visual recording has not been given up. Two methods of visual recording which meet at least a portion of the requirements are being used quite successfully by RCA Communications, Inc., and its associated company, RCA-Victor. RCA-Victor has carried on the development of the so-called "carbon recorder" whereas RCA Communications, Inc., have been active with an arrangement called "ink vapor recording." At the present time, RCA Communications, Inc., makes use of the "ink vapor recording" to monitor all incoming and outgoing photoradio business. This provides an instantaneous check on the circuit operation.

The two recording methods used by RCA Communications, Inc., will be described in the following paragraphs.

1. Visual

The arrangement used for visual recording is illustrated in schematic form by Fig. 20.

The vacuum tube assembly which delivers the signal energy to the vapor control is shown at the top of the figure. The signal furnished by the radio receiving station is amplified by the vacuum tube VT_1 and its associated input and output transformers T_1 and T_2. The diode-triode vacuum tube VT_2 rectifies the signal and, also, combined with vacuum tube VT_3 provides a degree of limiting and stage reversing required to excite properly the power tubes VT_4 and VT_5. The output of the power tubes is connected to the signal winding of an electrodynamic unit, shown mounted on the lead screw carriage of a photoradio machine. The field *FS* of the gun unit can be either of the fixed magnet type or energized from available bus voltage. The armature *ST* of this unit is arranged to form a moving shutter or deflection plate for the ink vapor directed by pressure through nozzle *NZ*. Valve *V* and its associated waste bottle, is provided in order to furnish con-

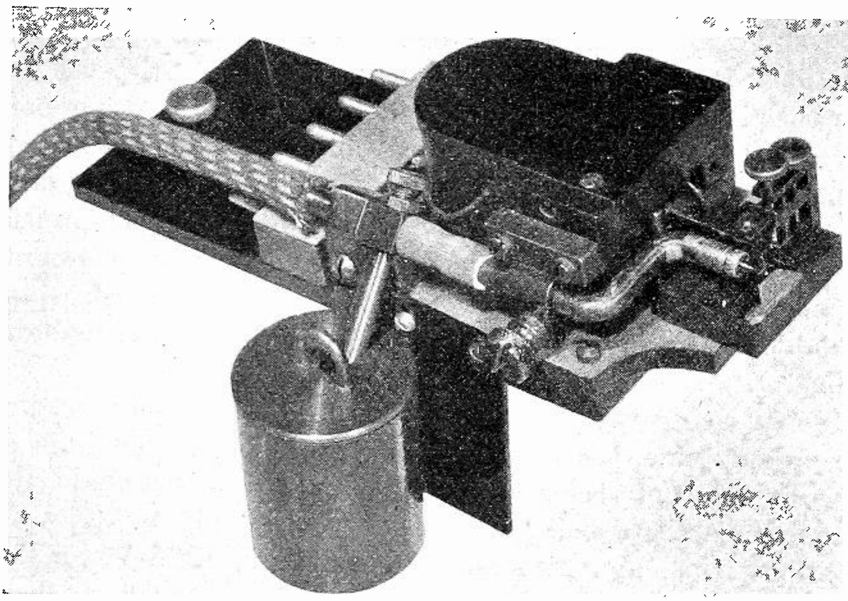

Fig. 19—Visual recording schematic layout.

stant pressure without ink splatter throughout a given drum revolution. Approximately eight to ten pounds of pressure is maintained into the atomizer container.

With the system in operation, a series of black marks will be recorded with every turn of the drum. The width of the mark, fixed by the nozzle size, is made to correspond to the line advance, and the number and length per drum revolution are dependent upon the CFVD keying being transmitted. Tests have indicated that the gun movement will follow screen frequencies as high as 1000 cycles with sufficient amplitude.

Fig. 19 is a photograph of a typical gun unit. Close examination will reveal the parts mentioned in the description of the unit.

The ink used is a black dye combination soluble in alcohol. The ordinary letterhead weight and quality paper is satisfactory for a recording surface, but a glossy surface enhances the appearance. With careful handling it is difficult at times to distinguish between visual and photographic recording unless a close examination is made. With ordinary handling the results obtained provide an excellent check on circuit operation.

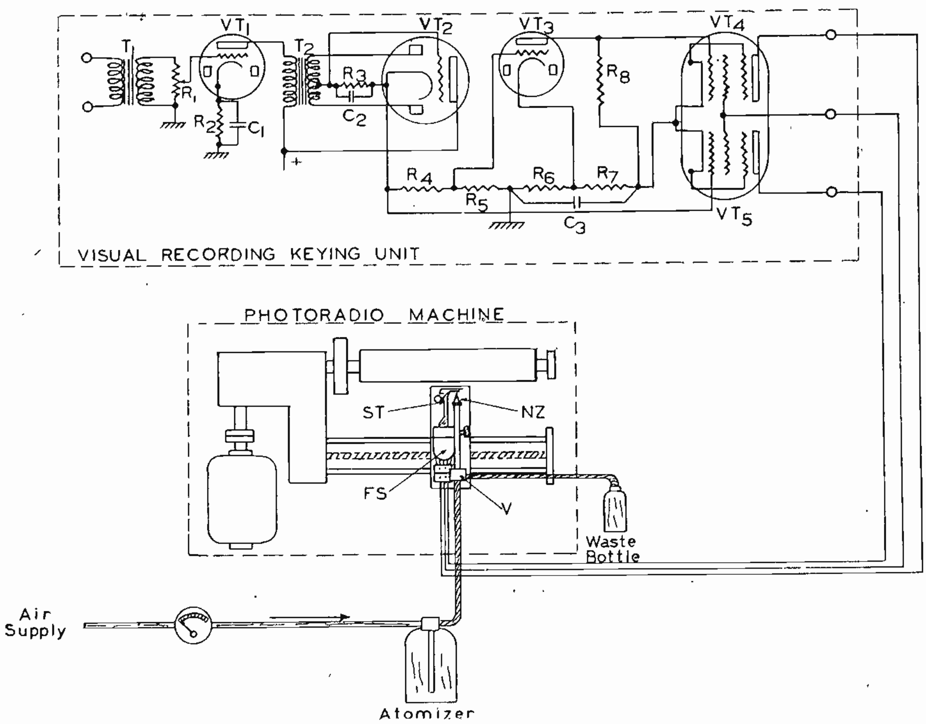

Fig. 20—Visual recording gun unit.

2. Photographic

The arrangement used for photographic recording is illustrated in schematic form by Fig. 21.

The vacuum tube stage which makes possible either positive or negative recording, is shown at the top of the figure. The tone signal received from the receiving station is rectified before connection to this unit. Switch S_1 allows the input polarity to be changed so that the vacuum tube VT_1 is either working down to or up from cutoff. The glow tubes N provide a steady anode supply for VT_1. The output of VT_1 is used to flash the recording glow tube NB on and off in response to the CFVD keying.

The recording glow tube has an argon-helium gas content. The light emitted is blue and contains a considerable amount of ultraviolet. The breakdown voltage is approximately 250 volts and the operating current ten milliamperes.

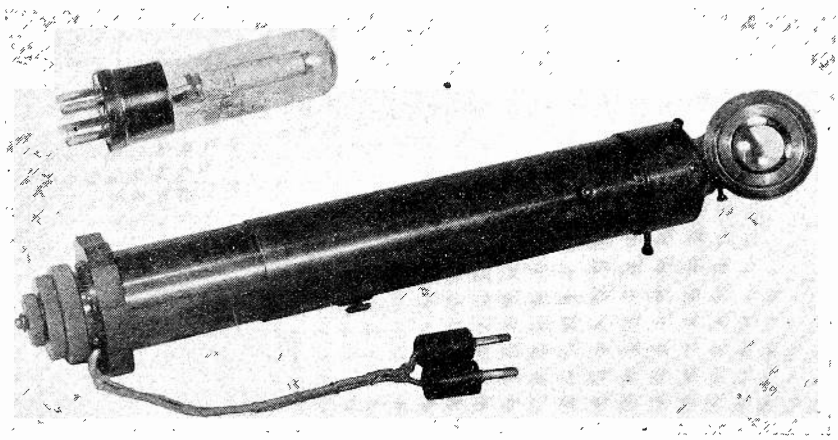

Fig. 21—Photographic recording neon and mounting barrel.

The optical system is similar to that used between the scanning lamp and the subject drum. A condensing lens is used between the

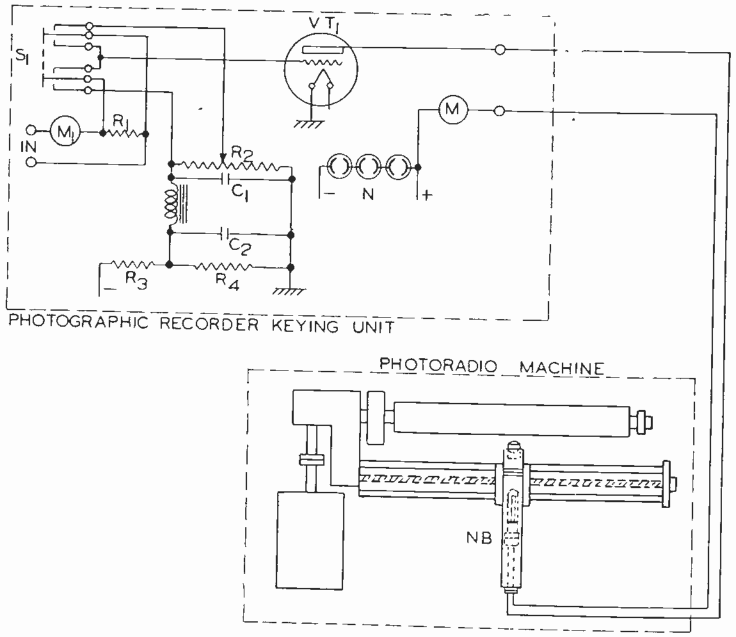

Fig. 22—Photographic recording schematic layout.

recording glow tube and the diaphragm which is a duplicate of that used on the transmitting scanner. The lens at the drum end of the barrel is arranged to focus the diaphragm image on the recording surface. A viewing lens is provided on the drum end of the barrel to facilitate adjustment of the optical system. Fig. 22 is a photograph of the lens barrel assembly and the recording glow tube.

The recording surface used is either news bromide paper or printon film. Both can be used under red or yellow light without fogging difficulties.

Fig. 23—Riverhead photoradio monitor equipment.

VIII. Monitor

Any type of communication service where twenty-four-hour operation is required must be provided with an efficient trouble checking and monitoring arrangement. If checking and monitoring methods are important for normal radiotelegraph and telephone service, they naturally assume a still greater importance when the signaling speed is several times faster, as is the case for photoradio operation. Rapidly changing radio circuit conditions require close monitoring of all outgoing and incoming pictures in order to avoid delay.

RCA Communications, Inc., make use of the visual ink vapor recorder at the radio transmitting and receiving stations as well as the central office when photoradio operation is in progress. This means of monitoring is used in addition to normal meter and oscilloscope technique. Monitor arrangements of the type described assure maximum equipment efficiency, with the radio circuit the remaining variable.

If reference is made to Fig. 1 and Fig. 23, typical monitoring arrangements can be observed. In Fig. 1, which is a view of the New York office, the visual recorder is on the right end of the machine table. The apparatus rack at the right has available a volume indicator and other meter checking combinations normally used. Fig. 23 is a view of some of the monitoring arrangements in use at the Riverhead, L. I., radio receiving station. The visual ink vapor recorder and the cathode-ray oscilloscope can be observed. Similar arrangements are used at our radio transmitting stations.

IX. Operating Technique

1. General

With improvements in mechanical and electrical features, it has become possible to develop an operating technique which minimizes the personal factor in setting up the CFVD converter for transmission. The method of making converter adjustments as the result of meter readings and calibration charts, enables the operator to transmit pictures with great fidelity and ease. At the same time, it becomes possible to adjust the converter accurately to take into account circuit conditions.

It is well to review the general influence of the radio circuit distortion on CFVD keying before discussing further the operating technique.

In general, assuming a perfect emitted signal from the radio transmitter, the radio circuit tends to decrease the contrast range of a picture. This is brought about by multipath transmission which adds the equivalent of a tail to the signal. This tail, then, increases the apparent weight of marking so that a transmitted twenty per cent dot, for example, would be received as a thirty per cent dot. For the same conditions, an eighty per cent dot would be received as ninety per cent. Since contrast is defined as the ratio of the heaviest dot to the lightest, the transmitted contrast would be four while the received contrast would be three. It is apparent that the elongation is particularly disastrous when the marking is on the order of ninety per cent, because at this point filling may increase the mark time to 100 per cent at the receiving station.

In order to overcome the radio circuit distortion just described, the actual marking range at the transmitter must not exceed a value such that at the receiving station the longest dot transmitted will just mark 100 per cent. Distortion control may also be increased by the proper selection of screen or keying frequency. If the actual duration in time of marking is made large, the effect of multipath tailing is not important.

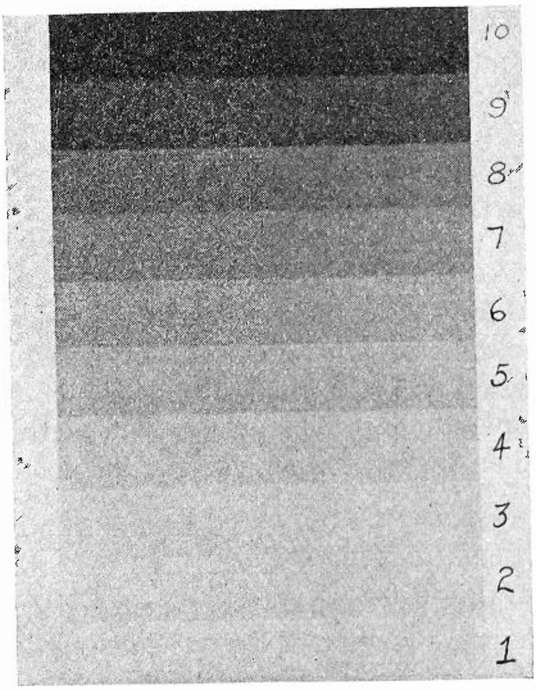

Fig. 24—Density wedge.

With the effect of the radio circuit distortion appreciated, it can readily be understood that the operator must have means under control for expanding and contracting the contrast range transmitted.

2. Pretransmission Requirements

A standard test subject, calibration of essential controls, and assurance of over-all linear response is required before correction for radio circuit distortion can be applied.

The test standard requirement is met by providing a ten-step density wedge arranged as a photographic positive print with the steps ranging from white to solid black, the step spacing arranged approximately ten per cent apart. Fig. 24 illustrates the arrangement described. Care in the construction of the wedge was taken to insure a linear reflected light response when progressing through the density range. Care was also taken to provide uniform density for each step. The density wedge then provides a definite standard for calibration and test transmission.

The calibration of the photoradio transmitting equipment is dependent upon three variables; the scanner output as indicated by its output voltmeter, screen tone input to the converter mixing tube as indicated by V, Fig. 15, and the converter low-pass keying filter output as indicated by M, Fig. 15. The output value of interest for the scanner is that noted when white or wedge step 1 is being scanned. With the proper white output from the scanner, the converter standardization is accomplished by the manipulation of the screen voltage and low-pass filter output to specified standard values as indicated by meters V and M respectively and then adjusting P_2, Fig. 15, until the converter output just "breaks white."

3. Contrast Control

With either a picture or the test wedge on the subject drum, the desired converter contrast control can be obtained by referring the M meter readings noted on black and white or the 10-1 test wedge steps, to a series of prepared charts. The charts will dictate readings which must be observed on V and M when varying the potentiometers P_3 and P_1 respectively.

4. Radio Circuit Condition Information

With the converter set for 0–100 per cent mark, the test wedge is transmitted, assigning a portion of the transmitting time to each of the several screen frequencies normally used. When the transmission has been completed, the receiving station reports on wedge step failures noted, such as 90 cycles failing steps 2 and 8, 135 cycles failing steps 2 and 8, 200 cycles failing steps 3 and 7. With this information available at the transmitting office, reference to the proper chart will supply the required readings for V and M in order to transmit 0 to 100 per cent mark. Any report indicating failures on steps 3 and 7 show lack of proper circuit contrast and therefore transmission should not be attempted.

5. Picture Acceptance Data

In order to avoid cancellation of pictures and loss of time due to poor quality of the received transmissions, it is exceedingly important that no picture be accepted in which the detail required exceeds that which it is possible to receive under average circuit conditions. The limitation of acceptance can be stated as follows.

"No picture, in which the point of interest is less than one-half by one-half inch in size should be guaranteed delivery with suitable detail."

The following Table III, of drum speeds and screen tone frequencies, gives the conditions which will provide suitable detail for the minimum area of interest of one-half by one-half inch.

TABLE III

Drum Speed	Screen Tone Frequency
20 r.p.m.	90 cycles
30 "	162 "
40 "	202.5 "
60 "	270 "

Under certain conditions, if a higher frequency than that indicated in the table can be used for a given drum speed, greater detail will result, but in order to protect against excessive transmission time and subsequent cancellations, the minimum area is generally adhered to.

Black-and-white acceptance is governed by the fact that type and line combinations smaller than twelve-point type formation cannot be guaranteed delivery.

Conclusion

The various improvements which have been reported in this paper are endeavors toward the desired goal of dependability, stability, fidelity, and ease of operation. It is felt that the CFVD principle of photoradio transmission is an important advance in the art of facsimile transmission over long-distance radio circuits. Close attention to detail is the major reason for the successful development and operation of the present-day photoradio service.

Acknowledgment

The authors wish to express their appreciation for the coöperation extended by their photoradio associates in Cables and Wireless, Ltd., London; Transradio Internacional Compania Radiotelegrafica Argentine, Buenos Aires; and Deutsches Haupttelegraphenamt, Berlin.

The associated subsidiary companies in the RCA organization, especially the group under the direction of Mr. C. J. Young in the RCA Victor Company, have contributed to the photoradio improvements.

Within RCA Communications, Inc., Mr. W. A. Winterbottom, Vice-President and General Manager, the Engineering and Traffic Operating Groups under the direction of Messrs. C. H. Taylor and J. B. Rostron, respectively, have all furnished wholehearted support and assistance.

APPENDIX A

Analysis of Facsimile Keying

In considering the CFVD facsimile system, it is of importance to know the effect of finite band width of the receiver on the quality of the reproduced signal. It is necessary also to know the effect of the ether path; i.e., fading and multipath transmission, on the signal reproduction at the receiving end.

All these factors impose distinct limitations on the maximum speed of transmission, but in spite of their combined effect commercial facsimile is entirely feasible. The individual limitations are discussed in detail below.

First, we shall consider the case of the dot envelope used to key the radio transmitter. Since the dots follow each other at a constant rate, assuming constant weight of marking, the square wave can be resolved into a Fourier series, as is well known, thus,

$$f(p) = \frac{2}{\pi}\left[\frac{\alpha}{2} + \sin\alpha\cos pt + \frac{\sin 2\alpha\cos 2pt}{2} + \frac{\sin 3\alpha\cos 3pt}{3} + \cdots\right] \tag{1}$$

where α equals one half of the marking period in angular measure and p equals the angular velocity of the screen tone.

If a carrier of angular velocity, ω, is then modulated by such a wave as given by (1), we obtain,

$$f(\omega) = \frac{2}{\pi}\cos\omega t\left[\frac{\alpha}{2} + \sin\alpha\cos pt + \frac{\sin 2\alpha\cos 2pt}{2} + \frac{\sin 3\alpha\cos 3pt}{3} + \cdots\right]. \tag{2}$$

Expanding (2) we get the well-known result,

$$f(\omega) = \frac{2}{\pi}\left[\frac{\alpha}{2}\cos\omega t + \frac{\sin\alpha}{2}\{\cos(\omega - p)t + \cos(\omega + p)t\} + \frac{\sin 2\alpha}{4}\{\cos(\omega - 2p)t + \cos(\omega + 2p)t\} + \frac{\sin 3\alpha}{6}\{\cos(\omega - 3p)t + \cos(\omega + 3p)t\} + \cdots\right]. \tag{3}$$

Equation (3) simply tells us that the transmitter emits a plurality of frequencies comprised of the carrier frequency and a number of side bands made up of the sums and differences of the carrier and harmonics of the screen frequency. It further shows that the amplitudes of the component frequencies are a function of the marking period, α.

This well-known result, may be expanded by considering the effect of the variable marking period produced in the CVFD system of transmission. Suppose we place on the record drum of the scanner a subject which varies from white to black to white sinusoidally. Then, as the drum rotates the marking period will vary from zero to 100 per cent; i.e., α will vary from 0 to $\pi/2$ and its value will be given by

$$\alpha = \frac{\pi}{2}(1 - \cos st) \tag{4}$$

where s is the angular velocity of the drum and consequently small compared to ρ and ω. Replacing in (3) by its value given in (4) and noting that

$$\left.\begin{aligned} \sin\frac{\pi}{2}(1-\cos st) &= J_0\left(\frac{\pi}{2}\right) - 2J_2\left(\frac{\pi}{2}\right)\cos 2st \\ &\quad + 2J_4\left(\frac{\pi}{2}\right)\cos 4st - \cdots \\ \sin\frac{2\pi}{2}(1-\cos st) &= 2J_1(\pi)\cos st - 2J_3(\pi)\cos 3st \\ &\quad + 2J_5(\pi)\cos 5st - \cdots \\ \sin\frac{3\pi}{2}(1-\cos st) &= -J_0\left(\frac{3\pi}{2}\right) + 2J_2\left(\frac{3\pi}{2}\right)\cos 2st \\ &\quad - 2J_4\left(\frac{3\pi}{2}\right)\cos 4st + \cdots \\ &\text{etc.} \end{aligned}\right\} \tag{5}$$

where J_n equals the Bessel function of the order n we obtain

$$f(\omega) = \frac{2}{\pi}\left[\frac{\pi}{2}(1 - \cos st)\cos\omega t\right.$$

$$+ \frac{1}{2}\left\{J_0\left(\frac{\pi}{2}\right) - 2J_2\left(\frac{\pi}{2}\right)\cos 2st + \cdots\right\}\left\{\cos(\omega - p)t\right.$$

$$+ \cos(\omega + p)t\} + \frac{1}{2}\{J_1(\pi)\cos st$$

$$- J_3(\pi)\cos 3st + \cdots\}\{\cos(\omega - 2p)t + \cos(\omega + 2p)t\}$$

$$+ \frac{1}{6}\left\{-J_0\left(\frac{3\pi}{2}\right) + 2J_2\left(\frac{3\pi}{2}\right)\cos 3st - \cdots\right\}$$

$$\{\cos(\omega - 3p)t + \cos(\omega + 3p)t\} + \cdots\Big]. \qquad (6)$$

If (6) is expanded, we find that the angular velocities present are

$$\begin{array}{lllll}
\omega & \omega - p & \omega + p & & \\
\omega - s & \omega - p \pm 2s & \omega + p \pm 2s & \omega - 2p \pm s & \omega + 2p \pm s \\
\omega + s & \omega - p \pm 4s & \omega + p \pm 4s & \omega - 2p \pm 3s & \omega + 2p \pm 3s \\
 & \omega - p \pm 6s & \omega + p \pm 6s & \omega - 2p \pm 5s & \omega + 2p \pm 5s \\
 & \vdots & \vdots & \vdots & \vdots
\end{array}$$

$$\begin{array}{lll}
\omega - 3p & \omega + 3p & \\
\omega - 3p \pm 2s & \omega + 3p \pm 2s & \\
\omega - 3p \pm 4s & \omega + 3p \pm 4s & \\
\omega - 3p \pm 6s & \omega + 3p \pm 6s & \text{etc.} \\
\vdots & \vdots &
\end{array} \qquad (7)$$

Of course, in the actual case of scanning, α does not vary sinusoidally but it does vary in a quasi-periodic fashion, so that its variation can be described by a Fourier series in terms of the angular velocity, s. The analysis is then straightforward, the series being used in (4) to derive an expression similar to that of (6). The net result would be merely to change the magnitude of the component angular velocities.

Equation (7) can be interpreted as meaning that the single discrete sum and difference frequencies, appearing as side bands, are replaced by groups of frequencies, i.e., each side band has its own satellite of side-band frequencies whose spacing is given by $s/2\pi$. If we assume that the fiftieth harmonic of $s/2\pi$ is necessary to describe faithfully the variation of α, then the band width requirement is only increased 100 cycles over the requirement based on a fixed marking period. Consequently, in the discussion to follow on the band width requirements for facsimile keying under operating conditions, the marking period will be regarded as constant.

Present-day diversity reception with its associated limiter-keyer figuratively uses only the main body of the signal, that is, the minimum and maximum values of the signal are rejected by the threshold value and limiting action, respectively. Under these conditions, the signal exhibits an essentially linear slope from a point $12\frac{1}{2}$ per cent above its minimum value to a point $12\frac{1}{2}$ per cent below its maximum value. Accordingly, the actual signal envelope may be replaced by a trapezoidal one.

It is evident that fading will cause varying weight of marking for such an envelope since for a strong signal, the threshold value of the limiter-keyer is reached earlier than in the case of a weaker signal.

The slope of the equivalent signal at α can be determined by taking the derivative of the Fourier series, representing the square dot, and substituting α for pt.

Thus, the slope, S, is

$$S = -\frac{2}{\pi}[\sin\alpha\sin pt + \sin 2\alpha \sin 2pt + \sin 3\alpha \sin 3pt + \cdots]. \quad (8)$$

Substituting in (8)

$$S = -\frac{2}{\pi}[\sin^2\alpha + \sin^2 2\alpha + \sin^2 3\alpha + \cdots]. \quad (9)$$

Note the identity

$$\sin^2 m\alpha = \tfrac{1}{2}(1 - \cos 2m\alpha) \quad (10)$$

and substituting (10) in (9), we get, after some algebraic manipulation,

$$S = -\frac{1}{\pi}[1+1+1+\cdots-(\cos 2\alpha + \cos 4\alpha + \cos 6\alpha + \cdots]. \quad (11)$$

Now,

$$\cos 2\alpha + \cos 4\alpha + \cos 6\alpha + \cdots + \cos 2n\alpha = \frac{\cos(n+1)\alpha \sin n\alpha}{\sin\alpha}. \quad (12)$$

Substituting (12) in (11) we obtain the slope S as a function of n retained terms,

$$S = -\frac{1}{\pi}\left[n - \frac{\cos(n+1)\alpha\sin n\alpha}{\sin\alpha}\right]. \quad (13)$$

Table I gives the values of S for ten per cent dot ($\alpha = \pi/10$) and fifty per cent dot ($\alpha = \pi/2$) for the various numbers of harmonic terms retained.

TABLE I

Number of Terms Retained	Slope for 10% Mark	Slope for 50% Mark
1	0.06	0.64
2	0.28	0.64
3	0.70	1.28
4	1.28	1.28
5	1.91	1.91
6	2.50	1.91
7	2.90	2.55
8	3.12	2.55
9	3.19	3.19
10	3.19	3.19
11	3.24	3.82
12	3.46	3.82
13	3.89	4.45
14	4.45	4.45
15	5.10	5.10
16	5.68	5.10
17	6.10	5.72
18	6.32	5.72
19	6.37	6.37

Distortion As a Function of Signal Slope

We shall now consider the effect of a sloping signal in producing distortion, by noting that with such a signal, the marking interval at half amplitude has the same weight as a perfect straight-sided signal. Consider Fig. 25.

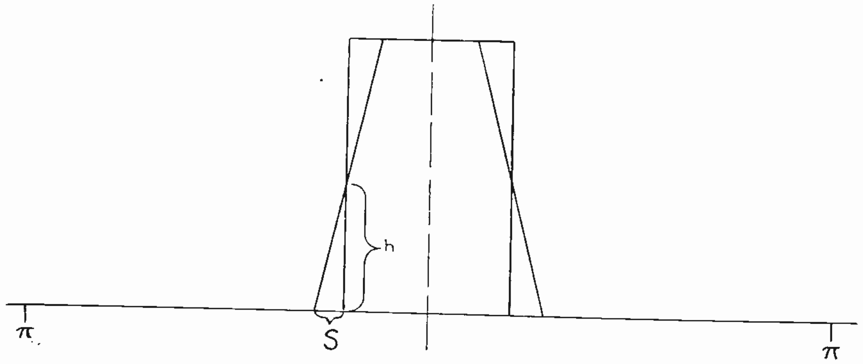

Fig. 25—Idealized signal envelopes.

Now note that

$$S = \frac{h}{\delta}. \tag{14}$$

Since h is equal to half of the signal amplitude,

$$h = \tfrac{1}{2} \tag{15}$$

so that,

$$S = \frac{1}{2\delta} \tag{16}$$

or,

$$\delta = \frac{1}{2S}. \tag{17}$$

Now the absolute distortion is defined as the elongation of the signal at the base divided by the length of the cycle. Consequently,

$$D = \frac{2\delta}{2\pi} = \frac{\delta}{\pi}. \tag{18}$$

Substituting for S, we get,

$$D = \frac{1}{2\pi S} \tag{19}$$

or,

$$S = \frac{1}{2\pi D}. \tag{20}$$

Thus from this equation we can determine the slope necessary to meet the permissible distortion and having the slope, we can find, from Table I, the number of terms which must be retained. The band width necessary is given by $2nf$, where f is the screen frequency; i.e., $\rho/2\pi$.

Our next step is to determine the effect of fading on distortion from which we can find the band width requirement to cope with this difficulty.

Distortion as a Function of Fading and Signal-to-Noise Ratio

Consider the idealized form of signal received with a finite number of harmonics as shown in Fig. 26, where the envelopes for two conditions of fading are represented.

Let,

Σ = maximum signal
σ = minimum signal
θ = threshold keying value
D = absolute distortion of signal at base
d = distortion at threshold for Σ
d' = distortion at threshold for σ
$\Delta = d - d'$, distortion at threshold due to fading
N = noise level
$\Sigma/\sigma = \phi$, fading ratio
$\Sigma/N = \mu$, signal-to-noise ratio.

From similar triangles

$$\frac{D-d}{D} = \frac{\theta}{\Sigma} \tag{21}$$

whence,

$$d = D\left(1 - \frac{\theta}{\Sigma}\right) \tag{22}$$

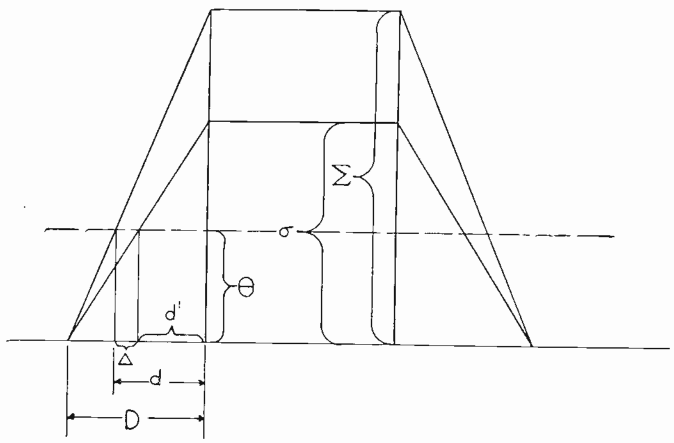

Fig. 26—Effect of fading on signal weight.

similarly,

$$d' = D\left(1 - \frac{\theta}{\sigma}\right) \tag{23}$$

$$\Delta = d - d' = D\left(1 - \frac{\theta}{\Sigma}\right) - D\left(1 - \frac{\theta}{\sigma}\right) \tag{24}$$

$$\Delta = \theta D\left(\frac{1}{\sigma} - \frac{1}{\Sigma}\right) = \frac{\theta D}{\Sigma}\left(\frac{\Sigma}{\sigma} - 1\right] \tag{25}$$

but since,

$$\frac{\Sigma}{\sigma} = \phi \tag{26}$$

we have,

$$\Delta = \frac{\theta D}{\Sigma}(\phi - 1). \tag{27}$$

Now $\theta_{\min} = N$, so that for extreme conditions

$$\frac{\Sigma}{N} = \frac{\Sigma}{\theta_{\min}} = \mu \tag{28}$$

whence,

$$\Delta = \frac{D(\phi - 1)}{\mu}. \tag{29}$$

Now Δ represents the distortion due to fading while D represents the distortion due to finite band width of the receiver. Rearranging terms in the last equation gives

$$D = \frac{\mu\Delta}{\phi - 1}. \tag{30}$$

This equation gives the distortion allowable due to band width for a given distortion due to fading in terms of the signal-to-noise ratio and fading ratio.

Since,

$$D = \frac{1}{2\pi S} \tag{19}$$

we can write,

$$S = \frac{1}{2\pi D} \tag{20}$$

and substituting for D in (30) we have

$$S = \frac{\phi - 1}{2\mu\pi\Delta} \tag{31}$$

which gives the required slope for given conditions of maximum permissible distortion, fading ratio, and signal-to-noise ratio.

Distortion As a Function of Multipath Transmission

The effect of multipath transmission is, in general, to add a constant amount of working time to the signal, although it is possible for cancellations also to take place. The additional signal duration so produced is dependent on the transmitter carrier frequency and the signal intensity, low frequencies producing greater time delay than high frequencies, and weak signals producing little or no delay.

Since the signal duration, due to multipath transmission, is independent of the band width transmitted, its effect is to determine the maximum keying frequency permissible to meet a given tolerance of distortion. That is, the actual duration of multipath contribution to the signal divided by the time of one cycle of the keying frequency must not exceed the permissible distortion.

Thus,

$$D_m = f_{max} t_m \tag{32}$$

where,

D_m = distortion due to multipath transmission

f_{max} = maximum screen frequency

t_m = time duration of signal contributed by multipath.

From this, we see that the maximum screen frequency which can be used is

$$f_{\max} = \frac{D_m}{t_m}. \tag{33}$$

The maximum distortion permissible is 0.06, a value which has been determined by inspection of large numbers of test subjects, while the time delay varies from 0.04 to 1.0 millisecond on long-distance East and West circuits from New York. On long-distance North and South circuits, experience has shown that the multipath effect is only one half to one third that of the East-West circuits. Using these figures we should expect that screen frequencies from 60 to 150 cycles would be usable.

However, since multipath transmissions occur during the time of large signal-to-noise ratio, the full effect of it is reduced by the setting of the threshold value of the limiter-keyer. As a result, it has been found possible to use screen frequencies from 90 to 175 cycles.

Band Width Requirement

Under extreme conditions of fading, the rectified output to the keyer will vary three to one, while the signal-to-noise ratio is four. With a distortion of 0.06 permissible and for a ten per cent dot we can determine the band width requirement at the receiver, by assuming that one half of the distortion may be caused by fading and that fluctuations in multipath transmission may contribute the other half. Under these conditions, $\Delta = 0.03$ and from (31) the required slope is

$$S = \frac{\phi - 1}{2\mu\pi\Delta} = \frac{3 - 1}{2 \times 4 \times 3.14 \times 0.03}.$$

$$S = 2.65.$$

Consulting Table I, we see that in order for ten per cent dots to have a slope of this value n must equal seven. From this we determine the band width requirement must be $14f$. Assuming that 175 cycles per second is the highest frequency to be used, a band width of $14 \times 175 = 2450$ cycles is necessary at the receiver.

This value has been experimentally verified by using receivers of 2800 and 10,000 cycles width and observing their outputs simultaneously. No practical difference in the signals was observed except that the narrow band receiver showed less background noise, as would be expected.

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NOTES ON INTERMEDIATE-FREQUENCY TRANSFORMER DESIGN*

By
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Summary—*A method of determining coil or condenser merit is cited. The more usual types of intermediate-frequency transformer assemblies are described. Formulas are given for predicting gain and selectivity. Frequent difficulties in composite amplifiers are mentioned. A method of obtaining high fidelity is described in detail.*

The merit of coils or condensers may be conveniently expressed by the sharpness of resonance of a circuit in which they constitute one reactive element. Sharpness of resonance, may be defined as the fractional change in current in an oscillatory circuit for a given change in either capacity, inductance, or frequency. The quantity "Q" of a circuit is equal to the ratio of the inductive or capacitative reactance to the resistance

$$Q = \frac{\omega L}{R} = \frac{1}{\omega C R}.$$

One common way of measuring Q is known as the capacity variation method. Consider that a coil is to be measured. It is tuned to resonance with a condenser of negligible losses. A vacuum tube voltmeter is used in shunt as an indicator. The capacity is then changed to values above and below resonance so that the reading of the voltmeter becomes 0.71 of the resonant value. The following simple formula may be used except for very low values of Q or for frequencies too near the natural resonance:

$$Q = \frac{2C_0}{\Delta C}.$$

C_0 is the total capacity in the circuit at resonance, and ΔC is the total change between the 0.71 readings.

Coils

Coils for intermediate-frequency transformers have several common types of winding. "Universal" windings (multilayer self-supporting coils) are commonly used for the low and medium frequencies,

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from 125 to 480 kilocycles. Several narrow "pies" in series give higher Q coils at the high-frequency end of this range. Bank windings and single layer solenoids are used for the higher frequencies. Litzendraht, especially when composed of many fine strands and silk covered, has a decided advantage over solid wire at medium frequencies.

Space requirements determine largely the form of winding which can be used and limit the value of Q which can be reached with an air-core coil. All coils suffer greatly from losses due to the placing of metal in their fields.

TABLE I

Type of Winding	Wire	Diam. of Core	L mh	Q	
				$1\frac{3}{8}$ inch □ Can	$1\frac{7}{8}$ inch ○ Can
1 Pie	No. 36 S.S. En.	$\frac{1}{2}$ inch	1.24	46	—
1 Pie	3/40 Litz	$\frac{3}{8}$ inch	1.15	75	—
3 Pie	7/41 Litz	$\frac{1}{2}$ inch	1.7	94	103
3 Pie	7/41 Litz	$\frac{5}{8}$ inch	1.8	87	108
4 Pie	7/41 Litz	$\frac{1}{2}$ inch	1.5	99	106
Iron Core 1 Pie	7/41 Litz	$\frac{3}{8}$ inch	1.5	136	—

Coils exposed to atmospheric humidity will slowly absorb moisture and suffer loss in Q. Treatment with certain waxes seems to be the best means of preventing such absorption.

Condensers

The Q of condensers may be found with the apparatus previously described for coils. The condenser under observation is substituted for a part of the tuning capacitance. As the Q of the condensers commonly used will be several times as large as the coil, best accuracy is obtained by choosing a very high Q coil in the measurement circuit.

$$Q_x = \frac{Q_1 Q_2}{Q_1 - Q_2} \cdot \frac{C_x}{C_{\text{total}}}$$

where Q_1, is the value first measured, Q_2 is the one using the unknown condenser and C_x/C_{total} is the ratio of the capacity of the added condenser to the total capacity in the circuit.

Variable capacitors are usually used in intermediate-frequency transformers so that the circuits may be accurately aligned to the proper frequency. If mica is used as the dielectric, it must be clean and have no flaws.

Air dielectric trimmers will average better power factor than the mica ones, but their greatest advantage is better stability with temperature variation. Greater cost and larger size has limited the use of air condensers.

Fig. 1 shows a double air trimmer designed for intermediate-frequency transformers. Each capacitor consists of a fixed and a variable section. The assembly of coils and trimmer fits into a shield can $1\frac{3}{8}$ inches $\times 1\frac{7}{8}$ inches and 4 inches long.

Coupling

Field linkage between transformer coils is a combination of electromagnetic and electrostatic coupling. If the coils are wound in the same direction and either the starting or finishing leads are connected to grid and plate, the magnetic coupling will oppose the capacity coupling. Reversing the connections or direction of winding of either coil will cause these couplings to become additive.

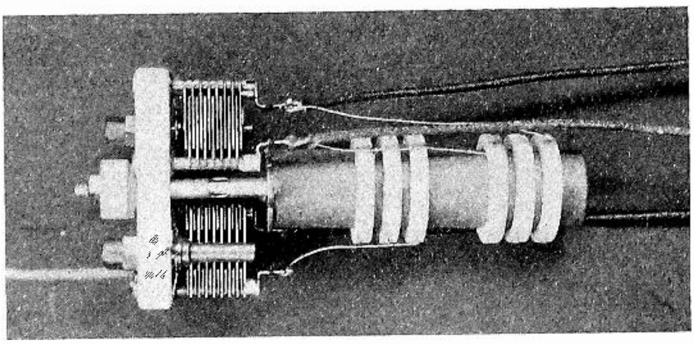

Fig. 1

It makes little difference in performance which type predominates as long as the effective coupling is of the right amount. Commercial practice is to keep the capacity coupling to a low value so as to make coil spacing noncritical.

Optimum coupling is the amount necessary for highest gain. Undercoupling gives better selectivity. Overcoupling will cause double humped curves and may cause tuning difficulties.

Shielding

Shielding of intermediate-frequency assemblies is necessary to confine the stray magnetic and static fields and prevent interaction between amplifier stages. Large sized cans of nonmagnetic, good-conducting material introduce the least losses at radio frequencies.

Stage Gain

High Q antiresonant circuits (having capacity and inductance in parallel) are commonly used in radio to obtain high impedances at a desired frequency of transmission. Such an impedance in the plate circuit of a modern radio-frequency pentode will give high voltage amplification.

Approximate formulas for stage gain are readily derived. The case of the single tuned circuit in the plate (impedance coupling) gives gain as the product of the tube's mutual conductance and the tuned impedance of the circuit.

$$\text{Gain} = G_m\left(\frac{1}{\dfrac{1}{r_p} + \dfrac{1}{Q\omega L}}\right)$$

where r_p is the tube plate resistance.

In the case of a double tuned transformer, the output voltage is taken across the secondary. The gain will be affected by the combined electromagnetic and electrostatic couplings between circuits. As it is difficult to separate and measure these couplings at radio frequencies, most formulas for gain are derived for critical coupling and make a further simplification in assuming it to be all of one kind.

$$\text{Maximum gain} = \frac{G_m}{2}\sqrt{\frac{1}{\dfrac{1}{r_p} + \dfrac{1}{Q_1\omega L_1}}}\sqrt{Q_2\omega L_2}.$$

Subscripts 1 and 2 refer to primary and secondary circuits respectively.

The product of Q and ωL represents an impedance. With constant Q more gain can be obtained by increasing the inductance. But L should not be increased to the point that stray wiring and tube capacity are too disturbing.

Selectivity

An antiresonant circuit offers less than resonant impedance to off-resonant frequencies. Such a device put in the plate circuit of a vacuum tube will cause the amplification to vary with frequency, as the output potential, μe_g, divides between the tube plate resistance and the load. The band width of a single isolated tuned circuit may be found to be approximately

$$\Delta f = \frac{f_r}{Q}\sqrt{\left(\frac{E_r}{E}\right)^2 - 1}.$$

If it is loaded with a vacuum tube plate resistance, the expression becomes

$$\Delta f = f_r\left(\frac{1}{Q} + \frac{\omega L}{r_p}\right)\sqrt{\left(\frac{E_r}{E}\right)^2 - 1}.$$

Δf is the band width corresponding to the ratio of resonant voltage, E_r, to off resonant voltage, E. f_r is the resonant frequency.

Band width is thus seen to vary inversely with the Q of the circuit and to vary directly with the resonant frequency, hence the frequent difficulty of obtaining sufficient selectivity at the higher intermediate frequencies.

The selectivity of two loosely coupled tuned circuits may be given as

$$\Delta f = \frac{f_r}{\sqrt{Q_1 Q_2}} \sqrt{\frac{E_r}{E} - 1}.$$

If the circuits have the critical coupling value and the primary is fed from a vacuum tube plate,

$$\Delta f = \frac{f_r}{\sqrt{Q_2}} \sqrt{2\left(\frac{1}{Q_1} + \frac{\omega L}{r_p}\right)} \sqrt{\left(\frac{E_r}{E}\right) - 1}.$$

Subscripts 1 and 2 refer to primary and secondary, as before.

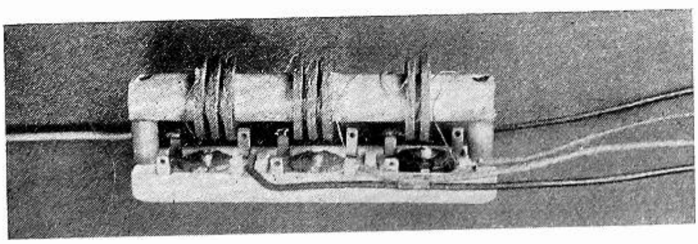

Fig. 2

Greater selectivity can be obtained in a single transformer by combining three tuned circuits rather than the conventional two. However, there is some sacrifice in gain.

Fig. 2 shows a triple tuned intermediate-frequency transformer. Each coil is tuned with a mica dielectric condenser. The two-pie winding is the plate coil. The middle coil is the coupling link between plate and grid; it has one external lead for the purpose of grounding.

Over-All Intermediate-Frequency Characteristics

The components of one stage of amplification are an amplifier tube and two transformers. The two-stages are comprised of two tubes and three transformers. The two stage amplifier may be used to obtain added gain but more often is used for better selectivity. Gain is a function of tuned circuit impedance, $Q\omega L$, while selectivity is a function of Q. Thus two-stage amplifiers often have low L but high Q circuits.

Feedback is always present to some extent in an intermediate-frequency amplifier, usually occurring in the common source of high potential and automatic volume control. Regeneration overcomes circuit losses and may effectively overcouple transformers which are adjusted for maximum nonregenerative gain, as $\omega M=\sqrt{R_1R_2}$ for critical

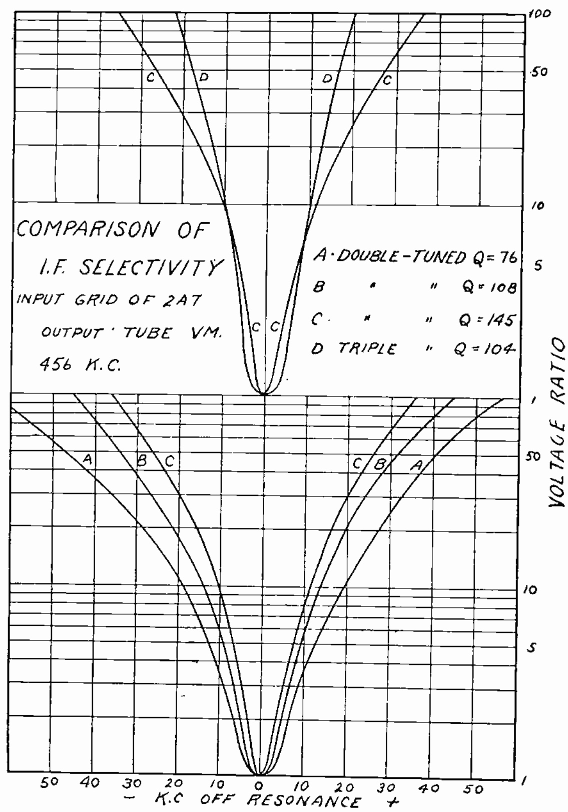

Fig. 3

coupling, where R_1 and R_2 represent the total effective values of resistance for primary and secondary.

The more common transformer connections cause a reversal in voltage phase. This, together with the reversal which takes place in the tube, puts the currents in the plate supply source in phase. A change in one of the coil connections will cause the stray currents to be out of phase.

Fig. 4 shows an intermediate-frequency selectivity curve in which the transformers were individually adjusted for maximum gain. The original curve was labeled "A." "B" was taken after a successful attempt to minimize reaction by improved filtering. "*C*" represents a further improvement and was taken after the coupling in each transformer had been reduced a little.

High Fidelity

All the radio industry is now interested in improved fidelity. It is generally conceded that maximum fidelity requires three major corrections in both transmitters and receivers; the auditory perspective

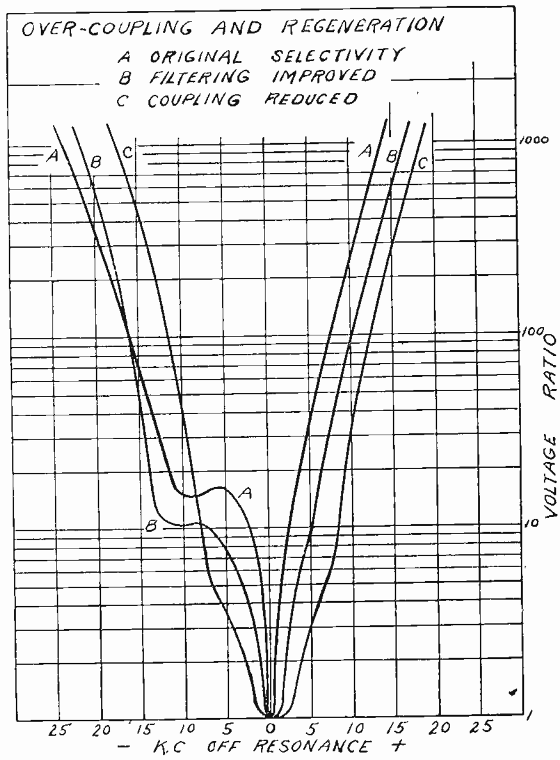

Fig. 4

should be improved, the volume range extended, and the frequency response should be leveled from 20 to 13,000 cycles. The latter is the only one which concerns intermediate-frequency transformer design. Side-band cutting or discrimination against the higher audio frequencies has been attendant to the use of higher *Q* circuits.

The extension in audio response to 13,000 cycles is not practicable with the present allocation of broadcast frequencies. A separation of ten kilocycles between carriers allows only 5000-cycle modulation before side-band frequencies overlap. Some semblance of high fidelity may be realized in the reception of a strong local transmitter, as receiver sensitivity is reduced to a point where there is no response to

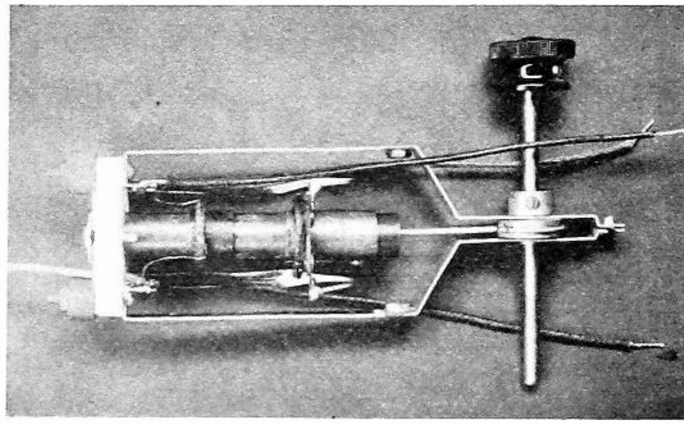

Fig. 5

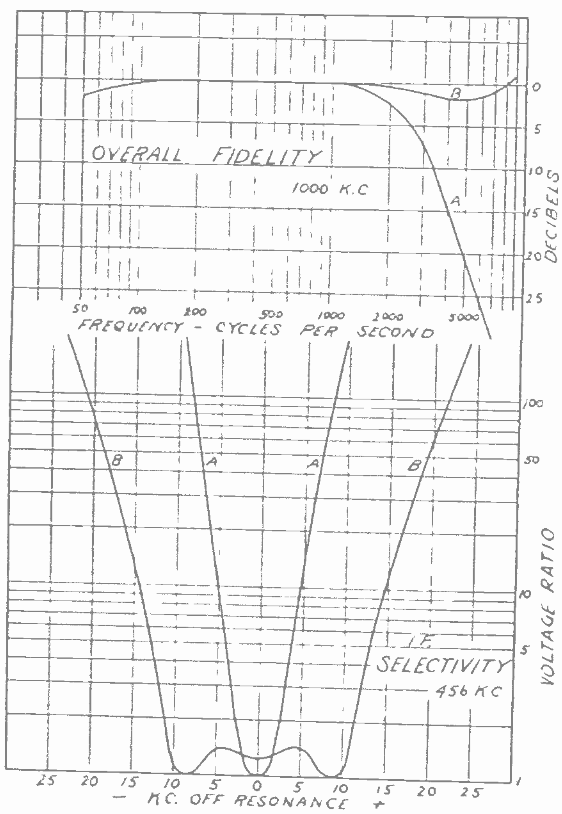

Fig. 6

weak interference. To give the best fidelity consistant with any receiving conditions variable band width intermediate-frequency amplifiers have been devised.

Three methods of band expansion are in vogue. One employs variable loss, absorption circuits coupled to several intermediate-frequency transformer secondary coils. Another has a means of staggering the tuning. A third varies the coupling of two transformers by moving the coils. A receiver was built up using the third method. The first two intermediate-frequency transformers had sliding primary coils moved by a simple cam mechanism capable of easy ganging. The third transformer, feeding the diode, had fixed coupling.[1]

Fig. 5 shows the sliding coil mechanism. The connections from the moving coil to the condenser are made through two long phosphor-bronze springs. Fig. 6 gives the electrical characteristics of the expanding band receiver, so equipped.

Band expansion gives additional problems in receiver design such as variation in sensitivity and difficulties in tuning.

[1] Wheeler and Johnson, "High fidelity receivers with expanding selectors," Proc. I.R.E., vol. 23, pp. 594–609; June, (1935).

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THE IONOSPHERE, SKIP DISTANCES OF RADIO WAVES, AND THE PROPAGATION OF MICROWAVES*

By
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Summary—*A detailed description of the ionosphere is given based on the recent measurements of the National Bureau of Standards and the Department of Terrestrial Magnetism of the Carnegie Institution.*

From the ionosphere data the skip distances of radio waves are calculated for temperature and tropical zones, with the diurnal and seasonal changes. The values agree with the observed values as far as they are known. It is concluded that daylight short-wave communication is controlled by the E region in summer and by the F_2 region in winter.

Theory indicates that the observed bending of microwaves over the horizon, and hence the successful microray communication to a distance of 200 kilometers, is due to diffraction of the microwaves over the bulge of the earth and to temperature gradients in the lower atmosphere.

INTRODUCTION

THE ionized regions of the upper atmosphere, known as the "ionosphere," are important in the long-distance propagation of radio waves, for the waves pass over the curve of the earth by successive reflections between the ionosphere and the surface of the earth. The skip distances of the shorter waves depend on the amount of ionization in the ionosphere and upon its height above the surface of the earth. Thus knowledge of the ionosphere leads to information about radio communication. It is only recently that extensive quantitative data of the ionosphere have become available. A program of measurements by means of the radio echo method of Breit and Tuve[1] was entered upon by the Department of Terrestrial Magnetism of the Carnegie Institution in coöperation with the United States National Bureau of Standards, with stations at Washington, D. C., latitude 38°48′ north, and at Huancayo, Peru, latitude 12° 3′ south. Results have appeared in a series of papers[2] which may be regarded as a first approach to a world-wide survey of the ionosphere. The data cover three years' observations at Washington and one year at Huancayo.

* Decimal classification: R113.61. Original manuscript received by the Institute, June 24, 1935.

[1] Breit and Tuve, *Phys. Rev.*, vol. 28, p. 554, (1926).

[2] Kirby, Berkner, and Stuart, PROC. I.R.E., vol. 22, pp. 481–521; April, (1934); Berkner and Wells, PROC. I.R.E., vol. 22, pp. 1102–1123; September, (1934); *Terr. Mag. and Atmos. Elec.*, vol. 39, p. 215, (1934); Wells, *Terr. Mag. and Atmos. Elec.*, vol. 39, p. 209, (1934); Gilliland, *Bur. Stan. Jour. Res.*, vol. 14, p. 283, (1935); Kirby and Judson, *Bur. Stan. Jour. Res.*, vol. 14, p. 469, (1935).

In the present paper the main features of the ionosphere are described. From the measured values of the ionization in the ionosphere the skip distances for Washington and Huancayo are calculated throughout the day and the season. The bending of waves below ten meters over the bulge of the earth is discussed.

The Ionosphere

In any region of ionization the electron density, i.e., the number of electrons per cubic centimeter, increases with the height above sea level to a maximum value. At this point the height is denoted by z and the electron density by y; z is termed the "height of the region above sea level" and y the "density of ionization of a region," although it is realized that the region of ionization may extend many kilometers above and below z. The radio echo method gives the "virtual," or optically equivalent, height z' of a region; this is always greater than the true height z of the region, but usually not very much greater. Although the exact value of $z'-z$ cannot be determined from the experimental data, one can estimate whether it is large or small from the manner in which z' changes with the radio frequency of the echo.

During the day there are three main regions of ionization in the ionosphere and two at night. These are designated by E, F_1, and F_2 for day and by E and F for night, the daytime regions F_1 and F_2 coalescing to form the night F region.

The Heights of the Ionized Regions

The observed virtual heights z' of all the regions are given in Figs. 1 and 2 for Huancayo from November, 1933, to August, 1934, and for Washington from April, 1933, to March, 1934. z' is probably not more than twenty kilometers greater than z for E, F_1, and F and not more than fifty kilometers greater than z for F_2. From the data of Figs. 1 and 2 the following conclusions may be stated: The height of E above sea level is about 100 kilometers and remains constant within twenty kilometers for day, night, latitude, and season. F_1 is at about 200 kilometers above sea level and its height varies less than about thirty kilometers with the latitude and season. The night F region is at about 250 kilometers above sea level, any variance with latitude and season being less than about thirty kilometers. The height of F_2, however, exhibits pronounced changes with the time of day, the latitude, and with the season in temperate latitudes. It increases to a maximum value at about noon at both Washington and Huancayo, the noon values at Washington being about 350 and 250 kilometers for summer and winter, respectively; the height at Huancayo the year round was about the same as for summer conditions at Washington.

The Density of Ionization of the Ionized Regions.

It may be stated without giving the details that experiment has shown that the ionization in F_1, F_2, and F is predominately electronic as far as the refraction and absorption of radio waves is concerned. Whether E is electronic or ionic has not yet been determined; some experiments indicate mainly ions and some show the presence of

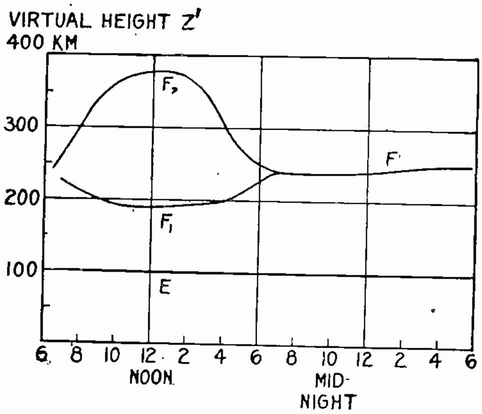

Fig. 1—Virtual height of the ionized regions, during 1933 and 1934 at Huancayo,[2] latitude 12° 3′ south.

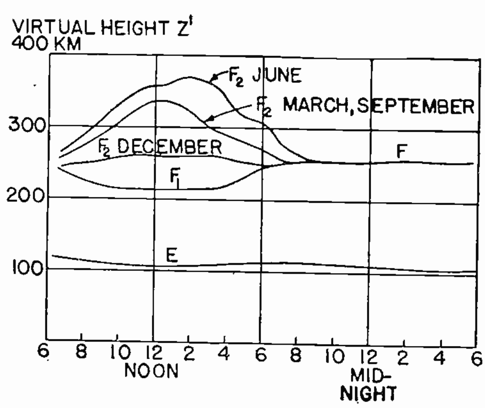

Fig. 2—Virtual height of the ionized regions during 1933 and 1934 at Washington,[2] latitude 39° 48′ north.

electrons. However, for considerations of radio-wave propagation, there is no error in treating the ionization in all regions as though it were electronic.

The observational data of y at Washington[2] are in the form of monthly averages of hourly values through the day and night over the year from April, 1933, to March, 1934, and at Huancayo[2] from November, 1933, to August, 1934. A portion of the measurements is given in Figs. 3 to 7. The data show that y of E and F_1 increases during the day to a maximum value at about noon, being greater in summer than in winter for Washington. At Huancayo y of E and F, exhibited no marked variations with the season of the year, but as shown in Fig. 3 y of F_2 was on the whole greater during the southern summer, November to March, than it was in the winter, April to August. Huancayo is so near the equator that large seasonal variations would not be ex-

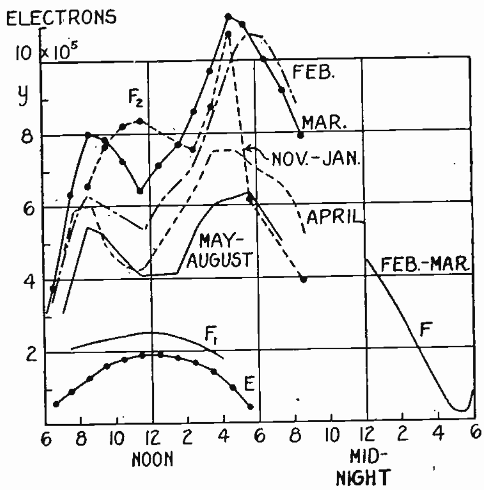

Fig. 3—The ionosphere at Huancayo,[2] November, 1933, to August, 1934.

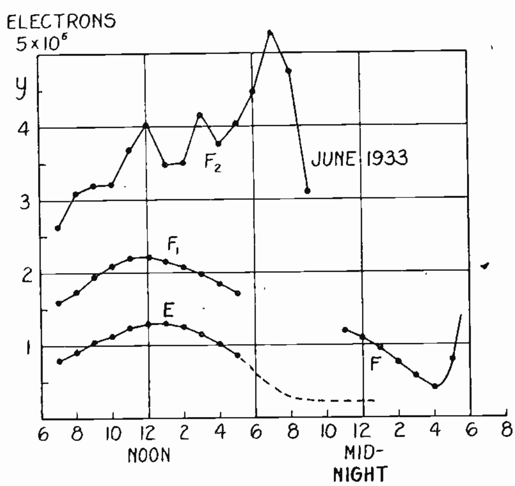

Fig. 4—The ionosphere at Washington,[2] June, 1933.

pected. At night y of E decreases to values below 0.3×10^5, which are two low to be measured within the range of the present radio echo equipment. Sudden and erratic increases were often observed in y of E night which might persist for a few minutes or an hour or more.

The diurnal and seasonal variations in y of F_2 are complex. At Huancayo and at Washington in the summer y of F_2 has two maxima, a ragged and erratic maximum in the morning, a minimum near noon and a higher, broader maximum in the afternoon. In winter at Washington y of F_2 is a maximum around noon. y of F_2 is at all times more variable than y of E or F_1, its actual value at any particular time may vary as much as forty per cent from day to day. At Huancayo and for summer and equinox at Washington, y of F at night decreases through the night with occasional temporary increases of as much as fifty per cent. For midwinter nights at Washington, y of F diminishes during the early part of the night, then it increases to a maximum between 2 and 4 A.M., and diminishes again until the rise at dawn.

General Description of the Ionosphere

From the data of Figs 1 to 7, we may describe the main world-wide features of the ionosphere as follows:

The E and F_1 regions are fairly simple, they lie at about 100- and 200-kilometer levels, respectively. The density of ionization of each is

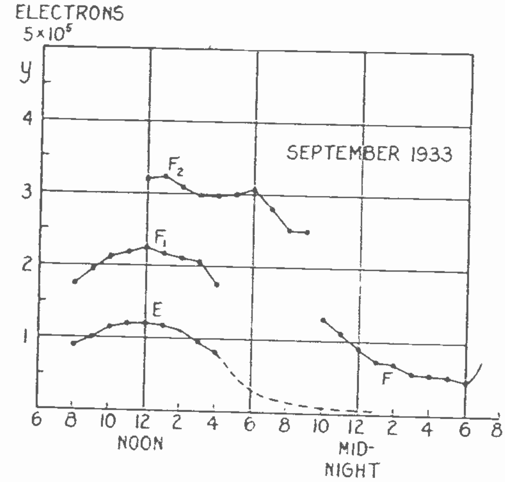

Fig. 5—The ionosphere at Washington,[2] September, 1933.

a maximum directly under the sun, so that the E region fades away in the night and the F_1 region is swallowed up on the twilight meridian by the descending F_2 region. The F_2 region is more complicated. It bulges out toward the sun, being 300 to 400 kilometers away from the surface of the earth at its most lofty point, and sinks back to a 250-kilometer level in the twilight zone. Its density of ionization is always greater and more variable than that of the E and F_1 regions. In tropical latitudes its density of ionization has two maxima, an irregular one around 10A.M. and a larger, smoother one near sunset; in temperate latitudes there is a single maximum in winter and in summer there are two maxima similar to the tropical case except that often the morning maximum is ill defined. During the night, apart from irregular ups and downs, the F ionization density diminishes at all latitudes except for a recrudescence in the small hours of winter morning in temperate latitudes.

A general theory of the ionosphere based on the hypothesis that the ionization is caused by the ultra-violet light of the sun has recently been sketched out by Hulburt.[3]

The Ionosphere and Solar Activity

The data of Figs. 1 to 7 were obtained during 1933 and 1934, an epoch of sunspot minimum and few magnetic disturbances. They there-

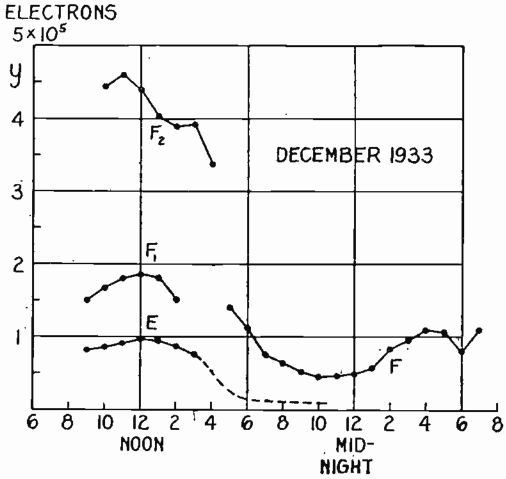

Fig. 6—The ionosphere at Washington,[2] December, 1933.

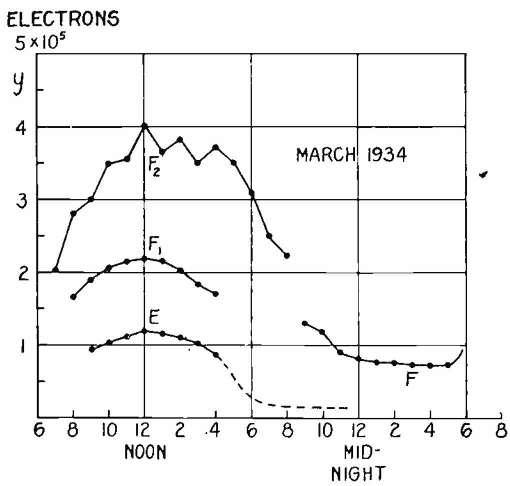

Fig. 7—The ionosphere at Washington,[2] March, 1934.

fore refer fairly well to the period of solar quiescence. Maximum solar activity is expected to occur about 1938. The observed changes in the skip distances during a sunspot cycle and a few ionosphere measurements indicate that the values of y should increase by about fifty to one hundred per cent from 1933 to 1938.

[3] *Terr. Mag. and Atmos. Elect.*, vol. 40, p. 193, (1935).

Detailed investigation of the ionosphere during magnetic storms has not yet been carried out. A number of observations indicate that the ionized regions, particularly the upper ones, are much confused during strong magnetic disturbances.

Skip Distances of Radio Waves

Approximate Formula for Skip Distance

The skip distance $2s$ of a radio wave of wavelength λ is related to y and z by the relation[4]

$$\frac{ye^2\lambda^2}{\pi m} = 1 - \frac{\sin^2 s/r}{\sin^2 s/r + (1 + z/r - \cos s/r)^2} \tag{1}$$

where r is the radius of the earth, and e and m are the charge and mass of the electron; c.g.s. electromagnetic units are used. The skip distance varies with the state of polarization of the radio wave and with the direction of propagation with respect to the magnetic field of the earth. The amount of the variation is not large, being in most cases less than fifteen per cent of $2s$, and (1), which refers to the polarization or propagation such that the electric vector of the wave is parallel to the magnetic field of the earth, is a fair average among the possible variations. (1) is approximate in that it is based on the assumption that the waves experience reflection from the ionized region, as if the region were a mirror. The approximation has been shown to be valid[4] in certain cases, although the waves probably are bent back to the earth by refraction in the region. In applying (1) to the case of several ionized regions, the assumption is made that the density of ionization between the regions is small; the assumption probably leads to no serious error.

Skip Distances During Minimum Solar Activity

The values of the skip distances were calculated by means of (1) with the values of y and z of Figs. 1 to 7, and are plotted in the curves of Figs. 8 to 11 for various times of day and seasons of the year 1933–1934. The period was one of minimum solar activity and it is of interest to compare the calculated skip distances with the average daytime temperate zone skip distances observed[5] in 1923, also a period of minimum solar activity. The comparison is shown in Table I; it is seen that the average of the last three columns, which are the calculated values taken from Figs. 8, 9, and 10, agree fairly well with the observed values in the second column.

[4] Taylor and Hulburt, *Phys. Rev.*, vol. 27, p. 189, (1926).
[5] Taylor, Proc. I.R.E., vol. 13, pp. 677–684; December, (1925).

It may be emphasized that the curves of Figs. 8 to 11 offer a more complete world-wide survey of the skip distances than has heretofore

TABLE I

Average Day Skip Distances

λ	Observed 1923–24	From Fig. 8 Jan., 1934	From Fig. 9 March, 1934	From Fig. 10 June, 1933
40 meters	400 km	400 km	500 km	400 km
30	700	800	800	700
20	1200	1500	1600	1400
16	2000	2300	2200	2200

been available. Fig. 8 shows that the winter skip distances reach their greatest lengths around midnight and then become shorter from 2 to 4 a.m. It is not certain that this effect has been observed. The effect is probably erratic, for the recrudescence of ionization in the small hours of winter night to which the effect is due, is erratic. It is a matter of general experience that the skip distances in the early morning hours are variable and ill defined.

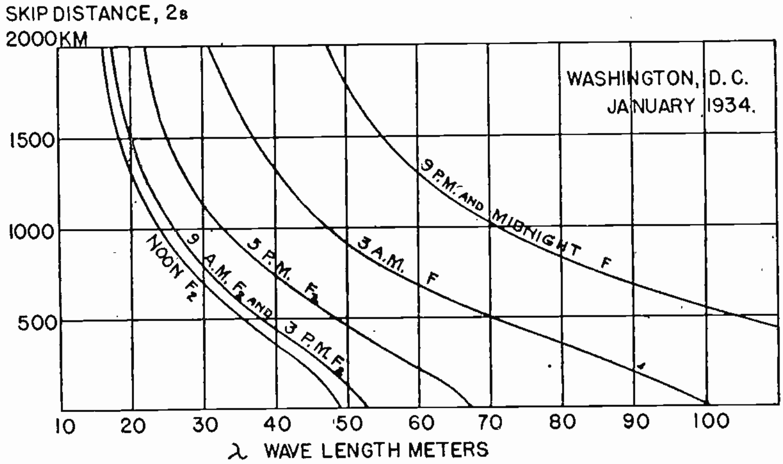

Fig. 8—Skip distances at Washington, January, 1934.

In Figs. 8 to 10 no skip-distance curves are given for the sunrise period between 3 A.M. and 9 A.M., because the ionosphere data for all regions are not sufficiently complete during these hours. The data indicate roughly that the average skip distances do not change much between 3 A.M. and sunrise, and that after sunrise they shorten rapidly.

It is hardly necessary to remark that in the case of the longer waves, for which the skip distances are less than 200 or 300 kilometers, the ground wave may often fill in the skipped region. The skip distance is essentially the range within which the sky wave cannot be refracted to the earth.

Effect of the Various Regions of Ionization on Short-Wave Propagation

Since there are several regions of ionization, the skip distances for each epoch were calculated for each region by the use of y and z for the region. The region which gave the shortest values of $2s$ was of course the region which controlled the skip distances for the epoch in ques-

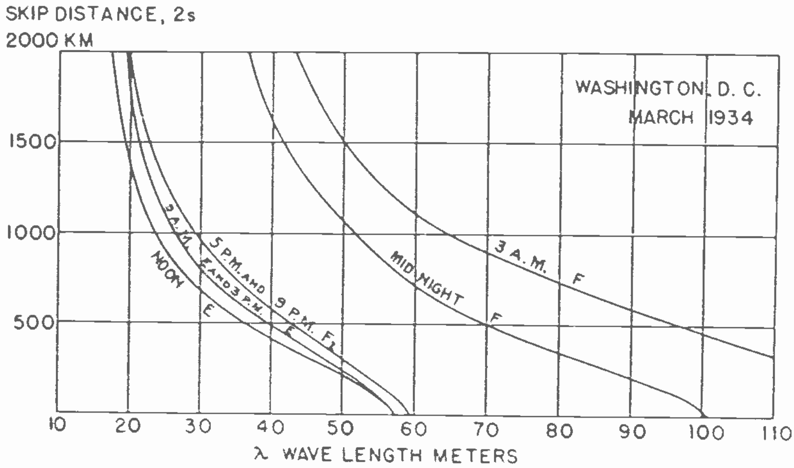

Fig. 9—Skip distances at Washington, March, 1934.

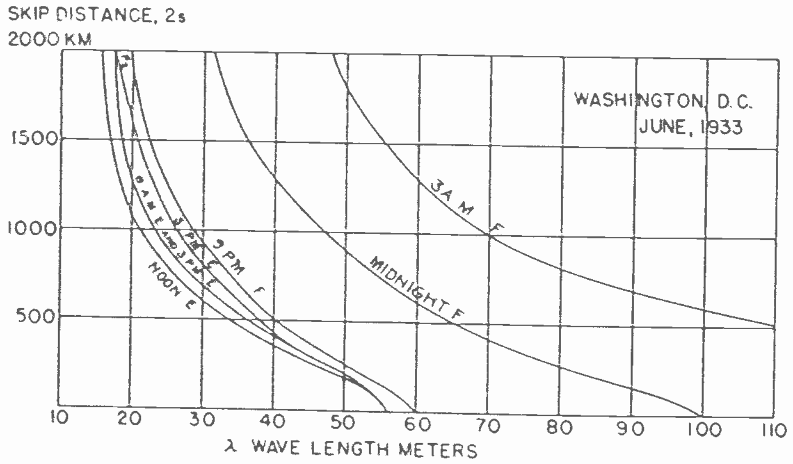

Fig. 10—Skip distances at Washington, June, 1933.

tion. In the curves of Figs. 8 to 11 the region which controlled the skip distances is marked on each curve. From this certain new facts stand out. For example, it is seen from Fig. 10 that at Washington in summer the skip distances during the day are controlled by the E region. This means that low angle rays, which are those of long-distance propagation, are turned down by E and do not get through to F_1 or F_2 at all. Whereas in winter, Fig. 8, the skip distances are controlled by

F_2, and hence low angle rays pierce through E and F_1 to be refracted by F_2. Therefore long-distance short-wave communication in winter is by rays which experience less than half the number of earth and ionosphere reflections which they experience in summer, since F_2 is more than twice as high as E. Further, absorption of energy from the radio wave is greater in E than in F_2, since the molecular density in E is greater than in F_2. Thus we find an explanation of the well-known fact that in general long-distance short-wave communication is better in winter than in summer.

The summer afternoon case, as shown by the 5 P.M. curve of June at Washington, Fig. 10, presented the complexity that for waves from about twenty to forty meters the skip distances were controlled by E, and for waves below twenty meters the skip distances were controlled by F_2. One might conclude that better temperate zone summer afternoon short-wave communication would be achieved by decreasing the wavelength until penetration of E was secured. However, such a conclusion is hardly safe in view of the fact that the values of y of Figs. 3 to 7 are monthly average curves of ionization which fluctuated within fairly wide limits from day to day, and any conclusion, as the foregoing, which demands high accuracy of the ionization curves, is not very dependable.

Other possibilities appear from the ionization curves. For example, a ray of waves of the correct wavelength and correct angle to the horizontal to pass through E transmitted southward from Washington might be refracted downward by F_1 or F_2 to E again over, say, Florida, and because of the increased y of E, due to the lower latitude, the ray might not penetrate E but be refracted upward again to F_1 or F_2; and so on, until the ray finally reached the latitude where it could get through E down to earth again. Such a ray would have an abnormally long skip; similarly for east-west propagation, where the change in y of the regions of ionization with longitude at any specified instant of time might introduce various complicated effects.

The Shortest Wavelength Useful for Long-Distance Communications

The skip distance curves of Figs. 8 to 10 rise rapidly with decreasing wavelength and soon become imaginary. The value of λ at this point, denoted by λ_1, is the shortest wavelength useful for long-distance communication, since waves shorter than λ_1 are not refracted back to the earth again. λ_1 may be determined from the $2s$, λ curves of Figs. 8 to 11, or, what amounts to the same thing, may be calculated directly from the relation[6]

[6] Hulburt, *Phys. Rev.*, vol. 31, p. 1018, (1928).

$$\frac{ye^2\lambda_1^2}{\pi m} = 1 - [r/(r + z)]^2, \tag{2}$$

where λ_1 refers to the region of height z and ionization density y.

λ_1 is tabulated in Table II. The values are open to considerable uncertainty, as much as twenty per cent, for (2), being a limiting case of (1), depends on the same assumptions as (1), but is more sensitive than (1) to errors in the assumptions. Table II agrees in general with

TABLE II

Values of λ_1

	Washington, 1933–1934 January	March	June	Huancayo, 1933–1934 Year average
6 A.M.	—	—	—	18 m
9 A.M.	15 m	19 m	17 m	13
Noon	14	17	16	13
3 P.M.	15	19	17	13
5 P.M.	19	17	16	10
9 P.M.	42	17	17	12
Midnight	42	32	28	13
3 A.M.	28	38	42	20

the observed values as far as they are known. A possible discrepancy is that according to the table λ_1 for noon at Washington is slightly greater in June than in January, whereas qualitative experience appears to indicate the reverse.

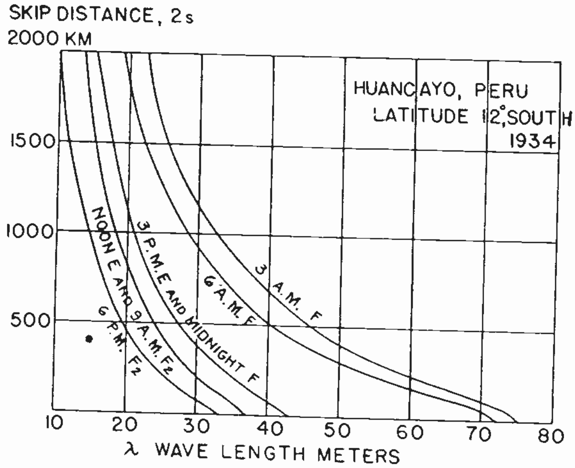

Fig. 11—Skip distances at Huancayo, 1933–1934.

Skip Distances, the Shortest Useful Wavelengths and the Sunspot Cycle

The skip distances $2s$ of Figs. 8 to 11 and the shortest useful wavelengths λ_1 of Table II refer to 1933 and 1934, a period of approximately sunspot minimum. Fragmentary data indicate that the ionization y of the regions of the ionosphere increases by about fifty to one hundred per cent from the minimum to the maximum of the sunspot cycle.

Therefore at sunspot maximum in 1938, $2s$ and λ_1 should be twenty to forty per cent less than the values of Figs. 8 to 11 and Table II. It will be of interest to calculate $2s$ and λ_1 each year as the ionosphere data of the year become available. Comparison will be possible between the calculated values for 1938 and the skip distances observed[7] during 1927–1928, the last epoch of maximum solar activity.

The Bending of Microwaves Over the Horizon

Experiments have shown that radio waves below ten meters in length, at the present time termed "microwaves," may be received at moderate distances over the horizon from the transmitter. Available evidence indicates that the microrays are not sufficiently refracted by the ionosphere to turn them back to the earth again. Therefore to account for the wave energy over the bulge of the earth one must turn to other causes. Refraction of the waves in the lower atmosphere and diffraction are obvious causes which come to mind, and, as shown in the following paragraphs, appear adequate to account for the observed effects.

Diffraction of Microwaves Over the Curve of the Earth

A complete solution of the diffraction of electromagnetic waves over the curve of the earth is complicated. An approximate solution was obtained by Epstein[8] for moderate distances neglecting effects of the atmosphere. Denote the distance of the transmitter from its horizon by b and the great circle distance from the transmitter to the receiver by x. For x not much greater than b the expression for the energy at the receiver involved a number of integrals and for x large with respect to b, and not too large, the energy E at the distance x was given by

$$E = E_0 r^2/\pi^2\lambda(x - b)^3(x + b)x, \tag{3}$$

where E_0 is the energy at the transmitter, r is the radius of the earth, and λ is the wavelength. Approximately (3) becomes

$$E = E_0 r^2/\pi^2\lambda x^5. \tag{4}$$

This yields a decline of the energy inversely as the fifth power of the distance.

Marconi[9] transmitted 50- to 60-centimeter waves from an elevation of 750 meters and found that reception over water was usually good up to about 200 kilometers and occasionally to 230 kilometers. The dis-

[7] Taylor and Young, in "Radio Engineering Handbook," McGraw-Hill, (1933).
[8] Epstein, *Proc. Nat. Acad. Sci.*, vol. 21, p. 62, (1935).
[9] Marconi, *Proc. Roy. Soc.* (London), vol. 27, p. 509, (1933).

tance b to the optical horizon was 98 kilometers. Epstein evaluated the expression for E at various distances as given in Table III, putting $E=1$ at 98 kilometers. From the table it is seen that at $x=200$ kilometers, or 102 kilometers from the optical horizon of the transmitter, the energy was about 1/200 of the energy at the optical horizon. Such an intensity appeared to be detectable and Epstein concluded that "the experimental range of the waves did not materially exceed the expectations of a theory disregarding all atmospheric influences."

TABLE III

x	E	x	E
98 km	1.00	180 km	0.012
100	0.96	200	0.0052
120	0.22	220	0.0026
140	0.057	240	0.0014
160	0.020	250	0.0011

It is noted that (3) or (4) yields an energy inversely proportional to the wavelength. Therefore the intensity of longer waves bent over the horizon is less than that of shorter waves, although the rate of attenuation with the fifth power of the distance is the same for all wavelengths.

Refraction of Electromagnetic Waves in the Lower Atmosphere

If the refractive index of air decreases upward, electromagnetic rays passing horizontally through the air will be bent downward. In order for the rays to be bent with the curvature of the earth the refractive index μ must satisfy the relation

$$\frac{d\mu}{dr} = -\frac{\mu}{r}. \tag{5}$$

where r is the radius of the earth, dr being positive away from the earth.

An upward decrease in μ may be caused by (a) an upward increase in temperature, (b) an upward decrease in water vapor, (c) an upward decrease in pressure, and (d) an upward increase in ionization. Of these four possible causes it will be shown that geophysical conditions are such that probably (a) is the most important, (b) is doubtful, and (c) and (d) are inappreciable.

The refractive index of air for waves longer than the infrared is $\mu=1.00027$, and with $r=6.36\times 10^8$ cm, (5) becomes

$$d\mu = -1.57 \times 10^{-9}\, dr. \tag{6}$$

The change in μ with the temperature t is given by

$$d\mu = -5.5 \times 10^{-7}\, dt. \tag{7}$$

Whence from (6) and (7)

$$\frac{dt}{dr} = 0.0029° \text{ centigrade cm}^{-1}. \tag{8}$$

Therefore t must increase upward 0.0029 degree centigrade per centimeter, or about one degree per ten feet, in order that the rays be bent with the curvature of the earth. Temperature gradients of this magnitude due to warm and cold streaks in the air occur and would be expected to cause fading and strengthening of microray signals.

The conclusion is in accord with the recent results of Hull[10] in his five-meter transmissions over a distance of 148 kilometers, one terminal being 200 meters and the other 100 meters above sea level. In 239 days of communication schedules only five days were missed because of insufficient signal level. We may conclude that the reliable communication was due to diffraction. Further, the strongest signals were obtained when the temperature increased upward as observed in airplanes to an altitude of 2000 meters. The highest signal levels were obtained on occasions when a warm tropical air was overrunning a cold polar air. This is in keeping with the temperature gradient refraction calculation.

The refraction calculation based on (5), (6), and (7) is independent of the wavelength; it is the same for all wavelengths from the infrared to the longest radio wave. It is similar to the calculation underlying the explanation of the twinkling of lights and stars. However, in considering radio waves one must keep in mind the relative size of the radio wave and the regions of the atmosphere where the temperature gradients exist. For example, a warm streak of air thirty feet across would be large enough to bend downward a five-meter wave but would have no noticeable effect on a 100-meter wave.

The effect of water vapor, case (b), may be calculated from the dielectric constant 1.0072 for pure water vapor at 17 degrees centigrade and atmospheric pressure, measured by Zahn[11] for frequencies of the order of one kilocycle, assuming that the same value is true for the frequencies of microwaves. It turns out that to bend the rays downward with the curvature of the earth there is required a water vapor gradient such that there is saturated vapor at 20 degrees centigrade at one point and zero vapor at a point one kilometer higher. Gradients of this amount, and greater, probably occur in the lower atmosphere. *A priori*, the gradients might be either plus or minus, that is, the water vapor might increase or decrease upward, depending on the previous

[10] Hull, *QST*, vol. 19, p. 13; June, (1935).
[11] Zahn, *Phys. Rev.*, vol. 27, p. 329, (1926).

history of the atmosphere. Thus, the water vapor might bend the rays either up or down.

However, Hull[10] found good agreement between the five-meter reception and temperature inversion. We may conclude that in his experiments either the water vapor bending was usually small compared to the temperature inversion bending, or that the water vapor bending was usually in the same direction as the temperature inversion bending.

Referring to (c), in order for refraction due to an upward decrease in pressure to satisfy (6), the change in pressure p with r must be

$$\frac{dp}{dr} = 10.5 \text{ dynes cm}^{-1} \tag{9}$$

which amounts to a decrease of 1050 dynes per meter. This is about ten times the normal decrease in pressure with altitude in the lower atmosphere. Such large pressure gradients do not occur except in very abnormal meteorological conditions as tornadoes and waterspouts. We may conclude that pressure gradients in the atmosphere are very rarely large enough to cause appreciable refraction of radio waves, long or short. Similar calculations led to the same conclusion for (d) gradients in the ionization of the lower atmosphere.

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ULTRA-SHORT-WAVE PROPAGATION OVER LAND*

BY

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Summary—*From theoretical considerations it is found that for ultra-short-wave propagation over level terrain, the received field should equal 4π times the product of the antenna heights divided by the product of the wavelength and the distance times the field that would be received for transmission in free space.*

This equation has been checked experimentally for horizontal polarization, antenna heights between 2 and 25 meters and frequencies between 17 and 150 megacycles for the two distances 9.4 and 26.3 kilometers. The results indicate that in the absence of detailed information regarding the transmission path, this formula gives the probable value of the received field. The deviations of an actual path from the ideal should cause corresponding deviations in the received field from that calculated by the above formula. For these two paths, the mean of the deviations was found to be about three or four decibels.

At 45 kilometers fading was observed with low antennas, and at greater distances there were rather large variations in the received field at all antenna heights available.

At higher frequencies and longer distances the curvature of the earth introduces additional attenuation. At the distance $d=5\times10^4k^{2/3}\lambda^{1/3}$ meters, theoretical considerations indicate that the curvature of the earth will reduce the field strength by a factor of about two, and beyond this "shadow distance" by the factor $2ka/\pi\sqrt{d^3/\lambda}$, where ka is the effective radius of the earth with refraction and λ is the wavelength, all measured in meters. This results in the received field being inversely proportional to the seven-halves power of the distance.

The method of calculating the received field in hilly country, which was developed on a theoretical basis in an earlier paper, has been confirmed experimentally both for an optical path and for a nonoptical path in the frequency range from 17 to about 100 megacycles. The discrepancies between the experimental and theoretical frequency characteristics between 100 and 200 megacycles indicate that some element of the phenomenon of ultra-short-wave propagation that has been omitted from the theoretical formula becomes important in this frequency range.

In these experiments it was found that comparatively small sloping areas of ground near the terminals reflected an appreciable fraction of the incident wave at the higher frequencies.

IN THIS paper, an approximate theory[1] of ultra-short-wave propagation over land is extended. For the case of propagation over level terrain, an approximation that greatly simplifies the expression for the received field has been made. Some experimental limits to the

* Decimal classification: R113.7. Original manuscript received by the Institute, May 3, 1935. Presented before Tenth Annual Convention, Detroit, Mich., July 3, 1935.

[1] J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, "Ultra-short-wave propagation," PROC. I.R.E., vol. 21, pp. 427–463; March, (1933), and *Bell Sys. Tech. Jour.*, vol. 12, pp. 126–161; April, (1933).

range of validity of this expression have been determined. The effect of placing one of the terminals on a hill has been examined theoretically. Theoretical frequency characteristics have been checked experimentally for transmission over two paths with elevated terminals; one optical and one nonoptical. In these experiments it was found that comparatively small sloping areas of ground near the terminals reflected an appreciable fraction of the incident wave.

Level Terrain

There are good experimental indications that the propagation of ultra-short waves over unobstructed paths may be interpreted on the basis of plane optics[1,2] and, that in the meter wavelength range, there is a well-defined reflected wave. Consequently for propagation over level terrain, the explanation is as follows (Fig. 1): Energy is propa-

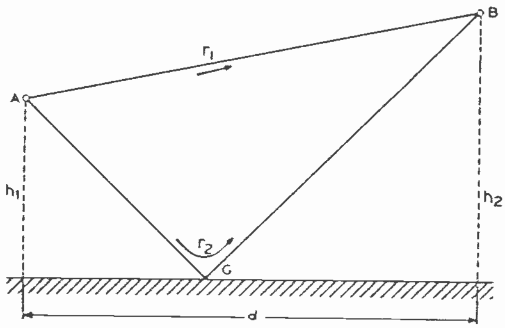

Fig. 1—Propagation over level terrain.

gated from a transmitter at A, at a height of h_1 above the ground, to a receiver at B, at a height of h_2 above the ground, both directly, as represented by r_1, and by reflection at G, as represented by r_2, the distance between transmitter and receiver being represented by d. For the practical case where h_1 and h_2 are small compared with d the reflected wave impinges upon the ground at nearly grazing incidence, so that a negative reflection coefficient the magnitude of which is unity[3] for ordinary ground (not water) is obtained. This results in the field at B being the difference between two vectors of approximately equal magnitude and differing in phase by an amount corresponding to the difference in path lengths, r_2 and r_1. For the case under consideration,

$$r_2 - r_1 = 2h_1h_2/d \tag{1}$$

[2] C. B. Feldman, "The optical behavior of the ground for short radio waves," Proc. I.R.E., vol. 21, pp. 764–801; June, (1933).

[3] For the case where one of the antennas is near the ground the small difference between the magnitude of the reflection coefficient and unity becomes important. This case together with that for which the conductivity is not negligible is considered in the Appendix.

and the angle between the vectors is

$$2\pi(r_2 - r_1)/\lambda = 4\pi h_1 h_2/\lambda d. \tag{2}$$

Except in the immediate proximity of the transmitter this angle is small[4] and the resultant field is

$$E = E_0(4\pi h_1 h_2/\lambda d) \tag{3}$$

where E_0 is the free space field. Substituting the value of the received field in free space,[5]

$$E_0 = 60\pi HI/\lambda d \tag{4}$$

in (3), the resultant field becomes

$$E = 240\pi^2 HIh_1h_2/\lambda^2 d^2. \tag{5}$$

Equation (5) says that the field over level terrain is inversely proportional to the square of the distance from the transmitter.

If instead of substituting the value of the free space field from (4) the equivalent equation in terms of radiated power[6] is used,

$$E_0 = \frac{3\sqrt{5}\sqrt{P_t}}{d} \tag{6}$$

and the received field is expressed in terms of the useful[7] received power,

$$P_r = \left(\frac{E\lambda}{8\pi\sqrt{5}}\right)^2 \tag{7}$$

the transmission loss may be expressed by

$$\frac{P_r}{P_t} = \left(\frac{3h_1h_2}{2d^2}\right)^2. \tag{8}$$

This says that the transmission loss for ultra-short-wave propagation over level terrain with nondirective antennas is independent of the frequency.

It was recognized that the range of validity of the above equations for the received field strength over level terrain would be limited at the short distances by the presence of the so-called "surface wave" and at

[4] When the angle is not small enough to make twice the sine of half the angle equal to the angle, (3) should be multiplied by $\sin(2\pi h_1h_2/\lambda d)/(2\pi h_1h_2/\lambda d)$.

[5] If I is in amperes, d in meters, and H the effective height of the antenna, and λ the wavelength in the same units, E_0 is given in volts per meter.

[6] The power is expressed in watts. Equations (6) and (7) apply to short linear antennas. Half-wave antennas have power gains of 0.4 decibel over short antennas.

[7] Useful power is used to mean the maximum power that can be transferred from the receiving antenna to the first circuit of the receiver.

the longer distances by the curvature of the earth. A series of experiments was conducted in an effort to determine the distances, antenna heights, and frequencies for which this equation is valid and also to determine the magnitude of the variations from the calculated values which might be expected to result from deviations of the actual terrain from a plane. The short-distance limit was found to be less than the

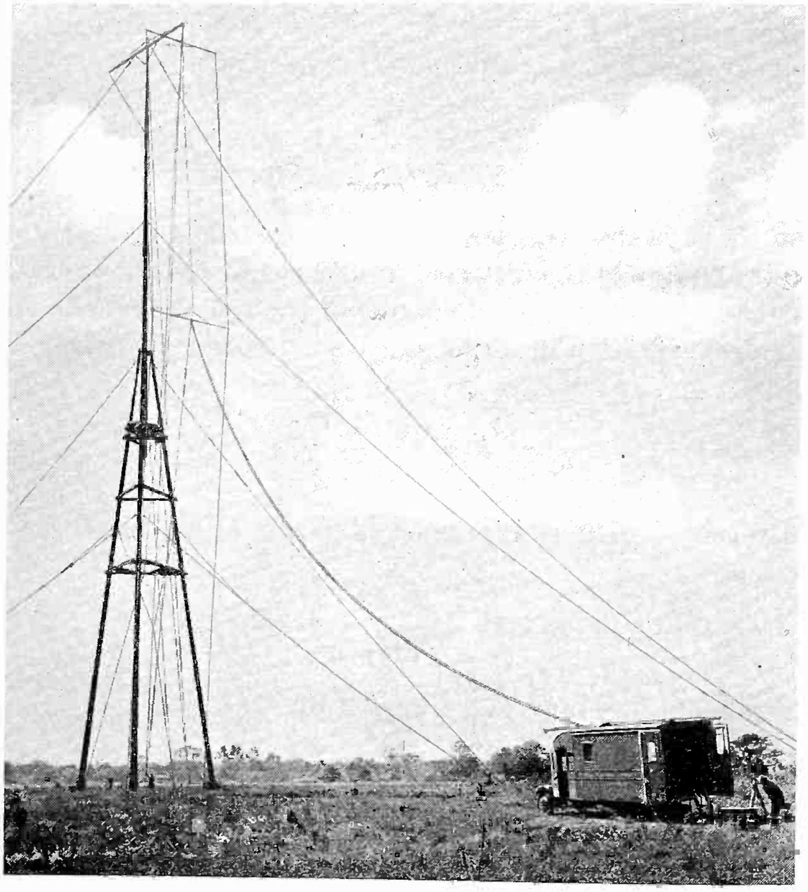

Fig. 2—Fifty-meter wooden mast and mobile transmitting laboratory. A transit was used to adjust the antenna height and to read the antenna current.

shortest path over which measurements were made (six miles). The inception of fading prevented the determination of the long-distance limit.

Locations

Data were obtained over the frequency range from 17 to 150 megacycles with antenna heights up to 25 meters over four paths employing horizontal polarization. Paths 1 and 2 of 9.42 kilometers (5.85 miles) and 26.3 kilometers (16.35 miles) were chosen as representing the nearest approach to a nonwooded level terrain path that could be found in New Jersey. When using paths 3 and 4 of 44.5 kilometers (27.6 miles) and 72.5 kilometers (45.0 miles) a rather large deviation from level terrain was tolerated in the vicinity of one of the terminals (*D* of Fig. 3) in order to take advantage of the gain which would result from using a fifty-meter pole already erected (see Fig. 2) and from the greater power available at Deal, N. J. An idea of the deviations of these paths from a plane may be obtained from the profiles given in Fig. 3

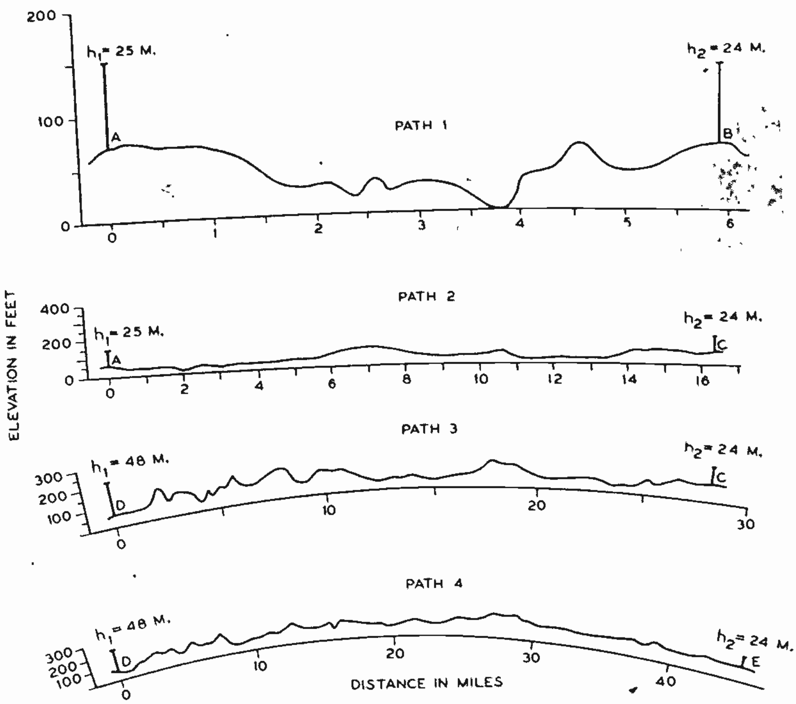

Fig. 3—Profiles of the level-terrain paths.

Locations of terminals.

A =	lat. 40°06′11″N,	long. 74°07′43″W.
B =	40°01′41″	74°10′52″
C =	39°55′36″	74°18′20″
D =	40°15′39″	74°02′00″
E =	39°50′28″	74°41′10″

in which the irregularities have been magnified Actually the irregularities on path 1 are smaller than those on path 2 which in turn are smaller than those on paths 3 and 4.[8] It should also be remembered that the contours on either side of the narrow paths indicated by the profiles affect the transmission.[9]

[8] Inspection of these profiles may lead the reader to question whether this terrain should be called "level." In common parlance it would be considered extraordinarily so. Its justification here naturally depends on whether the outcome of the experiment is in accord with what would be expected over truly level ground.

[9] Due to the impossibility of reproducing the desired detail, contour maps showing these transmission paths have not been included. The reader who is

The terminals were selected so as to have a minimum deviation from the ideal plane in their immediate vicinities. For each terminal that was located at some distance from the laboratories a portable antenna mast such as the one shown in Fig. 4 was employed.

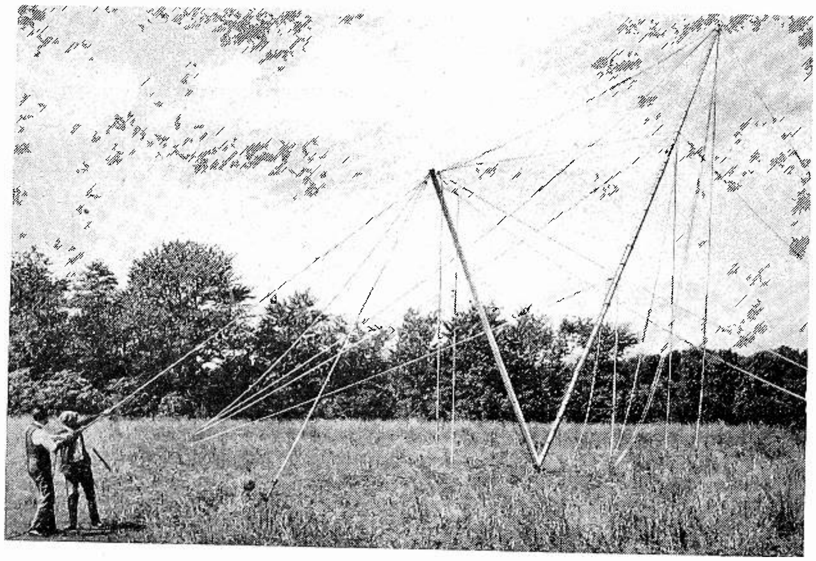

Fig. 4—Erecting one of the portable antenna masts. These masts consist of three 28-foot sections that can be transported on the side of a small truck. Two men can assemble and erect one of them in one hour. (At a new location, an additional hour is required to set the stakes.)

Method

The attenuation over each of these paths was obtained by comparing the field strength received over the path in question with the field strength measured near the transmitter. From the measurements made near the transmitter it was possible to determine the free space field as explained in the following paragraph. Because of the impossibility of measuring the field both near the transmitter and over the transmission path without distrurbing the receiving equipment, these fields were compared with an auxiliary field generated by a standard field generator.[1] Since the same standard field generator with the same loop and the same current indicating device were used at the same distance from the receiving antenna, the ratio of the received field strength to the free space field strength was obtained independent of absolute measurements of transmitted power and received field strength, thus eliminating any uncertainty that might exist in these measurements.

desirous of examining these paths more critically may locate the terminals from the information given on Fig. 3 on Atlas Sheets Nos. 28–29–32–33, published by the State of New Jersey.

The standard field generator was placed at such distances from the ground and receiving antennas as had been previously experimentally determined would eliminate any effect that might be due to the ground.

In order to determine the magnitude of the field that the transmitter would produce in free space at the distance of the receiving antenna, measurements were made locally at Deal. For this purpose, both the transmitter and the receiver were set up about a hundred meters apart on particularly level ground. Using the field from the standard field generator as a reference, the field from the transmitter

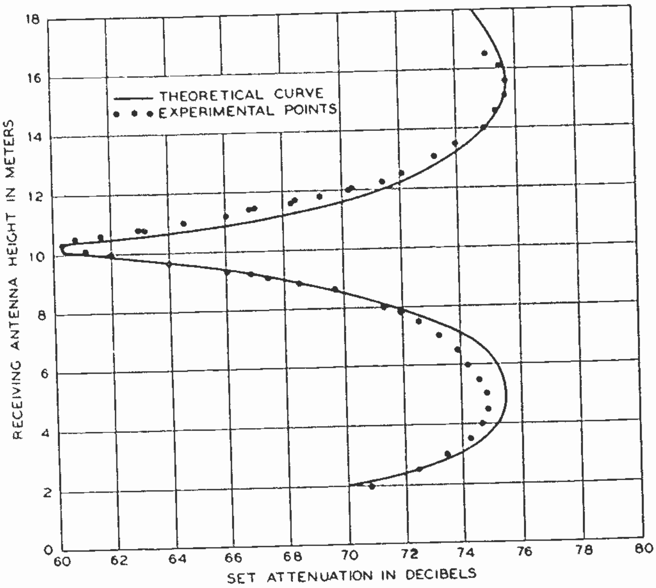

Fig. 5—Variation of received field with antenna height.
Frequency = 34.6 megacycles.
Path length = 108 meters.
Height of transmitting antenna = 55 meters.
Horizontal polarization.

was measured as a function of receiving antenna height, as shown in Fig. 5. By changing the antenna height the phase relations between the direct and reflected waves were varied so that both the sum and difference of these two components were obtained. This allowed the determination of the free space field as illustrated in footnote 10. To

[10] Method of calculating the free space field radiated by the transmitting antenna from the data of Fig. 5.

Average maximum attenuation = 75.2 decibels
Minimum attenuation = 60.4 decibels

Ratio $= 14.8$ decibels or 5.5 times $= \dfrac{1+K}{1-K}$

where K = ratio of incident to reflected field strength.

obtain the free space field for any path, it was necessary merely to divide by the ratio of the length of the path in question to the length of the path used in the calibration.

Experimental Results

A sufficient number of measurements[11] were made to allow their analysis on a statistical basis. In an effort to determine the best equation that would represent the variation of received field with antenna height, the best straight line was drawn through each set of data similar to that shown in Fig. 6. These lines then, gave the best curves of the form

$$E = kh^n \tag{9}$$

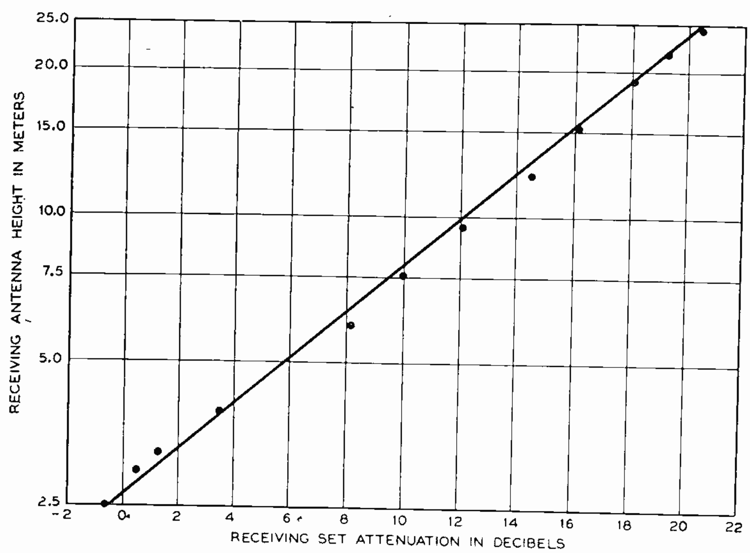

Fig. 6—Variation of received field with antenna height for 26.3-kilometer path (No. 2) on 34.6 megacycles with a horizontal half-wave transmitting antenna 24 meters above the ground.

through these data. An analysis of these curves showed that the mean value of the exponent (n) was unity well within the probable error, and that half of the slopes differed from unity by something less than ten per cent.

Since there is theoretical justification for the belief that the slope should be unity, the best straight line with that slope has been drawn

Therefore $K = 0.692$ = reflection coefficient.

Ratio of maximum to incident field strength $= 1.692$ or 4.6 decibels.

Therefore, the set attenuation for the incident field strength $= 75.2 - 4.6 = 70.6$ decibels. The set attenuation may be converted into the received field strength in decibels above one microvolt per meter by adding a number which is obtained from the receiving set calibration by means of the standard field generator.

[11] These include at least two variable height runs for each path and frequency in Table I from about one to 25 meters.

through each set of data. The deviations of individual points from these straight lines have been determined and plotted on probability paper as shown in Fig. 7. The fact that the curve is almost a straight line indicates that the deviations of individual points from the relation

$$E = kh \tag{10}$$

is approximately random. The curve also shows that the average deviation is 0.3 decibel (25 per cent of the points have deviations of 0.3 decibel or more in the negative direction and also 25 per cent have deviations of 0.3 decibel or more in the positive direction).

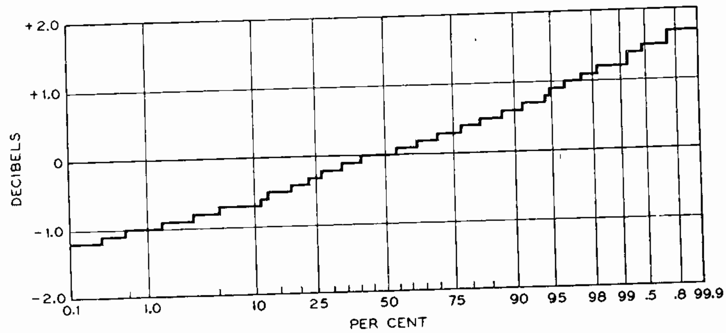

Fig. 7—Deviations of individual points from $E = kh$. The abscissa indicates the per cent of points whose algebraic deviation is less than the ordinate.

An important engineering conclusion which may be drawn from these data is that the relation $E = kh$ gives the *relative advantage* of different antenna heights for ultra-short-wave propagation over level terrain with a probable error of the individual observations of 0.3 decibel for the antenna heights, distances, and frequencies represented by these data.

The above discussion applies only to the relative field; the bearing of the data on the absolute field will now be considered.

The experimental determination of the dependence of field strength upon frequency and path length is not as simple as the experimental determination of the dependence upon antenna heights which has just been discussed. This is due to the fact that when either the frequency or receiving location is changed a new experimental setup[12] is required so that the measurements must be referred to a common base rather

[12] With change of frequency, the antennas and receiver tuning changes, necessitating calibration. Small variations in the distance and orientation of the standard field generator cause commensurate changes in the calibration. Possibilities of error are introduced such as would be caused by calibrating on a slightly different frequency, personal error in accuracy of tuning, etc. The same considerations apply to the change of receiving location. The time element and the possibility of the transmitted frequency being slightly different from that used for local calibration introduce other uncertainties.

than compared among themselves as is possible in the case of the variation with height. The base chosen for these comparisons was equation (3).

In the interpretation of Table I, the reader should bear in mind that fading was observed on the two longer paths, indicating instability of the transmitting medium (see later section). For path 3 the magnitude of the fading at the higher antenna heights was not sufficient to affect the tabulated results. For path 4, however, the fading magnitude at all antenna heights was sufficient to cause a rather large uncertainty in the tabulated results. Hence at best the theoretical equations which have been based upon a transmitting medium which does not vary with time can be expected to give only the order of magnitude of the received field strengths for these conditions.

TABLE I

DEVIATIONS (IN DECIBELS) OF MEASURED FIELDS FROM THE VALUES CALCULATED BY MEANS OF (3)

Frequency Megacycles	Path 1 5.85 miles	Path 2 16.35 miles	Path 3 27.6 miles	Path 4 45.0 miles
17.3	+1.6	−0.6 −5.0		−18.9
25.6	−3.6			
34.6		+5.7 −1.3	− 3.6	−15
51.4	−4.5	−1.4	−10.6	+ 2.4
75	+1.3	−4.0	− 7.2 −12.0	−12.5
100	+1.2	−3.6		
150	−5.5	−7.8		
Av.	+0.2	+0.7	− 8.4	−15.5

The deviations of the averages in Table I for paths 1 and 2 which are representative of level terrain paths, are within the experimental error for this set of data. Hence, it may be concluded that (3) has been verified for two representative level terrain paths of lengths 9.42 and 26.3 kilometers, respectively. It will be recalled that paths 3 and 4 deviate considerably from level terrain paths in the immediate vicinity of terminal D in that it is located behind a 150-foot rise (see Fig. 3). The measured fields over these paths averaged eight and fifteen decibels below those values calculated from (3).

The causes of the individual variations in the above table have not been determined. Part of the variation undoubtedly is the result of experimental error as indicated in the preceding footnote. The variations, however, are larger than should be expected from the conditions of the experiment as a result of this cause alone, so that it is the opinion of the experimenters that they are real.

These variations between the measured and calculated fields for paths 1 and 2 have been plotted on probability paper in Fig. 8. In so far as these data are representative of transmission conditions over level terrain, Fig. 8 indicates that the actual field received over level terrain will on the average be three to four decibels different from that calculated by (3). Some difference might be expected because of the deviations between the experimental path and an ideally level path. For the first two paths these deviations might be expected to cause oscillation in the frequency characteristic about that calculated from (3). In so far as the precision of measurement allows, the deviations tabulated above indicate the magnitudes of these oscillations. There also may be some fading whose period is long compared with the time required to make

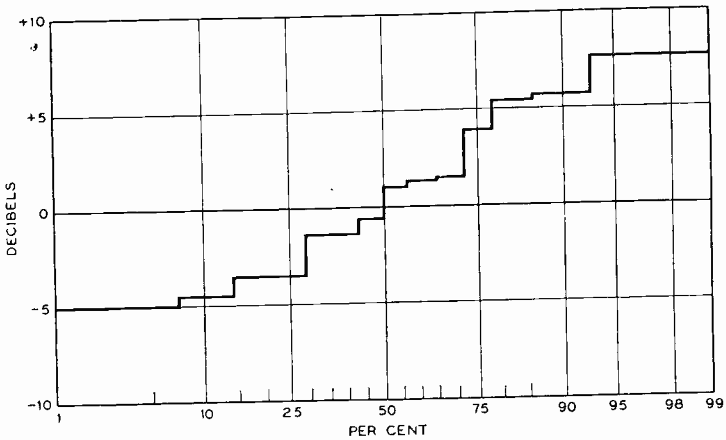

Fig. 8—Distribution of ratios, between measured and calculated fields over paths 1 and 2. The abscissa indicates the per cent of points whose algebraic deviation is less than the ordinate.

a measurement such as was found for the two longer paths. On the two longer paths both departures from level terrain and fading undoubtedly contributed to the recorded deviations.

While these data verify (3), which is based upon a theoretical picture that does not include the curvature of the earth, it cannot be concluded that for all frequencies the shadow or diffraction effect due to the earth's curvature is negligible, compared to the negative reflection effect. It is well known that at very much higher frequencies, such as those in the visible spectrum as an extreme case, the diffraction effect predominates.

Effect of Earth's Curvature

In a recent article, Epstein[13] has calculated the field to be expected for ultra-short-wave propagation over a perfectly absorbing earth and found the received field in this case to be the same as that which would

[13] Paul S. Epstein, "On the bending of electromagnetic micro-waves below the horizon," *Proc. Nat. Acad. Sci.*, vol. 21, pp. 62–68, January, (1935).

result from diffraction by a straight edge at the line of intersection of the planes tangent to the earth from the transmitter and the receiver. A similar result was obtained several years ago by F. B. Llewellyn but the results were not published because it was found that the earth could not be treated as a perfect absorber for ultra-short waves.[14] In accordance with the experimental evidence presented above, the earth definitely produces a "reflected" wave which near the surface is approximately out of phase with the direct wave and results in a received field very much less than that which would result if the earth were a perfect absorber. In this case we may think of a transmitter at

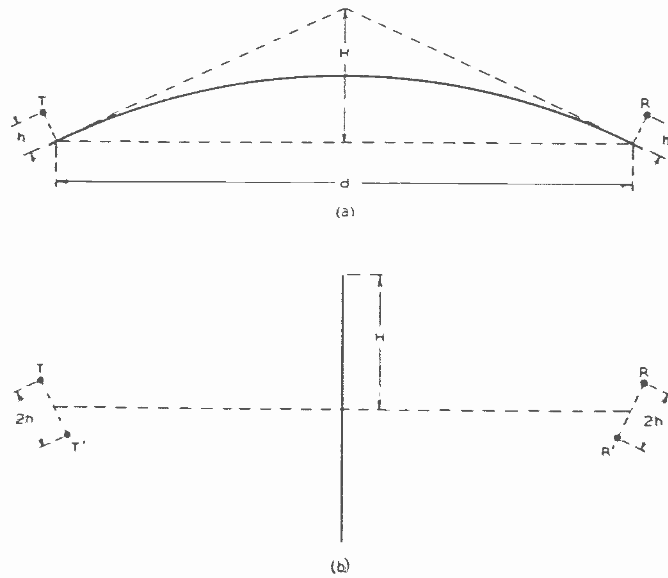

Fig. 9—Propagation over spherical earth.

T of Fig. 9(a) at a height *h* above spherical earth and a receiver at *R* at the same height. For the case of the perfectly absorbing earth the effect of the earth is the same as if a semi-infinite screen were introduced between the transmitter and the receiver as indicated in Fig. 9(a). In the actual case the effect of the earth near the transmitter appears experimentally to be the same as if a negative image of the transmitter were located[15] at *T′* (Fig. 9(b). A similar condition exists

[14] At some shorter wavelength, the irregularities of the ground may be sufficient to scatter the waves so that no well-defined reflected wave results. At this wavelength the assumption of a perfectly absorbing earth might be justifiable.

[15] This implies that it is the ground near the terminals which is most important. While a satisfactory theoretical proof of this is lacking, no other assumption seems consistent with the experiment employing variable antenna elevations. It is sometimes convenient to think of the ground near the radiator as part of the antenna system, the whole of which, in the case of level terrain, directs a null in the horizontal direction.

at the receiver. Hence, in addition to the shadowing effect of the earth which is taken into account by the semi-infinite screen, the earth produces negative images at the transmitter and receiver. This results in the received field at R being the sum of four components. These components may be calculated as the fields that would be diffracted from (1) T to R, (2) T' to R, (3) T to R', and (4) T' to R'. At the longer distances where an appreciable reduction in the field as a result of the earth's shadow might be expected, the height of the straight edge is large compared with the height of the antenna above the earth so that the magnitudes of these four components are approximately the same.[16] The first and fourth components are approximately equal in magnitude and equal in phase and the second and third components

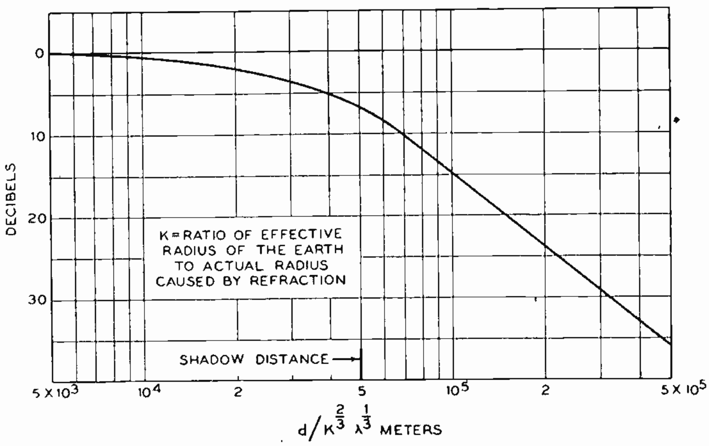

Fig. 10—Additional attenuation caused by the earth's shadow.

are equal in both magnitude and phase. Furthermore, the phase relation between the sum of the first and fourth components and the sum of the second and third components is exactly that which would have been calculated on the basis of plane earth. Accordingly, the received field for the case of ultra-short-wave propagation over spherical earth is equal to that which would result over plane earth multiplied by an attenuation factor which is equal to twice that which would result by diffraction over perfectly absorbing spherical earth. This factor is plotted in Fig. 10. For short distances the effect of the earth's curvature is negligible. For large distances this factor is inversely proportional to the three-halves power of the distance so that the received field varies inversely with the seven-halves power of the distance. Since it is known[1] that the usual effect of atmospheric refraction is to increase the effective radius of the earth, this same curve may be used to cover

[16] As the distance is increased the differences between the magnitudes of these components decreases without limit.

the case of refraction by the lower atmosphere by assigning the proper value to k, the ratio of the effective radius of the earth to its actual radius.

As a partial check on the correctness of the theoretical interpretation represented in Fig. 9, it is interesting to note that for the case of plane earth, this analysis gives the correct result. Here, instead of having the two components that result from the reflection picture, Fig. 1, there are four components. The first and fourth add up identically to the direct wave and the second and third are each equal to one half of the reflected wave. Furthermore, the phase relations are identical with those obtained in Fig. 1.

As the frequency is increased the level terrain attenuation becomes less and less until the wavelength becomes comparable with the path difference between the direct and reflected waves at which frequency the effect of reflection from the ground is to introduce interference fringes. At these frequencies the received field strengths for perfectly absorbing spherical earth are exactly one fourth the maxima of these fringes. Even for these frequencies as the distance is increased, the negative reflection effect again becomes important and finally the received field varies inversely as the seven-halves power of the distance. Of course, at some high frequency, the irregularities of the earth will be sufficient to cause scattering rather than regular reflection. At these frequencies the results obtained on the basis of a perfectly absorbing earth might apply. Moreover, the absorbing earth picture may give the correct order of magnitude at lower frequencies if sufficiently high antennas are used.[17]

To date, these theoretical calculations of the effect of the earth's curvature have not been checked experimentally. Even for the highest frequency and longest path for which experimental data are given above, the calculated additional attenuation caused by the earth's curvature is only six decibels. Here, the instability of the lower atmosphere caused such deep fading that the experimental data obtained could hardly be used to check a theory which is based on a transmitting medium that does not vary with time. The reader should be cautioned that while this theoretical formula is based on assumptions that appear reasonable, their validity has not been proved and although this formula probably represents the best estimation of the effect of the earth's curvature, at the present time, implicit faith should not be placed in it until it has been verified experimentally.[18]

[17] These heights may be impractically high.

[18] It, of course, does not apply for conditions under which fading is observed unless the fading is caused by variations of the refractivity of the lower atmosphere that may be accounted for by a variation in the magnitude of k. It does

The limit of the optical paths for spherical earth, as defined by the geometric shadow, is given by

$$d = 3570(\sqrt{h_1} + \sqrt{h_2}) \text{ meters} \tag{11}$$

Except for very high frequencies and antennas this distance occurs within the range of distances for which the factor plotted in Fig. 10 is substantially unity, indicating that the effect of diffraction is to mask the geometric shadow.

The distance at which the "shadow effect" becomes appreciable might better be chosen as $5\times10^4k^{2/3}\lambda^{1/3}$ meters because at this point the earth's curvature has reduced the field strength to approximately half the value it would have above plane earth and at this distance the shadow begins to deepen at approximately its maximum rate. While the ratios of the path lengths to these "shadow distances" is the preferable basis on which to compare the relative "shadow effects" caused by the earth on two paths,[19] they do not, in general, give the correct indication of the relative total attenuations over the two paths because the additional attenuation caused by negative reflection as given in (3) usually predominates. In practice, an effort is usually made to place the antennas as high above the ground as possible. This is done, however, for the purpose of reducing the attenuation caused by negative reflection rather than to obtain what is known as an "optical path"; i.e., a path where the straight line between transmitter and receiver does not pass through an obstruction.

Data presented as Fig. 8 of reference 1 may be used to verify the level terrain formula and to illustrate the rôle of the "shadow distance." These data have been replotted as Fig. 11. The fact that the ground in the immediate vicinity of the measurement locations was more irregular than that for the new data presented in this paper, undoubtedly contributes to the deviations of the points from the mean curve. While data are shown for distances about eight times the longest optical path, the effect of the earth's shadow is masked by the negative reflection effect for the entire range of distances for which data are given. This could have been inferred from the fact that all of the measurements were made at distances less than the "shadow distance."

When the effect of the curvature of the earth as estimated in this section is taken into consideration, the received field strength may be expressed by

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not apply for any signals that may be returned from discontinuities in the lower atmosphere such as have been referred to by Ross A. Hull in "Air-mass conditions and the bending of ultra-high-frequency waves," *QST*, vol. 19, pp. 13–18; June, (1935).

[19] It is important to note that this is not a function of the geometry of the paths alone, but also depends upon the wavelength employed.

$$E = \left(\frac{3\sqrt{5}\sqrt{Pt}}{d}\right)\left(\frac{4\pi h_1 h_2}{\lambda d}\right)\left(\frac{\sin\,(2\pi h_1 h_2/\lambda d)}{2\pi h_1 h_2/\lambda d}\right)F. \qquad (12)$$

The first factor corresponds to the free space field of (4) or (6), the second and third factors together give the reduction caused by negative reflection and the last factor, F, gives the reduction caused by the curvature of the earth as plotted in Fig. 10. For sufficiently low antennas the third factor is unity, and for sufficiently short distances the fourth factor is unity. Under these conditions the field strength is given

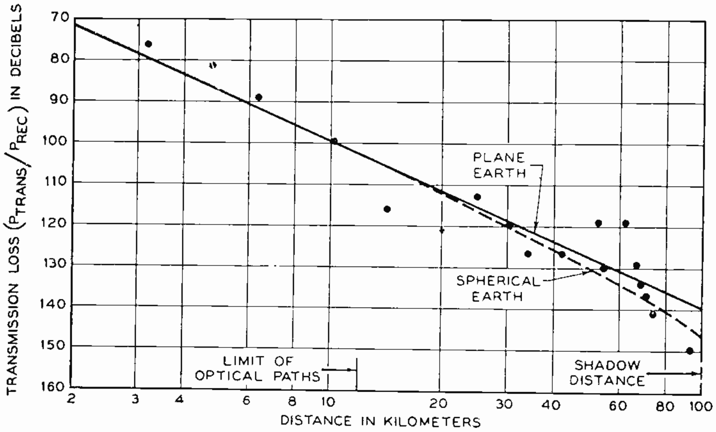

Fig. 11—Replot of experimental data presented as Fig. 8 of reference 1, together with the theoretical curves for ultra-short-wave propagation over plane and spherical earth. The experimental points represent measurements made on a horizontally polarized wave at a frequency of 69 megacycles over moderately level terrain.

by the first two factors which together correspond to equations (5) and (8).

Stability of Signals

Previous experiments[1] have indicated that fading may be encountered in the reception of ultra-short waves when the attenuation in addition to inverse distance attenuation reaches the order of thirty to forty decibels. Similar results were found while making the tests reported herein.

No fading was observed on any frequency over the two shorter paths 1 and 2, while fading was observed on all frequencies over the two longer paths 3 and 4. For all frequencies on path 3, the fading magnitude was greater for the lower antenna heights, indicating that the fading increases with the attenuation (in excess of the free space attenuation). This would be true if the fading were caused by the interference of a wave propagated along the surface of the earth with a wave reflected or refracted from above; e.g., from the under surface of clouds. At present it is impossible to say whether or not the energy is propagated by more than one path. It would, however, be possible to settle this question beyond doubt by employing either of the two methods ("pulse" and "frequency-variation") that have been used generally for determining the effective height on short waves. Arrangements are now being made for conducting these experiments.

In an effort to determine the diurnal variation of ultra-short-wave transmission, if any, one sixteen-hour test was made under the conditions which had seemed to be the least stable. Path 4, 45 miles long, was used with a 34.6-megacycle wave. The tests included sunrise and sunset. Half-wave horizontal antennas were employed, 48 and 25 meters above the ground at the transmitter and receiver, respectively.

For this test the day was divided into eight half-hour periods centered about sunrise, eight one-hour periods during the day and eight half-hour periods centered about sunset. During the first fifteen minutes of each period, the field strength was continuously recorded. The remaining part of the period was devoted to calibrating the receiving system and making such observations as the vagaries of the field made pertinent. The average field during these fifteen-minute periods varied by as much as twelve decibels throughout the tests. These variations apparently were not systematic.

While it is not safe to draw general conclusions from one day's test, it may be significant that the average field was weaker for the first four hours after sunrise than for the remainder of the test. The variations within the fifteen-minute intervals are shown in Fig. 12. Throughout the day there appear to be small variations of about a three-minute period superimposed upon larger variations with a period two or three times as long. The deepest fade within a fifteen-minute interval had an amplitude of about twenty decibels and occurred about an hour after sunrise. The range between maximum and minimum fields during the test was about thirty decibels. Except for a fifteen-minute period about a half hour after sunrise there were no variations so rapid that they could not be shown on Fig. 12. A copy of the original record taken during this period is shown in Fig. 13(a). The record shows fades of about one- or two-decibel amplitude at the rate of about five per minute, increasing in rate to about sixty per minute. This rapid fading commenced about fifteen minutes after sunrise and disappeared about ten minutes later not to appear again during the test. Fig. 13(b) is a sample of the record showing the more prevalent type of fading observed. Fig. 13(c) shows how steady the field is at times.

During the day some field strength records were taken with the receiving antenna at different elevations. The records for 12- and 18-

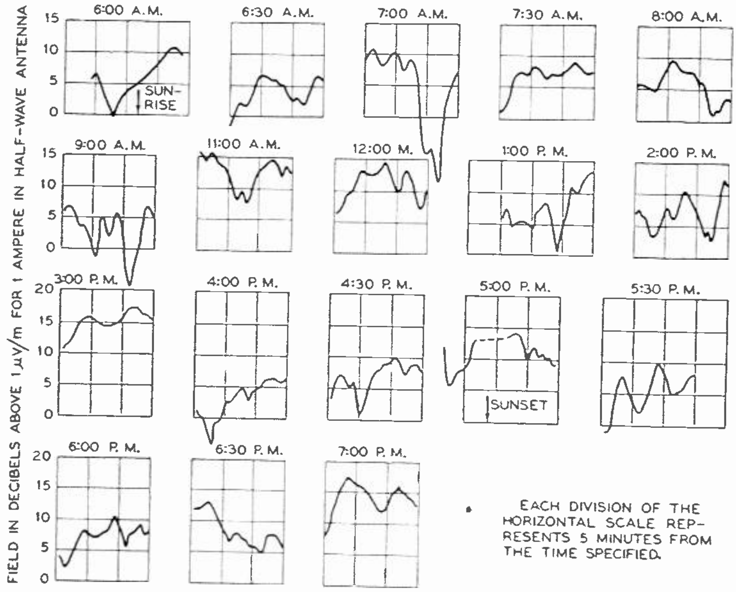

Fig. 12—Fading record over 72.5-kilometer path on 34. 6 megacycles with horizontal polarization. Transmitting height 48 meters, receiving antenna height 25 meters.

meter heights do not show any significant difference in either the type of fading or average field strength from those taken at the 24-meter

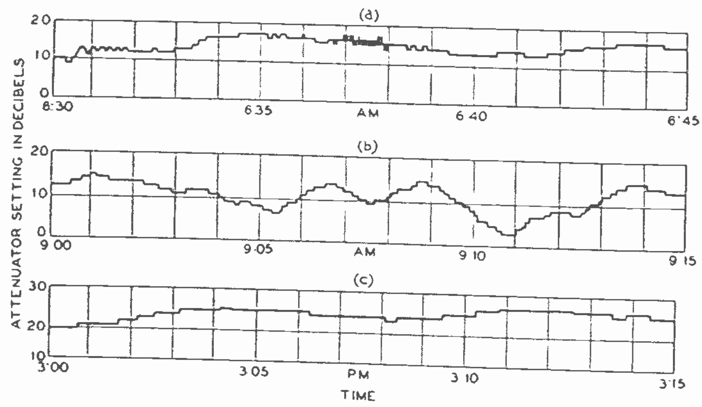

Fig. 13—Samples of the fading record.

height. The average fields at the six- and three-meter heights were less than those at the 24-meter height approximately in the ratio of heights, as would be expected from (3).

Level Terrain with Elevated Terminals

In locating ultra-short-wave terminals, hills may be employed to increase the antenna height. If the ground between the transmitter and receiver is substantially level except for the slope of the hill supporting the antenna the effect of the hill on the received field may be estimated as follows:

The received field may be considered as a vectorial sum of a direct wave and a wave reflected from the average slope in front of the antenna.[1,20] The magnitude of the direct wave may be expected to increase proportionally with the height above the level valley in accordance with (3). The reflected wave would be less than the direct

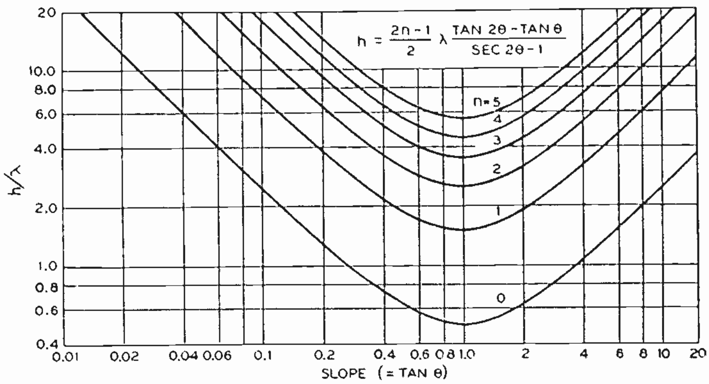

Fig. 14—Optimum heights for antennas above ground having a slope equal to $\tan\theta$. *Caution:* For large slopes, this applies only to horizontal antennas.

wave by the loss at reflection and the ratio of the heights of the antenna and of the point of reflection above the level ground. The relative phase between the direct and reflected components would produce maxima and minima as the antenna height is varied. On the basis of geometric optics all the maxima that occur at heights less than twice the height of the hill above the valley would be approximately equal. This is more exactly true for the smaller slopes. For slopes of more than 45 degrees the maxima would be expected to increase with antenna height but then the antenna must be located on the side of the hill to obtain full advantage of the slope. Actually, the maximum fields would be somewhat less than this because of the finite extent of the reflecting plane, but there probably would be little advantage in increasing the antenna height above the first maximum unless the height of the antenna structure is made greater than the height of the hill above the level ground.

The heights for which these maxima occur are given in Fig. 14. Even for wavelengths in the meter range, slopes which would permit

[20] Both of these waves are in themselves composed of two components as indicated in Fig. 1.

supporting the antenna at the first optimum height with ordinary poles are not uncommon.

Hilly Terrain

Experiments were made on the propagation of ultra-short waves over two paths of this type. The first one, an optical path, was very similar to those already considered, the main difference being the location of the antennas on hills to increase their height above the intervening terrain. In the other path an intervening hill obstructed rectilinear propagation. (See Fig. 15.)

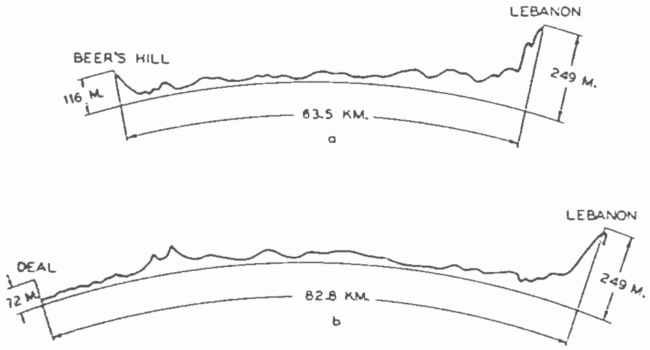

Fig. 15—Profiles of (a) optical path between Beer's Hill and Lebanon, (b) nonoptical path between Deal and Lebanon, (New Jersey).

The attenuation over these paths was obtained in the same manner as over the level terrain paths. The attenuation exclusive of that produced by local terminal effects was obtained from the variable antenna height characteristics in a manner similar to that described in Fig. 5 for separating the incident and reflected waves. The variable height characteristics for the receiving site at Lebanon, N. J., are shown in Fig. 16. The solid circles represent the experimental data. The dotted lines show the characteristics that would result if the received field were composed of an incident wave and one reflected from the local slope with a 180-degree phase shift. This slope was determined from topographical maps and local surveys. The lack of agreement between these two sets of characteristics is sufficient to indicate that this picture is far from complete. In an effort to improve the agreement between theory and experiment, a third component such as might be reflected from any other slope near the receiver was assumed. Preliminary calculations showed that in order to fit the experimental curves, the third component should have a magnitude of approximately 0.8. Assuming a magnitude of 0.8, the phase relation between the third component and the other two that would give the best fit between the experimental and theoretical curves was determined. It was found

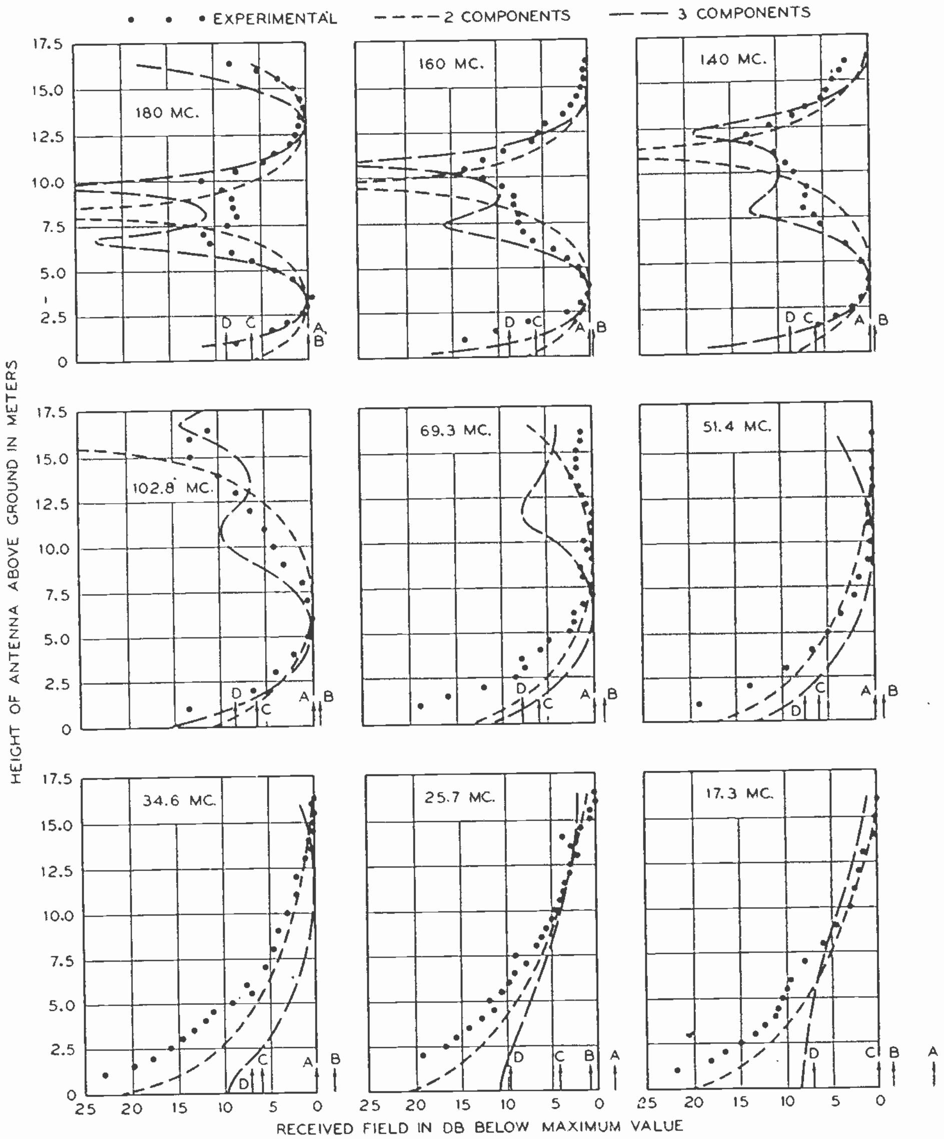

Fig. 16—Variation of received field strength with antenna height at Lebanon, N. J. Horizontal polarization.

The experimental points are shown together with theoretical curves based on:

1. Complete reflection from the major slope (labeled 2 components)
2. Complete reflection from the major slope and an additional reflection of eighty per cent from the secondary slope (labeled 3 components).

The arrows indicate the field strength values as follows:

A = maximum value of the received field assuming reflection from only the major slope.

B = maximum value of the received field assuming reflection from two slopes (all three components in phase).

C = value of the received field corrected to eliminate the effect of reflection from the major slope near the receiver on the assumption that the secondary slope does not affect the received field.

D = value of the received field corrected to eliminate the effect of local reflections assuming them to come from both the major and secondary slopes near the receiver.

that the required phase relation very closely approximated that which would be produced by a reflection from the ground immediately underneath the receiving antenna. (See Fig. 17.) This reflecting plane is of such small dimensions that at first it was thought that the reflection from it would be of minor importance. The improvement in the agreement between the theoretical and experimental curves obtained when the theoretical curves were based on additional reflection from the second plane over that obtained by assuming reflection from only the major slope is sufficient to indicate that the minor slope did play an important rôle in determining the received field for frequencies higher than about 100 megacycles. This improvement is most prominent for antenna heights of about ten meters on the three highest frequencies

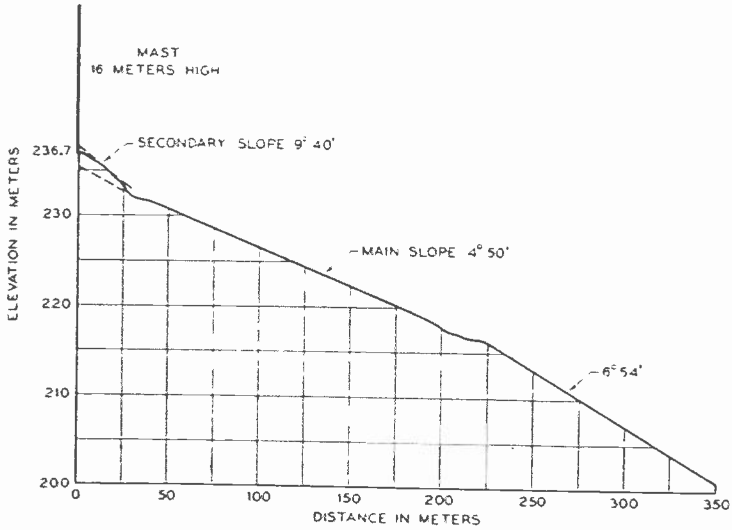

Fig. 17—Profile of receiving site, Lebanon, N. J.

because of the pronounced difference between the two sets of calculated curves in this region. For frequencies lower than about 100 megacycles the area of the secondary slope was not great enough, apparently, to contribute appreciably to the received field. For these frequencies, the secondary slope probably should be averaged in with the major slope. This would have very little effect on the theoretical curve except near the ground. At the lower antenna heights the effective distance that the antenna is above the reflecting plane should be more nearly equal to the height above the ground under the antenna. This would raise the theoretical curves approximately 1.6 meters resulting in a substantial improvement in the agreement between the calculated and measures curves. For the higher antenna heights, the points of geometric reflection move away from a point directly beneath the antenna so that the secondary slope contributes less to the resultant field. This reasoning is substantiated by the curves of Fig. 16 where if the change were made for all frequencies, the calculated curve for one reflection would fit the experimental curve better for the lower antenna heights.

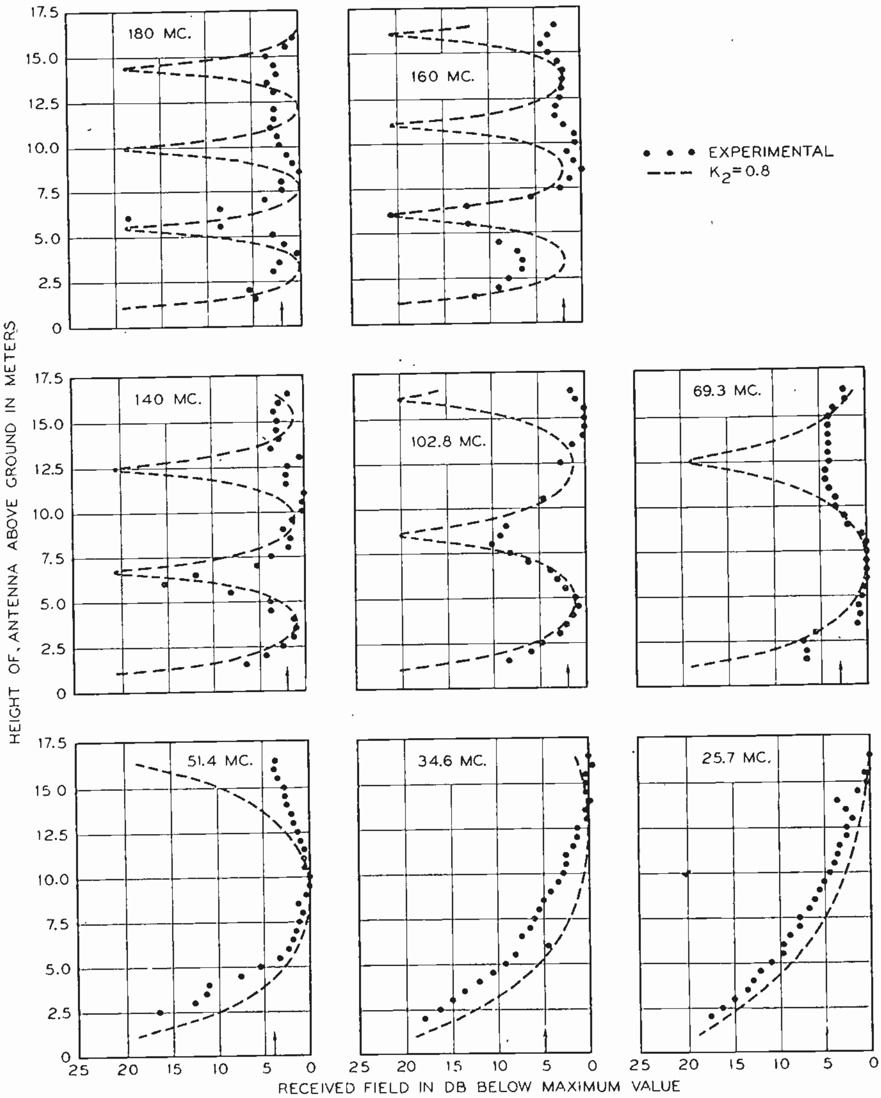

Fig. 18—Variation of received field with antenna height at Beer's Hill, N. J. The broken curve is based upon the addition of two components one of which is eighty per cent of the other, with a phase difference determined by assuming regular reflection from the average slope of the ground near the antenna in the direction of transmission.

In accordance with the considerations of this paragraph, the measured values plotted in Figs. 19 and 20 correspond to the values indicated in Fig. 16 by C for frequencies lower than 100 megacycles and by D for frequencies higher than 100 megacycles.

Fig. 18 shows the variations of the received field with antenna height at the Beer's Hill site for a number of frequencies both as determined experimentally and as calculated on the assumption that there is a wave reflected from the average slope near the antenna whose magnitude is eighty per cent of the incident wave. The assumption that there is a regularly reflected wave from the local slope is not as justifiable in this case as it was at Lebanon because of the rather large irregularities that existed where local reflections would occur. The experimental curves do not show any indication of the presence of two or more components whose phases vary regularly with height except at the lower antenna heights. In fact, when the antenna was anywhere above half of the available antenna height, the received field at the three highest frequencies appeared to be constant except for small random variations. This indicates that the resulting reflected component for these conditions should be of small magnitude. Hence in the absence of any better information, it has been assumed that the effect of the local conditions when the antenna is above half the available height may be neglected for the three highest frequencies. For the lower frequencies the effect of the local conditions was determined on the basis of a wave regularly reflected from the ground in the neighborhood of the antenna. This results in the incident fields having the magnitudes indicated by the arrows in Fig. 18.

Since the ground in the vicinity of the Deal terminal is level, there should be no local reflections in the same sense as there were at Lebanon and Beer's Hill. Consequently in arriving at the values for the received field over the Deal-Lebanon path, corrections for local effects were applied only at the Lebanon terminal.

The results of the transmission tests between Deal and Lebanon and between Beer's Hill and Lebanon are shown by the frequency characteristic curves of Figs. 19 and 20[21] together with those calculated upon the basis of the theory of ultra-short-wave propagation presented in an earlier paper.[1,22] The experimental characteristic for the Deal-Lebanon path (Fig. 20) has been dotted for frequencies above 160 megacycles to show the fact that the measurements on these frequencies indicated different values for the received field on different days. The experimental points below 130 megacycles showed no large variations from day to day and the measurements between 130 and

[21] This frequency characteristic was presented before the New York Meeting of the Institute of Radio Engineers, November 2, 1933, as part of a discussion of the paper referred to in reference 1.

[22] For the Beer's Hill-Lebanon path the reflection coefficient of -0.8 used in the earlier paper was retained. For the Deal-Lebanon path an improvement in agreement between experiment and theory resulted from using a reflection coefficient of -0.85.

160 megacycles were made during a rapid frequency run on the same day.

For both of these paths there is good agreement between theory and experiment for the lower frequencies. The irregularities in the experimental characteristics begin, however, at 100 megacycles while those for the theoretical characteristic do not begin until about 1000

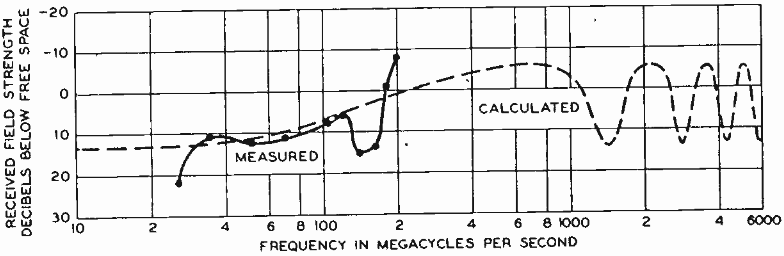

Fig. 19—Frequency characteristic of optical path Beer's Hill to Lebanon, N. J. The experimental points have been corrected to eliminate the effect of local reflections near the terminals as explained in the text.

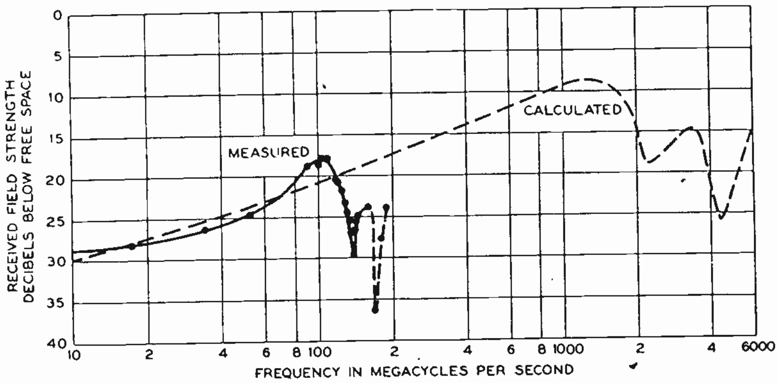

Fig. 20—Frequency characteristic of nonoptical path Deal to Lebanon, N. J. The experimental points have been corrected to eliminate the effect of local reflections near the terminals as explained in the text.

megacycles. This indicates that the approximations of the theory are not valid for these paths at frequencies higher than about 100 megacycles. It may be significant that where there is a discrepancy between the approximate theory and experimental results, variations in the received field have been observed indicating instability in the transmitting medium.

Acknowledgment

The authors wish to express their thanks for the encouragement, advice, and criticism given by Mr. J. C. Schelleng, under whose general supervision this work was done.

Appendix

For ultra-short-wave propagation over land $2\sigma/f$ may be neglected in comparison with $(\epsilon-1)$ so that for the practical case where h_1 and h_2 are small compared with d the magnitude of the reflection coefficient is given by the following expressions:[23]

$$1 - 2\epsilon(h_1+h_2)/d\sqrt{\epsilon-1} \tag{13}$$

for vertical polarization and

$$1 - 2(h_1+h_2)/d\sqrt{\epsilon-1} \tag{14}$$

for horizontal polarization.[24]

Where these more exact expressions for the magnitude of the reflection coefficient are used the resultant field instead of being given by (3) becomes

$$E_V = E_0 \frac{4\pi h_1 h_2}{\lambda d}\sqrt{1 + \frac{\epsilon^2(h_1+h_2)^2\lambda^2}{(\epsilon-1)4\pi^2 h_1^2 h_2^2}} \tag{15}$$

and,

$$E_H = E_0 \frac{4\pi h_1 h_2}{\lambda d}\sqrt{1 + \frac{(h_1+h_2)^2\lambda^2}{(\epsilon-1)4\pi^2 h_1^2 h_2^2}} \tag{16}$$

for vertical and horizontal polarization, respectively. When the lower of the two antennas is more than a couple of wavelengths off the ground (not fresh water), as would be true in most practical cases except for mobile transmission,[25] the radicals are substantially unity and these equations reduce to (3).

The variation of the received field strength with antenna height for both polarizations and a series of values of dielectric constant and ratios of antennas heights is given[26] in Fig. 21 which is a plot of (15) and (16). The ground constants for which any curve applies depend

[23] Both ϵ and σ in electrostatic units.

[24] These expressions follow from Appendix I of reference 1. In accordance with the conventions used in this reference, vertical polarization refers to a vertical electric field and horizontal polarization refers to a horizontal electric field. The convention used for the phase change at reflection is the change in phase of the vertical component in the case of vertical polarization, and the change in phase of the horizontal component in the case of horizontal polarization. These conventions differ in part from those of optics.

[25] For a discussion of this case see Charles R. Burrows, Loyd E. Hunt, and Alfred Decino, "Ultra-short waves in urban territory," *Elec. Eng.*, vol. 54, pp. 115–124; January, (1935), and *Bell Sys. Tech. Jour.*, vol. 14, pp. 253–272; April, (1935).

[26] The fact found experimentally by Feldman[2] that for vertical polarization the received field is substantially independent of the antenna height when the antenna is within a half wavelength of the ground may be used as a partial confirmation of the horizontal portion of these curves.

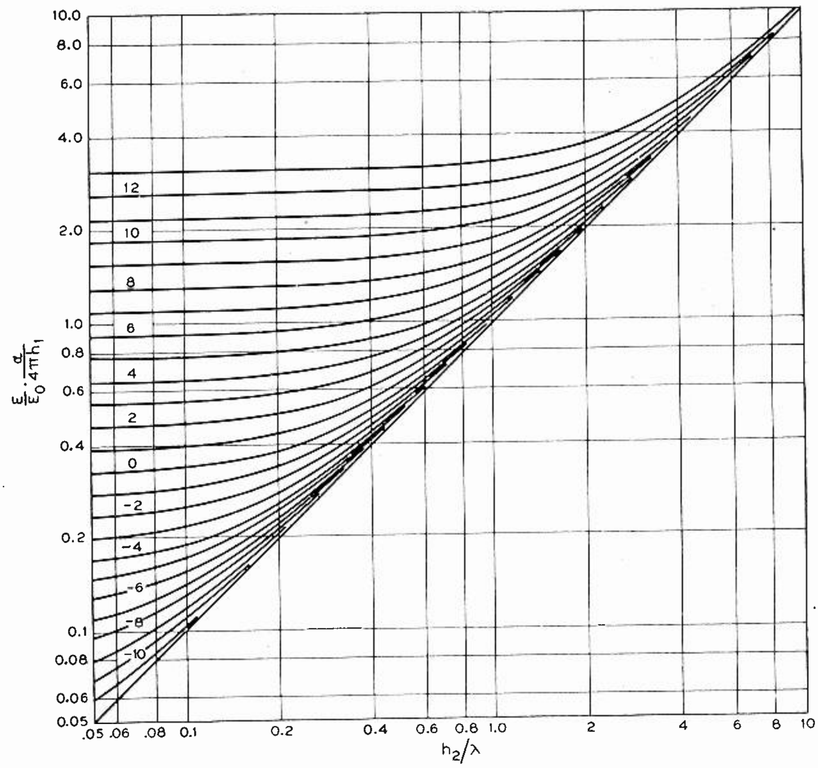

Fig. 21—Variation of received field with antenna height for ultra-short-wave propagation over level earth. The curves apply to the values of ϵ given by (17) and (18) and Table II.

TABLE II

VALUES OF DIELECTRIC CONSTANT FOR WHICH THE CURVES OF FIG. 21 APPLY

	Horizontal Polarization		Vertical Polarization	
n	$h_2 \ll h_1$	$h_2 = h_1$	$h_2 < h_1$	$h_2 = h_1$
13	1.00276	1.01105	359 or 1.0029	89.6 or 1.0111
12	1.00390	1.01562	255 1.0039	62.9 1.0161
11	1.00555	1.02222	180 1.0055	43.2 1.0230
10	1.00781	1.03125	127 1.0078	31.0 1.0325
9	1.0110	1.0443	89.6 1.0111	21.6 1.0482
8	1.0156	1.0625	62.9 1.0161	14.93 1.0679
7	1.0221	1.0885	43.2 1.0230	10.20 1.1041
6	1.031	1.125	31.0 1.0325	6.82 1.1562
5	1.044	1.177	21.6 1.0482	4.26 1.2395
4	1.062	1.250	14.9 1.0679	2.00 2.0000
3	1.088	1.354	10.2 1.1041	
2	1.125	1.500	6.82 1.1562	
1	1.177	1.707	4.26 1.2395	
0	1.250	2.000	2.00 2.0000	
−1	1.353	2.414		
−2	1.500	3.000		
−3	1.706	3.825		
−4	2.000	5.000		
−5	2.415	6.667		
−6	3.000	9.000		
−7	3.825	12.3		
−8	5.000	17.0		
−9	6.650	23.6		
−10	9.000	33.0		
−11	12.325	46.3		

upon the polarization and ratio of antenna heights in accordance with the following equations where n is the number on the curves.

$$\epsilon = 1 + (1 + h_2/h_1)^2/2^{2+n/2} \tag{17}$$

for horizontal polarization, and

$$\epsilon = 2^{1+n/2}\left\{1 \pm \sqrt{1 - (1 + h_2/h_1)^2/2^{n/2}}\right\}/(1 + h_2/h_1)^2 \tag{18}$$

for vertical polarization. These curves apply when

$$2\sigma/f \ll \epsilon - 1, \quad \frac{h_1^2 + h_2^2}{2d^2} \ll 1, \quad \text{and} \quad \frac{4\pi h_1 h_2}{\lambda d} \ll 1.$$

The appropriate value of dielectric constant for both polarizations and two ratios of antenna heights are given in Table II. Comparison with the straight line which is a plot of (3) shows the error introduced by using this relationship when one of the antennas is near the ground.

When $2\sigma/f$ is small but not negligible with respect to $\epsilon - 1$ the last term of (13) must be multiplied by

$$1 + \frac{(2\sigma/f)^2}{8\epsilon(\epsilon - 1)} \frac{\epsilon - 4}{\epsilon - 1} \tag{19}$$

and the last term of (14) must be multiplied by

$$1 - \frac{3}{8} \frac{(2\sigma/f)^2}{(\epsilon - 1)^2}. \tag{20}$$

Accordingly, the phase shift at reflection differs from a complete phase reversal in the sense that the resultant angle between the two components is decreased by

$$\frac{(2\sigma/f)}{\sqrt{\epsilon - 1}} \frac{(\epsilon - 2)}{(\epsilon - 1)} \frac{(h_1 + h_2)}{d} \text{ radians} \tag{21}$$

and,

$$-\frac{(2\sigma/f)}{(\epsilon - 1)^{3/2}} \frac{(h_1 + h_2)}{d} \tag{22}$$

for vertical and horizontal polarization respectively. This results in the last term under the radical of (15) being multiplied by

$$1 - \frac{(2\sigma/f)}{\epsilon\sqrt{\epsilon - 1}} \frac{(\epsilon - 2)}{\epsilon} \frac{2\pi}{\lambda} \frac{h_1 h_2}{(h_1 + h_2)} \tag{23}$$

and the last term in (16) being multiplied by

$$1+\frac{(2\sigma/f)}{\sqrt{\epsilon-1}}\frac{2\pi}{\lambda}\frac{h_1h_2}{(h_1+h_2)}. \tag{24}$$

When (23) and (24) are substantially unity the ground in question behaves as a perfect dielectric.

The effect of finite conductivity for horizontal polarization is to make the magnitude of the reflection coefficient more nearly unity. The change in phase at reflection is a complete reversal both for the case of zero and infinite conductivity and has its greatest departure from a complete phase reversal when $2\sigma/f=\sqrt{3}(\epsilon-1)$. For this case when the first correction term is applied to (3) the received field is given by

$$\frac{E}{E_0}=\frac{4\pi h_1h_2}{\lambda d}\sqrt{1+\frac{\lambda}{\sqrt{2(\epsilon-1)}}\frac{h_1+h_2}{2\pi h_1h_2}}. \tag{25}$$

Comparison of (16) and (25) shows that for horizontal polarization the received field is given very accurately by (3) for all values of the ground constants.

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A QUANTITATIVE STUDY OF THE DYNATRON*

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Summary—*This paper describes two methods of predicting the performance of a dynatron oscillator. The use of the Massachusetts Institute of Technology differential analyzer is the first method described. This machine has been used to determine the amplitude of oscillation and harmonic content of a type '24 tube dynatron oscillator for various conditions. A corresponding experimental study is presented in the Appendix for comparison.*

The second method described is the use of the output characteristics. These are curves of the fundamental component of plate current produced by a sinusoidal plate voltage. They may be computed from the tube charcteristics or measured experimentally. An analysis by this method of the type '65 tube oscillator is given. An experimental study is presented for comparison. One case of interest is the use of a crystal as the tuned circuit.

INTRODUCTION AND STATEMENT OF THE PROBLEM

THE vacuum tube has become a most important device in the fields of pure and applied physics. For this reason it is highly desirable to have accurate means for analyzing and predicting its performance. It is the object of this paper to solve one special case of this more general problem, namely, the performance of vacuum tube oscillators dependent for their operation upon secondary electron emission from one element.

Pure theoretical deduction leads one to conclude that any circuit involving a vacuum tube can be rigorously treated by utilizing the characteristic surfaces and the mathematical surface relationships. Vacuum tube surfaces are not of a simple nature, however, and their complexity often requires the use of infinite series. In resorting to such a practice one is not only involved in cumbersome methods but oftentimes the significance of the several steps is shadowed by monotonous computation Refuge is usually sought in some assumptions whereby a more simple attack can be exercised, and in consequence the results are often incomplete and in some cases startling but erroneous

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There is one oscillatory system however, though it may contain a multielement tube, where all but one of the element potentials are independent of current since no series impedances are involved with these elements. This is the dynatron oscillator.

A dynatron[1] is a vacuum tube whose operating voltages are so chosen that use is made of a region in its characteristic of plate current vs. plate voltage where the slope of the characteristic is negative, or has an effective negative resistance and acts as a seat of electromotive

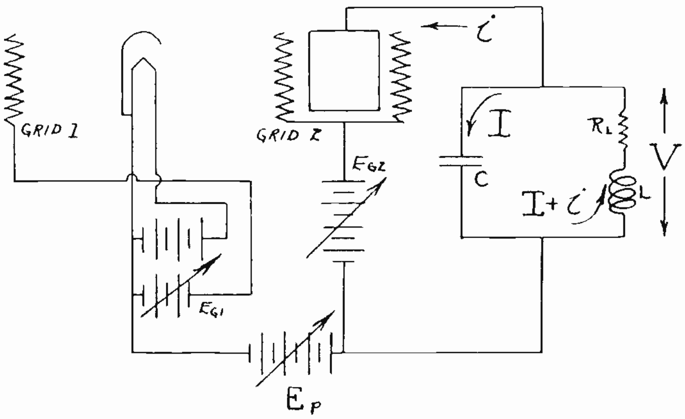

Fig. 1—The tetrode dynatron oscillator circuit.

force. It exhibits a two-dimensional characteristic; i.e., the plate current can be expressed as a function of the plate voltage alone. The typical circuit wherein a dynatron characteristic is employed with a four-element tube is represented by Fig. 1, and it is this circuit with which the present studies were made.

The Static and Dynamic Characteristics of the Tube

A typical set of characteristics of the '24 type tube obtained by the point-by-point method is shown in Fig. 2. These curves were obtained for grid 1 and cathode connected together and they were used in the '24 oscillator machine solutions discussed later. A more complete set of curves for the '65 as a dynatron, obtained by the use of a bifilar oscillograph[2] is shown in Figs. 3 and 4.

With any tube it will be found that the amplitude of the characteristic is controlled by the voltage of grid 1, since it controls the total cathode current which is practically independent of the plate voltage.

[1] A. W. Hull, "The dynatron, a vacuum tube possessing negative resistance," Proc. I. R. E., vol. 6, pp. 5–36; February, (1918).

[2] These curves were obtained from oscillograms of plate current and applied sinusoidal plate voltages of thirty and sixty cycles. Identical response was obtained for thirty and sixty cycles.

The voltage of grid 2 controls the general shape of the curve. The plate voltage, however, acts to change the distribution of the cathode current between grid 2 and the plate. These principles are advantageously used in choosing the operating point on the family of characteristics.

The tube characteristics can be determined in several ways. The simplest way is by the usual point-by-point method using a voltmeter and ammeter. This method is available to all. However, resulting values for the higher impressed voltages depend upon the manner and order of taking the data. It is necessary to allow the system to reach

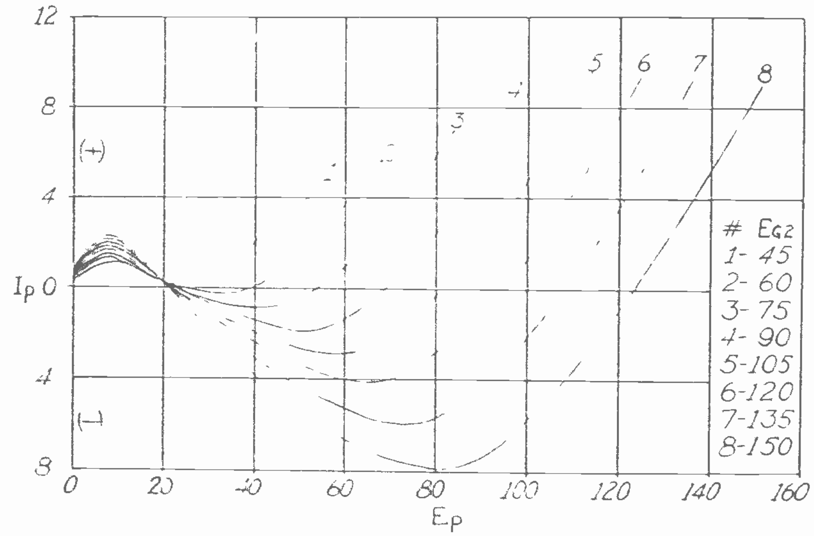

Fig. 2—Plate characteristics of the '24 tube as a dynatron showing the effect of the grid 2 voltage.

equilibrium at each point in order to obtain consistent results for the higher impressed voltages.[3] If the data are taken too rapidly they will depend upon the internal conditions of the tube controlling the secondary emission at the previous settings of voltage and current. These irregularities are most marked at the higher voltages where total cathode current is excessive and where the condition of a slow drift of the plate current exists after the application of plate potential. These discrepancies indicate that the point-by-point method does not simulate the internal tube conditions in the dynamic state for the higher voltages.

In order to obtain characteristics in which the plate and grid 2 are in the proper condition, it is necessary to obtain by a more elaborate

[3] Aging a tube at excessive voltage for 48 hours will often improve the consistency of the characteristics.

means the external characteristics while the tube is in a dynamic condition. This may be accomplished by the use of the cathode-ray or other oscillograph. The cathode-ray oscillograph will give the desired cur-

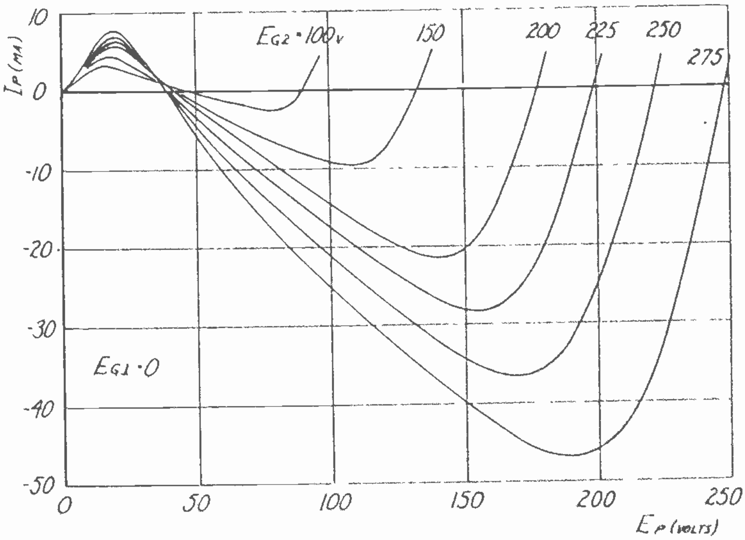

Fig. 3—Plate characteristics of the '65 tube as a dynatron showing the effect of the grid 2 voltage.

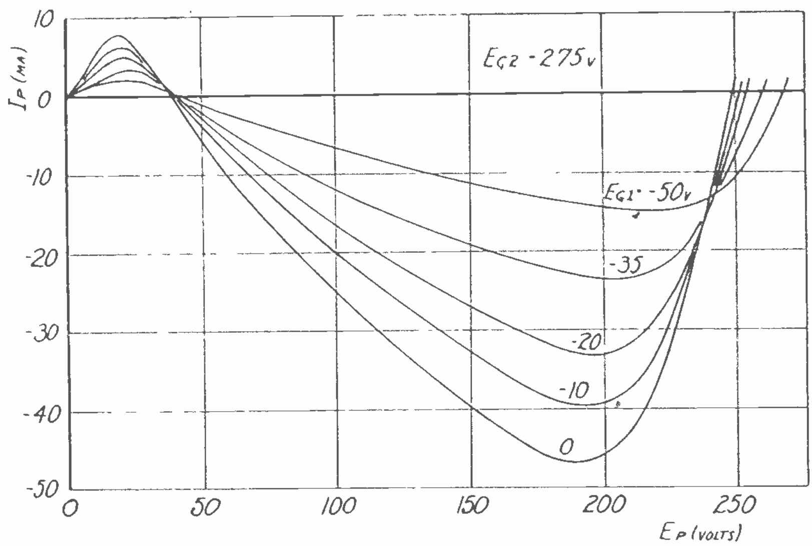

Fig. 4—Plate characteristics of the '65 tube as a dynatron showing the effect of the grid 1 voltage.

rent-voltage curves directly and in addition they may be obtained in two ways; one, while the tube and circuit are in a self-oscillating condition, the other, when the tube is driven by some external alternating-current source. The latter case is the more desirable because the range of conditions is unlimited. The cathode-ray tube, however, does not always give the desired accuracy. The usual bifilar oscillograph will be found very satisfactory with the disadvantage that the desired currents and voltages will be obtained as functions of time and must be replotted against each other to obtain the desired characteristics. In either case, however, the conditions internal to the tube are practically the same as those for the normal oscillatory state and the curves determined by the oscillographic methods more properly represent the dynamic characteristics of the tube.

A supplementary observation in regard to the dynamic characteristics should be mentioned. Since the characteristics depend upon surface conditions of the elements they will vary slowly as the tube ages with use. In some cases this change is quite slow, while in others it is very noticeable over a few hours use. The rapidity of the change depends on the load demanded of the tube.

These plate-current—plate-voltage characteristics are all that is necessary to determine the performance of the dynatron in any circuit.

Analysis by Means of the Differential Analyzer

A mechanical means has been developed at the Massachusetts Institute of Technology whereby the variable parameters of a system, representing a set of differential equations, can be properly accounted for and exact solutions obtained for various conditions. The solution by mechanical means has for its basis that the characteristic of the nonlinear device or devices is known and can be represented in graphical form. The Massachusetts Institute of Technology differential analyzer[4] offers an approach to vacuum tube problems since it is a machine which will solve differential equations containing nonlinear coefficients and it will accommodate high order equations of a considerable degree of complexity.

In brief statement, given the plots of the nonlinear parameters in the form $Y = f(x)$, the machine will provide solutions in the form of mechanically drawn curves or the data may be obtained in the form of a printed tape where, for example, ordinate and abscissa are stamped and preserved for reference at any time. The machine, its versatility, and its details have been adequately treated by Vanevar Bush and the details of the mechanical units which perform the processes of integration, algebraic operations, etc., will not be discussed here.

The exact solutions of the dynatron oscillator problem obtained with the aid of the differential analyzer not only determined the

[4] V. Bush, "The differential analyzer," *Jour. Frank. Inst.*, vol. 212, p. 447; October, (1931).

angular velocity and conditions for oscillation but the amplitude and wave form for various circuit and tube conditions. In addition they give the voltage regulation of the oscillatory system under various loads.

The equations of the oscillatory system shown in Fig. 1 and in suitable form for the differential analyzer are developed as follows. The voltage across the capacitance of the L, R_L, C portion of the circuit is given by

$$V = (R_C + 1/C_P)I.$$

The voltage across the inductive portion is similarly related with impedance and current by $-V = (R_L + L_P)(I + i)$. The necessary linking relation is given by the nonlinear tube characteristic shown in Fig. 2, in the form $i = f(V+E)$. Utilizing the last expression and assuming that the resistance of the condenser is negligible compared with that of the inductance, the equations represented in integral form are

$$V = (1/C)\int Idt$$

and,

$$I = -f(V+E) - 1/L\int [R_L f(V+E) + V + R_L I]dt.$$

These two equations were solved simultaneously and their adaptability to the analyzer allowed a rapid solution for I and V using various circuit parameters, vacuum tube electrode potentials, and numerous starting transient conditions. The functioning of the analyzer while in the transient or steady-state process of the solutions showed very nicely, among other things, the "flywheel action" of the tuned circuit since energy to the circuit from the tube is not continuous over the cycle for all voltages or values of L, R_L, and C. By stopping the machine at any time during the transient or steady-state portion of the solution, the results on the output table could be examined. It is quite interesting to be able to stop a transient in a circuit, examine it, and then allow it to continue to steady state.

The type of solution obtained is shown in Fig. 5 as a reduced plot of the transient and steady-state condition when the plate voltage is applied to the system of Fig. 1. This solution is of expected shape and required forty-five cycles to reach steady state. This represented the equivalent of one hundred and twenty-eight minutes of output table traverse. Although the transient state is interesting the steady-state solutions were the final goal. In consequence a method of assuming a charge on the condenser greatly reduced the running time for each solution and the average time for each solution was about sixteen minutes.

Rather than treat individual cases of the numerous solutions obtained by the foregoing method, representative solutions will be considered. All solutions were obtained for the '24 type tube at a fundamental frequency of forty-five cycles and a constant (X/R_L) ratio except one set obtained with (R_L) as a variable.

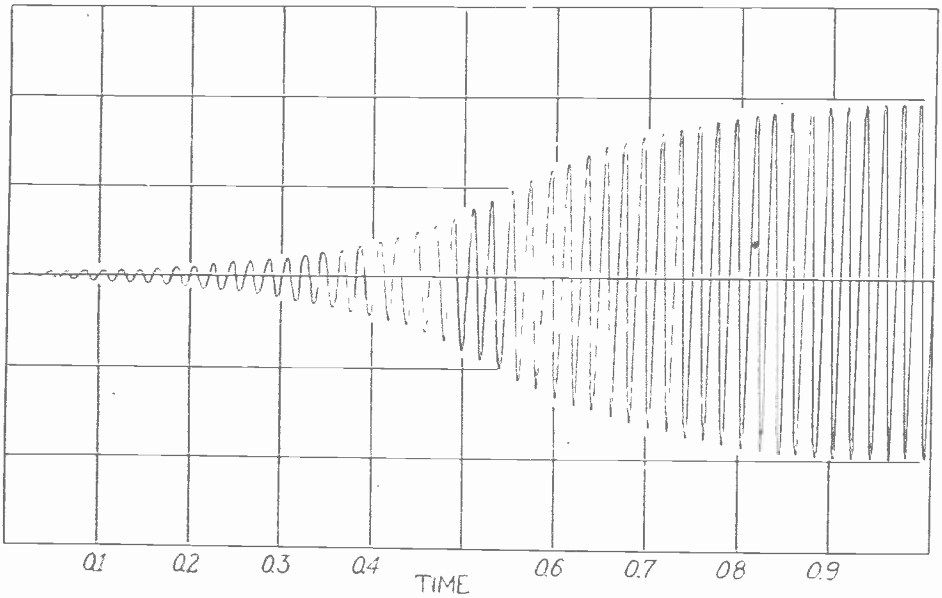

Fig. 5—Typical differential analyzer solution.

Solutions Obtained for Positive Plate Bias

The normal region of positive plate bias is here considered to be substantially bounded by the points of first positive and negative plate current maxima. Operation in this region gives rise to a self-starting system provided the effective negative resistance at the operating point is equal to or less than the absolute value of the oscillatory circuit impedance $(L/R_L C)$.

If the conditions for sustained oscillations obtain the largest amplitudes of oscillation will result as the first positive plate current maximum is approached as an operating point. By analyzer solutions it was found that for any operating point in the positive plate bias region a decrease in amplitude resulted if the plate voltage was increased. This is especially noticeable as the negative peak value of plate current is approached. This behavior is reasonable from the standpoint that the average energy per cycle to the circuit from the tube becomes smaller due to the pronounced double peaked plate current variation occurring in this region of oscillation.

When the positive and negative plate current maximum points are approached as operating points (the circuit previously oscillating at some operating point between these points) the slope of the characteristic is approaching zero and a distinction must be made for operating conditions at these extremes. The negative current maximum point has been discussed. At the positive plate current maximum point conditions become a speculation in circuit dissipation and energy storage in the oscillatory circuit before the operating point is shifted. For moderate values of circuit resistance the behavior at either extreme is about the same. However, there are some conditions where abnormal amplitudes can be maintained for operating points within the region of the first positive maximum of plate current as well as some giving zero value of plate current.

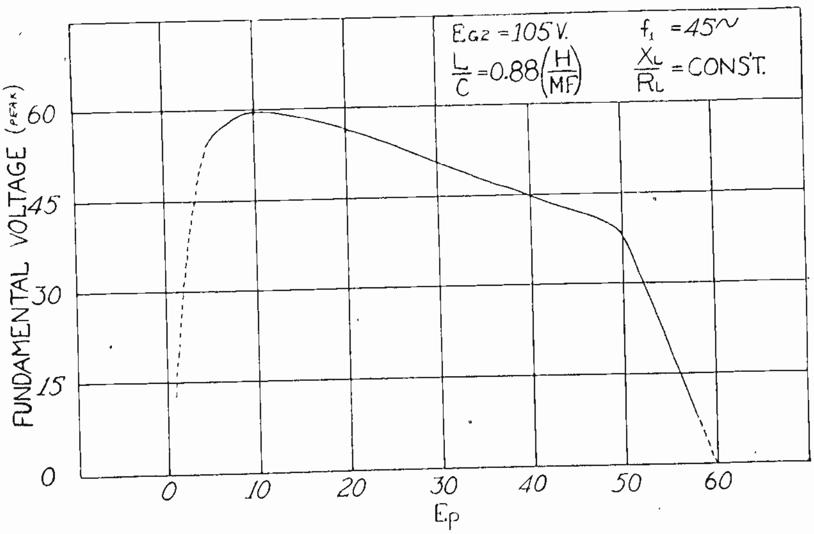

Fig. 6—Fundamental oscillation voltage for various direct plate voltages with the '24 (differential analyzer).

The effect of increasing the voltage of grid 2 is to change the negative resistance in the normal bias region and in addition to increase the abscissa region over which this remains substantially constant with voltage. This is indicative of a greater difference between the effective negative resistance at the operating point and the $(L/R_L C)$ of the tuned circuit. In consequence a larger amplitude will develop before circuit losses become equal to the available power supply from the tube.

Fig. 6 depicts the results of the solutions plotted against the position of the operating point in plate volts with the ordinate in fundamental volts. The points for this curve were obtained from schedule analysis of the oscillatory voltage wave of the analyzer solutions. Figs. 7, 8, and 9 show the per cent harmonic distribution for the fundamental voltages of Fig. 6 plotted against the same abscissa. It will be noted that the odd harmonics are a maximum when the dynamic excursion allowed the oscillatory voltage to sweep between the maximum points on the plate current curve. On the other hand the even harmonics are a minimum for this region and they approach maximum

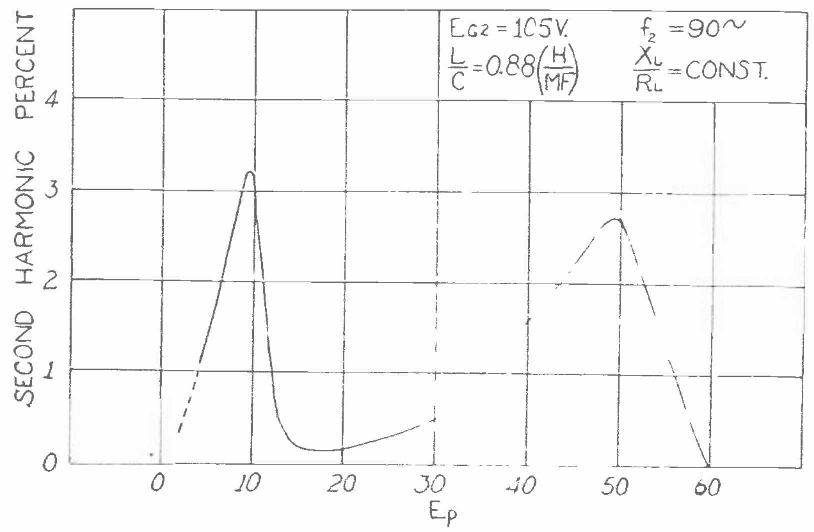

Fig. 7—Second harmonic voltage for various direct plate voltages with the '24 (differential analyzer).

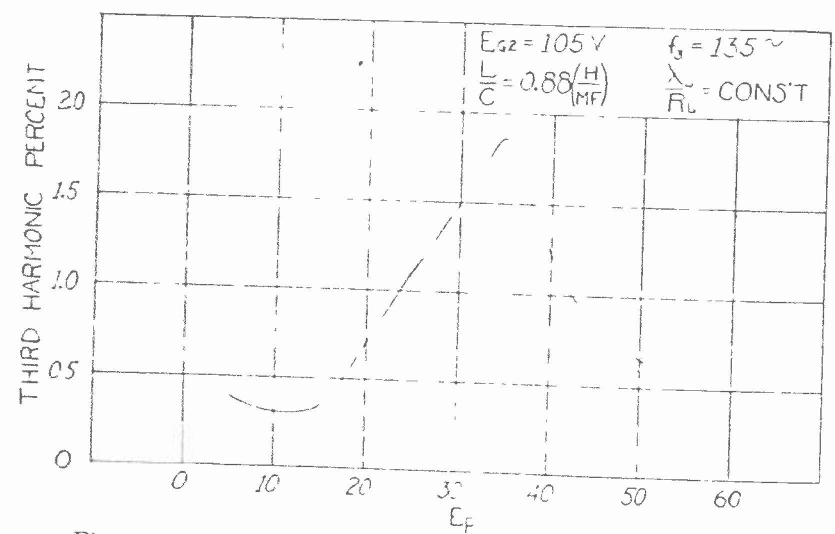

Fig. 8—Third harmonic voltage for various direct plate voltages with the '24 (differential analyzer).

percentages of the fundamental as the plate current maximum points are approached as operating points.

The shape of the harmonic distribution curves shown by Figs. 6, 7, 8, and 9 are shown[5] to represent in good accord slices taken through

[5] See Appendix A.

harmonic surfaces which were obtained experimentally with the same tube and circuit conditions as those used on the differential analyzer.

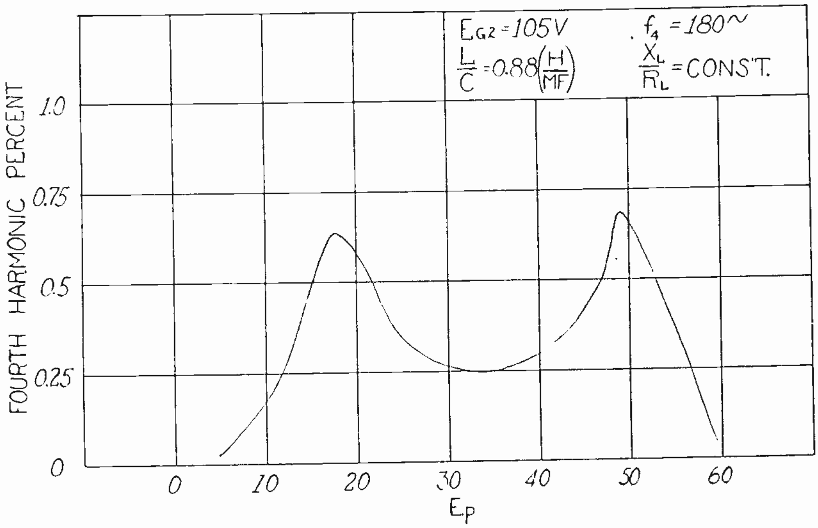

Fig. 9—Fourth harmonic voltage for various direct plate voltages with the '24 (differential analyzer).

Solutions Simulating a Loaded Oscillatory System

The next group of solutions to be discussed fall in the same classification as the first group. However, in this group the operating point was selected in the middle of the straight portion of the characteristic and R_L, the resistance of the tuned circuit, was taken as the variable. The variable parameter R_L was changed over a wide enough range to allow the absolute value of the impedance of the tuned circuit to vary from the maximum possible value down to one sufficiently low to throw the system out of oscillation. Such a procedure allows a check on the conditions for oscillation and a study of the effects of resistance in the tuned circuit on harmonic content. The results of these solutions are plotted in Fig. 10. As depicted, this plot represents steady-state amplitude as a function of the tuned circuit resistance for the fundamental, second, third, and fourth harmonic. The irregularity of these curves and the general tendency for the harmonic percentage to fall with an increase in the variable parameter is due to the fact that a variation in (R_L) not only changes the ratio between the value of the negative resistance at the operating point and the absolute value of the impedance of the tuned circuit, but it also limits the excursion over the dynamic characteristic, thus limiting the fundamental which controls the amount of energy fed from the tube to the circuit. This plot is interesting inasmuch as it indicates that as a practical matter the best circuit from the standpoint of (X_L/R_L) ratio is not always the circuit which results in a low harmonic content.

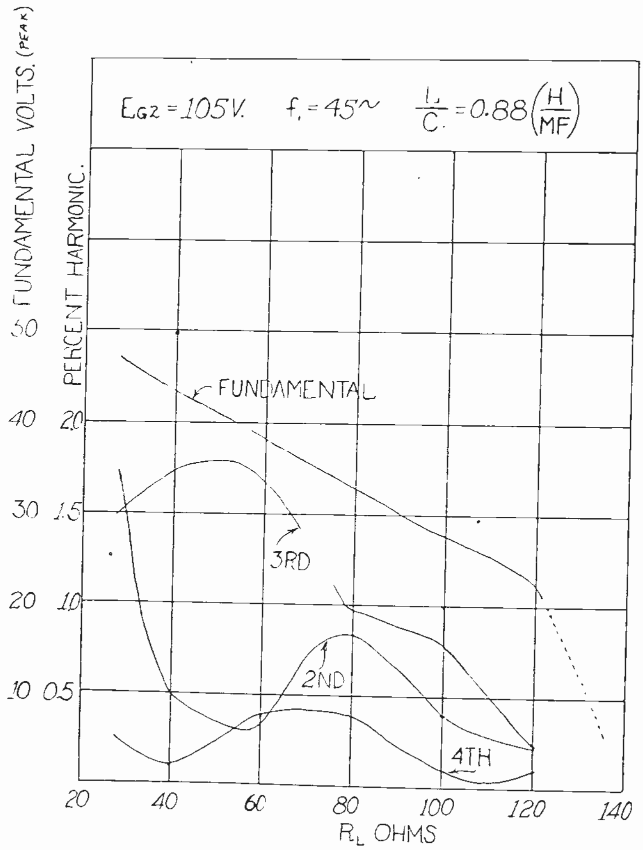

Fig. 10—Fundamental and harmonic voltages for various loads with the '24 (differential analyzer).

Solutions Obtained for Zero and Negative Plate Biases

In 1930, while experimenting with the dynatron tube and an oscillatory system, the authors noticed that if the system were oscillating and the $(L/R_L\ C)$ of the tuned circuit were high in value, a quick shift of the plate voltage to zero not only allowed the system to continue in oscillation but the steady-state amplitude became very much larger. This condition was found to exist for large values of grid 2 voltage but a substantial increase of this potential was not necessary if the tuned circuit impedance was high. An alternative was found where the plate potential was made initially zero and energy storage in the condenser was switched into the tuned circuit to produce the initial transient. In order to confirm these earlier experiments, zero plate voltage solutions were attempted on the differential analyzer with the result that they confirmed the original experiments. As an

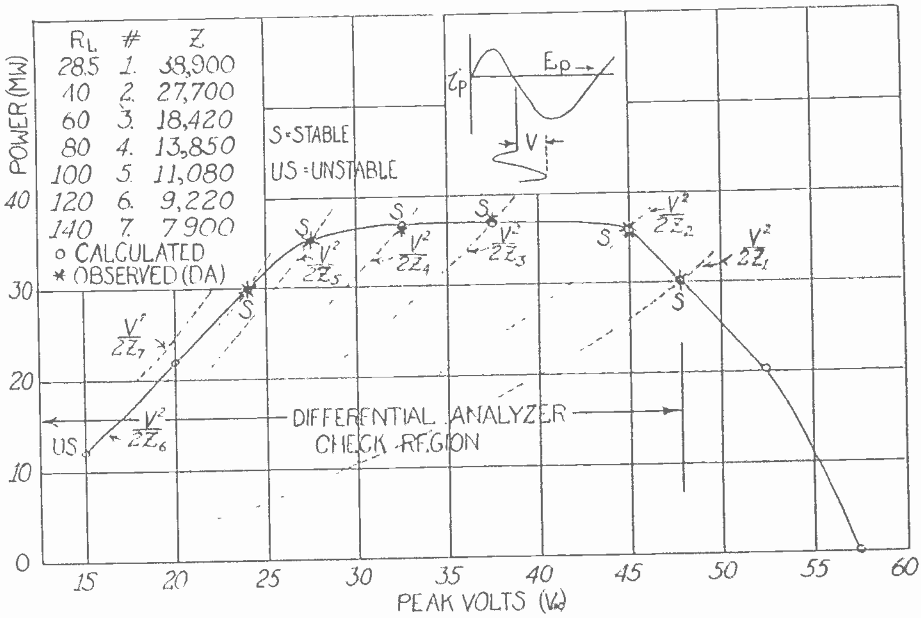

Fig. 11—Power output characteristic for the '24 with positive direct plate voltage (differential analyzer).

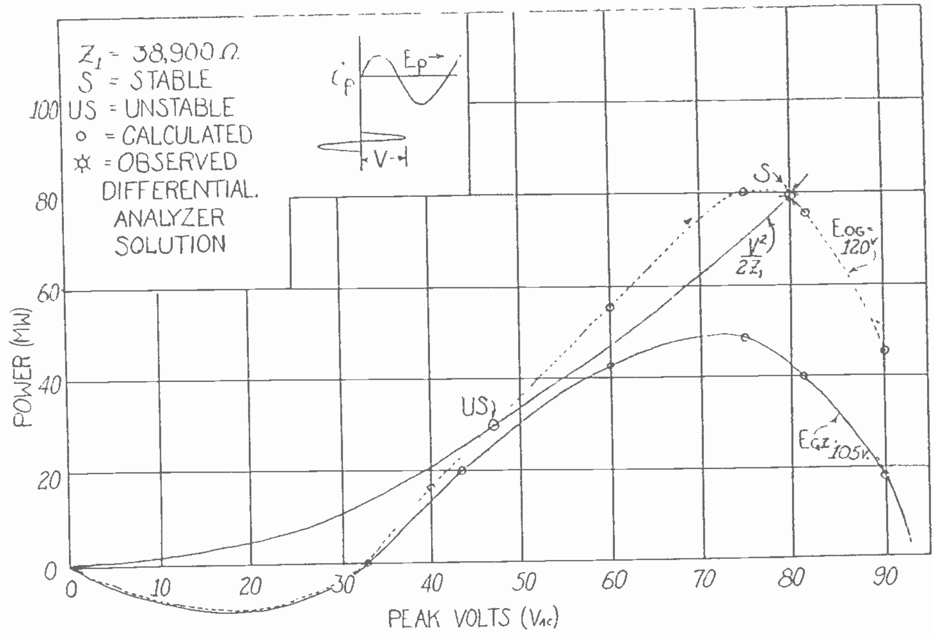

Fig. 12—Power output characteristic for the '24 with zero direct plate voltage (differential analyzer).

example one of these solutions is checked in amplitude by a graphical method of determining the amplitude and is shown in Fig. 12.

At the time the authors experimentally obtained the oscillatory steady state for zero plate bias, experiments were carried on for negative plate voltages. The experimental results indicated a still higher amplitude of oscillation and that the same circuit conditions necessary for steady-state oscillations resulted as those found for zero plate bias. Here again the differential analyzer was used to check the experimental results and several solutions were obtained for negative plate voltages under the same procedure as those obtained for zero plate bias. As an example one of these solutions is checked in amplitude as shown in Fig. 13 where this figure is part of a method of predicting the oscillatory amplitude.

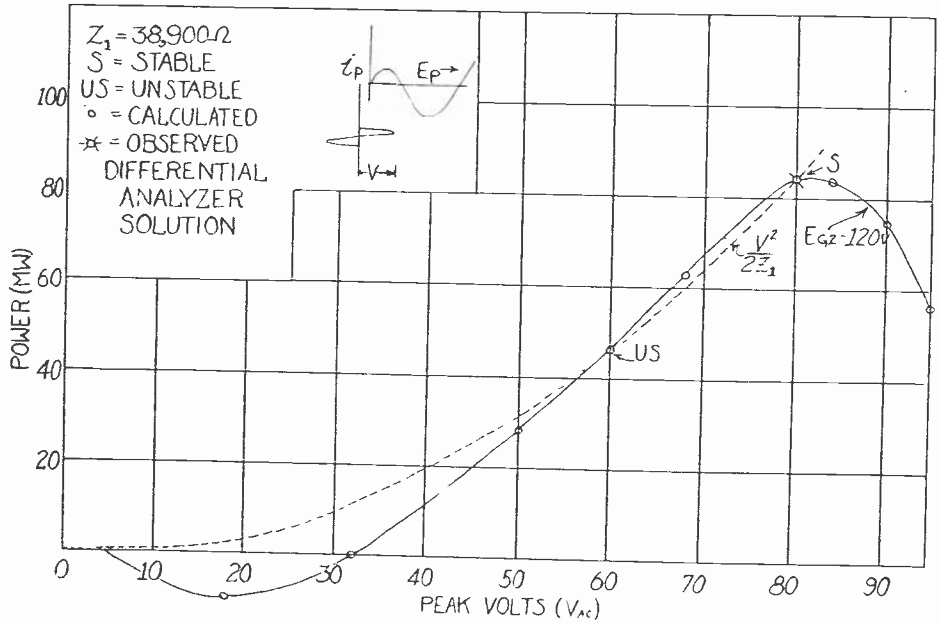

Fig. 13—Power output characteristic for the '24 with negative direct plate voltage (differential analyzer).

The physical explanation of the abnormal steady-state amplitudes obtained for zero and negative plate voltages is based upon the conservation of energy at equilibrium amplitude. As the plate bias point is brought to zero or moved into the negative region, the peak oscillatory voltages occur at correspondingly higher tube currents and thus the total energy per cycle from the tube to the circuit is increased. For a given circuit and fixed potentials there is an optimum point in the negative plate bias region beyond which the results are a diminishing return. This is explained by the fact that such operating points increase the "free oscillation" portion of the cycle, thus diminishing the time per cycle wherein the tube can supply energy to the circuit and in consequence resulting in a diminution in amplitude. The maximum amplitude for any set of fixed conditions will occur when the plate bias allows the peak of said maximum amplitude to occur in the vicinity of the point of maximum negative plate current.

Tube and Circuit Power Relations

The differential analyzer solutions, previously referred to, demonstrate that this machine can be used to obtain both the transient and steady-state solutions of a problem of this character. The method, however, has the disadvantage of requiring the differential analyzer which must be used in Cambridge and, therefore, is not easily available to all.

A general method, substantiated by experimental results and the differential analyzer solutions, predicting the performance of the dynatron oscillator can be developed from the law of conservation of energy. The net energy output of the tube must be that dissipated in the oscillatory circuit. In the general case each such equilibrium point in the steady state must be arrived at by means of successive approximation. A sine wave of voltage across the tuned circuit is assumed of such magnitude that the fundamental tube current produced is equal to the current produced in the circuit. Since these currents are in phase[6] with the voltage, this is the same as saying that the fundamental power output of the tube is equal to that received by the circuit. (A convenient method of determining this amplitude will be discussed later.)

Since the tube current produced by this voltage is not a pure sine wave, the harmonic currents resulting must pass through the tuned circuit and produce harmonic voltage drops. The original voltage can be corrected to include these harmonic voltages. The corresponding current can then be found and the process repeated until the desired accuracy results. By carrying out a number of these calculations for typical cases it will be found that, with oscillatory circuits with a Q of approximately 10 or better, the amplitude effects of the harmonics are negligible. If this condition is maintained, the harmonic power is negligible and the equilibrium point for the fundamental current and voltage is thus practically independent of the harmonic voltages.

This is substantiated completely by the results of the differential

[6] No loops in the characteristics were observed with bifilar and cathode-ray oscillograph studies of the dynamic characteristics of the dynatron with sinusoidal applied plate voltages. This indicates the absence of out-of-phase components of fundamental plate current and this has a direct bearing on the exceptional frequency stability of the dynatron oscillator.

analyzer solution and the experimental studies that follow. Let us refer later to the experimental evidence and consider now Figs. 11,12, and 13 where the power available from the tube to the circuit was calculated and plotted, assuming a sinusoidal variation of plate voltage. It will be noted that the calculated amplitudes and those obtained by the analyzer are in very close agreement. The available power curves depicted in these figures were computed by the familiar method of counting squares.

Fig. 11 is a plot of power versus the voltage across the oscillatory circuit. The full line curve represents the calculated power available from the tube to the circuit for corresponding oscillatory voltages. The dotted curves represent, as designated, the power loss curves for particular oscillatory circuits of designated impedances plotted against the voltage across the oscillatory circuit. These parabolic loss curves intersect the available power curve at several points. It will be noted on this figure that the amplitude of oscillation predicted by the intersection points were substantially those amplitudes obtained by the differential analyzer solutions for the same circuit conditions.

Fig. 12 depicts the same type of power-voltage curve as Fig. 11 but the plate bias potential for these curves was zero. The calculated available power is shown for two grid 2 potentials, 105 volts and 120 volts. Superimposed on these two calculated available power curves are certain designated parabolic circuit loss curves. The intersections predicted by calculation are exactly checked by the differential analyzer solution obtained for the same circuit conditions.

Fig. 13 is similar to Figs. 11 and 12; however, in this figure the plot is made for a plate bias potential of minus five volts. Here again the intersection between the calculated curves represent exactly the stable amplitudes of oscillation determined by the differential analyzer solution obtained for the same circuit conditions.

With the preceding substantiation along with further experimental evidence, the performance of the dynatron oscillator can be predicted by a knowledge of the dynamic characteristic, the oscillatory circuit, and fundamental voltages. This greatly simplifies the problem.

The Output Characteristics

The previously mentioned graphical method of presenting the power relations between the tube and circuit are obtainable by one possessing equipment capable of measuring the dynamic characteristic of the tube alone. The graphical method involves calculation and to those more fully equipped to observe the circuit and tube conditions the following method, based upon the same law of conservation of energy and a measured output characteristic, will be found to give a very excellent picture of the performance of the oscillator.

The output characteristics are curves of the fundamental component of plate current plotted against the sinusoidal plate voltage producing them. If they are multiplied by the sinusoidal plate voltage the result would be the available power output produced by the tube for each particular voltage. In other words, the ordinate of the output characteristics is power output divided by voltage. These current curves may be more convenient to use than the power curves. However, the results will be the same, since when the fundamental power output of the tube equals that received by the circuit the fundamental components of the two currents are also equal.

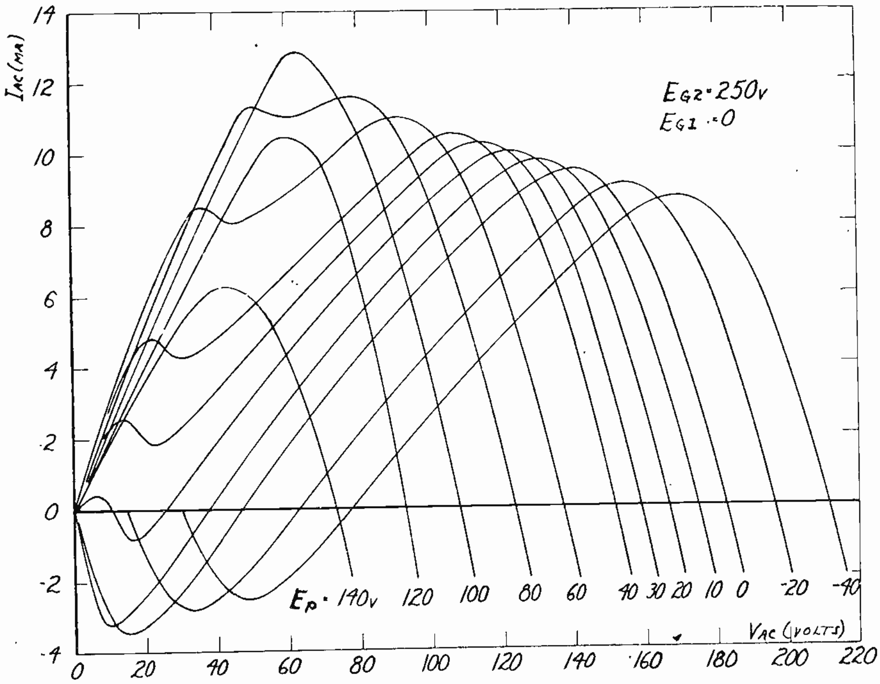

Fig. 14—Measured output characteristics for the '65 for various direct plate voltages.

The output characteristics may be determined in several ways. The most obvious is to assume a sinusoidal plate voltage of a given magnitude, determine the resulting plate current variation from the tube static characteristic and find the fundamental component of current by an harmonic analysis. This method is, of course, subject to the errors sometimes inherent in the static characteristic. However, an improvement can be obtained by using the tube characteristic obtained by means of the oscillograph.

The best means is to determine the output characteristics by experimental measurement. The internal tube conditions will then be the same as if the tube were actually oscillating with the same amplitude of plate voltage as that applied for measurement. This is an improvement over using the characteristics determined by an oscillograph since there are some cases where the characteristic is a function of the amplitude of oscillation. These variations are, however, not very large. The experimental measurement is simply an harmonic analysis of the tube current produced by an applied sinusoidal voltage.[7]

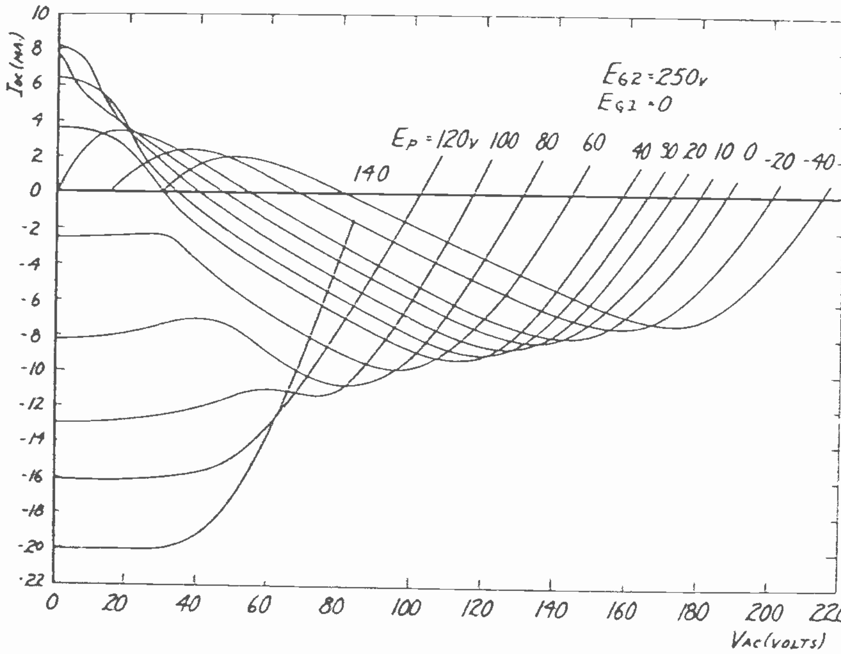

Fig. 15—Measured direct-current output characteristics for the '65 for various direct plate voltages.

A set of output characteristics obtained by experimental measurement of the '65 tube are shown in Figs. 14 and 16. The direct-current output curves are also shown in Fig. 15. These latter curves can be obtained by any of the methods described at the same time as the alternating-current output curves. They are typical curves and show the effects of the direct plate voltage and grid 1 voltage. Similar curves, somewhat changed in shape, may be obtained for other grid 2 voltages. The general shape of these curves can be derived from the static characteristics and need not be discussed here. They have been found to be very stable and measurements can easily be repeated. They will, however, change slowly as the tube ages with use.

[7] See Appendix B for details.

The Various Regions of Stable Oscillations

By means of the output characteristics one can predict the amplitude of oscillation. The output characteristics give the fundamental tube current for each alternating voltage. The current characteristic for the oscillatory circuit is a straight line the slope of which is the admittance of the circuit at its resonant frequency. If the latter characteristic is superimposed upon the output characteristics the intersections will represent points of equilibrium. This is the same as saying that the average power output of the tube is equal to the power input to the circuit at such points of equilibrium.

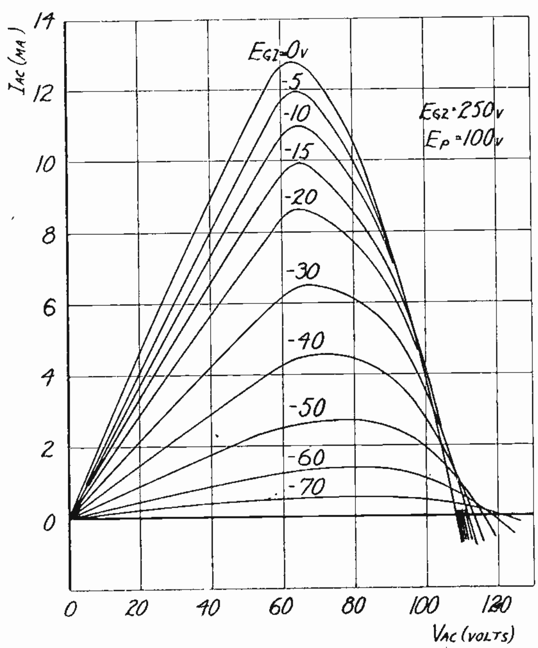

Fig. 16—Measured output characteristics for the '65 for various grid 1 voltages.

Let us consider two typical output curves for two different direct plate voltages and a circuit line as shown by Fig. 17. With curve (A) there will be an operating point at (0), the origin, and at (1). At alternating plate voltages between those corresponding to (0) and (1) the power delivered by the tube to the circuit is greater than that dissipated by the circuit in normal steady state at that voltage. This surplus power is used to increase the average stored energy in the oscillatory circuit and thus to increase the amplitude of oscillation and dissipation. For voltages above that corresponding to (1) the reverse is true. Therefore, the equilibrium point (1) is stable since any change in amplitude of oscillation produces a change in the surplus power delivered by the tube which tends to restore the equilibrium. If the reverse is true the operating point is unstable, as for example point (0).

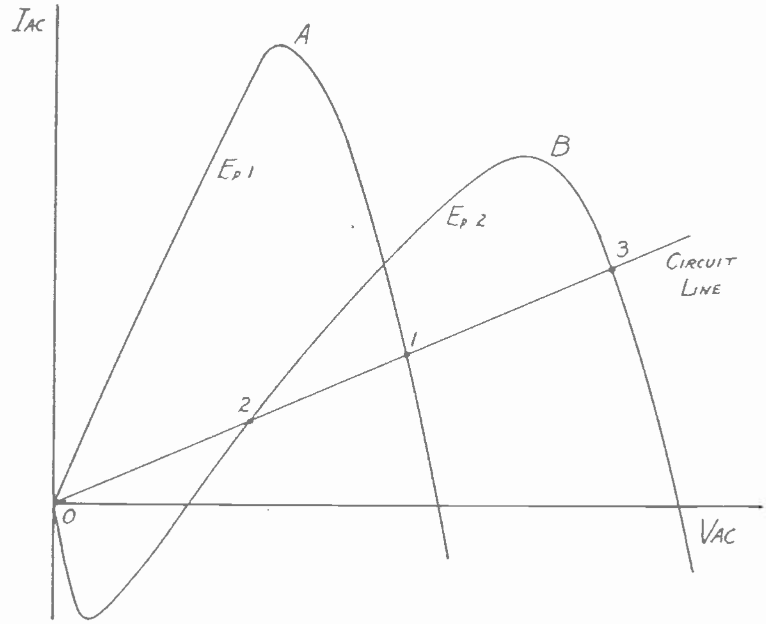

Fig. 17

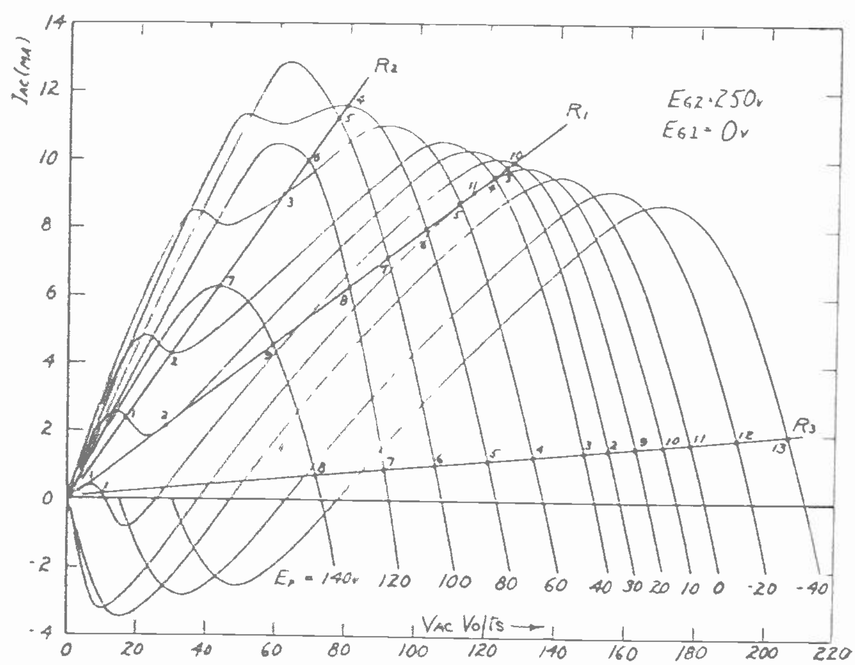

Fig. 18

Considering curve (B) Fig. 17, one notes that the origin, unlike that for curve (A), is a stable operating point. From this it is obvious that the amplitude of oscillation cannot cross this unstable point (2) on curve (B) without the aid of some external energy source. Oscillations in this case are not self-starting as in the previous case. If, however, the amplitude of oscillation is increased beyond point (2) by some outside means the oscillations will continue and become stable at the point (3). Similar stable and unstable points of operation are noted in Figs. 11, 12, 13, and 18 at the intersection points.

This method of predicting the amplitude of the dynatron oscillations is illustrated more fully in Fig. 18 wherein is shown the variation of amplitude with E_p for three different circuit impedances. The various regions of oscillation should be noted and particularly the high

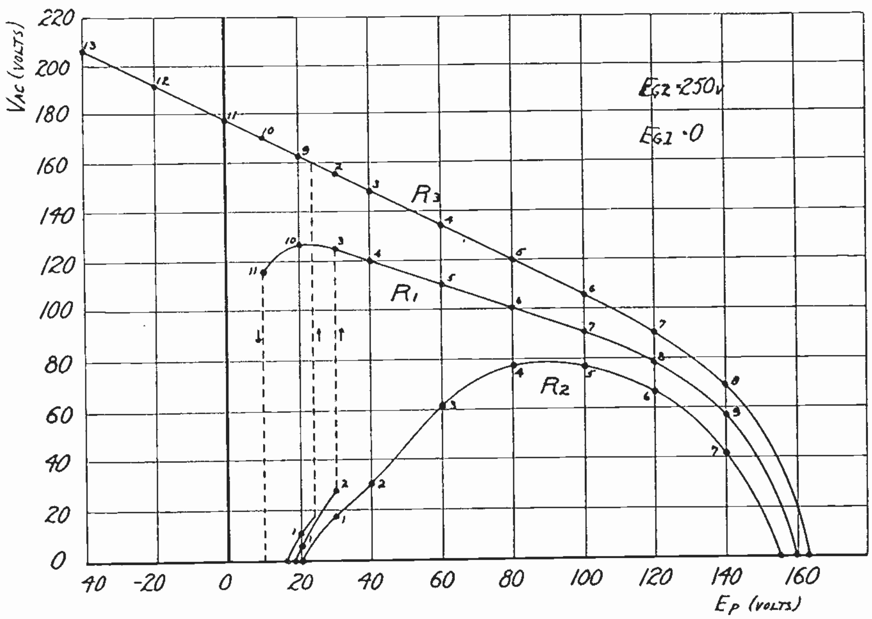

Fig. 19—Variation of oscillation voltage of the '65 for various direct plate voltages. (From Fig. 18.)

amplitude possible for negative direct plate voltages. This suggests modulation possibilities by varying the direct plate voltage. However, a nonlinear impedance is presented by this circuit to the modulator and considerable power is also required. Fig. 19 shows the various amplitudes of oscillation plotted V_{AC} versus E_p for the three load circuits represented by circuit lines 1, 2, and 3 in Fig. 18.

The variation of amplitude with oscillatory circuit impedance is shown by Fig. 20 wherein the load circuit lines (1) through (11) make intersections with the (P/V) curve. Similar amplitude tuned circuit impedance relations are shown by Figs. 11, 12, and 13 wherein the differential analyzer solutions are related with graphically determined circuit loss and available power curves. In all of these figures it will be noted that not only the amplitude of oscillation but the regulation with

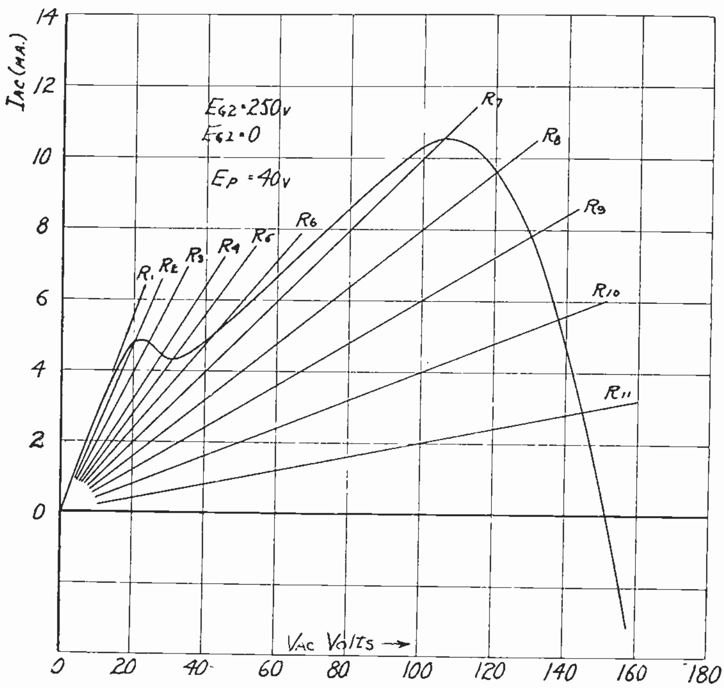

Fig. 20

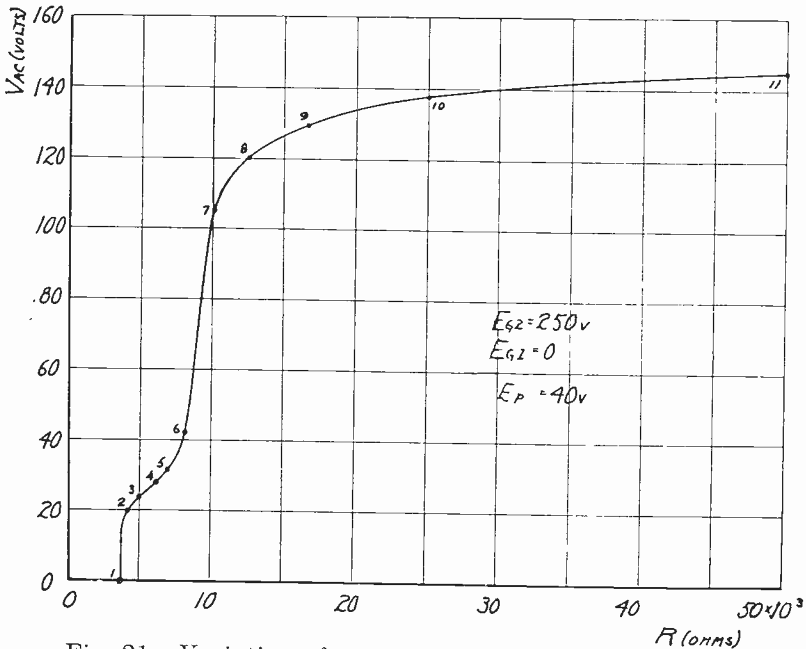

Fig. 21—Variation of oscillation voltage for the '65 for various plate loads. (From Fig. 20.)

load is available. Fig. 21 gives the V_{AC} versus resistance and this plot taken from Fig. 20 shows, very nicely, the effect on the oscillatory circuit of varying the load.

The variation of amplitude of oscillation with inner-grid voltage is shown in Figs. 22 and 23. These curves give the justification for the usual method of using the dynatron to measure the losses in a high-

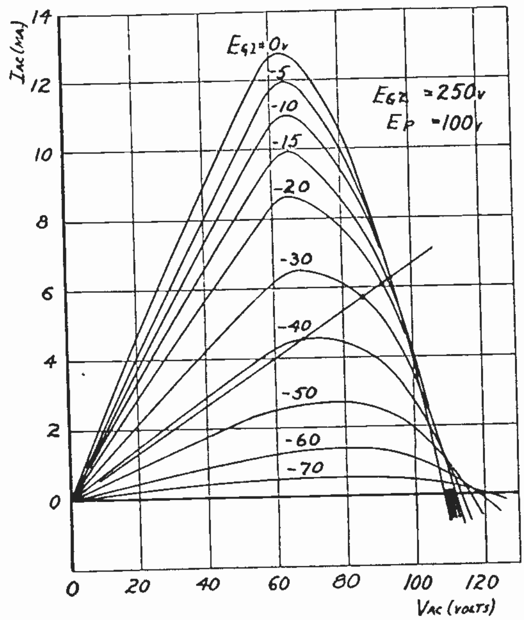

Fig. 22

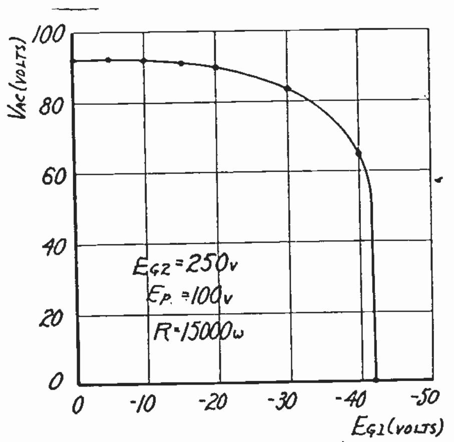

Fig. 23—Variation of oscillation voltage for the '65 for various grid 1 voltages. (From Fig. 22.)

frequency tuned circuit. Due to the very definite and steep slope of the curve the control or grid 1 voltage can be accurately determined. The final resistance obtained in the conventional manner is that determined from the slope of the static characteristic if the amplitude of oscillation be reduced very slowly.

The Output Characteristics for Zero Average Plate Current

From the general set of output characteristics, Figs. 14 and 15, it can be seen that it is possible to have oscillations with a zero average plate current. This suggests the possibility of oscillations with, for example, a crystal placed directly in the plate circuit of the dynatron. The points of operation with such a circuit can be derived from the output characteristics by finding the locus of the points at which the direct plate current is zero and this can also be done easily by measurement. The measured output characteristics for zero average plate current are given in Figs. 24 and 25, and these are presented for the '65 type of tube. The E_p curves give the direct plate voltage necessary to produce zero average plate current for each condition. They represent the voltage which will appear across the crystal or blocking condenser at this particular amplitude of oscillation.

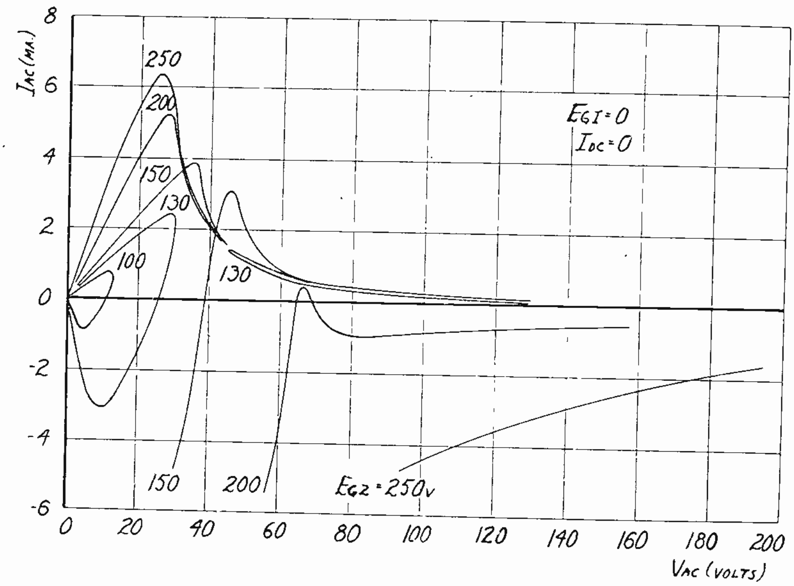

Fig. 24—Measured output characteristics with zero direct plate current for the '65.

There are two sets of curves, one derived from the right-hand zero of the I_{DC} curves and one derived from the middle, Figs. 24 and 25. As will be seen, each set of curves will give stable points of operation as far as the oscillations are concerned. However, another question of stability arises with these curves; namely, the direct voltage necessary to maintain the zero I_{DC} must be stable. Any slight change in this voltage will cause a direct current to flow; this current must be in such a direction as to tend to counteract the original voltage change. This is accomplished by the current placing the proper charge on the blocking condenser or crystal. It will be found that for the set of curves derived from middle zero point the direct plate voltage is unstable. Even if a stable alternating-current operating point is found the direct plate voltage will not be stable. The set of curves derived from the right-hand zero point will give, on the other hand, stable direct plate voltages. Therefore, the only usable operating points for such a load circuit are depicted on this set of curves.

It should be noticed that the region of operation next to the origin is unstable. This means that the oscillations for zero I_{DC} are not self-

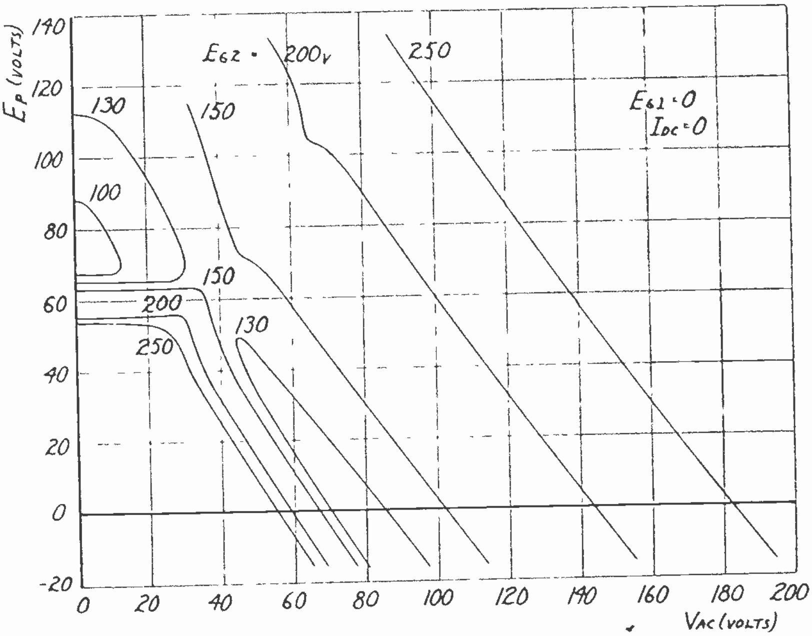

Fig. 25—Direct plate voltage variations on the '65 necessary to produce the zero direct plate current output characteristics.

starting. The oscillator must be excited by some other means until the amplitude of the alternating voltage is great enough to carry it into the the region of stable operation. This excitation may be obtained by self-starting oscillations which will result if the proper direct plate voltage is supplied through a choke shunting the blocking condenser or crystal. It should also be noted that the screen voltage necessary to give suitable oscillations is critical. The phenomena explained above have also been observed experimentally in an independent study made by Mackinnon,[8] and it will be found that his results are explained by the above discussion.

[8] K. A. MacKinnon, "Crystal control applied to the dynatron oscillator," Proc. I.R.E., vol. 20, pp. 1689–1714; November, (1932).

Comparison with Observed Oscillations

In order to show the accuracy of the measured output characteristics of the dynatron they were compared directly with the actual oscillation conditions. In each case the actual output characteristic was obtained by computing the fundamental component of the plate current from the observed value of the self-oscillatory current in a

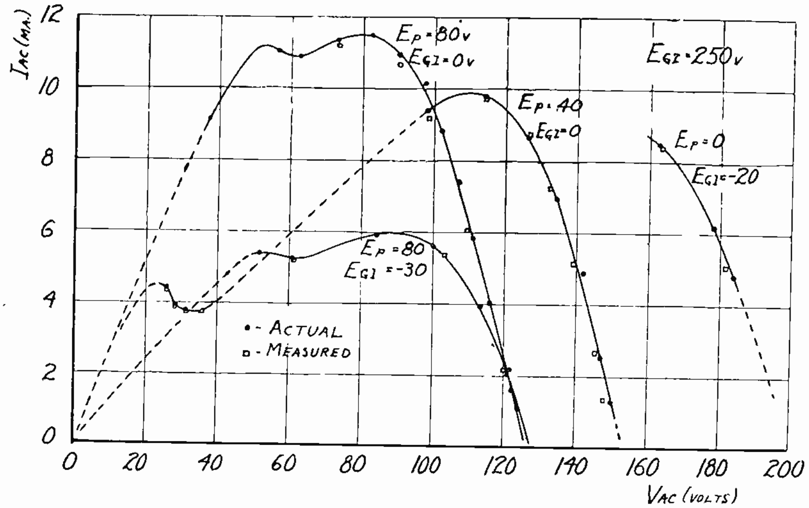

Fig. 26—Actual oscillation voltages with the '65.

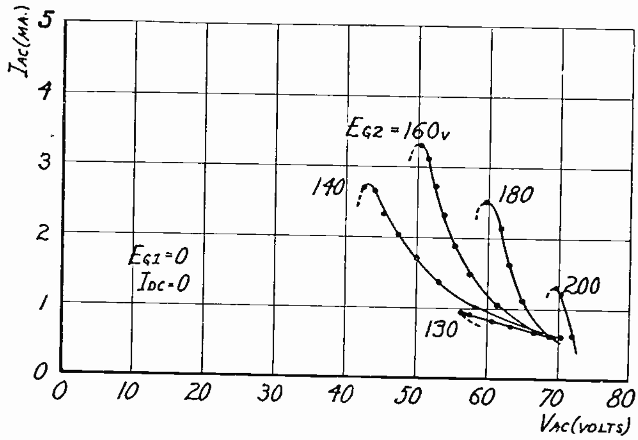

Fig. 27—Actual oscillation voltages with the '65 for zero direct plate current.

known oscillatory circuit.[9] The cscillatory voltage was also computed by this means. Immediately after observing the amplitudes of oscillation for various conditions the output characteristics were measured by the means already described.

The comparison of the actual and measured characteristics are given in Fig. 26. The check is exceedingly good. A similar set of actual

[9] See Appendix C.

curves for the case of zero I_{DC} is given in Fig. 27. The curves represented in these last two figures demonstrate an experimental check on the method. A similar check has been mentioned previously in regard to the work on the '24 with the differential analyzer and this definitely shows the validity of neglecting the harmonic power.

General Discussion and Conclusions

If one looks upon the dynatron oscillator as a circuit problem alone the methods discussed in this paper will take care of any of the normal conditions arising in practice. That is, if one knows the characteristic of the tube the problem is handled without difficulty. Unfortunately the tube characteristic is not too stable for high values of grid 2 potential. It will be found to change slowly with use. If one is interested in short intervals of use one can predict and use quantitatively the methods discussed well within the usual engineering accuracy. In any event the output characteristic of a particular tube will allow one to predict quantitatively the operation of the tube.

Many observers have reported on the variations of the dynatron characteristics between tubes of the same type and on the life of the tubes acting as dynatron oscillators. The latter is due to the past history of the tube either in manufacture or use. The reasons for such variations and the physics of secondary emission have been and are at present the subject for study by various investigators.

A life study of a dynatron oscillator indicates that such a device has a reasonable life. Like any other electronic device the voltages and element currents must be kept within reasonable limits. The desired type of characteristic is controlled mainly by the grid 2 voltage while the amplitude of the currents involved is controlled by the grid 1 voltage. These two factors are practically independent and may be varied to give reasonable operation. Experience has shown that a tube operated as a dynatron can be expected to have a life in excess of five hundred hours.

The differential analyzer solutions and the methods used in obtaining these solutions represent one method in determining the complete performance of an oscillatory system involving nonlinear parameters. This general method is being applied to other oscillatory systems for further investigation of such problems as separately excited class C amplifiers or self-excited oscillators.

Appendix A

Details of the Experimental Harmonic Surfaces of the '24

The harmonic surfaces for the '24 type tube were obtained by the use of a regenerative vacuum tube voltmeter. The oscillatory system was set in operation and the regenerative voltmeter was tuned to the desired harmonic of the first and the deflection noted. The voltmeter was then thrown to a calibrating circuit fed from a sine wave vacuum tube oscillator. In this manner the data for the harmonic surfaces were obtained and plotted in per cent of the fundamental or first harmonic.

Those experienced in the use of a regenerative voltmeter and those familiar with its inherent instability will appreciate the necessary precaution of terminating the input terminals of the voltmeter to the same magnitude and fórm of impedance for both the measuring and calibrat-

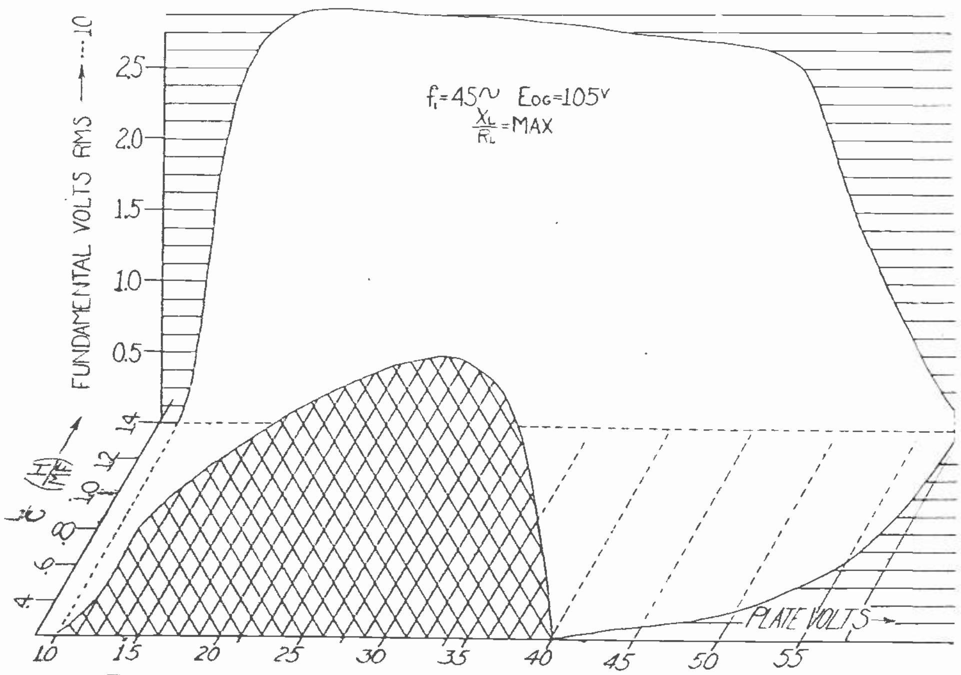

Fig. 28—Actual fundamental oscillation voltages for the '24.

ing circuits. This precaution was taken for the harmonic surfaces presented here and it would have been quite impossible to check a slice through these surfaces by differential analyzer solutions if this precaution were not taken.

The oscillatory system was first set into oscillation at a plate bias potential at or near the point where one secondary electron is obtained on the average for every primary electron impinging. From this point the plate bias was increased and then decreased until the dotted line boundaries were obtained.

Figs. 28, 29, 30, and 31 show the harmonic surfaces of the '24 type oscillator plotted against plate bias, or operating point and the (L/C) ratio of the oscillatory circuit. These surfaces were obtained experimentally for a fixed screen and inner-grid voltage but sufficient experimental evidence was obtained to show that the general shapes of these surfaces were approximately the same for different inner- and outer-

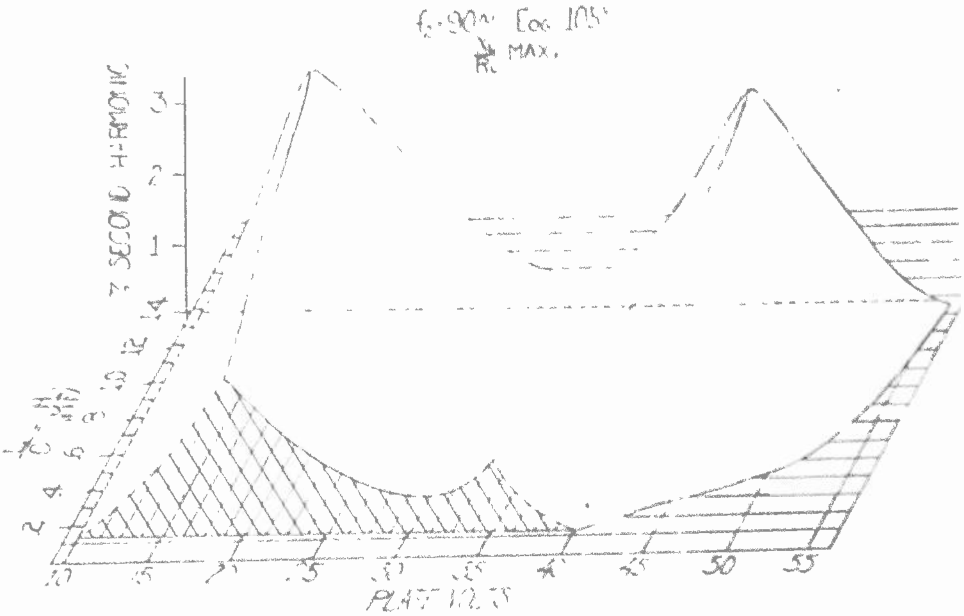

Fig. 29—Actual second harmonic voltages for the '24.

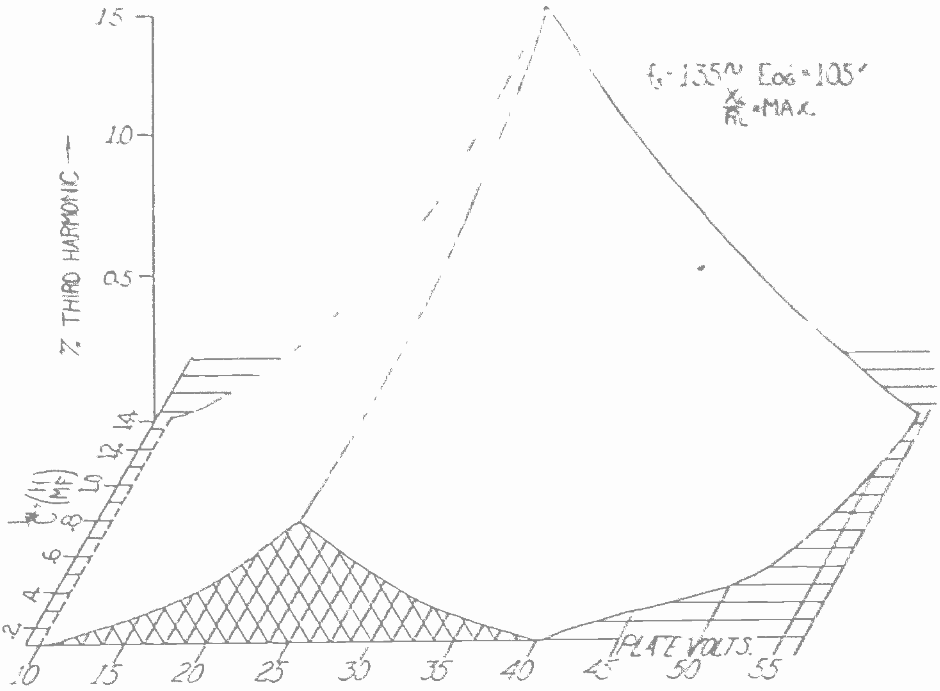

Fig. 30—Actual third harmonic voltages for the '24.

grid potentials only that they increase and decrease in over-all dimensions with slight irregularities at the boundaries.

Over a considerable range of (L/C) ratio it can be said that the even harmonics are a maximum when the odd harmonics (exception on the fundamental) are a minimum and vice versa. There is a discrepancy at the boundary points for voltages other than normal due to the past history of energy storage in the system, and in addition if these surfaces are calculated by the method shown in this paper the boundary contour may vary depending upon whether or not the operating point was brought up from zero or from some maximum plate voltage down to zero.

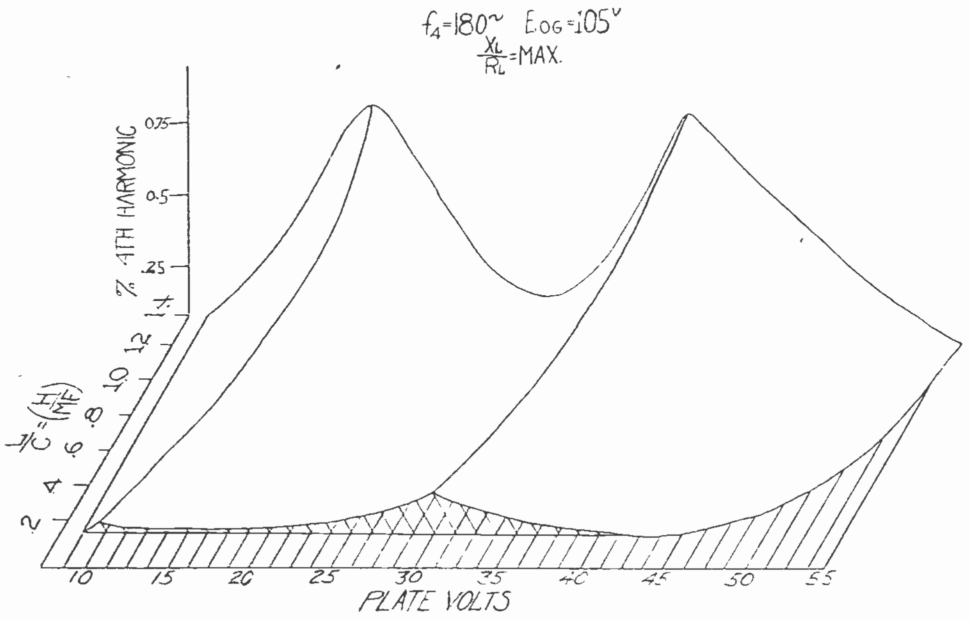

Fig. 31—Actual fourth harmonic voltages for the '24.

The general shape of these surfaces are verified by the differential analyzer solutions shown in Figs. 6, 7, 8, and 9. These solutions, as plotted, represent slices taken through the harmonic surfaces in a plane parallel to the plane of plate bias voltage and amplitude. The slices obtained by the machine solutions are in good agreement with the general shape of the harmonic surfaces. It is not necessary to show in detail that the harmonic surfaces or the machine solution slices should relate to each other as they do from a consideration of the operating point and its relation to the two knees of the dynamic characteristic of (i_p) as a function of (E_p).

Appendix B

The Determination of the '65 Output Characteristics

The output characteristics were determined by applying a sinusoidal voltage to the plate of the electronic device and measuring the fundamental component of plate current. It will be found that any commercial frequency which is sufficiently high is applicable in this method. The circuit diagram used to measure the output characteristics is shown in Fig. 32. Variable voltages were obtained by the use of a low impedance slide-wire potentiometer by-passed by a condenser as indicated. The fundamental component of alternating plate current was measured by means of a ten-ohm shunt across a sixty-cycle vibration galvanometer arranged with a simple filter to remove the higher order harmonic and protect the galvanometer. It was found that the deflection produced by the remaining harmonic currents was negligible. The galvanometer calibration was constant and was determined by the comparison method, as shown in Fig. 32, with a standard thermocouple meter.

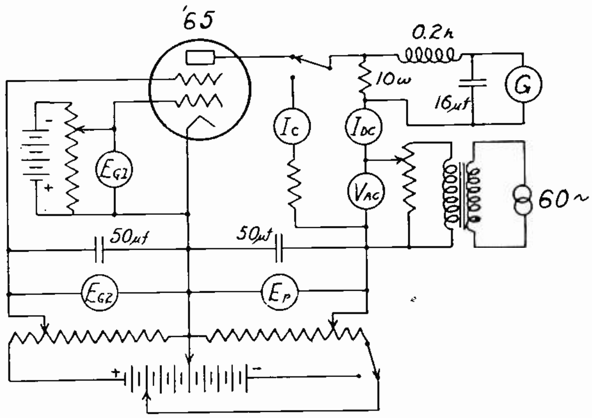

Fig. 32—Circuit for measuring the output characteristics of the '65.

The output characteristics for zero plate current were determined with the above method by varying the proper voltages until the direct-current plate ammeter registered zero deflection.

Appendix C

The Observed Output Characteristics of the '65

The observed output characteristics were obtained by determining the fundamental component of plate current and voltage when the tube and circuit were in the oscillatory state.

The circuit used to determine these characteristics is shown in Fig. 33. The current in the inductance was observed by means of a thermocouple ammeter. If the direct current was found large enough by comparison the inductance current was corrected for this factor though for the most part it was found to be negligible. The placement of the ammeter in the inductance leg of the circuit allowed the harmonic currents to be neglected. The alternating plate voltage was computed by the relation

$$V_{AC} = \omega L I_L$$

and the fundamental alternating-current component of plate current was computed by the relation

$$I_{AC} = I_L \frac{R}{\omega L}.$$

The observed characteristics for zero direct plate current were obtained in the same manner except that the plate circuit was blocked

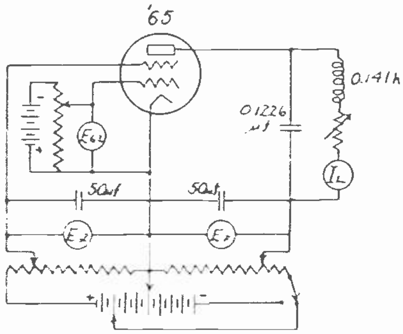

Fig. 33—Circuit used for the actual oscillation voltages with the '65.

off by a two-microfarad condenser. Oscillations were started by shorting the condenser either directly or through a choke. In order to eliminate the effect of slow variations in the characteristics of the tube the observed and measured output characteristics were taken within one hour of each other.

Acknowledgment

The authors wish to express their appreciation to Lieutenant James E. McGraw, United States Army, for his invaluable aid in the experimental work on the harmonic surfaces of the '24 and his coöperation in the work on the differential analyzer; for the interest and timely suggestions of Dr. Vanevar Bush, Vice President and Dean of Engineering, The Massachusetts Institute of Technology; and for the helpful suggestions of Mr. S. H. Caldwell of the same Institution. Finally the authors are indeed indebted to the departments of their respective institutions, Boston College and Columbia University, for their approval of this work as a contribution from the Department of Physics, Boston College, and the Department of Electrical Engineering, Columbia University.

BOOK REVIEWS

Measurements in Radio Engineering by Frederick E. Terman, Stanford University. Published by McGraw-Hill Book Co., New York. 400 pages. Price $4.00.

This volume is a companion book to the "Radio Engineering" by the same author, and in a practical manner contains timely information on measuring methods and equipment. It will doubtless come to be considered as a standard reference to radio laboratory practice.

In many places much equipment (especially if it looks neat and "commercial") is used without regard to any possible limitations and the indications used without corrections. In this text one finds a complete and up-to-date treatise on practically every type of radio measurement with details as to preferred methods, possible accuracy, and in most cases, equipment design arrangements. In the author's words, "Success in making measurements in radio work is primarily a matter of having available a satisfactory technique that is thoroughly understood rather than having available innumerable alternatives." This work will fill a dual purpose; as a reference work for the practicing engineer and as a text for college laboratory courses in radio measurements and experimental sciences. Toward this end, a well-arranged laboratory course is outlined in the appendix, which not only outlines the experiments, but utilizes equipment and methods which have been previously described in the text. A great deal of the apparatus can be constructed with student aid from the data so that any laboratory should be able to provide a well-equipped course even on a limited budget.

A complete list of the subjects covered would be quite extensive. The following résumé of chapter headings does not give a complete picture but will indicate the trend: voltage, current, power, resistance, inductance, and capacity measurements under different conditions and ranges, are taken up in the first four chapters. These are followed by chapters on frequency measurements, wave form and phase determination, and then vacuum tube characteristics. After these are chapters on measurements of audio-frequency amplifiers, radio receiving sets, transmitters, and antennas. A chapter on measurements of radio waves, field strength, and transmission lines follows. Separate chapters are included on two basic essentials to radio laboratories: laboratory oscillators and cathode-ray oscillographs.

The circuits in all cases are given, although in some cases the field of usefulness would be increased to service men and amateurs if a few additional circuit constants had been suggested on some of the apparatus. The various chapters are well supplied with selected references for additional study.

This book is recommended without reservation to anyone interested in radio engineering as a fundamental laboratory "accessory."

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The Cathode-Ray Tube at Work, by John F. Rider; 1935, 322 pages, published by the author, New York. Price $2.50.

This book, as the author clearly states, is intended as an instruction book for service men, amateurs, and students who are interested in the elementary theory and operation of the cathode-ray tube, and its practical application to electrical problems. In such regard, the text of the book is excellent.

The first three chapters are devoted to a nonmathematical, clear description of the manner in which a cathode-ray tube functions. The principal available commercial tubes are listed and described. Numerous diagrams are included to explain their operation. The theory of the most common types of sweep circuits is fully and simply explained.

Each of a large variety of available commercial cathode-ray oscillographs is described and its circuit diagram is given, with instructions for operation in general. The discussion and interpretation of Lissajous patterns, modulated waves, audio distortion wave forms, and the visual study of the output potential waves of various pieces of commercial apparatus is quite detailed.

The information, diagrams, and oscillograms in the book are so elaborate that they will be valuable to engineers for reference. The text should prove of interest and utility to engineers not familiar with the construction and application of cathode-ray tubes, and is as complete as a nonmathematical treatise on this subject could be.

†MADISON CAWEIN

Practical Radio Communication, by Arthur R. Nilson and J. L. Horning. First edition 1935. Published by McGraw-Hill Book Co., Inc., New York City. 754 pages. Price $5.00.

The object of this book is to provide in one volume most of the technical information required by the practical radio operator and technician. The first six chapters treat on principles of the following general subjects: direct-current electricity and magnetism, alternating-current electricity, vacuum tubes, transmitting circuits, receiving circuits, antennas, and wave propagation. The remaining chapters are devoted to the practical application, installation, operation, and servicing of radio communication equipment. The chapters take up the following subjects: studio acoustics and apparatus, control-room equipment and operation, broadcast and communication transmitters, radio receivers, radio aids to navigation, rectifier units, dynamoelectric machinery, meters, and storage batteries.

Two appendixes are given including such subjects as the vacuum tube numbering system, characteristics of transmitting and rectifying tubes and Western Electric tubes, wavelength-frequency tables, and other useful operating information.

There are many wiring diagrams and other drawings in which the various parts are carefully marked, thus shortening the required explanations in the text. Examples of some of the problems met in radio communication are worked out for the student. In treating the flow of electric current, the electron viewpoint of negative to positive is used. The book is written in a very clear, interesting manner. It is the belief of the reviewer that the book will be of value not only to the radio operator and technician but also to the radio engineer desiring practical information on the latest equipment now in use in radio communication systems.

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BOOKLETS, CATALOGS, AND PAMPHLETS

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

National Union Laboratories of 365 Ogden Street, Newark, N. J., has issued data sheets on the 6L7G, externally excited electron-coupled converter and on resistance-coupled voltage amplifier design.

A series of bulletins on direct-current automatic motor starters of various types has been issued by the Ward Leonard Electric Company of Mt. Vernon, N. Y.

Several booklets in German describing cathode-ray tubes and associated equipment have been issued by Leybold and von Ardenne, Köln-Bayental, Bonners Strasse 500, Germany.

Engineering data on magnet wire and coils are contained in Catalog No. 352 issued by the Anaconda Wire and Cable Company of 25 Broadway, New York City, and 20 N. Wacker Drive, Chicago, Ill.

Radio coils and products for the manufacturer, amateur, custom-set builder, and experimenter are covered in several leaflets issued by J. W. Miller Company, 5917 S. Main Street, Los Angeles, Calif.

The Type 148 cathode-ray oscillograph is described in a leaflet issued by Allen B. DuMont Laboratories of Upper Montclair, N. J.

Volume 1, No. 5 of "Electrical Measurements" issued by Sensitive Research Instrument Corporation of 4545 Bronx Blvd., New York City, is devoted to the universal polyranger.

Catalog No. 59 covering a large range of radio parts and assembled equipment has been issued by Wholesale Radio Service Company of 100 Sixth Avenue, New York City; 901 West Jackson Blvd., Chicago, Ill., and 430 W. Peachtree Street, N.W., Atlanta, Ga.

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1935

An Index covering the years 1909 through 1930 is available in pamphlet form, and may be obtained from the Institute at $1.00 per copy. A very few copies of the 1931 and 1933 Indexes, which were published in pamphlet form, are offered for distribution. The Indexes for 1932 and 1934 are to be found in the December issues of the PROCEEDINGS, preceding the advertising section.

The Index for 1935 is a continuation of the previous ones. It is numbered chronologically, and the numbers at the left of the titles are keys for use in referring to it from the Authors Index and the Cross Index.

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AUTHOR INDEX

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U

INCORRECT ADDRESSES

Listed below are the names and the last-known addresses of one hundred and thirty members of the Institute whose correct addresses are unknown. It will be appreciated if anyone having information concerning the present addresses of any of the persons listed will communicate with the Secretary of the Institute.

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Sparks, Neil R.	2462 Bancroft Way, Berkeley, Calif.
Spitalny, Max	235 West 102nd Street, New York, N. Y.
Spratt, George G.	116 So. 11th Ave., Coatesville, Pa.
Stevens, E. F.	314 West 94th St., New York, N. Y.
Straw, F. D.	222 West 53rd St., Ashtabula, Ohio.
Strommen, Einar	M-S Harboe Jensen, c/o Tropical Radio Tel. Company, Pier 3, N.R., New York, N. Y.
Swanson, Merrill J.	Radio Station WRAK, Williamsport, Pa.
Tickus, Louis J.	3001 St. Jean Ave., Detroit, Michigan.
Tingey, Robert	45 Nevern Square, Earl's Court, Rd., S. W. 5, England.
Toby, Arthur S. R.	c/o Edwards, 94 Ladywell Rd., Lewisham, London, S. E. 13, England.
Tumleson, James E.	4 South Portland Ave., Brooklyn, N. Y.
Turnquist, Arvid G.	628 Grand Ave., St. Paul, Minn.
Vaill, Edward W.	80 Maiden Lane, New York, N. Y.
Van Duyne, Eugene D.	41 Waldo Ave., E. Rockaway, L. I., N. Y.
Weagant, Roy A.	101 Forest Rd., Douglas Manor, L. I., N. Y.
Weaver, Karl S.	27 Howe Ave., Montclair, N. J.
Weeden, William N.	1 Alden Pl., Bronxville, N. Y.
Wiebach, William I.	207 Ave. C., Rochester, N. Y.
Williams, Aaron F.	RCA Victor Company, Inc., Building 6, Camden, N. J.
Williams, Henry L.	281 Carlton Ave., Brooklyn, N. Y.
Wohlfart, Emil	Wohlfart's Radio Service, 25-12 Steinway St., Astoria, L. I., N. Y.
Wolters, Brum J.	S. S. Haiti, Colombian Line, 17 Battery Pl., New York, N. Y.
Woodward, V. M.	808 Alamitos Ave., Long Beach, Calif.
Yeh, Yung Ching	The Institute of Military Communication. 18 Tung-Chieng Road, Nanking, China.

The Institute of Radio Engineers

Incorporated

330 West 42nd Street, New York, N.Y.

APPLICATION FOR ASSOCIATE MEMBERSHIP

(Application forms for other grades of membership are obtainable from the Institute)

To the Board of Directors

Gentlemen:

I hereby make application for Associate membership in the Institute of Radio Engineers on the basis of my training and professional experience given herewith, and refer to the members named below who are personally familiar with my work.

I certify that the statements made in the record of my training and professional experience are correct, and agree if elected, that I will be governed by the constitution of the Institute as long as I continue a member. Furthermore I agree to promote the objects of the Institute so far as shall be in my power, and if my membership shall be discontinued will return my membership badge.

....

(Sign with pen)

....

(Address for mail)

....

(Date) (City and State)

Sponsors:

(Signature of references not required here)

Mr. .... Mr. ....

Address .... Address ....

City and State .... City and State ....

Mr. ....

Address ....

City and State ....

The following extracts from the Constitution govern applications for admission to the Institute in the Associate grade:

ARTICLE II—MEMBERSHIP

Sec. 1: The membership of the Institute shall consist of: * * * (c) Associates, who shall be entitled to all the rights and privileges of the Institute except the right to hold any elective office specified in Article V. * * *

Sec. 4: An Associate shall be not less than twenty-one years of age and shall be a person who is interested in and connected with the study or application of radio science or the radio arts.

ARTICLE III—ADMISSION AND EXPULSIONS

Sec. 2: * * * Applicants shall give references to members of the Institute as follows: * * * for the grade of Associate, to three Fellows, Members, or Associates; * * * Each application for admission * * * shall embody a full record of the general technical education of the applicant and of his professional career.

ARTICLE IV—ENTRANCE FEE AND DUES

Sec. 1: * * * Entrance fee for the Associate grade of membership is $3.00 and annual dues are $6.00.

ENTRANCE FEE SHOULD ACCOMPANY APPLICATION

(Typewriting preferred in filling in this form) No. ....

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

Name ....
(Give full name, last name first)

Present Occupation ....
(Title and name of concern)

Business Address ....

Permanent Home Address ....

Place of Birth ....Date of Birth ....Age ......

Education ....

Degree ....
(college) (Date received)

TRAINING AND PROFESSIONAL EXPERIENCE

DATES	

Record may be continued on other sheets of this size if space is insufficient.

Receipt Acknowledged ....Elected ....Deferred............

Grade ....Advised of Election ....This Record Filed...........

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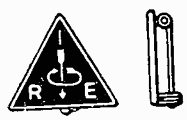

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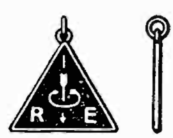

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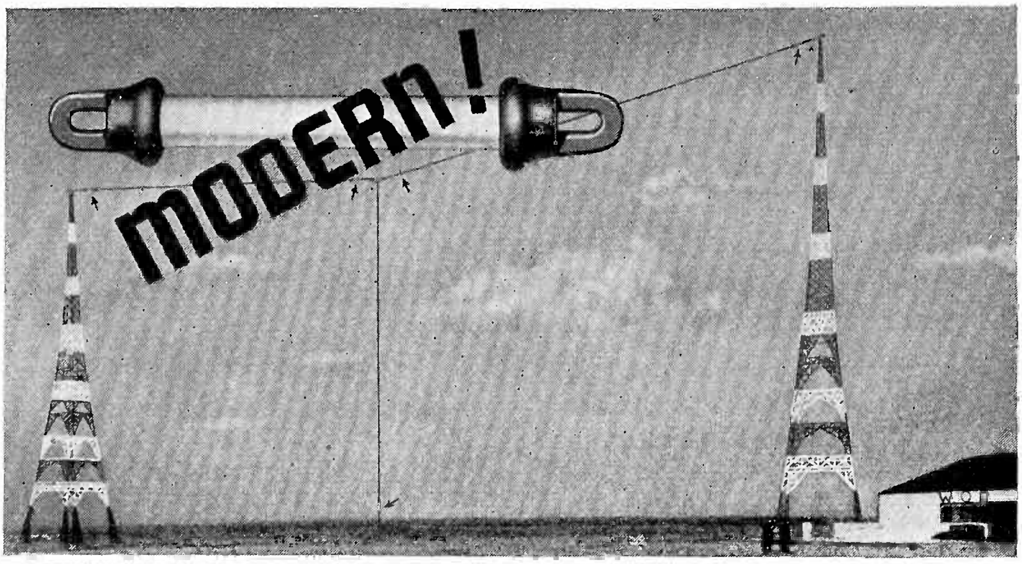

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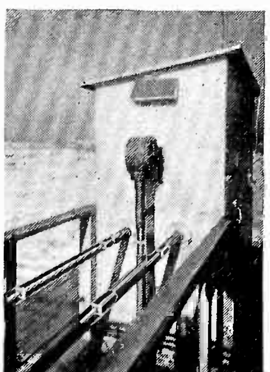

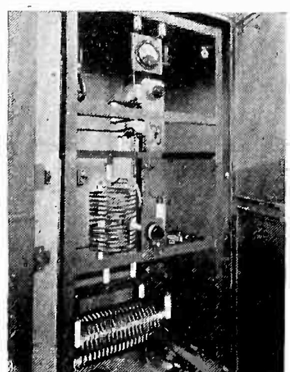

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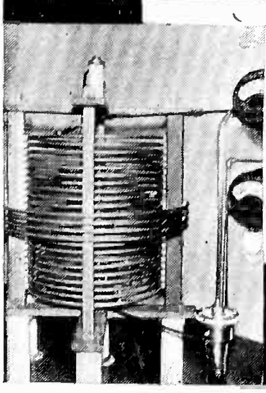

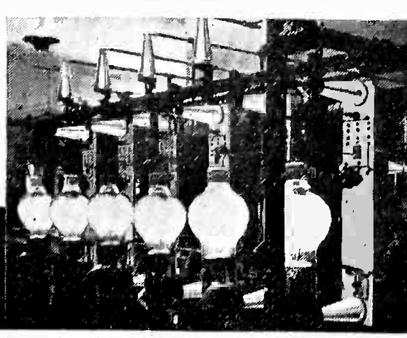

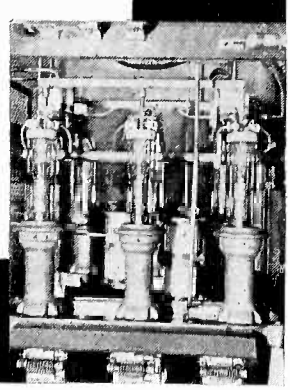

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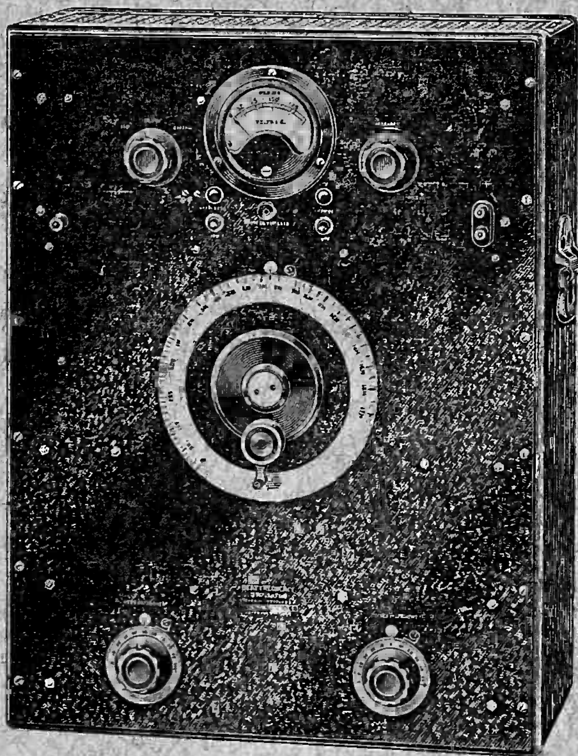

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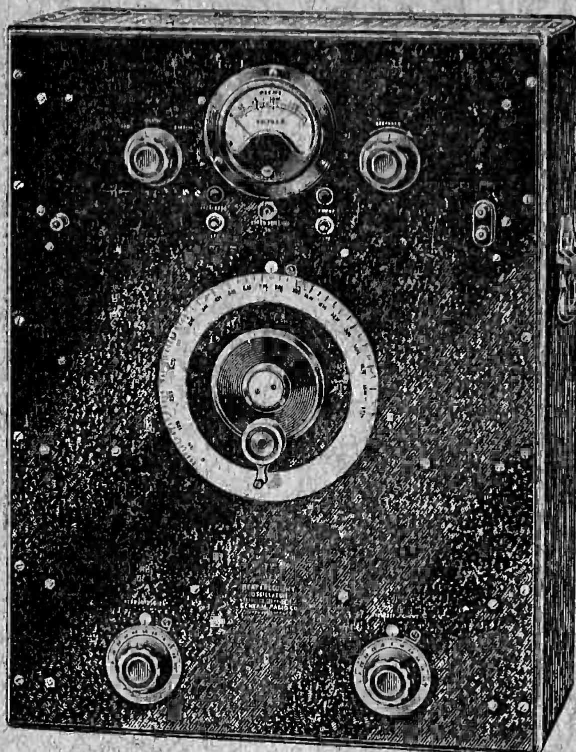

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