PROCEEDINGS

of

The Institute of Radio Engineers

Institute of Radio Engineers
Forthcoming Meetings

LOS ANGELES SECTION

September 18, 1934

PHILADELPHIA SECTION

October 4, 1934

WASHINGTON SECTION

September 10, 1934

CONTENTS

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.

REPRINTING PROCEEDINGS MATERIAL. The right to reprint portions or abstracts of the papers, discussions, or editorial notes in the PROCEEDINGS is granted on the express condition that specific reference shall be made to the source of such material. Diagrams and photographs published in the PROCEEDINGS may not be reproduced without making specific arrangements with the Institute through the Secretary.

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INSTITUTE SECTIONS

ATLANTA—Chairman, Henry L. Reid; Secretary, Philip C. Bangs, 23 Kensington Road, Avondale Estates, Ga.

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PITTSBURGH—Chairman, Lee Sutherlin; Secretary, A. P. Sunnergren, West Penn Power Company, 14 Wood St., Pittsburgh, Pa.

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WASHINGTON—Chairman, T. McL. Davis; Secretary, C. L. Davis, Washington Building, 15th and G Sts. N. W., Washington, D. C.

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED
JUNE 7, 1934

Transferred to the Member Grade

District of Columbia	Washington, 4550 Connecticut Ave.	Jackson, W. E.
Michigan	Grand Rapids, 1500 S. Union Ave.	Andres, L. J.
Mississippi	Greenwood, 209 Virginia St.	Fausett, F.
New York	New York City, 222 W. 39th St.	Cockaday, L. M.
	New York City, 2902 Grand Concourse	Hornung, J. L.

Elected to the Member Grade

Canada	Toronto, Ont., University of Toronto	Bayly, B. D. F.
England	London S. E. 7, 28 Sundorne Rd.	Meyer, L. W.
India	Bangalore, Indian Institute of Science	Sreenivasan, K.

Elected to the Associate Grade

California	Los Angeles, 1117 Venice Blvd.	Martin, C. S.
District of Columbia	Washington, 2500 K St. N.W.	Heatwole, E. P.
Georgia	Atlanta, 328 Mell Ave. N.E.	Hoke, G. W.
	Atlanta, 375 Blvd. N.E.	Howard, C. W
Illinois	Chicago, The Wil-Mar, 11 W. Division St.	Stevens, C. A.
Kentucky	Prestonsburg	Davis, R. B.
Maryland	Mt. Rainier, 3732–29th St.	Chapman, R. W.
Massachusetts	Concord, Ripley Hill Rd.	Morse, J. E.
	Dorchester, 485 Washington St.	Lane, A.
	Springfield, 130 Belmont Ave.	Johannessen, H. R.
New Jersey	Passaic, 70 Jackson St.	Siklosi, A.
New York	New York City, 1443 Wilkens Ave.	Berger, B.
	New York City, 400 E. 57th St.	Libbey, B. E.
	New York City, Columbia Broadcasting System, 485 Madison Ave.	Lodge, W. B.
North Carolina	Asheville, 40 Edgemont Rd.	Brimley, A. H.
	Charlotte, 1804 Wilmore Dr.	Rosekrans, P.
Pennsylvania	Middletown, 435 N. Spring St.	Arnold, C. H.
	Pottsville, 112 E. Market St.	Neiman, W. E.
Washington	Keyport, U. S. Naval Radio Station	Byer, E. D.
	White Swan	Freeland, C.
Australia	Casino, N.S.W., Box 27 P.O.	Cradick, R. L.
	Melbourne, Vic., 2 Wattle Grove, Moreland N. 13	Woodward, C.
Ceylon	Jaela, Jaela Estate	Soysa, N.
England	Anlaby, Nr. Hull, Yorks., 7 Tranby Lane	Witty, R.
	Dudley, Worcs., 28 Grange Rd.	Graham, A.
	Kenilworth, Warwickshire, Ivycote, Rosemary Hill	Lovelock, R. T.
	London W.9, 3 Landerdale Rd.	Clark, F. E. T.
	Shipley, Yorks., Avondale Bldgs., Bradford Rd.	Kinsey, C.
Territory of Hawaii	Paauilo, Hamakua Mill Co.	Fabius, A.
Irish Free State	Tralee, Co. Kerry, Basin View	Fennell, J. A.
Japan	Tokyo, Japan Broadcasting Corp., Hibiya	Hukugita, H.
New Zealand	Christchurch, 99 Abberley Rd.	Simpson, A. E. H.

Elected to the Junior Grade

California	Coronado, Box 333	Labowitch, D. M.
District of Columbia	Washington, 3226-13th St. N.W.	Langenegger, J. R.
Florida	Tampa, 6700 Elizabeth Ave.	White, J. D., Jr.
North Carolina	Greenville, Radio Station WEED	Humphrey, W. H., Jr.
Pennsylvania	Jeannette, 109 N. 1st St.	Kihchel, O. D.

Elected to the Student Grade

New Jersey	Grantwood, 6 Everett Pl.	Eichhorn, C. W.
Pennsylvania	Bethlehem, 448 S. New St.	Geiger, F. E.
	Carlisle, Beta Theta Pi House	Woodward, W. R.
Washington	Seattle, 4548-47th Ave. N.E.	Frost, N.
Porto Rico	Mayaguez, P.O. Box 26	Marti, H.

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 July 31, 1934. Final action will be taken on these applications August 1, 1934.

For Election to the Associate Grade

California	Hollywood, 5011 Ambrose Ave.	Mueller, W. A.
	Los Angeles, 1358 S. Grand Ave.	Cole, F. H., Jr.
	Los Angeles, 746 S. Hauser Blvd.	Lipschutz, B.
	Pasadena, 1637 Oakdale St.	Cobine, J. D.
Connecticut	Greenwich, 32 W. Putnam Ave.	Delage, G. C.
District of Columbia	Bellevue, Anacostia, R.M.S. Naval Research Lab.	Baldwin, J. H.
	Bellevue, Anacostia, R.M.S. Naval Research Lab.	Cooke, N. M.
	Bellevue, Anacostia, R.M.S. Naval Research Lab.	Hieronymus, J.
	Bellevue, Anacostia, R.M.S. Naval Research Lab.	Jenkins, A.
	Washington, 1109 Maryland Ave., N.E.	Pirtle, J. H.
Massachusetts	Boston, 75 S. Huntington Ave.	Goldfarb, P. M.
	Cambridge, 1572 Moss Ave.	Buckingham, S. A.
	Cambridge, Cruft Lab., Harvard University	Muzzey, D. S., Jr.
	Melrose, 40 Garland St.	Clark, G. L.
New York	Brooklyn, 281 Carlton Ave.	Williams, H. L.
	Mt. Vernon, 146 Union Ave.	Ferres, A. M.
	New York City, c/o American Tel. & Tel. Co., 32-6th Ave.	Harrison, P. W.
	New York City, 463 West St.	Knoop, W. A.
	New York City, 800 West End Ave.	Levenberg, M. H.
	New York City, 1225 Seneca Ave.	Salzberg, A.
North Carolina	Greenville, c/o Station WEED	Wallace, R. M.
Ohio	Steubenville, 724 Adams St.	McFeely, H. H.
Pennsylvania	Kingston, P.O. Box 111	Whitesell, F. L.
	Mt. Airy, Philadelphia, 438 W. Durham St.	Bingley, F. J.
	Philadelphia, 1515 W. Tioga St.	Lowe, L.
	Wilkinsburg, 1312 Elm St.	Mouromtseff, I. E.
Cuba	Havana, Maximo Gomez No. 53, Dept. 6	Verson, R. D.
England	Brondesbury, London N.W.6, 4 Plympton Ave.	Pinsent, J. T.
	Lewisham, London S.E.13, c/o Edwards, 94 Ladywell Rd.	Toby, A. S. R.
	N. Wembley, Middlesex, 108 The Fairway	Pollock, R.
	Rugby, 37 Hillmorton Rd.	Hartwright, J. F.
Italy	Milan, 4 Via A. Mussolini	Massarelli, F.
	Genova-Rivarolo	Testaverde, A.
Japan	Kyoto, Suehirokoji, Shijo-Otabicho	Kumagai, T.
	Tokyo, Tokyo Wireless Central Office (Tokyo Musen) Ote-Machi, Kojimachi-Ku	Kikutani, H.
New Zealand	Christchurch, 100 Rugby St.	Henderson, B. G.
	Christchurch, N.Z.B.B.'s Station 3YA, Gebbies Pass	Littlejohn, A.
Palestine	Haifa, Poste-Restante	Matussoff, G. M.

For Election to the Junior Grade

Florida	Key West, 1021 Washington St.	Wiggin, A. A.

For Election to the Student Grade

Connecticut	Cos Cob, 41 Grove St.	Landa, A. J.
Kansas	Ottawa, 1107 S. Main St.	Wolgast, L. H.
Michigan	Hart.	Hilbourn, C.
Ohio	Columbus, 2084½ N. High St., Apt. B	Beetham, R. B.
	Columbus, 229 W. Frambes	Moore, W. M.
	Delaware, R.R. No. 5,	Ross, A. J.
	Galion, 638 N. Columbus St.	Shumaker, H. A.
Tennessee	Ripley, 461 Jefferson St.	Rice, T. B.
Wisconsin	Madison, 843 S. Shore Dr.	Lund, A. O.
	Racine, 1915 Slauson Ave.	Tuttle, L.
Puerto Rico	Mayaguez, Central Igualdad	Villafone, O.
	Rio Piedras, Arsuaga No. 32	Lang, C.

CONVENTION COMMITTEE CHAIRMEN

Chairmen of the Ninth Annual Convention of the Institute of Radio Engineers, at the Benjamin Franklin Hotel, Philadelphia, May 28, 29, 30.

Standing left to right — E. B. Patterson, Entertainment, A. F. Murray, Exhibition, Knox McIlwain, Registration Information; seated left to right — E. W. Engstrom, Papers Technical Session, H. W. Byler, Treasurer, W. F. Diehl, Chairman, Jesse Haydock, Publicity, E. L. Forstall, Program. Not in picture: Mrs. W. H. W. Skerrett, Ladies Committee, R. L. Snyder, Treasurer, Harry Sadenwater, Entertainment.

June Meeting of the Board of Directors

The regular monthly meeting of the Board of Directors was held on June 7 at the Institute office. Those present were C. M. Jansky, president; Balth. van der Pol, vice president; Arthur Batcheller, O. H. Caldwell, Alfred N. Goldsmith, R. A. Heising, J. V. L. Hogan, L. C. F. Horle, L. M. Hull, F. A. Kolster, E. L. Nelson, E. R. Shute, H. M. Turner, A. F. Van Dyck, H. A. Wheeler, William Wilson, and H. P. Westman, secretary.

L. J. Andres, L. M. Cockaday, Floyd Fausett, J. L. Hornung, and W. E. Jackson were transferred to the grade of Member and B. de F. Bayley, L. W. Meyer, and K. Sreenivasan, were admitted to the Member grade. Thirty-two Associates, five Juniors, and five Students were elected to membership.

Deep appreciation was expressed for the very fine arrangements which had been made by the Philadelphia Section in the handling of the Ninth Annual Convention.

An invitation from the Detroit Section to hold the 1935 Convention in that city was tabled for consideration at a fall meeting of the Board.

President Jansky was designated the Institute's representative on the Executive Committee of the American Section of the International Scientific Radio Union.

A broad interpretation was placed on the term "engineering" appearing in Article II, Section 6 of the Constitution, prescribing the qualifications necessary for Student membership to include courses in physics and other sciences basic to engineering work.

Nine members were known to be placed by the Emergency Employment Service during May. New registrations have brought that figure to 667 and the service is continuing in its efforts to obtain employment for Institute members.

Ninth Annual Convention

The Ninth Annual Convention of the Institute was held in Philadelphia on May 28, 29, and 30 with headquarters at the Hotel Benjamin Franklin.

Eight technical sessions were necessary for the presentation of the thirty-two papers which were delivered. The summaries of these papers appeared in the May PROCEEDINGS.

Two inspection trips were made by the men. The first of these was to the Franklin Institute and the Fels Planetarium where a special lecture was given describing the installation. The other trip was to the RCA Victor plant in Camden where the manufacturing facilities of that organization were examined. In addition, a number of trips were arranged for the ladies.

Fifty-six organizations participated in the exhibition of engineering equipment and components for manufactured radio products. The exhibitions were held in hotel rooms and a complete floor of the hotel was given over to them.

The banquet which was held on the evening of the 29th was attended by 465. The Institute Medal of Honor was presented to S. C. Hooper, for many years Director of Naval Communications, for the orderly planning and systematic organization of radio communication in the government service with which he is associated, and the concomitant and resulting advances in the development of radio equipment and procedure.

The Morris Liebmann Memorial Prize was given to V. K. Zworykin of the RCA Victor Company for his contributions to the development of television. The awards were presented by President Jansky and Captain Hooper and Dr. Zworykin expressed their appreciation in their replies.

President Jansky then addressed the following statement to E. H. Armstrong who in 1918 received the first Institute Medal of Honor:

"Sixteen years ago you received from the Institute of Radio Engineers its Medal of Honor in recognition of your outstanding contributions to the radio art. Because of a chain of circumstances well known to many of us, you came to this convention with the intention of returning this medal to us.

"The impulse which prompted this decision on your part clearly demonstrates how deeply you feel your obligations to the Institute. The Board of Directors has been informed by me of your views to which it has given full and complete consideration.

"Major Armstrong, by unanimous opinion of the members of the Board, I have been directed to say to you

"*First:* That it is their belief that the Medal of Honor of the Institute was awarded to you by the Board in 1918 with a citation of substantially the following import; namely,

'That the Medal of Honor be awarded to Edwin Howard Armstrong for his engineering and scientific achievements in relation to regeneration and the generation of oscillations by vacuum tubes.'

"*Second:* That the present Board of Directors, with full consideration of the great value and outstanding quality of the original scientific work of yourself and of the present high esteem and repute in which you are held by the membership of the Institute and themselves, hereby strongly reaffirms the original award, and similarly reaffirms the sense of what it believes to have been the original citation."

Major Armstrong in a short reply expressed his deep appreciation of this act of the Board of Directors and also of the reception which those present at the banquet had given him.

The registration at the Convention totaled 940 which is the largest in many years.

Annual Meeting of the Sections Committee

The annual meeting of the Sections Committee was held during the Ninth Annual Convention in Philadelphia on May 28 and the following were in attendance: David Grimes, chairman; Austin Baily, (representing H. B. Coxhead), A. B. Buchanan, E. D. Cook, C. L. Davis, J. H. Dellinger, Samuel Firestone, H. C. Gawler, R. A. Hackbusch, G. W. Kenrick, H. S. Knowles, P. E. Lehde, O. S. McDaniel, J. H. Miller, R. S. Ould, B. E. Shackelford, W. A. Steel, A. F. Van Dyck, K. S. Van Dyke, and H. P. Westman, secretary. This group included representatives of the Boston, Chicago, Connecticut Valley, Detroit, New Orleans, Philadelphia, Toronto, and Washington Sections. O. H. McDaniel and W. A. Steel represented proposed sections in St. Louis and Ottawa, respectively.

An analysis of the financial operation of all sections during 1933 together with data on the number of meetings held and the membership figures for 1931, 1932, and 1933 were considered and discussed in detail.

Colonel Steel presented the thought of establishing an Institute section in Ottawa and discussed with the committee the various requirements that must be met so the group with which he is working in Ottawa might proceed most directly to a successful conclusion in their efforts in the establishment of such a section.

Mr. McDaniel of St. Louis also discussed with the committee the present possibilities of establishing a St. Louis Section. While the membership in the proposed territory is still rather small, a number of meetings have been held during the past couple of years and it is anticipated that an organization could be built up which at a later date can readily take over the operation of a section if such is approved by the Board of Directors.

The possibility of the necessity for a modification of the present system of rebates to sections was discussed in considerable detail. It is apparent that the largest of our sections are not obtaining sufficient income under the present system to permit their satisfactory operation. It is not entirely clear whether this situation is due to the change in rebate system which became effective two years ago or results from the substantial reduction in membership and its effect in reducing the section income. It was agreed that the matter be given further consideration at a later date when more data will be available.

Institute Meetings

Buffalo-Niagara Section

A meeting of the Buffalo-Niagara Section was held at the University of Buffalo on May 23 and was presided over by L. G. Hector, chairman. Eighteen members and guests were in attendance.

A paper on "Service Problems in Broadcast Receivers" was presented by C. L. Dirickson. It was opened with a brief history of the development of broadcast receivers since 1928 as regards servicing equipment for diagnosing failures. The increasing difficulties in servicing of all-wave receivers and the crowding of parts in the midget types were pointed out. There was then described a general procedure for diagnosing trouble starting with those failures most commonly met in the field and ending with those least likely to occur. Failures characteristic of certain makes of receivers due to poor design or ineffective materials were outlined. A general discussion followed.

Connecticut Valley Section

K. S. Van Dyke, chairman, presided at the April 26 meeting of the Connecticut Valley Section which was held in the Central High School auditorium, Springfield, Mass. The attendance was seventy-five.

A paper on "Developments in Police Radio Transmitters" was presented by D. G. Little, chief engineer of the Chicopee Falls plant of the Westinghouse Electric and Manufacturing Company. After recounting briefly the history of police radio, he proceeded through a general mechanical description of a Westinghouse police radio transmitter. A complete transmitter was displayed; it and lantern slides were used to illustrate the paper which was concluded with some views showing typical installations of both early and recent date.

The second paper on "Police Receiving Equipment" was presented by S. E. Benson, radio engineer on the staff of the United American Bosch Corporation. He established the considerations to be observed

in the design of police radio receiving equipment, including general limits as to power, economy, sensitivity, and power output. A circuit diagram of a standard police receiver was shown, representative models of old and new types were demonstrated as well as an ingenious installation on a motorcycle. Typical automobile installations were illustrated and the New York City Police Department equipment was used chiefly for this purpose.

CLEVELAND SECTION

F. T. Bowditch, chairman, presided at the May 31 meeting of the Cleveland Section held at the Case School of Applied Science. Forty-one members and guests were present.

The first paper was on "Factors Affecting Radiation from Broadcast Antennas" and was by J. F. Hill and C. C. Homeier. All the authors who presented papers were seniors at Case School and were introduced by Professor Martin. Mr. Hill discussed the effects of various structures upon the transmission pattern of a local broadcast station and Mr. Homeier told of fading in relation to the ionosphere.

The next paper was presented by P. F. Lange on the "Design and Measurement of Audio-Frequency Transformers" and he discussed various design factors and methods of measuring the over-all efficiency of such transformers.

A paper was then presented by A. I. Nace on "Sweep Circuits for Cathode Ray Oscillographs." The circuit developed for this purpose was described and its performance up to 70 kilocycles shown by graphs and demonstrated with actual equipment.

The final paper of the evening was on "Dielectric Constants of Liquid Apparatus" by J. Kissner who presented a résumé of tests of a precision variable condenser. Certain problems in its calibration were discussed and methods of accurately determining its capacitance outlined.

DETROIT SECTION

A meeting of the Detroit Section was held on April 20 at the University of Michigan in Ann Arbor. Samuel Firestone, chairman, presided and fifty members and guests were present. Ten attended the informal dinner which preceded the meeting.

A paper on "One- to Five-Centimeter Waves" was presented by N. H. Williams, Professor of Physics in the University of Michigan. He reviewed past experiments in the short-wave field which were based on electrical feed-back oscillator circuits. These could be used to about one and one-half meters below which Barkhausen-Kurtz or

Gill-Morrel arrangements were employed. After these circuits were described, work being done at the University of Michigan was discussed. This work was based on some experiments which showed that ammonia gas should absorb electromagnetic waves of approximately one and one-half centimeters length. To prove this experimentally, such waves had to be generated.

A magnetron oscillator was tried, and by successive reductions in the size of the elements of the tube, it was found possible to generate waves only one centimeter long. These waves were measured by means of mirrors and a grating which consisted of several aluminum shutters arranged similar to a window blind. The shutter faces remained parallel but could be moved apart. Points of maximum and minimum reflection from the shutter permitted the wavelength to be measured directly from the shutter constants. By inserting a container holding ammonia gas in the beam it was found that the ammonia molecule had a maximum absorption upon waves of one and one-half centimeters length. A number of those present participated in the discussion of the paper.

The May meeting of the Detroit Section was held on the 18th in the Detroit News Building and was presided over by Chairman Firestone.

Taintor Parkinson of the Physics Department of the University of Michigan presented a paper on "Radio Wave Phenomena and the Heaviside Layer." In it was traced the work leading up to the Kennelly-Heaviside layer theory. Records of reflections of radio waves from the E and F layers were shown and a method for making continuous records of the layer heights was described. Methods of calculating the density of the layers were outlined.

A number of the thirty-five members and guests in attendance participated in the discussion. The informal dinner which preceded the meeting was attended by ten.

WASHINGTON SECTION

A meeting of the Washington Section was held on May 14 at the Potomac Electric Power Company auditorium and was attended by 157. T. McL. Davis, chairman, presided. Thirty-eight attended the informal dinner which preceded the meeting.

The paper "WLW 500-Kilowatt Transmitter", presented at the Institute Convention in Philadelphia and abstracted in the May PROCEEDINGS, was presented by J. A. Chambers, technical supervisor of broadcasting of the Crosley Radio Corporation of Cincinnati, Ohio. It was discussed by a number of those present.

Personal Mention

E. K. Ackerman formerly with Amplivox Engineering, has become chief engineer of Sound Systems, Inc., Cleveland, Ohio.

Sidney Bloomenthal previously with the RCA Victor Company, has become associate physicist in the War Department, Frankford Arsenal at Philadelphia, Pa.

Ralph Bown has been transferred from the American Telephone and Telegraph Company to Bell Telephone Laboratories as associate radio research director.

L. A. Briggs formerly manager of the RCA Central Frequency Bureau, has been named European director of communications for RCA Communications with headquarters in London.

W. J. Brown has become chief engineer in the design department of Electrical and Musical Industries, Ltd., Hayes, Middlesex, England.

Previously with U.S. Radio and Television Corporation, E. C. Carlson has joined the engineering staff of General Household Utilities Corporation of Chicago.

Formerly at the University of Melbourne, R. O. Cherry has become a physicist for the Radio Corporation Party, Ltd., of Melbourne, Australia.

R. L. Davis previously with the Westinghouse Electric and Manufacturing Company has opened a consulting practice in Detroit, Mich.

Formerly with U.S. Radio and Television Corporation, H. C. Forbes has become chief engineer for automotive radio of General Household Utilities Company of Chicago.

F. R. Furth, Lieutenant, U.S.N., has been transferred from the U.S.S. Tennessee to the staff of the commander of the battleship division with headquarters in New York City.

J. H. Gough has left the Westinghouse Electric and Manufacturing Company to join the radio engineering staff of the Naval Research Laboratory, Bellevue, Anacostia, D.C.

W. S. Harmon has joined the radio engineering staff of General Household Utilities Company of Chicago having formerly been with the United Air Cleaner Corporation.

L. M. Harvey, Lieutenant, U.S.N., has been transferred from the U.S.S. Neches to the Navy Yard, Washington, D.C.

Guy Hill, Captain, U.S.A., has been transferred from Manila to Fort Monmouth, N.J.

M. W. Kenney formerly chief engineer of the Grunow Corporation, is now director of engineering for the General Household Utilities Company, Chicago.

Previously with the Westinghouse Electric and Manufacturing Company, G. R. Kilgore has joined the Research and Development Laboratories of RCA Radiotron Company, Harrison, N.J.

F. H. Kroger formerly with RCA is now with RCA Communications at Rocky Point, N.Y.

Now an engineer for Wired Wireless, Ltd., of London, England, Edward Lawrence was formerly with Gaumont British Picture Corporation.

R. D. LeMert formerly with Pioneer Mercantile Company is now a radio engineer for General Talking Pictures Corporation in New York City.

F. M. Link previously with the DeForest Radio Company has established the Fred M. Link Company of New York City.

Formerly with Les Laboratoires Standard in Paris, F. C. McLean has joined Standard Telephones and Cables of London, England.

H. G. Moran, Lieutenant, U.S.N., has been transferred from Cavite, P.I. to the Navy Yard, Brooklyn, N.Y.

H. V. Nielsen formerly with U.S. Radio and Television Company, had been made manager of the Spartan radio plant, Jackson, Mich.

Previously with Claude Neon National Laboratories, P. H. Nisley has been made superintendent of Voltarc Tubes, Newark, N.J.

Formerly with the Pacent Electric Company, O. B. Parker has established a consulting practice at Great Neck, N.Y.

C. E. Pfautz of the Radio Corporation of America has been transferred from Washington to New York as manager of the Central Frequency Bureau.

A. B. Pitts, Captain, U.S.A., has been transferred from Dayton, Ohio, to Langley Field, Va.

F. H. R. Pounsett previously with DeForest Crosley, is now on the radio engineering staff of Rogers-Majestic Corporation, Toronto, Canada.

R. C. Powell, Jr., formerly with R. C. Powell and Company, is now editor of *Broadcast Engineer*, New York City.

Formerly with E. H. Scott Radio Laboratory, E. R. Pfaff has become chief engineer for the Radiart Corporation, Cleveland, Ohio.

Sebastian Riccobono is now with Empire Electrical Products Company of New York City having formerly been with the United Scientific Laboratories.

Formerly in consulting practice, William Salt has joined the engineering staff of Radio Rentalo, Ltd., London, England.

C. E. Sargeant has been transferred from New York City to the Greenwood, Miss., headquarters of Supreme Instruments Corporation.

SOME NOTES ON THE PRACTICAL MEASUREMENT OF THE DEGREE OF AMPLITUDE MODULATION*

By

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Summary—*Some questions met with when measuring the degree of amplitude modulation are discussed. The degree of modulation is defined as the ratio of certain amplitude quantities, a clear distinction being made between what are defined as "up" and "down" values. The difference between the unmodulated carrier amplitude and the "mean amplitude during modulation" is stressed. It is shown that for an arbitrary amplitude modulation the effective value of the modulated current is dependent only on the amplitude of the harmonics, whereas the degree of modulation depends largely on the phase as well. From this it follows that any "effective" method of modulation degree measurement must be rejected as being not in accordance with the definition of the quantity to be measured. The effects of nonlinearity in a class B amplifier and of overmodulation in a class C amplifier are discussed, noting the rise of "mean high-frequency amplitude", the formation of harmonics, and the resulting degrees of modulation. After a short critical review of some existing methods, general requirements are proposed for a measuring device, and a description of the principle and performance of a direct reading modulation meter is given.*

A "ripple meter" constructed to measure exceedingly small degrees of modulation, and developed along lines similar to the modulation meter, is also briefly described with attention to its principle and its capabilities.

I. INTRODUCTION

ONE may perhaps use the term "modulation of fundamental form" or "fundamental modulation" for the now almost classical case of a pure sine wave of high frequency being modulated in amplitude by a low-frequency source, also of pure sine wave form. For this fundamental form of modulation the physical interpretation is relatively simple, the usual explanation being the well-known one which leads to two side frequencies being associated with modulation.[1]

The practical form of modulation is however rather more complex, first because the actual low-frequency source commonly has a broad frequency spectrum, and second because, due to nonlinearities, some distortion is always introduced in the low-frequency and high-frequency portions of the system.

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[1] Numbers refer to Bibliography.

Distortion in the low-frequency equipment is generally due to non-linearities of the valve's dynamic characteristics; in the worst case it is due to overloading or improper use of the tubes, and is most pronounced at the extremities of the low-frequency spectrum where load impedances may have unsatisfactory values. Distortion in the high-frequency part of the system is commonly caused by nonlinearities in the dynamic characteristics of the modulator class C amplifiers and the succeeding class B amplifiers. The resulting distortion is generally found to increase rapidly with the degree of modulation.

It is the main intention of this paper to discuss some typical forms of distortion frequently met with in practice, especially with regard to their influence on the degree of modulation and the effective value of the modulated high-frequency current. From this discussion some information may be had as to the specifications for a device for the measurement of the degree of amplitude modulation.

II. Definition of Per Cent Modulation

In a discussion of methods of measuring a quantity, it is desirable to start with a definition of the quantity to be measured.

We are considering a high-frequency current of cyclic frequency ω and of amplitude A:

$$i = A \sin \omega t. \tag{1}$$

Here we assume the amplitude A to be some arbitrary but periodic function of time. Let the period of this function A be T. By the magnitude,

$$A_0 = \frac{1}{T} \int_0^T A \, dt, \tag{2}$$

we define the mean value of the high-frequency amplitude *during modulation*. This distinction is of some importance, as it indicates that A_0 is not necessarily equal to the amplitude of the carrier current when not modulating. As will be shown later this case arises due to nonlinearities in class B amplifiers for high-frequency operation.

Further we write

$$A = A_0 + A(t) \tag{3}$$

where $A(t)$ represents some function with its mean value equal to zero. Further let A_{max} be the maximum value and A_{min} the minimum value of $A(t)$ in the period T. (See Fig. 1). Now, mathematically A_{min} may have any negative value; physically, however, no transmitter can

give less than zero amplitude, therefore we put A_{min} under the following restriction:
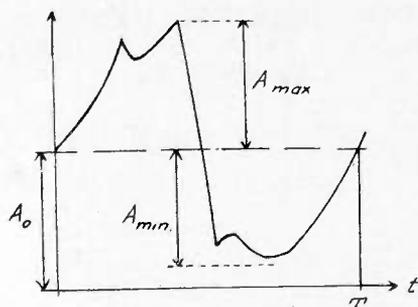
$$| A_{min} | \leqq A_0. \tag{4}$$

Fig.1—Illustration of an arbitrary but periodic amplitude function with the period T.

We then define "the percentage of modulation" in the following way:

(a) The percentage of "positive modulation" or shorter, "per cent up-modulation" we define by

$$m_u = \frac{| A_{max} |}{A_0} \cdot 100\%. \tag{5}$$

(b) The percentage of "negative modulation" or "per cent down-modulation" we define by†

$$m_d = \frac{| A_{min} |}{A_0} \cdot 100\%. \tag{6}$$

With restriction (4) for A_{min}, m_d cannot become greater than 100 per cent; however this restriction is not valid for the up-modulation m_u.
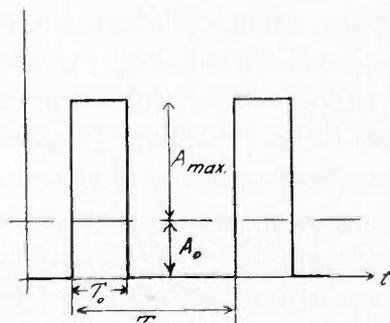
Fig. 2—Amplitude function illustrating a "sharp-dotted" modulation of a telegraph transmitter.

We have an example of this case when considering a telegraph transmitter which is giving "sharp" dots as illustrated in Fig. 2. If we assume rectangular curves and call T_0 the duration of one dot we have:

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† The terms "positive" and "negative peak modulation" have also been used.

$$A_0 \cdot T = T_0(A_0 + A_{max})$$

or,

$$
\left.
\begin{aligned}
m_u &= \frac{A_{max}}{A_0} = \frac{T - T_0}{T_0} \cdot 100\% \\[2mm]
m_d &= \frac{A_{min}}{A_0} = 100\%
\end{aligned}
\right\}
\qquad (7)
$$

thus,

$$m_u \gtreqqless 100\% \text{ if } T_0 \lesseqqgtr \frac{T}{2}. \qquad (8)$$

If $m_u = m_d$, we say that the modulation is symmetrical, but this does not guarantee the absence of low-frequency harmonics. A measurement showing the presence of symmetrical modulation is not conclusive evidence that the envelope is free from harmonics. It is, however, easily seen that the pure fundamental form of modulation is symmetrical.

Realizing that amplitude modulation means variation of amplitude, and that the peak amplitude which can be delivered with tolerable distortion limits the range of variation, it seems to be natural that any definition of per cent modulation must be based on relations between peak quantities.

III. The Effective Root-Mean-Square Value of an Arbitrary Amplitude Modulated Current

Since the introduction of modulation for practice purposes, well-known formulas have been established for the effective value of an amplitude modulated current. The "fundamental form" of modulation is especially well known in this respect. There are however some questions concerning the range of validity of these formulas which seem to be rather interesting from a physical point of view.

Taking the fundamental form, the following question may be raised: Is the validity of the formula unaffected by the ratio between the frequencies of the high-frequency and the low-frequency currents (ω/p)?

This seems to be of some importance in connection with badly distorted modulation where some harmonic of the modulation frequency may become of the same order as the high frequency. Another question is of interest: Is there some simple relation between the degree of modulation and the effective value of the modulated current when we are overmodulating a transmitter?

The following popular, though inexact, reasoning may be used to arrive quickly at the formula for the effective value of a modulated current:

Let us consider an arbitrary modulated current:

$$i = A \sin \omega t \tag{9}$$

A having the period T. The effective root-mean-square value (3) of the current i is then defined by:

$$J^2 T = \int_0^T A^2 \sin^2 \omega t \, dt. \tag{10}$$

Where we assume it to be correct to integrate over one low-frequency cycle, (10) becomes:

$$J^2 T = \frac{1}{2} \int_0^T A^2 dt - \frac{1}{2} \int_0^T A^2 \cos 2\omega t \, dt.$$

Making the rough assumption that A^2 is sensibly constant over the time π/ω which is the period of $\cos 2\omega t$, the last integral vanishes and we find

$$J^2 = \frac{1}{2T} \int_0^T A^2 dt \tag{11}$$

which means that the effective value of the modulated current is equal to $1/\sqrt{2}$ times the effective value of the envelope curve of the amplitudes.

Examples

(a)
$$A = A_0(1 + m \sin pt). \tag{12}$$

Here A is composed of a direct current A_0 and an alternating current of effective value $A_0 m/\sqrt{2}$, the effective value of this composed current being,

$$\sqrt{A_0^2 + \left(\frac{A_0 m}{\sqrt{2}}\right)^2}.$$

Thus we find:

$$J = \frac{A_0}{\sqrt{2}} \sqrt{1 + \frac{m^2}{2}}. \tag{13}$$

$A_0/\sqrt{2}$ is here the effective value of the mean amplitude during modulation (carrier).

(b) $\qquad A = A_0(1 + m_1 \sin pt + m_2 \sin 2pt).$ $\qquad$ (14)

The effective value of this envelope is equal to

$$\sqrt{A_0{}^2 + \left(\frac{A_0 m_1}{\sqrt{2}}\right)^2 + \left(\frac{A_0 m_2}{\sqrt{2}}\right)^2}.$$

Thus we find for the effective root-mean-square current:

$$J = \frac{A_0}{\sqrt{2}} \sqrt{1 + \frac{m_1{}^2 + m_2{}^2}{2}}. \qquad (15)$$

It is easily seen that the modulation given by (14) is symmetrical.
(c) The following amplitude function,

$$A = A_0(1 + m_1 \sin pt + m_2 \cos 2pt) \qquad (16)$$

gives the same effective current (15) as is easily verified, but here the modulation is unsymmetrical, the "up" value being

$$m_1 - m = m_u$$

and the "down" value,

$$m_1 + m_2 = m_d.$$

Thus, changing the phase of the second harmonic by 90 degrees changes the modulation from symmetrical to unsymmetrical, the effective current being the same.
(d) Assuming A to have the general form:

$$A = A_0\left\{1 + \sum m_n \sin (npt + \phi_n)\right\} \qquad (17)$$

and applying (11) one finds the general expression for the effective current,

$$J = \frac{A_0}{\sqrt{2}} \cdot \sqrt{1 + \frac{1}{2}\sum(m_n)^2}. \qquad (18)$$

The modulation given by (17) is an arbitrary one. It is interesting to note that no restriction whatever is put on the values of the fractions m_n. The two degrees of modulation resulting from (17) are determined by the values of m_n and of the phase angles ϕ_n. Let us now turn to the question of the influence of the ratio ω/p by considering in some detail the fundamental case of modulation.

$$\left. \begin{aligned} i &= A_0(1 + m \sin pt) \sin \omega t \\ &= A_0 \sin \omega t + \frac{mA_0}{2} \cos (\omega - p)t + \frac{mA_0}{2} \cos (\omega + p)t \end{aligned} \right\}. \qquad (19)$$

By taking the second power of the last expression one finds, after reducing,

$$i^2 = A_0^2 \left\{ \frac{1}{2} + \frac{m^2}{4} + \frac{m^2}{8} \cos 2(\omega - p)t + \frac{m^2}{8} \cos 2(\omega + p)t \right.$$

$$+ \frac{m}{2} \sin (2\omega - p)t - \frac{m}{2} \sin (2\omega + p)t$$

$$\left. + m \sin pt - \frac{m^2}{4} \cos 2pt - \frac{1}{2}\left(1 + \frac{m^2}{2}\right) \cos 2\omega t \right\}. \quad (20)$$

Thus the energy consists of a constant term plus a sum of trigonometric functions of cyclic frequencies:

$$\left. \begin{array}{c} 2\omega - 2p \\ 2\omega - p \\ 2\omega \\ 2\omega + p \\ 2\omega + 2p \\ p \\ 2p \end{array} \right\}. \quad (21)$$

By integrating (20) one finds the classical formula (13). Putting $p = \omega$ in (20) an interesting case arises; the lower side band assumes zero frequency and the effective value becomes

$$J = \frac{A_0}{\sqrt{2}} \sqrt{1 + \frac{3m^2}{4}}. \quad (22)$$

This is also easily seen by substituting for p directly in (19):

$$i = A_0 \sin \omega t + m A_0 \sin^2 \omega t$$

$$= A_0 \sin \omega t - \frac{m A_0}{2} \cos 2\omega t + \frac{m A_0}{2}. \quad (23)$$

The effective value of this composed current is

$$J = \sqrt{\left(\frac{A_0}{\sqrt{2}}\right)^2 + \left(\frac{m A_0}{2\sqrt{2}}\right)^2 + \left(\frac{m A_0}{2}\right)^2},$$

$$J = \frac{A_0}{\sqrt{2}} \sqrt{1 + \frac{m^2}{4} + \frac{m^2}{2}}$$

in accordance with (22).

This case, presumably, may be realized by modulating in the screen-grid circuit of a tetrode. We return now to the general modulation form given by the amplitude function (17) and consider the effective value from the side band method point of view.

We have,

$$i = A_0\{1 + \sum m_n \sin(npt + \phi_n)\} \sin \omega t. \tag{24}$$

By introducing the side frequencies in the conventional manner and taking the second power of i one finally finds:

$$\frac{i^2}{A_0^2} = \frac{1}{2} + \frac{1}{4} \sum m_n^2 + \frac{1}{8} \sum m_n^2 \cos\{(2\omega - 2np)t - 2\phi_n\}$$

$$+ \frac{1}{2} \sum m_n \sin\{(2\omega - np)t - \phi_n\}$$

$$- \frac{1}{2} \cos 2\omega t$$

$$- \frac{1}{2} \sum m_n \sin\{(2\omega + np)t + \phi_n\}$$

$$+ \frac{1}{8} \sum m_n^2 \cos\{(2\omega + 2np)t + 2\phi_n\}$$

$$- \frac{1}{4} \sum\sum m_n m_k \cos\{(2\omega + [k - n]p)t + \phi_k - \phi_n\}$$

$$+ \sum m_n \sin(npt + \phi_n)$$

$$- \frac{1}{4} \sum\sum m_n m_k \cos\{(n + k)pt + \phi_n + \phi_k\} \tag{25}$$

where n and k mean whole positive numbers from 1 to the highest order of harmonics present.

We find here also i^2 composed of a constant term

$$\frac{A_0^2}{2}\left(1 + \frac{1}{2} \sum m_n^2\right)$$

which constitutes the effective value found approximately in formula (18), plus a sum of trigonometric functions with the frequencies:

(a)	$l \cdot p$	$l = n,$	$l = n + k$	
(b)	$2\omega - lp$	$l = n,$	$l = 2n$	
(c)	2ω		$\cdot$	. (26)
(d)	$2\omega + lp$	$l = n,$	$l = 2n, \; l = k - n$	

Assuming none of these frequencies equal to zero, (18) is correct for the effective value.

Of the four frequencies in (26) only (b) and (d) can reach zero values, (b) for

$$n' = \frac{\omega}{p} \quad \text{and} \quad n'' = \frac{2\omega}{p} = 2n' \tag{27}$$

which means that ω must be a harmonic of p. The frequencies under (d) become zero for combinations of n and k which satisfy the relation:

$$n - k = \frac{2\omega}{p}. \tag{28}$$

To take a concrete example, assume that $\omega/p = 10$, then the following correction term is found from (25):

$$\Delta = \frac{1}{8}m_{10}{}^2, \ \cos 2\phi_{10} - \frac{1}{2}m_{20} \sin \phi_{20} - \frac{1}{4} \sum m_{20+k}m_k \cos (\phi_k - \phi_{20+k}).$$

Of course the fractions of so high an order as 10 are in general small quantities, so that the correction term Δ will generally be of no importance. Concluding, it may be stated that for the most general form of amplitude modulation there is no direct and simple relation between the two depths of modulation and the effective root-mean-square value of the modulated high-frequency current. In general the latter is dependent only on the fractions m_n, while the former depends largely on the phase angles ϕ_n as well.

IV. EFFECTS OF NONLINEARITY IN A CLASS B HIGH-FREQUENCY AMPLIFIER

We are considering the fundamental form of modulation

$$A_1 = A_0(1 + m \sin pt), \tag{29}$$

the effective value of the modulated current being

$$J = \frac{A_0}{\sqrt{2}}\sqrt{1 + \frac{m^2}{2}}. \tag{30}$$

We further assume this type of modulation to be applied to the grid of a class-B high-frequency amplifier having the following transfer-characteristic (excitation-curve):

$$A' = \alpha A_1 + \beta A_1{}^2. \tag{31}$$

This means that the plate circuit quantities are not strictly proportional to the grid circuit quantities. Putting $m = 0$ we find the unmodulated carrier amplitude in the plate circuit equal to

$$A_c' = \alpha A_0 (1 + \gamma) \tag{32}$$

where the abbreviation,

$$\gamma = \frac{\beta}{\alpha} \cdot A_0, \tag{33}$$

is introduced. By substituting A_1 from (29) in the transfer equation (31) one finds

$$A' = A_0'(1 + m_1 \sin pt - m_2 \cos 2pt) \tag{34}$$

which represents the amplitude function present in the plate circuit.

Here A_0' is the mean amplitude during modulation, m_1 is the first, and m_2 the second harmonic fraction. These quantities are related to the grid quantities and the transfer constants by the following equations:

$$A_0' = \alpha A_0 \left\{ 1 + \gamma \left(1 + \frac{m^2}{2} \right) \right\}$$

$$m_1 = \frac{m(1 + 2\gamma)}{1 + \gamma \left(1 + \dfrac{m^2}{2} \right)}$$

$$m_2 = \frac{m^2}{2} \cdot \frac{\gamma}{1 + \gamma \left(1 + \dfrac{m^2}{2} \right)}. \tag{35}$$

From this we see that the mean amplitude during modulation is not equal to the unmodulated carrier amplitude but differs by the quantity $\beta A_0^2 \, m^2/2$. For normal exitation curves β is positive and we get a certain increase in mean amplitude due to the modulation. This rise in amplitude increases with the input modulation m. From (32) and (35) it is seen that the ratio between the mean amplitude during modulation A_0', and the unmodulated carrier amplitude A_c', is given by

$$\frac{A_0'}{A_c'} = 1 + \frac{m^2}{2} \cdot \frac{\gamma}{1 + \gamma}. \tag{36}$$

For $m = 1$, i.e., 100 per cent modulation in the grid circuit, the rise in mean amplitude is computed as a function of γ, the result being

plotted in Fig. 3 (middle curve). The effective value of the modulated current is equal to

$$J = \frac{A_0'}{\sqrt{2}}\sqrt{1 + \frac{1}{2}(m_1{}^2 + m_2{}^2)}$$

$$= \frac{A_c'}{\sqrt{2}}\left(1 + \frac{m^2}{2}\frac{\gamma}{1+\gamma}\right)\sqrt{1 + \frac{1}{2}(m_1{}^2 + m_2{}^2)} \tag{37}$$
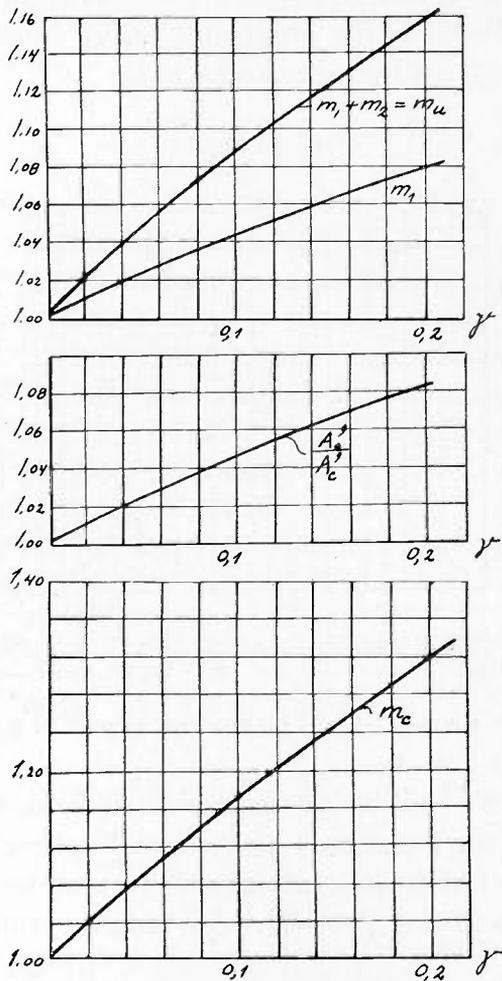

Fig. 3—Effects of nonlinearities in a class B, high-frequency amplifier, when $m = 100$ per cent input modulation in the grid circuit. The upper figure shows up modulation $m_u = m_1 + m_2$, first (or fundamental) harmonic fraction m_1 and down modulation m_d which is trivial $= 100$ per cent. The middle curve represent "rise" in mean amplitude due to modulation, and the lowest curve the degree of modulation m_c which wrongly would be calculated from the increase of root-mean-square valves due to modulation.

where m_1 and m_2 are to be substituted from (35). Assuming the nonlinearity to be unknown, one would calculate a wrong degree of modulation, m_c, from the rise in effective value by putting:

$$J = \frac{A_c'}{\sqrt{2}} \cdot \sqrt{1 + \frac{m_c^2}{2}}. \tag{38}$$

From (37) and (38) m_c can be computed as a function of m and γ. For $m = 100$ per cent, m_c as a function of γ is shown as the lower curve of Fig. 3.

For $\beta < 0$ which means that the transfer characteristic is concave towards the abscissa axis, γ is negative and the mean amplitude is decreased due to modulation. From (34) it is seen further that a second harmonic is introduced in the amplitude curve causing distortion. The corresponding distortion factor is found to be

$$D = m \cdot \frac{\gamma}{2 + 4\gamma} \tag{39}$$
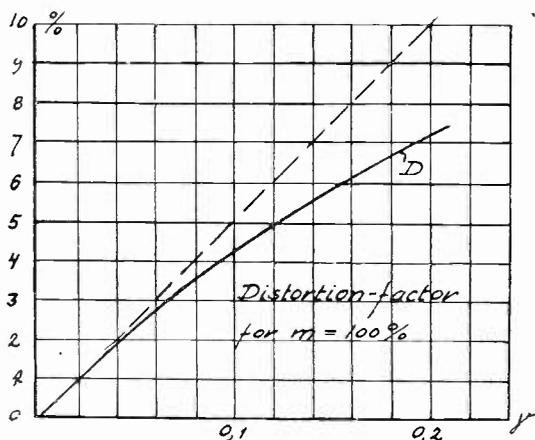
Fig. 4—Distortion factor due to nonlinearities in a class B, high-frequency amplifier for $m = 100$ per cent input modulation.

and is thus directly proportional to m. For $m = 100$ per cent. D is shown as a function of γ in Fig. 4.

This curve shows that a carrier deviation from linearity of 10 per cent is responsible for 4 per cent distortion. Further it is seen that the modulation is changed from symmetrical to unsymmetrical, the "up" and "down" values in the plate circuit being, respectively,

$$\left. \begin{aligned} m_u = m_1 + m_2 = m \cdot \frac{1 + 2\gamma + \gamma \dfrac{m}{2}}{1 + \gamma \left(1 + \dfrac{m^2}{2} \right)} \\[3em] m_d = m_1 - m_2 = m \cdot \frac{1 + 2\gamma - \gamma \dfrac{m}{2}}{1 + \gamma \left(1 + \dfrac{m^2}{2} \right)} \end{aligned} \right\} . \tag{40}$$

Here we note that for $m = 1$ in the input circuit, $m_d = 1$ also in the plate circuit for all values of γ. This, of course, is due to the fact that the assumptions postulate that zero input gives zero output.

We further note that the methods of measurement which are based on the measurement of the two quantities,

$$a = 1 + m_u$$

$$b = 1 - m_d$$

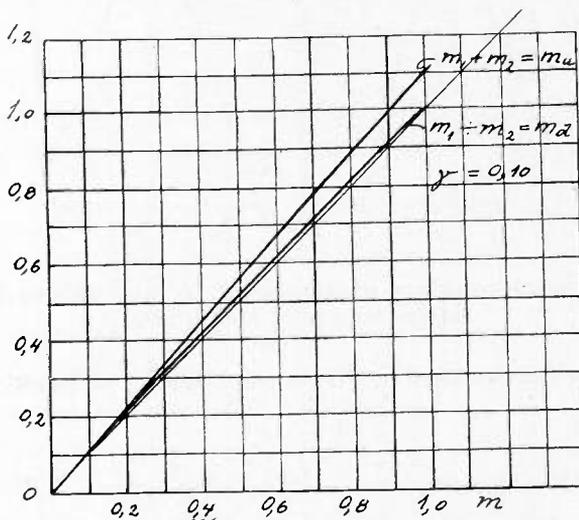without measuring the true mean amplitude during modulation, and which propose the ratio,

$$\frac{a - b}{a + b}$$

Fig. 5—"Up" and "down" modulations as functions of the degree of input modulation for $\gamma = 0.1$ in a class B, high-frequency amplifier.

for the "degree of modulation," in reality are measuring the quantity,

$$\frac{m_1}{1 + m_2}$$

which is less than the "up" value.

For $m = 100$ per cent the "up" and "down" values together with the first harmonic fraction m_1 are shown as functions of γ as the three upper curves of Fig. 3. It is seen that m_c shown in the lower curves is considerably in excess of the real "up" value.

For negative values of γ, it is readily seen from (40) that $m_1 < m$ and $m_2 < 0$. The "up" value is: $m_1 + m_2 < m$ and the "down" value: $m_1 - m_2 < m$, except for $m = 1$ when $m_1 - m_2 = 1$ also. Finally m_u and m_d are computed as functions of m for $\gamma = 0.1$, and the result is shown in the curves of Fig. 5. Note the characteristic form of the m_d curve.

V. Distortion Due to Overmodulation in a Class C Amplifier

We consider here a high-frequency amplifier of the class C type which is supposed to be plate modulated in the conventional way. Further we make the assumptions that the high-frequency output is zero for negative plate voltages, and that the characteristic follows linearly any "up" modulation.

By applying a sinusoidal plate voltage of modulating frequency p to the valve, we then get, in the overmodulated state, an amplitude function for the modulated high-frequency current as illustrated schematically in Fig. 6.
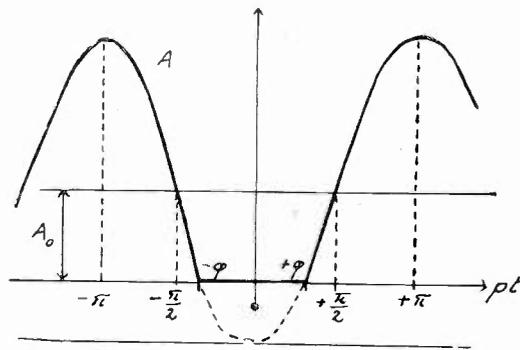
Fig. 6—Amplitude function illustrating overmodulation in a class C, high-frequency amplifier.

This amplitude function is mathematically defined by the following relations:

$$\text{where,} \quad \left. \begin{array}{ll} A = 0 & \text{for } +\phi > x > -\phi \\ A = A_0(1 - m \cos x) & \text{for } -\phi > x > -\pi \\ m > 1 & \text{for } +\pi > x > +\phi \end{array} \right\} . \quad (41)$$

The zero angle ϕ is then defined by:

$$\cos \phi = \frac{1}{m} \quad (42)$$

which also supposes $m > 1$ in order to give real values for ϕ. For the state of overmodulation we shall mainly consider the following quantities as functions of the fraction m:

(a) Rise in mean high-frequency amplitude due to the modulation.
(b) The real degree of "up" modulation.
(c) The content of harmonics and the distortion factor.

We develop the amplitude function defined by relations (41) in a Fourier series.

$$a(x) = a_0 + a_1 \cos x + a_2 \cos 2x + \cdots + a_n \cos nx \mp \cdots . \quad (43)$$

As the amplitude function is of the symmetrical type, satisfying $A(x) = A(-x)$, all sine terms vanish. For the mean amplitude during modulation a_0, we find,

$$\frac{a_0}{A_0} = 1 - \frac{\phi}{\pi} + \frac{tg\phi}{\pi}. \tag{44}$$

This relative "rise" in mean amplitude is plotted as a function of m in Fig. 7 as the curve marked "a_0/A_0 Rise." For $m = 2.0$ we find some 22 per cent rise. The maximum value of the amplitude function A (41)
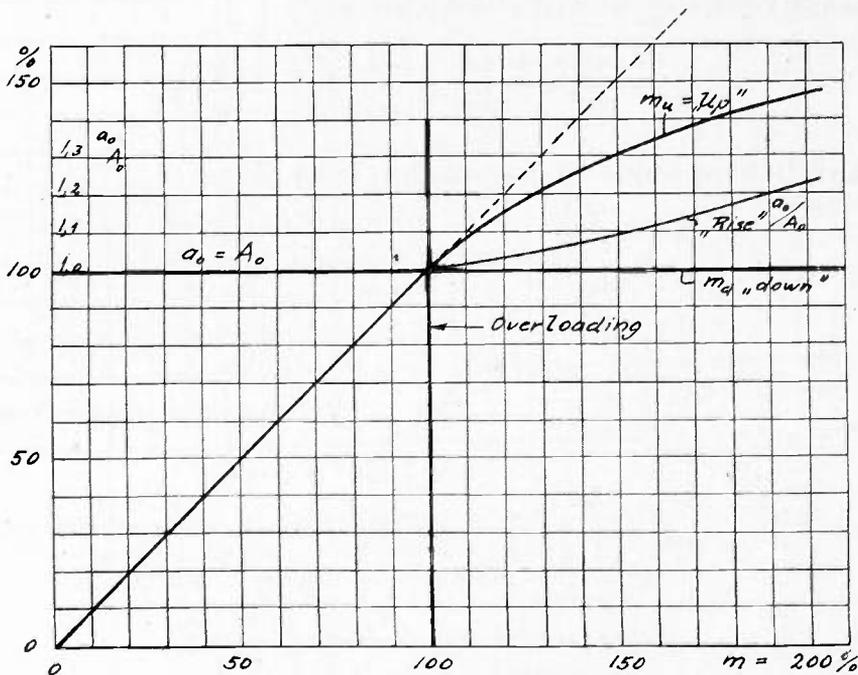

Fig. 7—Rise in mean amplitude and actual modulation degrees for overmodulation in a class C amplifier.

is equal to $A_{max} = A_0(1+m)$. The "up" modulation is thus found to be

$$m_n = \frac{A_{max} - a_0}{a_0} = \left[\frac{A_0}{a_0}(1 + m) - 1 \right]. \tag{45}$$

The "up" modulation is computed as a function of m, and the result is shown as the curve marked m_u of Fig. 7. The "down" modulation m_d is of course 100 per cent, the minimum amplitude being zero.

From this we see that 100 per cent overmodulation, i.e., giving an anode peak voltage equal to two times the voltage which is necessary for 100 per cent modulation, results in 22 per cent rise in mean high-frequency amplitude, an "up" modulation of 146 per cent, and a "down" modulation of 100 per cent.

Due to a considerable increase in mean amplitude the "up" modulation is less than what would be expected from the supplied low-frequency plate voltage.

Returning to the individual harmonics we find for the amplitude of the nth harmonic,

$$\frac{a_n}{A_0} = \frac{2}{\pi} \cdot \frac{1}{n^2 - 1} \left\{ \frac{\sin n\phi}{n} - tg\phi \cdot \cos n\phi \right\}. \tag{46}$$

This formula gives indefinite values for $n = 1$, the first harmonic, we have to compute directly, which gives,

$$-\frac{a_1}{A_0} = \frac{\sin \phi}{\pi} + \frac{1}{\cos \phi} \left(1 - \frac{\phi}{\pi} \right). \tag{47}$$

For the following harmonics up to the 5th we find,

$$\left. \begin{aligned}
\frac{a_2}{A_0} &= \frac{2}{3\pi} \left\{ tg\phi - \sin \phi \cos \phi \right\} \\[2mm]
\frac{a_3}{A_0} &= \frac{2}{3\pi} \cdot \sin^3 \phi \\[2mm]
\frac{a_4}{A_0} &= \frac{2}{15\pi} \left\{ 5 - tg^2\phi \right\} \sin^3 \phi \cos \phi \\[2mm]
\frac{a_5}{A_0} &= \frac{2}{\pi} \left\{ \frac{1}{3} \cos^2 \phi - \frac{1}{5} \sin^2 \phi \right\} \cdot \sin^3 \phi
\end{aligned} \right\}. \tag{48}$$

These harmonic amplitude ratios are plotted as functions of m in the curves of Fig. 8, where the trivial part for $m < 100$ per cent also is shown for the sake of illustration. It should be noted that the harmonics are plotted in ratio to A_0 and not to the fundamental a_1; this last ratio is however easily found from the data supplied. From the curves of Fig. 8 the magnitude of interferring side frequencies due to overmodulation may be found. Finally the distortion factor defined by

$$D = \frac{1}{a_1} \cdot \sqrt{\sum a_n^2}$$

is computed neglecting harmonics above the fifth.

The result is shown in Fig. 9, from which may be seen that if we assume 10 per cent to be the limit for perceivable distortion; $m = 140$ per cent modulation is allowable. However, interference in neighboring

channels may perhaps be the real limiting factor as regards overmodulation. For the effective root-mean-square value of the modulated current we find by applying (11).

$$I = \frac{A_0}{\sqrt{2}} \cdot \sqrt{\left(1 + \frac{1}{2\cos^2\phi}\right)\left(1 - \frac{\phi}{\pi}\right) + \frac{3}{2\pi} \cdot tg\phi}. \tag{49}$$
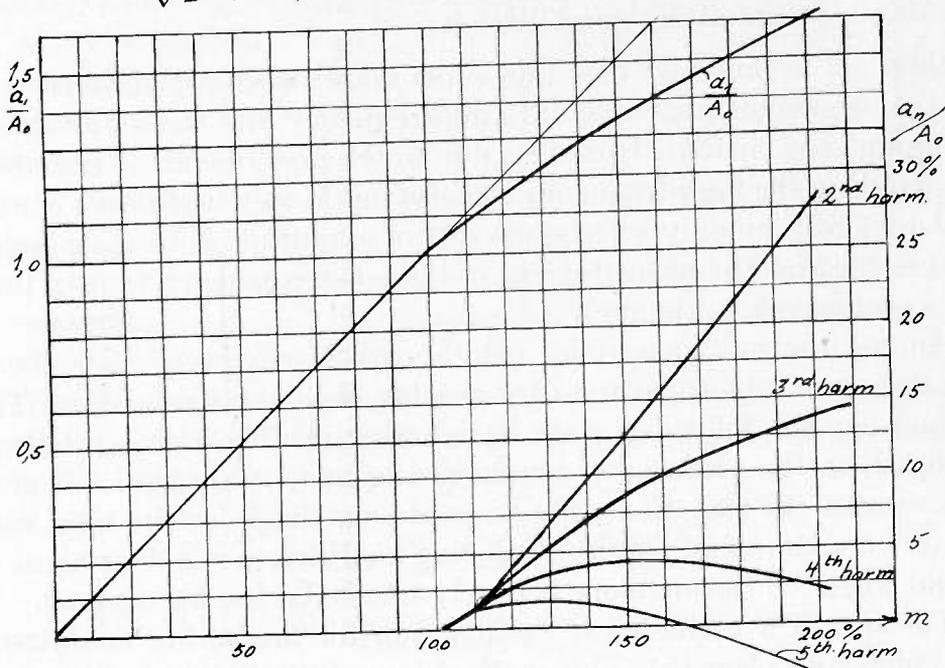

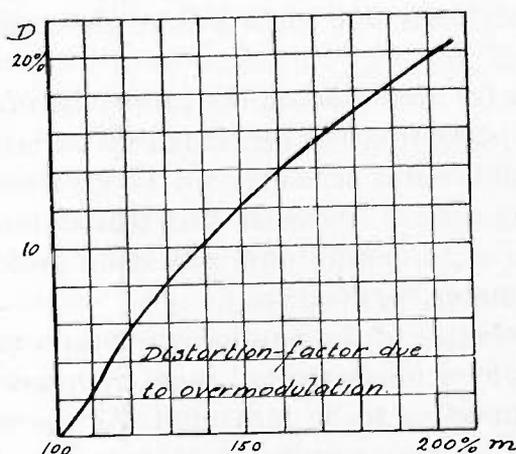Fig. 8—Harmonic distortion spectrum showing Fourier components up to the fifth for overmodulating a class C amplifier.

Fig. 9—Distortion factor due to overmodulation in a class C amplifier as a function of the low-frequency modulating voltage.

If we calculate the "degree of modulation" (m_c) from the rise in root-mean-square value, we also find in this case values in excess of the real degrees of modulation.

VI. DESCRIPTION OF A DIRECT READING MODULATION METER

The various proposed methods for the measurement of the degree of amplitude modulation may be classified as follows:

(a) Pure low-frequency measurements.
(b) Pure high-frequency measurements.
(c) Measurement after demodulation.

Method (a) is generally based on some known static relation between voltage or current of the modulating frequency and the amplitude of the modulated current. However, due to the possible fall or rise of the side bands in the high-frequency circuits this is only useful as a control method, and then only when some sort of amplitude indicating instrument is used for the measurement of the low-frequency quantity (usually a triode peak voltmeter).

Among the methods under (b) the rise of effective value due to modulation is sometimes used for quickly checking modulation. This method has the following main drawbacks: (1) The rise is relatively small, (2) in the presence of nonlinearities and/or harmonics it gives false results. (2) may thus only be used as a check for the ideal case, i.e., (i) when the mean amplitude during modulation is independent of m, and when (ii) the envelope is purely sinusoidal (no harmonics).

The use of a cathode-ray oscillograph for measuring modulation also belongs to class (b). This method is perhaps the one which most directly gives a qualitative picture of the modulation process. However due to the smallness of the screen picture and the lack of sharpness of the "spot," the accuracy for quantitative determinations is rather limited.

Methods under (c) are based on the properties of an ideal demodulator and give the possibility of observing any variation in mean high-frequency amplitude during modulation. To give correct results however the apparatus used to measure the demodulated low-frequency voltages should be of the amplitude indicating peak type (slide-back or peak triode voltmeter, oscillograph).

Following is a résumé of the requirements for a modulation meter:

(1) The principle of measurement must be in accordance with the definition of the quantity to be measured. As the modulation depth is defined as a ratio between amplitudes; this means that peak measuring instruments must be used. Measurement in the high-frequency field is necessary.

(2) Both the "up" and the "down" depth should be measurable, and also the mean high-frequency current or some quantity proportional to this value.

(3) The low-frequency response curve, i.e., measured modulation depth against modulation frequency, should be a straight line from 30 to 10,000 cycles, as well as for considerable low-frequency distortion in the amplitude envelope curve.

(4) For convenience the instrument should preferably be direct reading, a pointer indicating directly on a scale the measured value.

(5) The instrument should be self-contained and operated from the alternating-current circuit to simplify its use.

(6) The accuracy of the measured result should be within a few per cent.

(7) The high-frequency energy absorbed by the instrument should be as small as possible.

We shall now describe an instrument of type (c) which has been developed according to the above specifications and has been in use at this laboratory for some time.

Briefly, the instrument is built up of a linear high voltage diode demodulator D_1, a low-pass filter (L–C–C) with a terminating resist-
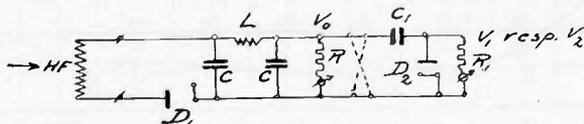
Fig. 10—Principle for measuring degrees of modulation after linear demodulation.

ance R which also serves as load resistance for the rectifier and a diode peak voltmeter $C_1R_1D_2$ for the peak measurement of the demodulated low-frequency voltages. (See Fig. 10.)

The measurement of degrees both "up" and "down" is made possible by commutating the low-frequency peak voltmeter as indicated by the dotted lines in the figure. Assuming that the diode and filter together with resistance R are rightly proportioned, we find in the voltage across the resistance R an exact reproduction of the amplitude function to which the high-frequency energy is modulated in the form of a direct voltage V_0 plus an alternating voltage with peaks V_1 and V_2. According to definition the two modulation degrees then are

$$m_1 = \frac{V_1}{V_0} \cdot 100\% \qquad\qquad m_2 = \frac{V_2}{V_0} \cdot 100\%.$$

If we keep V_0 constant V_1 and V_2 give directly the two degrees of modulation and with a direct reading low-frequency peak voltmeter the convenience of direct reading is readily obtained.

For this peak voltmeter a diode rectifier is used with a high CR constant, the input impedance of this voltmeter being so high that the

extra load on R due to the paralleling of the voltmeter may be neglected. This voltmeter was calibrated by means of an impulse voltage, of the form sketched in Fig. 11. The result of the calibration is shown in Fig. 11 from which it is seen that from 20 to 100 degrees (corresponding to 12 to 60 volts) this voltmeter reads the peak voltage within ± 1 per cent accuracy.

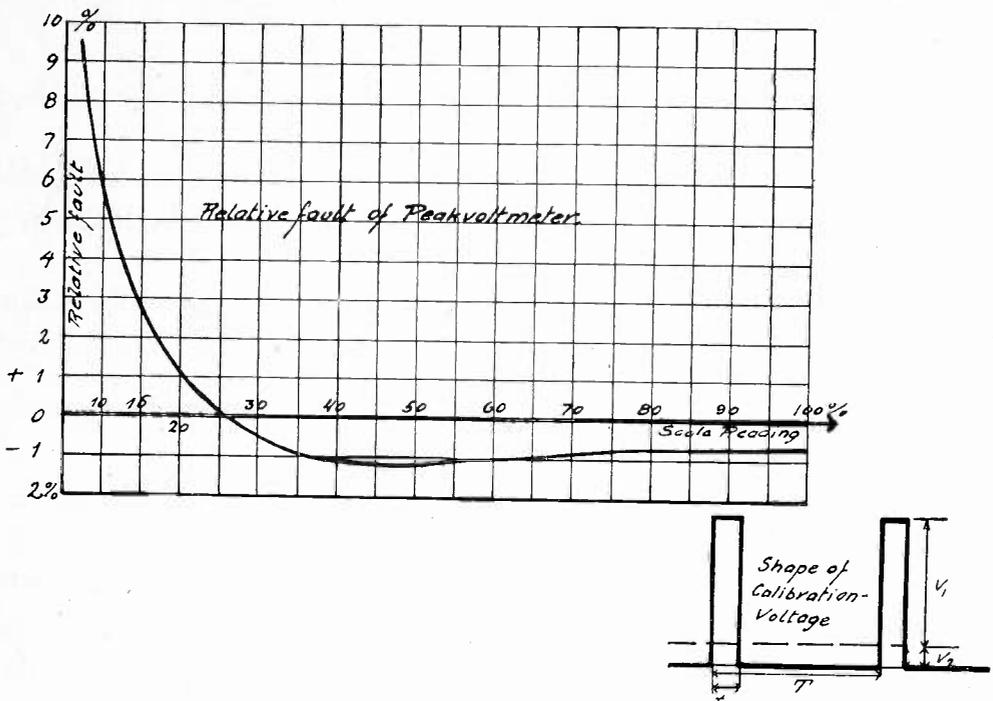The form of this correction curve is due to the fact that without external voltages applied to the valve some electrons are still flowing

Fig. 11—Curve showing relative errors of the low-frequency peak voltmeter adopted. The calibration voltage is of the impulse type shown schematically below in the right corner.

to the anode giving a small drop (1.1 volt) across R_1. This "bias" voltage gives to small voltages a positive error and to greater voltages it counteracts the normal error of such a voltmeter (reading less than peak). As to the design of the low-pass filter LC the following requirements are to be fulfilled:

(1) In order to keep high-frequency voltages out of the low-frequency peak voltmeter, the attenuation for the lowest "high-frequency" which is to be measured should be at least 40 decibels.

(2) Theoretically, the cut-off frequency could be placed at some frequency slightly above 10 kilocycles, as is shown by curve (a) of Fig. 12, where the cut-off occurs at 24 kilocycles. Measurements with such a filter where the "up" modulation was kept constant respectively at 39.2, 60.0, and 80 per cent are given in Fig. 13. As will be seen from these curves the readings are correct for the smaller degrees of modu-

lation where no distortion occurs, for greater degrees, however, where some distortion is to be expected we find increasing errors in the range of frequencies where the harmonics fall in the attenuation range of the
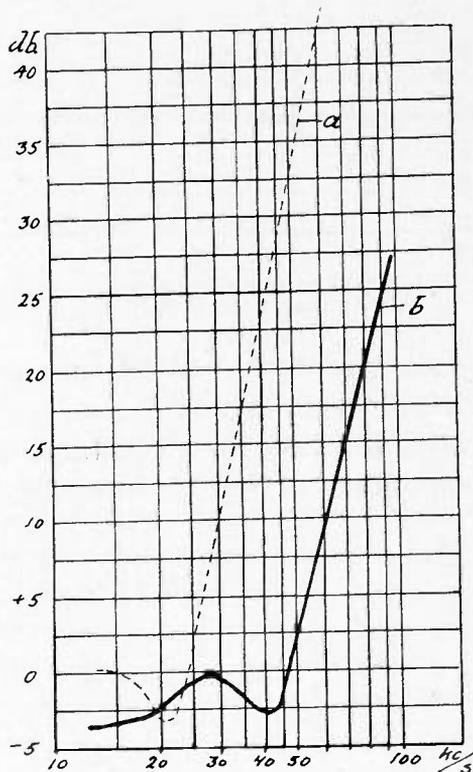

Fig. 12—Attenuation curves for the low-pass filter, *a* having a cut-off frequency too low, *b* the finally adopted one, leaving the 5th harmonic of 10 kilocycles free for transmission.
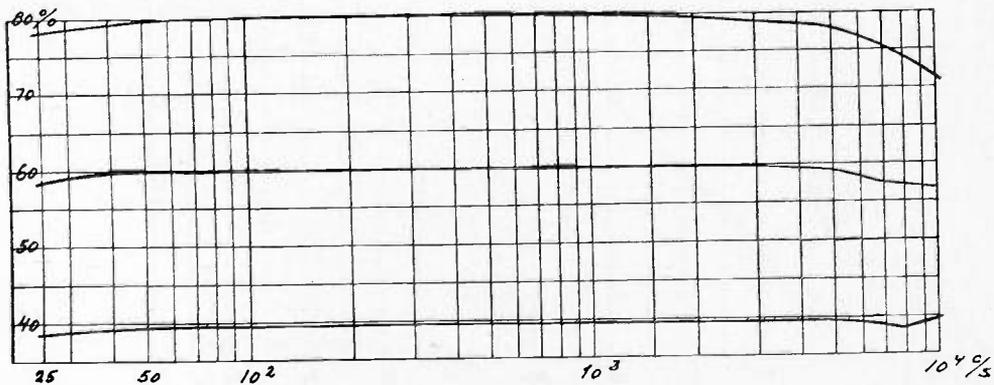

Fig. 13—Effects of too low cut-off frequency on the measuring results, showing increasing errors when harmonics are supplied but cut off by the low-pass filter.

filter (above cut-off). This is due to the fact that (according to definition) we measure the modulated low-frequency peak voltages, and these values are, as we have demonstrated, greatly dependent on the

amount of harmonics. From this it is seen, that the cut-off frequency must be placed safely above the important harmonics of the highest modulating frequency (10 kilocycles).
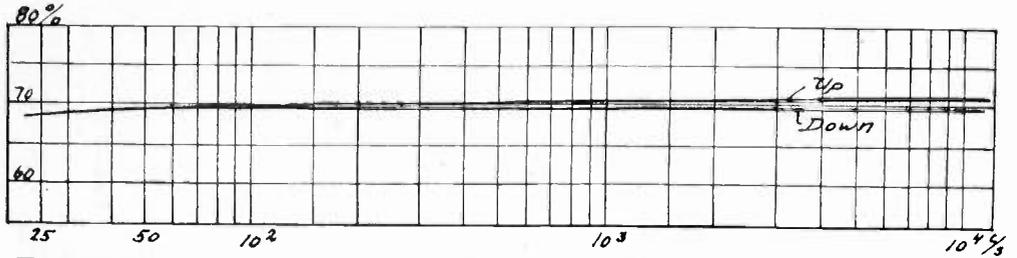
Fig. 14—Frequency response for 70 per cent up and down modulation by use of the filter finally adopted.

The filter finally adopted has a cut-off frequency of 50 kilocycles and the attenuation curve (*b*) in Fig. 12.

Measurements with this filter for a constant degree of modulation of 70 per cent are represented in Fig. 14, from which it is seen that the
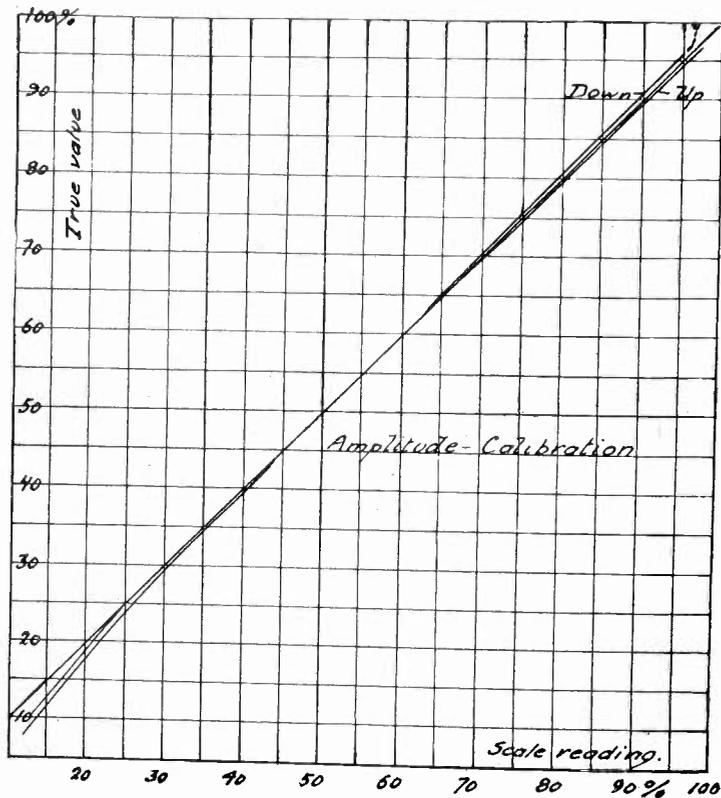
Fig. 15—Amplitude calibration of the described modulation meter.

frequency response from 25 to 12,000 cycles is linear within ± 0.25 decibels.

The filter is designed to work with a terminating resistance of $R = 8000$ ohms, the direct voltage V_0 across this resistance which is representative for the mean amplitude is chosen as 60 volts.

The energy absorbed from the high-frequency field is about 1 watt. The accuracy for the measured degree of modulation at 500 cycles can
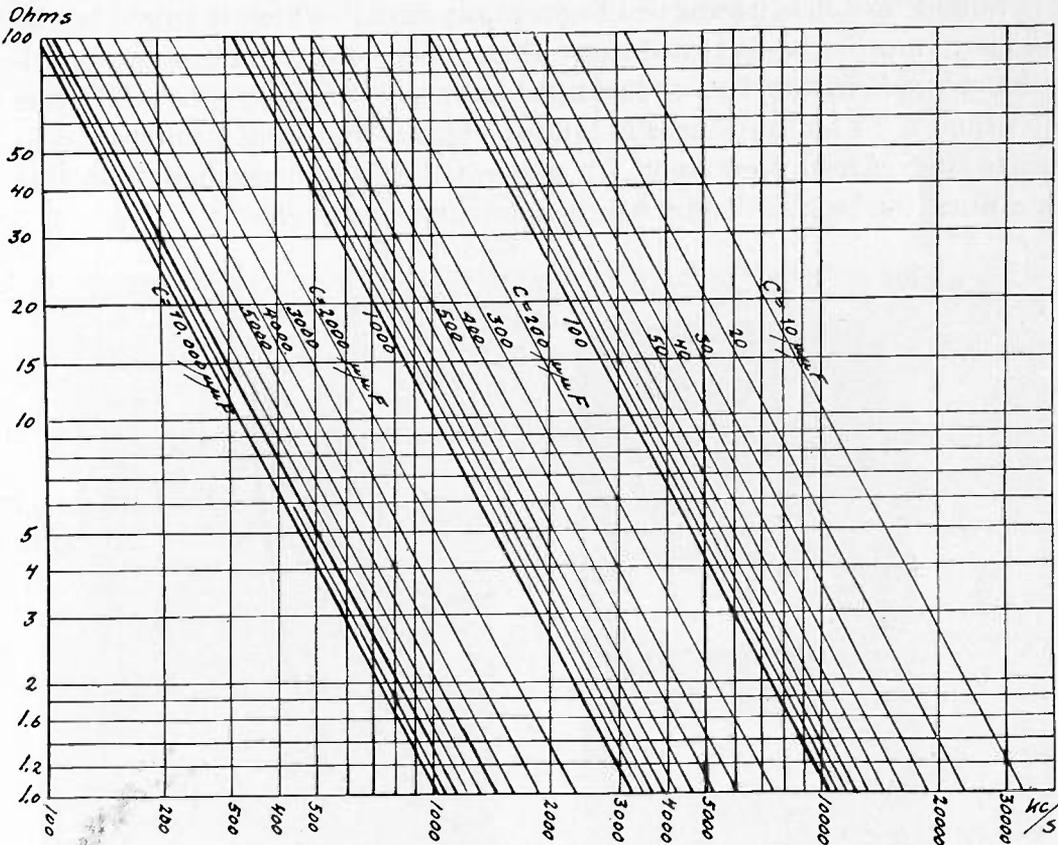
Fig. 16—Curves showing damping resistance to insert in a *L-C* circuit as a function of frequency and tuning capacitance to insure not more than 5 per cent loss of side bands by 10 kilocycles.
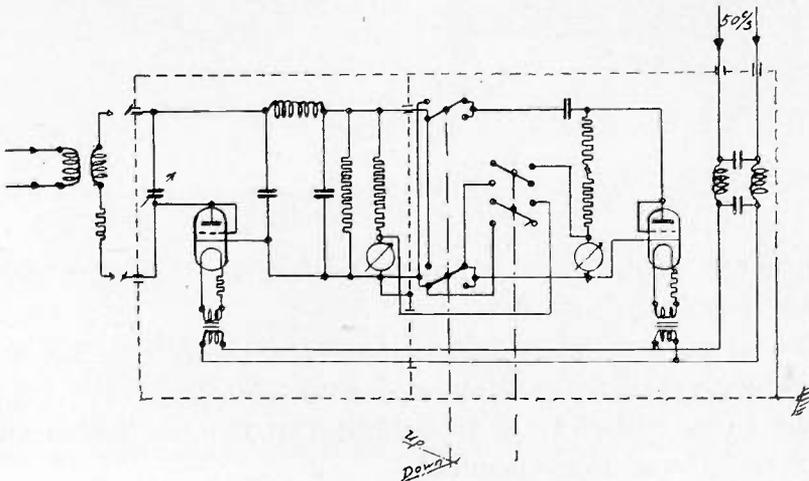

Fig. 17—Wiring diagram for the modulation meter.

be obtained from the calibration curve in Fig. 15. From 30 per cent to 100 per cent the relative error is well within 1.5 per cent, but increases

for smaller percentages due to the previously mentioned property of the low-frequency peak voltmeter.

Sometimes it is useful to use an input circuit which is tuned to the carrier. In order to have not more than 5 per cent reduction in the 10-kilocycle side bands, this circuit must have a certain damping, the value of which is to be found in Fig. 16 for various values of tuning capacitance and carrier frequency. In general it is recommended that this complication be avoided by using an untuned input circuit.
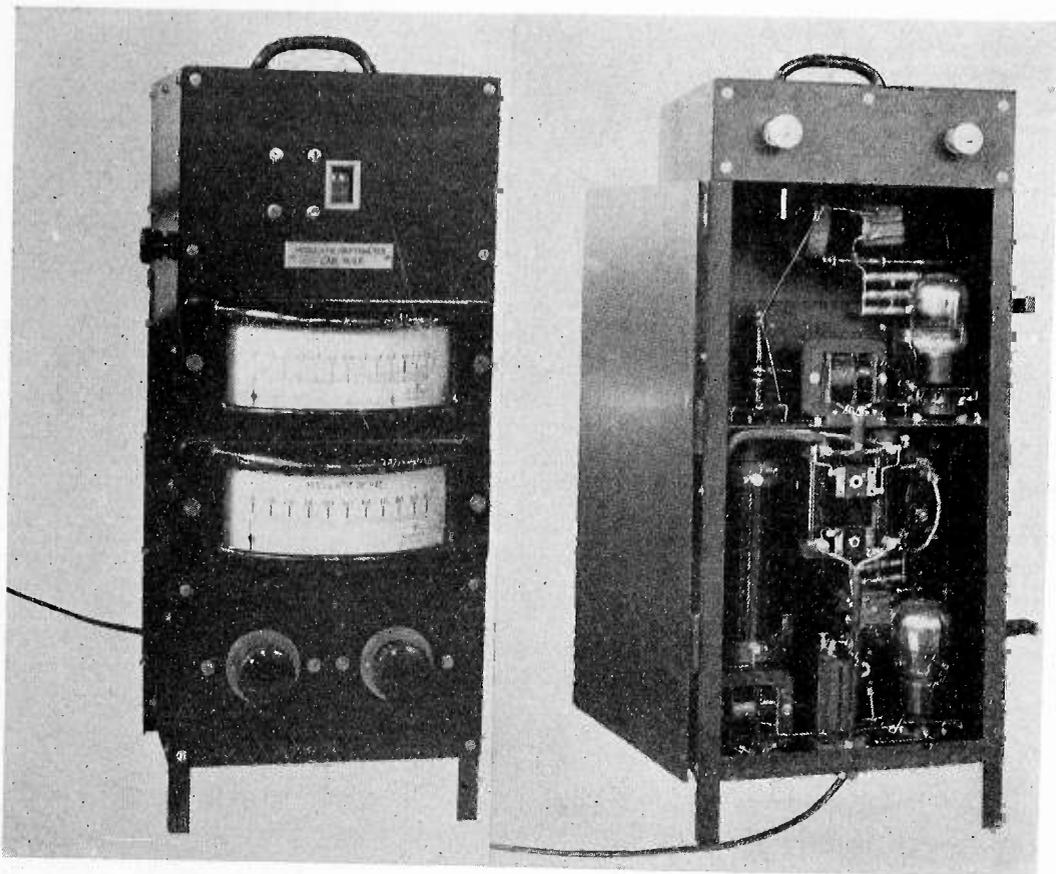
Fig. 18—Front and rear views (door opened) of a modulation meter as described.

A complete diagram of connections for a meter constructed according to these lines is given in Fig. 17.

A general idea of the constructional features may be had from Fig. 18 showing the front view and some details of the interior. A switch has been added to short-circuit the meters before switching from "up" to "down" measurement.

VII. Description of a Direct Reading Ripple Meter

Because of alternating-current components in power supply sources some extraneous modulation of small degree called "ripple modula-

tion" is present in almost every transmitter. A so-called "ripple meter" was constructed along the lines described in order to measure modulation degrees from 0.01 to 5 per cent. The principle (C) of measuring
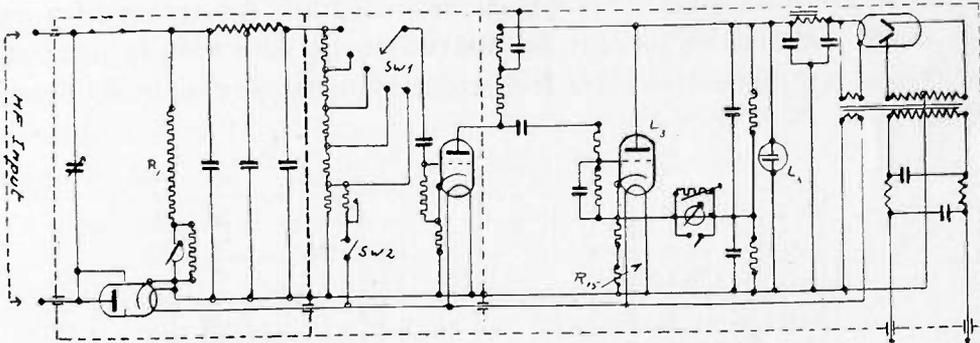

Fig. 19—Wiring diagram for a ripple meter used to measure exceedingly small degrees of ripple modulation.

after demodulation was adopted. However, in order not to arrive at an inconveniently high direct voltage after demodulation (V_0) a single
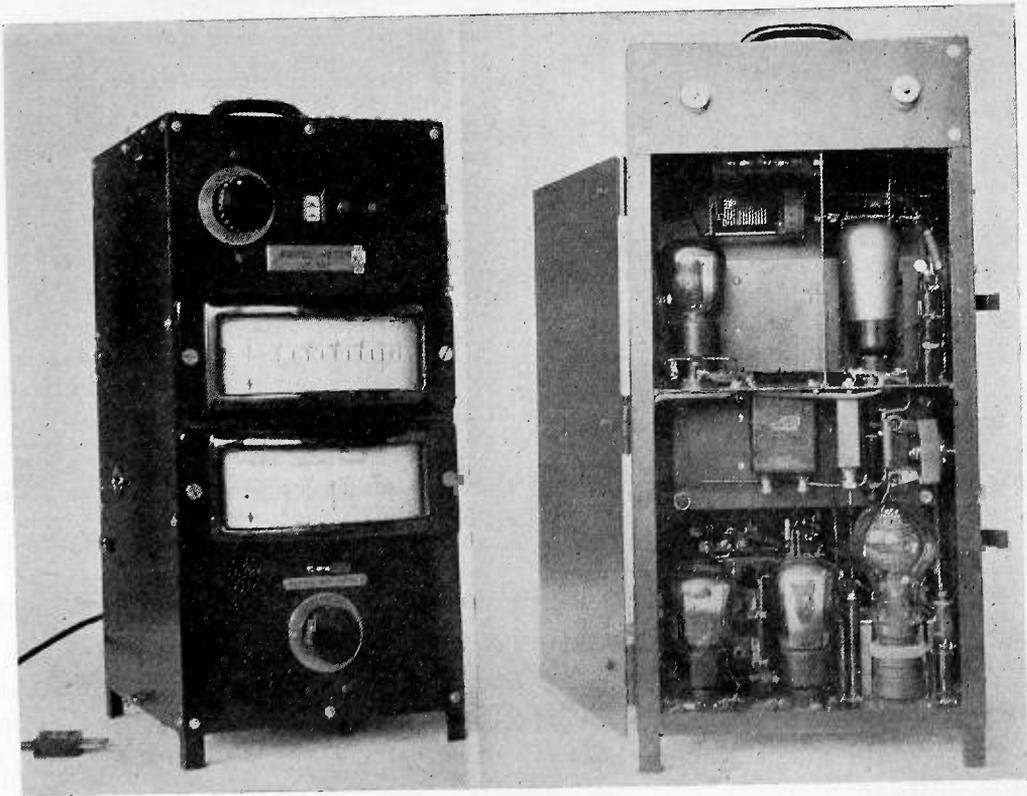

Fig. 20—Front and rear views of the ripple meter.

stage resistance coupled amplifier was inserted between the high-frequency filter and the low-frequency triode voltmeter.

This necessitated the addition of a calibration method, the one adopted being to apply a known 50-cycle voltage to the grid of the amplifying valve and to adjust the deflection of the triode voltmeter to a definite scale reading. This adjustment is made by means of a variable shunt across the indicating instrument of the triode voltmeter. The triode voltmeter used is of the bridge type employing grid demodulation.
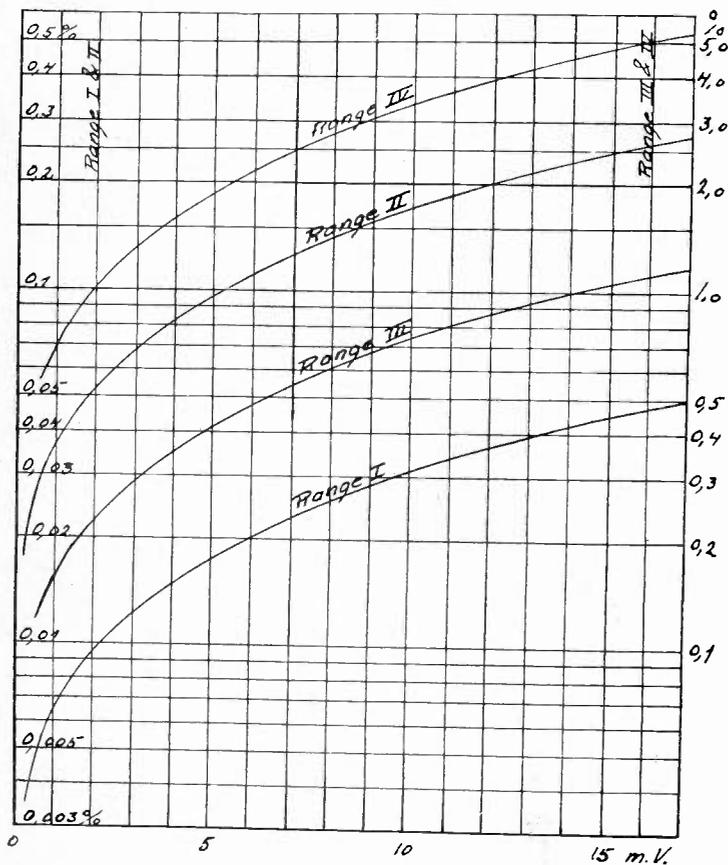

Fig. 21—Actual calibration curves for the four ranges of the ripple meter. The ordinates indicate per cent ripple modulation measured.

A complete wiring diagram is given in Fig. 19. The direct voltage across the resistor R_1 is 100, R_1 being 10^5 ohms.

The range from 0.01 to 5 per cent is covered by four positions of the switch Sw 1, the 5th position being for compensation and adjustment. The second switch Sw 2 is used for the application of a known calibration voltage.

A variable resistor R_{15} serves for compensation, the bridge being balanced with zero voltage at the grid of the triode L_3; L_4 is a stabilizing neon lamp. The external and internal appearance of the instrument is shown in Fig. 20.

The instrument was calibrated directly by introducing known ripple voltages into a battery-operated oscillator.

The calibration curves are given in Fig. 21. Fig. 22 gives the frequency response of the instrument, showing maximum deviation of ±2.5 decibels in the range of 40 to 1400 cycles. This relatively poor response is a result of the well-known difficulty in finding a compromise for exceedingly high attenuation for high-frequency, high working impedances (small energy), and linear frequency response.
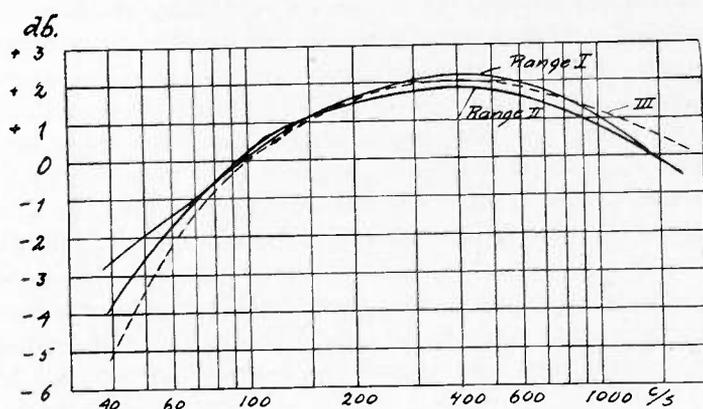

Fig. 22—Frequency responce of the ripple meter, measured for the ranges I, II, and III.

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MAINTAINING THE DIRECTIVITY OF ANTENNA ARRAYS*

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Summary—When a directive antenna array is used to maintain a certain minimum of signal in a given direction, or when a group of arrays are employed to provide intersecting space patterns, such as in the radio range beacon, it becomes necessary to maintain an accurate and constant relation between the phase and magnitude of the several antenna currents. Slight detuning effects in one antenna of a group will seriously alter the pattern. To overcome this trouble, a means of excitation has been developed which will hold a constant predetermined relationship between the various currents regardless of wide changes of antenna tuning. In brief, the system involves the use of constant current transmission lines built out with artificial sections to either (a) 90 degrees in length and connected in parallel, or (b) 180 degrees in length and connected in series.

Experimental data show the system to function satisfactorily and to be decidedly noncritical in adjustment. The new airways radio range beacon stations are using the arrangement with marked success and several broadcast stations have also applied its principles to their arrays.

I. INTRODUCTION

THE use of arrays of antenna structures to secure radiated space patterns of various shapes is a subject which has been considered in detail by various writers.[1] In general these arrays are intended to secure optimum transmission or reception along a given line or within a given arc, and in such cases it is not necessary that the pattern obtained be exactly correct. However, recently there has been considerable interest in the broadcast field, in antennas which produce no radiation in one or more directions.[2] Such antennas can be utilized to eliminate interference between adjacent transmitting stations and would increase the number of stations which could operate on a single channel. When so used, it is necessary that the minimum signal be as constant as possible at all times. This necessitates the maintenance of correct phase and amplitude relations between the currents in the various antennas of the array. Again, the development of the modified

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[1] R. M. Foster, "Directive diagrams of antenna arrays," *Bell Sys. Tech. Jour.*, vol. 5, p. 292, (1926); G. C. Southworth, "Certain factors affecting the gain of directive antennas," PROC. I.R.E., vol. 18, pp. 1502–1536; September, (1930).

[2] Radio Stations WFLA, in Florida, WKRC, in Ohio, and WORC, in Massachusetts, for example. The last two use a form of the stability methods described herein.

Adcock antenna system for the radio range beacon[3] presented another problem. In the radio range beacon, the on-course zone is determined by the intersection of two space patterns from two independent antenna groups. Any change in one antenna of a group would change the point of intersection and shift the course indication. Consequently it is necessary that the phase relations between the currents be maintained to much greater accuracy than with ordinary directive arrays.

During the progress of a research on the elimination of "night effect" in radio range beacons, conducted at the Round Hill Research Laboratories of the Massachusetts Institute of Technology, this shift of the antenna space pattern was found to handicap seriously the operation of the system in use at that period. At the same time the United States Bureau of Standards found similar difficulties in the operation of the experimental range beacon at Bellefonte, Pa., upon which experiments were also being conducted on the elimination of "night-effect." In order to overcome this difficulty, an investigation was begun which resulted in the development of an excitation system which maintains a stable space pattern regardless of changes in antenna tuning.

This system is also directly applicable to broadcast antenna arrays and several broadcast stations have now adopted it in order to insure the maintenance of minimum signal requirements where directive arrays are required by the Federal Radio Commission.

II. Effect of Change of Phase on the Space Pattern

The usual directive antenna array is composed of elements whose separation is a quarter wavelength, or multiples thereof, and in which the phasing is changed in similar fractions. In broadcast stations such separations are usually not possible, and in the radio range beacon installations the separation varies from 400 to 600 feet depending upon the ground available, which corresponds to separations of from one-eleventh to one-sixth wavelength for the frequency band employed. It is desirable, therefore, to investigate antenna arrays of this nature to determine the limits within which the phase may be allowed to vary.

Since the most rigorous requirements are probably those of the radio range beacon system, where the safety of aircraft is dependent upon a stable pattern, the excitation system was developed to meet these requirements.

Consider two vertical antennas whose height is less than one-quarter wavelength and which are suitably loaded at their bases. Omitting

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[3] H. Diamond, "The cause and elimination of night effects in radio raneg-beacon reception," *B. S. J. R.*, vol. 10, p. 7, (1933).

all unnecessary parameters, the general equation for the space pattern in the ground plane is

$$E = K \cos{(\pi m \cos{} + \pi n)}$$
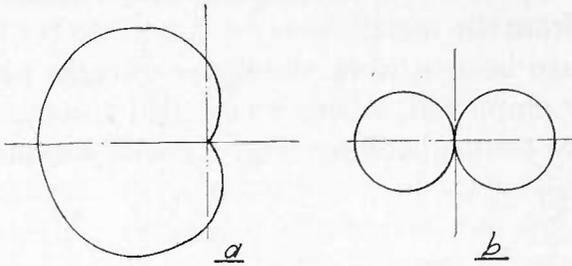
Fig. 1—Horizontal plane space patterns of two vertical antennas.

where E is the intensity at constant radius and at an angle of θ degrees from an arbitrary zero, m is the spacing of the two antennas in wave-
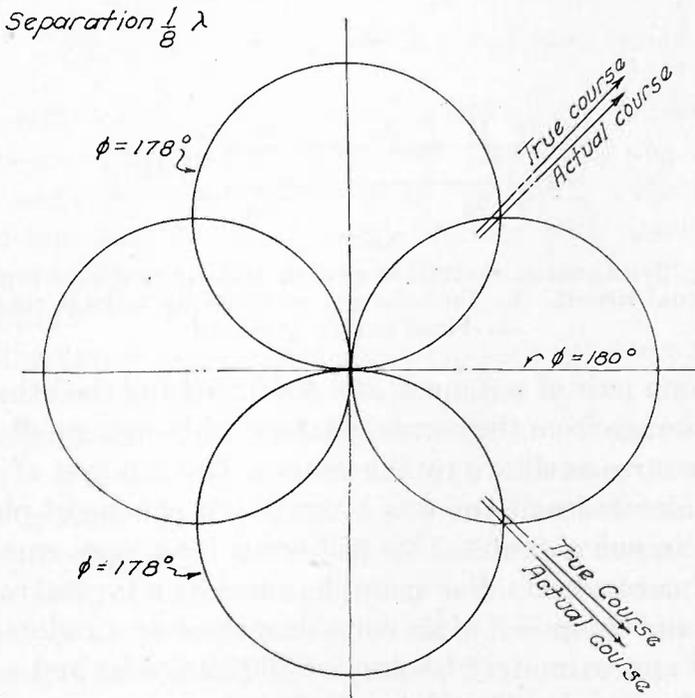

Fig. 2—Change of pattern of range beacon with slight phase shift.

lengths, or fractions thereof, while n is the time phase in fractions of a period. In general there are two sets of values of m and n commonly employed. First, $m+n = 1/2$, $m < 1/2$, which produces the cardioid or

single null form of pattern and second, $m < 1/2$, $n \cong 1/2$ which produces a figure-of-eight, or two-null form of pattern. These are illustrated in Fig. 1, (a) and (b). While this analysis is limited to these cases, the same principle applies to more complex arrays or group of arrays, as will be evident from the text.

In radio range beacon work the figure-of-eight pattern is the one most commonly employed, where n is varied through a certain small range to produce course bending. Fig. 2 shows a typical pattern with
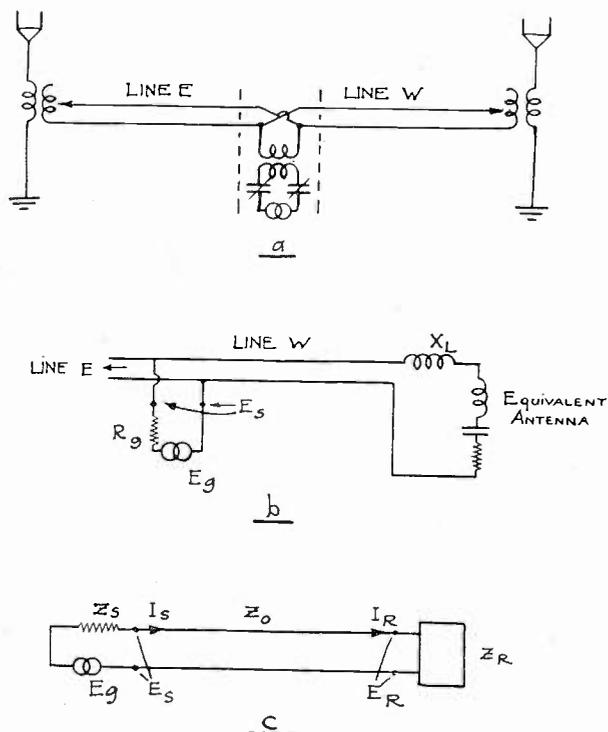
Fig. 3—Antenna excitation system and equivalent circuits.
a. Actual circuit. b. Transformer replaced by leakage reactance.
c. Final circuit analyzed.

$n = 1/2$ for one pair of antennas and $n = 31/64$ for the other pair. The amount of change from the correct pattern, although small, is sufficient to shift the course as shown by the arrows. This amount of course shift cannot be tolerated and yet was caused by a change of phase of only two degrees in one antenna. This will occur for a very small variation in the antenna constants. For example, consider a typical cage antenna 90 feet high and composed of six wires on a three-foot radius. This has a resistance of approximately 10 ohms at 300 kilocycles and a capacity of 500 micromicrofarads. The phase angle between current and voltage is arc tan X/R. For this to be equal to 2 degrees, tan $\theta = 0.035$. Assuming matched impedance in the exciting circuit, the effective resistance in the antenna circuit is 20 ohms. The net amount of reactance in the circuit would then be 0.7 ohm. Since $X = (X_L - X_c)$ and if we assume

X_L to be constant, then the change in capacitative reactance is 0.7 ohm. At 300 kilocycles this corresponds to a change of approximately 1 micromicrofarad. A cage antenna, or even a steel tower cannot be expected to remain constant within these limits and in consequence some means of preventing a change in space pattern is highly desirable.

III. EXCITATION SYSTEMS

It is the practice at the present time to employ some form of transmission line to transfer the power from the transmitter to the antenna system. The two-wire constant current line has certain advantages in that the radiated field is small and the length of line may be chosen to suit the physical conditions of the installation. A schematic circuit of one form of excitation is shown in Fig. 3(a). By adjustment of the
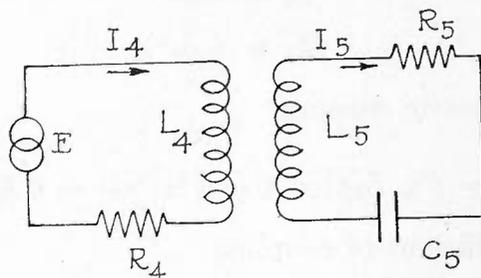

Fig. 4—Coupling transformer, equivalent circuit.

various circuit constants the antenna is suitably matched to the transmitting set for optimum power transfer. Change of any one of several constants will disturb this relationship and create a change of relative phase and magnitude of the antenna current with respect to the correct value. The first step is obviously to reduce the number of variables to a minimum.

Coupling transformers employed in radio work are customarily adjusted for the condition $\omega^2 M^2 = R_1 R_2$, under which maximum power transfer is secured. There is also, however, the possibility of using values of $M \gg \sqrt{R_1 R_2}/\omega$ and securing high efficiency. For example in Fig. 4 is shown a coupling transformer used to excite a load R_5 from a source of power. Now if the $\omega L/R$ of each coil is large compared with the load impedance and if the coupling coefficient is high, a transformer will act purely as an impedance-matching device in which the ratio of transformation (turns ratio) is equal to the square root of the ratio of impedances to be matched. For example, consider the circuit of Fig. 4. We may write the following relations:

$$E = Z_4 I_4 + Z_{45} I_5$$
$$O = Z_{45} I_4 + Z_5 I_5$$

whence,

$$I_4 = \frac{Z_5 E}{Z_4 Z_5 - Z_{45}^2}$$

$$I_5 = -\frac{Z_{45} E}{Z_4 Z_5 - Z_{45}^2}$$

$$\frac{I_5}{I_4} = -\frac{Z_{45}}{Z_5} = -\frac{j\omega M_{45}}{R_5 + j\omega L_5}$$

but,

$$\omega L_5 \gg R_5$$

$$L_4 = K n_4^2$$

$$L_5 = K n_5^2.$$

Where K is a geometric constant

$$M_{45} = C\sqrt{L_4 L_5} = C\sqrt{K^2 n_4^2 n_5^2} = C K n_4 n_5;$$

where C is the coefficient of coupling.

$$\frac{I_5}{I_4} = -\frac{j\omega M_{45}}{j\omega L_5}$$

$$= -\frac{M_{45}}{L_5} = -\frac{C K n_4 n_5}{K n_5^2} \cong -\frac{n_4}{n_5} \ (\text{as } C \cong 1).$$

Under these conditions, therefore, the current in the secondary is related to that in the primary solely by means of the turns ratio. Furthermore,

$$I_4 = \frac{(R_5 + jX_5)E}{(R_4 + jX_4)(R_5 + jX_5) + X_{45}^2}$$

$$= \frac{E}{(R_4 + jX_4) + \dfrac{X_{45}^2}{(R_5 + jX_5)}}$$

$$= \frac{E}{(R_4 + jX_4) + \dfrac{X_{45}^2}{R_5^2 + X_5^2}(R_5 - JX_5)}$$

$$= \frac{E}{(R_4 + \phi R_5) + j(X_4 - \phi X_5)}$$

where,

$$\phi = \frac{X_{45}^2}{R_5^2 + X_5^2} \quad \text{and since } X_5^2 \gg R_5^2$$

$$= \frac{X_{45}^2}{X_5^2}$$

$$= \frac{M_{45}^2}{L_5^2} = \frac{(CKn_4n_5)^2}{(Kn_5^2)^2} = \frac{C^2 n_4^2}{n_5^2} \cong \frac{n_4^2}{n_5^2}.$$

The impedances therefore are reflected through the circuit by a constant proportional to the square of the turns ratio. Now should the coefficient of coupling be unity ($C = 1$)

$$\phi X_5 = \frac{X_{45}^2}{X_5^2} \times X_5$$

$$= \frac{X_{45}^2}{X_5} \cong \frac{K^2 n_4^2 n_5^2}{K n_5^2}$$

$$= K n_4^2 = X_4$$

and,

$$j(X_4 - \phi X_5) = j(X_4 - X_4) = 0.$$

Consequently, for unity coupling the inductances do not affect the circuit. Should the coupling be less than unity there will remain some of the inductive terms which can be removed by making X_5 (or X_4) capacitative to the amount of such discrepancy. For high coefficients of coupling (90 per cent is secured on the coils in actual use) the ratio of transformation is not altered greatly and is compensated for by the method of tuning. This analysis neglects the effect of capacity coupling between circuits. For the lines in use, the voltages and frequencies were low enough so that the capacity could be neglected.

On the basis of the foregoing analysis, the coupling transformers in Fig. 3(a) may be replaced for purposes of this treatment by the circuit of 3(b), or more generally as 3(c).

Having fixed the relationships between primary and secondary transformer currents, the relation between the voltage of the source E and the current in the antenna is now independent of the transformer adjustments so long as the conditions specified in the mathematical treatment are fulfilled. We may now write the expression for the current in the load in terms of the voltage at the sending end of the line and the remaining variable parameters as follows:

$$\frac{E_s}{I_R} = Z_{TR} = Z_R \cosh \gamma l + Z_0 \sinh \gamma l$$

where γ is the propagation constant of the line per unit length, Z_0 is the surge impedance, and Z_R the load impedance.

Now, in general, a change of Z_R will result in a change of Z_{TR}. However, if $\cosh \gamma l = 0$, Z_{TR} is independent of the load, and the current-voltage relation is determined solely by Z_0. For a line with negligible attenuation $\cosh \gamma l \cong \cos \beta l$, a circular function which is zero for $\beta l =$ any odd multiple of $\pi/4$. Consequently, if the line is 90 electri-

Fig. 5—Series connection of transmission lines.
a. Actual circuit.   b. Equivalent circuit.

cal degrees long and has a low attenuation, $Z_{TR} = jZ_0$ and the load current will bear a constant relation to the sending voltage.

It is apparent that any number of lines may be connected to the source of voltage E_s. If the foregoing conditions are met, there will then be a constant relation with regard to phase and magnitude between the currents in all the antennas or other loads so connected, regardless of the value of the various load impedances. In this manner the two antennas of a pair in the range beacon array may be kept in absolute synchronism.

There is another method of attack, productive of the same results, which is advantageous in many cases. A loop antenna for example, is free from pattern distortion when detuned slightly, since the radiating elements form a continuous series circuit.

The transmission lines may also be connected in series as shown in Fig. 5(a). In this case the three winding transformer, or hybrid coil,

is used merely to balance the voltage to ground and prevent the flow of undesired earth currents. For purposes of analysis it may be considered as shown in Fig. 5(b). For such a circuit the current is given by the equation[4]

$$I_{s1} = \frac{E_g[Z_0 \cosh \gamma l + Z_{R1} \sinh \gamma l]}{Z_0(Z_{R1} + Z_{s2} + Z_g) \cosh \gamma l + [Z_0{}^2 + Z_{R1}(R_{s2} + Z_g) \sinh \gamma l}$$

where,

Z_0 is the surge impedance of the line

Z_{s2} is the input impedance of line 2

Z_g is the generator impedance.

Assume antenna 2 to be properly tuned. Then,

$$Z_{s2} = R_{s2}, \ Z_g = R_g, \text{ and } (R_g + R_{s2}) = Z_s \underline{|0°}.$$

We may now write

$$I_{R_1} = \frac{E_g}{(Z_R + Z_s) \cosh \gamma l + \left(Z_0 + \dfrac{Z_R Z_s}{Z_0} \right) \sinh \gamma l}$$

$$= \frac{E_g Z_0}{Z_0(Z_R + Z_s) \cosh \gamma l + (Z_0{}^2 + Z_R Z_s) \sinh \gamma l}$$

and since $I_{s1} = I_{s2} = I_g$

$$\frac{I_s}{I_{R_1}} = \frac{Z_0 \cosh \gamma l + Z_{R_1} \sinh \gamma l}{Z_0}$$

$$\frac{I_s}{I_{R_1}} = \cosh \gamma l + \frac{Z_{R_1}}{Z_0} \sinh \gamma l.$$

For the line which is properly terminated

$$\frac{I_s}{I_{R_2}} = \cosh \gamma l + \sinh \gamma l.$$

Hence,

$$\frac{I_{R_1}}{I_{R_2}} = \frac{\cosh \gamma l + \sinh \gamma l}{\cosh \gamma l + \dfrac{Z_{R_1}}{Z_0} \sinh \gamma l} .$$

[4] See treatment in Everitt, "Communication Engineering," McGraw-Hill Co., (1932).

That is to say, the phase and magnitude of I_{R_1} will change with respect to I_{R_2} as antenna 1 is detuned. Now if sinh $\gamma l \ll$ cosh γl this change will be negligible. This means a very short line (a simple series circuit) or again if the attenuation is small, any line which is a multiple of $\pi/2$ in length, since with that condition sinh $\gamma l = j$ sin $\beta l = 0$. From
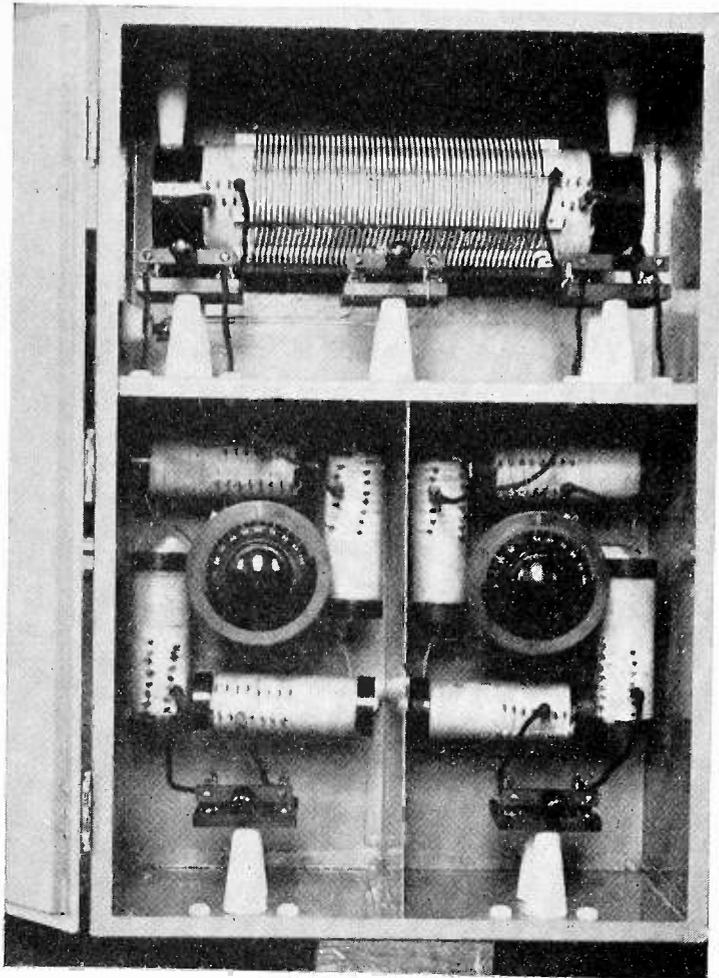
Fig. 6—Exciting transformer and building-out sections used at Station WKRC.

this it is evident that the series connection will also serve to maintain synchronism where the proper conditions are satisfied.

One of the advantages of the constant current transmission line already mentioned was the ability to use any desired length to suit the physical arrangement. If it is necessary to use quarter-wave or half-wave lines this feature is lost since the antennas are rarely spaced the desired amount, especially at the longer wavelengths. Where open wire lines are used, this would mean routing the line circuitously until the desired length was obtained. To avoid this, recourse was had to arti-

ficial lines or building-out sections. The line proper was extended from transmitter to antenna and its electrical length determined at the frequency to be used. An artificial line of identical surge impedance and of sufficient length to make up the deficiency was then added at the transmitter end of the actual line. This introduced no discontinuity and the lines functioned as well as those which were continuous. The use of building-out sections also adds flexibility to the system, since should a change of frequency be necessary, the section may be adjusted to secure synchronization at the new value without the necessity of rebuilding the actual line.

In Fig. 6 is illustrated the exciting transformer and building-out sections for two antennas as used at a radio broadcast station. The
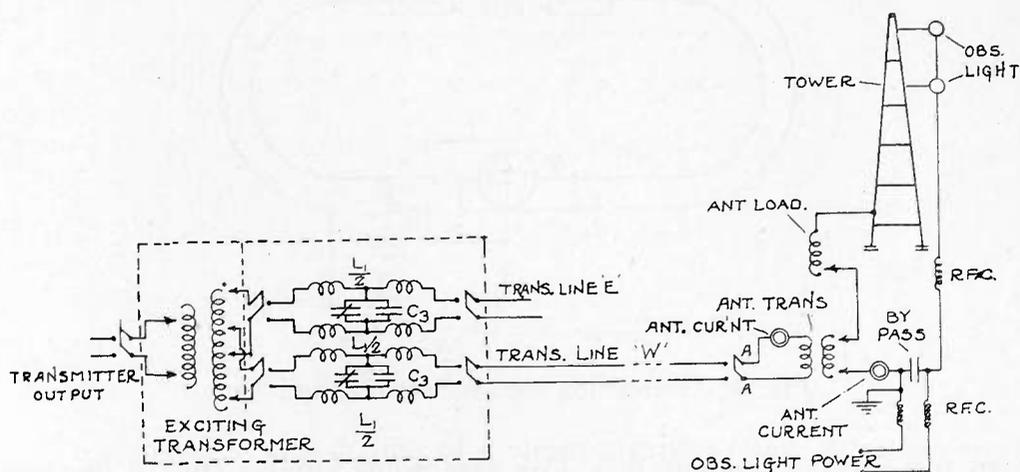
Fig. 7—Circuit diagram of Fig. 6.

circuit diagram is given in Fig. 7. The building-out sections are adjustable, by means of taps on the coils and the variable condenser, to the desired length. Taps on the primary of the exciting transformer permit matching the load to the transmitter output impedance while those on the secondary permit adjustment of the relative power delivered to each antenna. This particular unit is used to excite two towers 180 degrees out of phase, thereby maintaining a figure-of-eight pattern in space. One antenna is provided with slightly more current than the other to eliminate roughness in modulation near the points of minimum signal. The compact nature of the arrangement is well indicated by the illustration.

IV. EXPERIMENTAL WORK

1. Laboratory Tests.

In order to determine the practicability of those methods of excitation, a series of laboratory tests were conducted, prior to making any

changes on the actual antenna structure. This procedure had the advantage of permitting accurate control of the circuit elements. The transmission line used for the tests consisted of two number 14 rubber-covered wires enclosed in a lead sheath. The characteristics of this line are given in the appendix. Two sections of this line, each 120 feet long, were connected in series as shown in Fig. 8. The ends were terminated in an adjustable impedance and the plates of a cathode ray oscillograph connected across the resistance components of the load in each case. So long as the currents in the loads were in phase, the oscillograph pat-
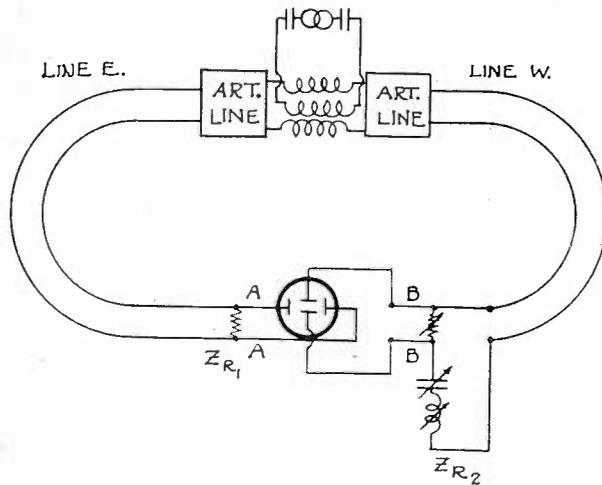
Fig. 8—Connection used for oscillograph tests.

tern remained a straight line. For deviations greater than 5 degrees, the line opened up into an elliptic shape. The accuracy of this test was not sufficient to be conclusive but it provided a means of securing qualitative data.

For the lines alone (no building-out sections), any change in load from a pure resistance changed the phase decidedly. This corresponded to two 40-degree sections. They were then built out to 180 degrees for each line by means of the artificial sections. The terminal impedance Z_R could now be varied from the correct value of $75 \overline{|0°}$ to $120 |33°$ and $120 |\overline{10°}$ before a shift of 5 degrees was observed. Variations beyond those limits caused the figure to open up. The fact that synchronization was limited was ascribed to the relatively high loss of this type of line and also to the presence of excessive sheath currents in the cable when standing waves were formed.

The lines were next connected in parallel and built out to 90 degrees. Observation on this connection showed that the synchronizing action was somewhat better in this case.

The experimental results showing that, in part at least, the theory

was conformed to, it was decided to apply the same arrangements to the actual antennas.

2. *Tests on the Antenna System.*

The antenna system available for these tests consisted of a pair of vertical antennas each 40 feet high and separated 220 feet. Coupling units were located at the base of each antenna to resonate them to the desired frequency (290 kilocycles). Because of the excessive loading required for this frequency, the antennas were especially unstable and furnished an excellent opportunity to test the method.

The coupling unit and building-out sections were similar to those shown in Fig. 6. The lines being lead encased were laid along the ground.
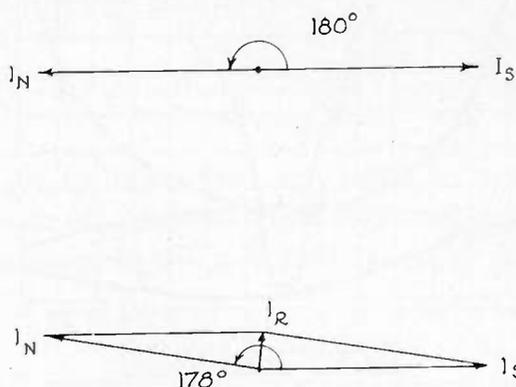
Fig. 9—Phase relations in test antenna.

To determine the degree of synchronization, two antennas were excited, and the voltage induced in a third antenna, equidistant from the other two, was measured by means of a milliammeter inserted in its base. If the phase is correct and the currents in the antennas are equal there will be no current apparent in the third antenna. However, a slight unbalance of magnitude or phase will cause the milliammeter to indicate. For example, if each antenna independently (I_n and I_s in Fig. 9) causes 100 milliamperes of current to flow in the test antenna, then a phase difference of 2 degrees from the true 180-degree relationship will cause a current of 3 milliamperes when both antennas are excited (I_R in Fig. 9). These currents could easily be read on the meter employed, and since 2 degrees is the variation which is chosen for the maximum allowable deviation, the method of observation was adequate. Previous experience with this method of adjustment had also shown it to be satisfactory.[3]

3. *Tests on Parallel Connection.*

With the antennas connected in parallel as shown in Fig. 3, a series of tests were made with various line lengths. The antennas were first

tuned to resonance individually. Both antennas were then excited, and
the current in the third, or test antenna, was measured as various
known amounts of reactance were inserted at the base of one of the
antennas of the array. In the graphs this reactance is expressed in
terms of per cent of total antenna resistance. Fig. 10 gives a compari-
son of stability between the 40-degree lines and the 90-degree lines. As
is evident from the graphs, in the case of the 40-degree lines, a small
amount of reactance in one antenna (less than 2 per cent) would shift
the phase sufficiently to alter seriously the space pattern. This was ac-
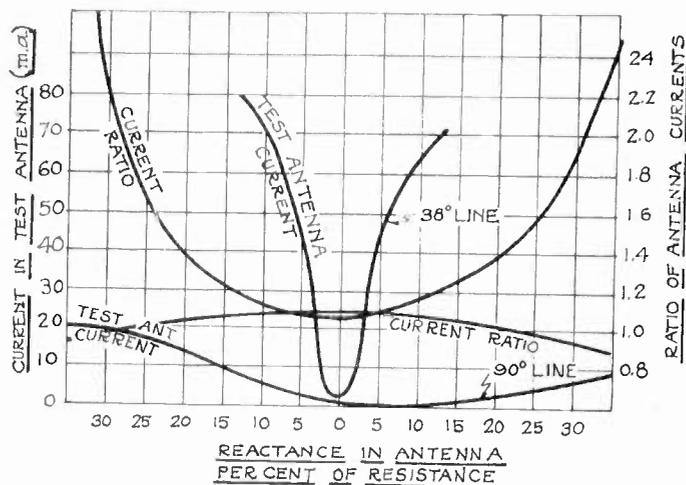
Fig. 10—Effect of building out-sections on phase stability,
(lines in parallel).

companied by a rise in current in the antenna still in tune and a drop
in the untuned antenna. The current ratio curve for the 40-degree line
shows this condition. For the 90-degree lines it was possible to intro-
duce 15 per cent of reactance before the allowable deviation of 2 de-
grees was exceeded. The antenna currents in this latter case rose and
fell together, keeping an almost constant ratio. Another interesting
comparison is the effect of incremental changes in reactance. For ex-
ample with the 40-degree lines, as the antennas swing in the wind, the
phase shifted over a large range of values. Using the 90-degree lines, the
wind had no effect on the phase as indicated by the constancy of cur-
rent in the test antenna, and furthermore even when the antenna was
detuned sufficiently to cause relatively large currents in the test an-
tenna, these currents were constant, although the antennas continued
to sway with the wind. The stability is therefore even more pronounced
than is at first indicated on the graphs.

In order to determine the optimum line length experimentally, a
series of tests were made for various line lengths from 79 to 101 de-
grees and the amount of detuning permissible for each length was de-

termined. This is shown in Fig. 11. Although the curve shows a decided peak in the vicinity of 90 degrees it is evident that any value of
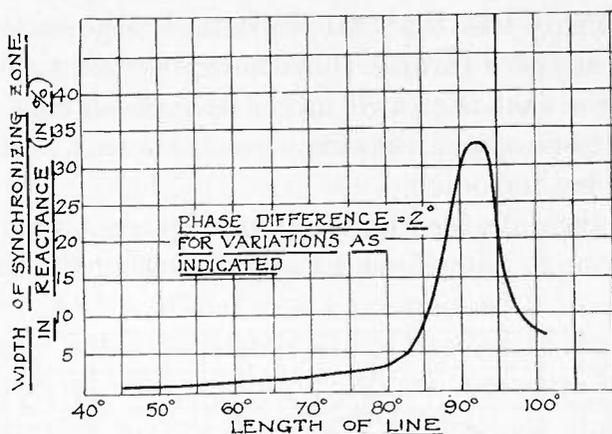
Fig. 11—Synchronizing effect for various line lengths,
(lines in parallel).

line length from 88 to 94 degrees will result in a decided synchronizing action. Consequently it is not necessary to determine the line length to laboratory accuracy. It will be noted also that the peak of the curve
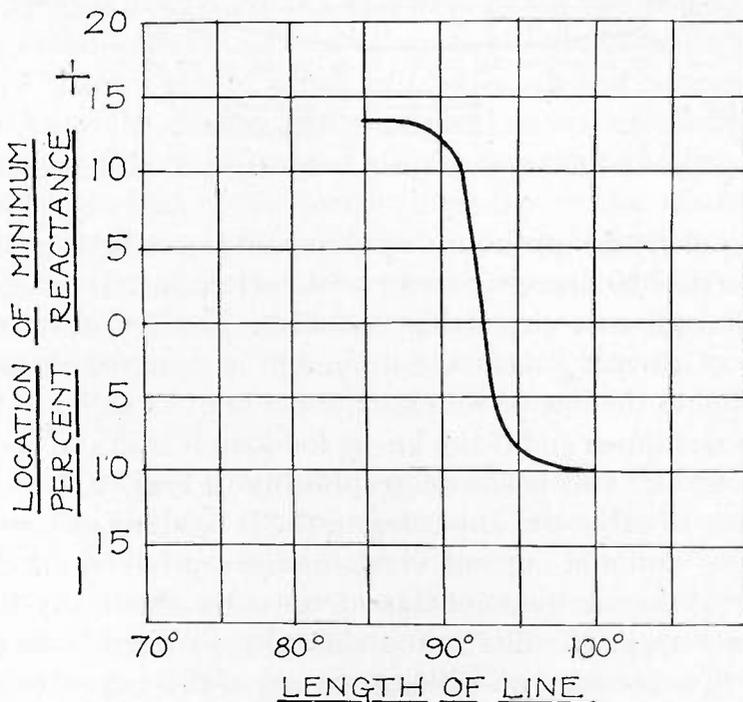

Fig. 12—Location of minimum vs. line length, (lines in parallel).

is at 92 instead of 90 degrees. The location of this peak is a function of the constants of the coupling transformer. As the coupling transformer approaches the perfect transformer, that is, the $\omega L/R$ of the primary

and secondary windings approaches infinity, the line length approaches 90 degrees. The correct length for any transformer exceeds 90 degrees by the same number of degrees that arc tan $\omega L/R$ of the primary winding is less than 90 degrees. In the case of the transformers finally adopted for use, this discrepancy amounts to 4 degrees. In any event, the fact that a region of several degrees width exists in which there is good synchronization prevents this additional feature from becoming bothersome.

In securing the data for Fig. 11, another phenomenon was observed which enables one to determine instantly whether the line in use is too
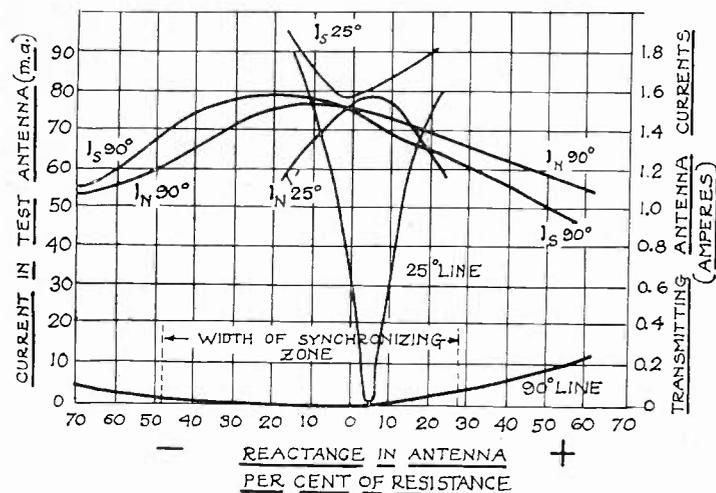
Fig. 13—Effect of low-loss lines on synchronization.

long or too short for optimum results. For any length of line in the neighborhood of 90 degrees the curve of current in test antenna versus reactance in antenna circuit has a definite location of its minimum. For a line of correct length the minimum is centered about zero reactance. Should the line be too short, the minimum shifts to the region of positive reactance and if the line is too long it shifts to the negative reactance region. This is shown graphically in Fig. 12. If it is desired to determine whether the line being tested is too long or too short, the curve of test antenna current versus reactance in the antenna is secured. Should the minimum of this curve center about, say 10 per cent positive reactance, the line is too short by $1\frac{1}{2}$ degrees as shown by Fig. 12. It thus becomes possible by the use of this curve, to secure the maximum amount of synchronization in every case where such accuracy is desired.

In making adjustments to secure the correct length of each transmission line, the one requirement which should be met in the case of all installations except the visual radio range beacon, is that both lines

have the same length. It is not advisable to distort the pattern by using lines of different length. Where a distorted pattern is desired a method to be described later is preferable.

All the foregoing data were secured with the type of line previously mentioned which had high losses. This was done as it was easier to work with mechanically. After the correct values were determined, a test was made on lines with low attenuation, (parkway cable). The results are shown in Fig. 13. This is the type of line used on the airways range beacon installation, and the wide range of synchronizing action makes the system comparable in stability to the loop antennas. A change of tuning of ± 40 degrees is possible with these lines, which is in excess of any detuning action which has been found to occur in an actual installation. The values of antenna currents in the north and south antennas are also plotted in this graph for various amounts of detuning. It will be noted that when synchronization is secured, the currents rise and fall together. When no synchronization is present the reverse is true, making the space pattern depart even further from the desired shape. The best rough check on the action of the system is observation of these currents. This may readily be done by the one making the installation and is a check which may be relied upon.

Tests to determine the limits to which the design must be held were also made. Using a line of which half its length had a surge impedance of 75 ohms and the other half 55 ohms had no apparent effect upon the synchronizing action. Mismatch of the antenna to the line within plus or minus 15 per cent of the correct impedance value made no measurable difference. From these results, it is safe to say that the system is decidedly noncritical in its action, and consequently installation may be made in the field under adverse conditions with reasonable certainty of successful operation.

4. Tests on Series Connection.

There are certain conditions under which the series connection is desirable. One of these occurs where the pattern is to be deformed. This will be dealt with in a later paragraph. Another instance occurs when it is desired to transmit nondirectional signals as well as directional signals on the same antenna group (the simultaneous radiophone and visual range beacon, for example). Consequently a series of tests were made on the antennas, similar to those already carried out for the parallel connection. It was found that the action was in accord with the theory where allowance was made for the relatively high attenuation of the lines used for the test. In addition to the actual attenuation acting to weaken the synchronizing effect, the presence of

sheath currents of relatively large magnitude also served to cause departure from complete control. The action of the half-wave lines is effectively to replace the transmission line with two connecting wires of zero length, and should any excessive flow of current occur from these wires to earth the substitution can only be partial. For series operation, therefore, the open-wire type of line is to be preferred.

The results of this test showed that synchronization was secured through a variation of 12.5 per cent reactance in the antenna circuit.
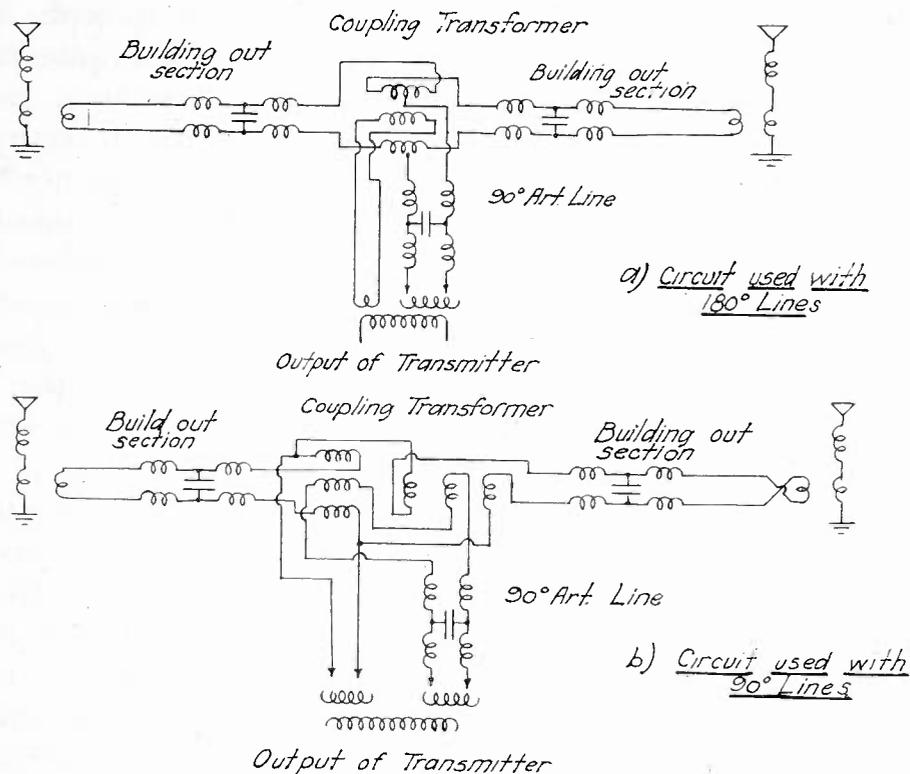
Fig. 14—Means for producing distorted space patterns.
a. Lines built out to 180 degrees.
b. Lines built out to 90 degrees.

This is somewhat less than that secured with lines of similar construction connected in parallel. It was sufficient, however, to make its use quite feasible even with high loss lines. Open-wire lines would increase this range by a factor of from four to five.

On airway installations it was decided to adopt the 90-degree parallel condition as standard, it being simpler as regards exciting means and requiring less building out.

5. *Distortion of Pattern.*

Since it is possible to secure a greatly distorted pattern with a very few degrees of phase shift in the antennas of a group, the obvious means of distorting the pattern is to employ lines of different length to the individual antennas, in which the net phase difference will pro-

duce the desired pattern. However, except for continuous-wave teleg-
raphy or the visual range beacon, this system is not recommended.
The side bands in a broadcast transmitter are sufficiently separated
from the carrier frequency to return to the sending end in various
time phase relations and the resultant pattern is continuously shifting.
When it is desired to distort the pattern, a circuit such as that of Fig. 14
should be employed.

In this circuit, radio-frequency energy is supplied to the tower by
two separate paths, recombination occurring in the hybrid coil. On
inspection of the circuit it will be seen that one path excites the an-
tennas in phase or nondirectionally and the other out of phase to pro-
duce a figure of eight. By suitably adjusting the relative phase and
magnitude of the two currents it is possible to secure any pattern from
a circular shape, through a cardioid, to a figure of eight. Where only
slight distortion is required, as is usually the case, the circuit is so de-
signed that the figure of eight is synchronized and the circle is not. Us-
ing this arrangement any change of tuning will have very slight effect
upon the pattern.

If the 90-degree lines are used, the figure-of-eight radiation is sup-
plied through the mid-tap of the hybrid coil, for 180-degree lines it is
supplied through the tertiary winding. In the former case it is neces-
sary that the individual parts of the hybrid coil be separated in so far
as their electromagnetic field is concerned. Otherwise a drop of current
in antenna W causing a rise in current in the sending end of its associ-
ated line will force a rise in current in the sending end of the E line by
means of its close coupling with the other half of the coil. If the coil is
split into two sections as in Fig. 14(b) synchronization can be main-
tained for correct adjustment of the tuning circuits.

When using 180-degree lines this precaution is not necessary and
for that reason 180-degree lines are recommended for this use. The cir-
cuit of Fig. 14(a) may also be used where it is desired to employ the
antenna directionally part of the time and nondirectionally the re-
mainder. By choosing the proper channel for supply of the radio fre-
quency the desired transmission may be accomplished. A simple
change-over switch is all that is required for such service, and since a
balanced hybrid coil makes the two circuits independent, no difficulties
are encountered in its application. This circuit may be especially use-
ful where it is desired to minimize interference during certain hours of
the day and operate with maximum coverage during the remainder.

6. Tuning Methods.

Almost any uniform tuning procedure may be adopted to adjust the
system for correct operation. Since it is noncritical, a slight misadjust-

ment will not affect its performance. The method which has been found to be most satisfactory is given here. It requires an impedance measuring device and a noninductive resistor of which the magnitude may be made equal to the surge impedance of the transmission line in use.

First determine the surge impedance of the line. This is used as the basis of all the future computations and must be reasonably accurate. By using the impedance measuring set, measure the line impedance open-circuited and also short-circuited. The surge impedance is then given by the relation

$$Z_0 = \sqrt{Z_{0c} Z_{sc}}.$$
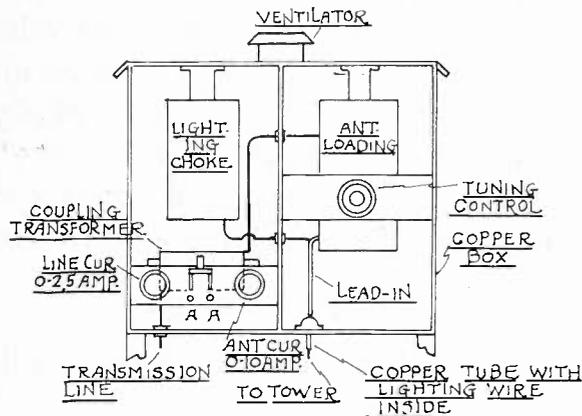
Fig. 15—Antenna tuning units.

The phase angle of the line may be computed from the formula

$$\tanh \phi = \sqrt{\frac{Z_{sc}}{Z_{0c}}}.$$

If the line has relatively low loss it is sufficiently accurate to use the equation

$$\tan \phi = \sqrt{\frac{Z_{sc}}{Z_{0c}}}.$$

In each case ϕ is the angle of the line.

Second, adjust the building-out sections. Knowing the angle of the line, determine the angle of the building-out section. This is 94 degrees $-\phi$ for the 90-degree lines or 184 degrees $-\phi$ for the 180-degree lines. The values of inductance and capacity necessary may then be computed from the usual filter circuit design formulas.

By means of the impedance bridge, the coils and condensers are adjusted to their computed values. Now terminate one section in its surge impedance with the noninductive resistor and measure its input impedance. If it is not a pure resistance adjust the condenser arm of the section until it is. Repeat this for each section.

Third, tune each antenna. A typical antenna tuning unit is illustrated in Fig. 15. Disconnect the line from the coupling transformer and connect the impedance bridge thereto. Next tune the antenna to resonance with the antenna loading inductance. If the resultant resistance at the input to the coupling transformer is not equal to the line impedance, adjust the transformer ratio until the correct value is secured. It will probably be necessary to make slight tuning adjustments each time the taps are changed on the transformer.
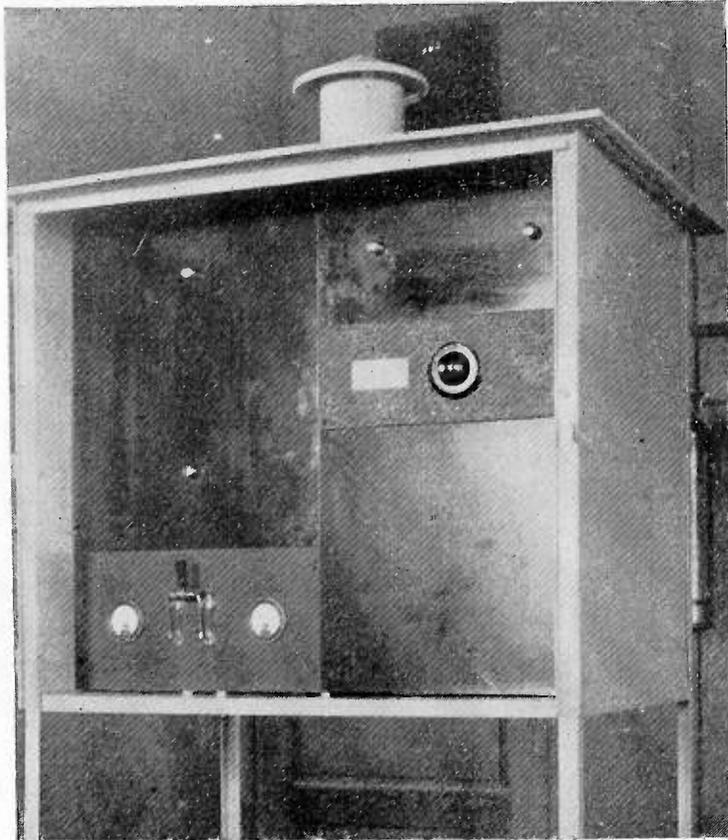
Fig. 16—Antenna tuning units.

Fourth, when the antennas are properly tuned, connect the lines to the tuning unit and measure the input impedance at each building-out section. It should equal the surge impedance of the line if all adjustments have been made correctly.

Fifth, connect the sections to the exciting transformer and make such adjustments at the transmitter as are necessary to excite the load properly.

The tuning of any antenna may be checked within the station at any time by opening the circuit of its respective building-out section and using the impedance bridge.

The operation of the system as a whole may be observed by detuning one antenna and observing the action of the currents in the other antennas. If proper phase stability is obtained, the currents will rise and fall together over a wide range of tuning variation.

V. Conclusions

The method of excitation described in this paper has been in use for a considerable period of time on the new airways radio range beacon stations. It provides a stable space pattern, free from the shifts previously caused by weather conditions. The fact that it is not difficult to calculate and install makes it a simple problem to secure the desired space pattern. Its application to broadcast station requirements is becoming of increasing importance and several stations have adopted it as a means for insuring the type of pattern required for the Federal Radio Commission. As the tendency toward directive antennas becomes more pronounced, it should find a wide field of usefulness on all frequency channels.

VI. Acknowledgment

This research forms a part of a general research program conducted by the author as a doctorate thesis for the Massachusetts Institute of Technology. Copies of the complete thesis are on file in the Institute Library.

The author wishes to acknowledge the assistance rendered by the Massachusetts Institute of Technology in providing the facilities of their Round Hill Research Laboratories where the research on this project was begun. The coöperation of the Round Hill staff greatly facilitated the solution of the problem.

The work was carried to completion and applied to the radio range beacon at the Experimental Laboratory of the Bureau of Standards at College Park, Maryland.[5]

Application of this system to radio broadcast stations was made at the laboratories of the Washington Institute of Technology, at College Park. The illustrations of broadcast station equipment were furnished by this last-named firm.

VII. Appendix

Transmission line constants measured at 290 kilocycles.
1. Two-wire No. 14 rubber-covered lead sheath:

[5] F. G. Kear, "Phase synchronization in directive antenna arrays with particular application to the radio range beacon." *B. S. J. R.*, vol. 11, p. 123, (1933); (Abstract) Proc. I.R.E., vol. 22, p. 116; January, (1934).

Surge-impedance $Z_0 = 53.5 \, \underline{|0°}$ ohm.

$\gamma = 0.00138 + j0.00577$ per foot of cable.

$\alpha = 0.138$ per 100 feet.

$\beta = 33°$ per 100 feet.

Phase velocity, 59,400 miles per second.

Attenuation, 3.76 db per 100 feet.

2. Parkway cable No. 12 rubber-covered wire, lead and steel sheaths:

Surge impedance $Z_0 = 69 \, \underline{|0°}$.

$\gamma = 0.00014 + j0.00362$.

$\alpha = 0.014$ per 100 feet.

$\beta = 20.7°$ per 100 feet.

Phase velocity, 92,800 miles per second.

SEVENTY-FIVE-CENTIMETER RADIO COMMUNICATION TESTS*

By

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(Signal Corps Laboratories, Fort Monmouth, N. J.)

Summary—A directional 75-centimeter Barkhausen-Kurz transmitting and receiving equipment for both telephone and code work has been developed. A simplified theory of the mechanism of oscillation is given. The range of 88 miles over water reported exceeded the distance from the transmitter to the horizon by a factor of five.

INTRODUCTION

SIGNALING by means of radio waves less than a meter in length offers advantages in some cases over methods employing longer waves. The advantages arise primarily from the smallness of the antenna array needed for directional work. This paper describes some experiments carried on at the Signal Corps Laboratories with a Barkhausen-Kurz equipment operating at a wavelength of 75 centimeters. A summary of previous work in this field has been given by Megaw[1] together with an extended bibliography, hence only a limited number of references are cited in this paper.

A diversity of opinions as to the actual mechanism of the Barkhausen-Kurz oscillations seems to be held.[2] The early view that the frequency of oscillation approximates the frequency of an electron oscillating through the meshes of the grid and acted on by electric forces depending on tube geometry and the steady potential differences applied to the tube electrodes has the merit of predicting quite satisfactorily the wavelength obtained in practice. However, it scarcely tells us why to expect oscillations in the first place nor the mechanism for the transfer of energy from the high voltage battery into high-frequency electromagnetic energy measured in a transmission line or radiated in an antenna. On the other hand, the contention that oscillations arise because of instability in the space charge found in the interlectrode space[3] does not appear to constitute a satisfactory solution to the problem. Benham[4] reduces the oscillator to an equivalent

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* Decimal classification: R423.5. Original manuscript received by the Institute, December 19, 1933. Published with permission of the War Department.
    [1] Megaw, *Jour. I.E.E.* (London), vol. 72, p. 313, (1933).
    [2] See in particular the discussion accompanying the paper by Megaw, footnote 1.
    [3] Tonks, *Phys. Rev.*, vol. 30, p. 501, (1927).
    [4] Benham, *Phil. Mag.*, vol. 11, p. 457, (1931).

shunt circuit in which one branch consists of an inductance in series with a resistance while the second branch consists of a capacity in series with a resistance. The writer prefers the simple view that oscillations are maintained in the steady state because the phase difference between the alternating component of the voltage on the tube electrodes and the alternating component of the electron current flowing in the interlectrode space is greater than ninety degrees for certain frequency ranges. This phase difference arises from electron inertia. According to this view, the sole function of the electron stream is to offer a negative resistance over certain ranges of frequency, thereby permitting oscillations once set up in the circuit, made up perhaps of the tube electrodes and leads only, or of the electrodes coupled to a
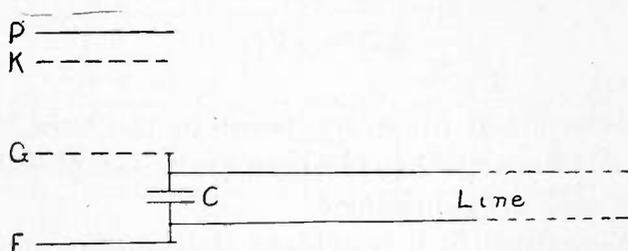
Fig. 1—Simplified Barkhausen-Kurz oscillator.

transmission line, to be maintained with constant amplitude. Losses in the circuit are made up by the negative resistance of the electron stream. An initial surge or impulse is needed to start oscillations. In its simplest form the analysis is as follows:

Consider a tube having plane electrodes as in Fig. 1 with filament F, grid G, plate P, and a virtual cathode whose position for symmetry may be taken as K. A transmission line is connected to F and G. The filament-grid capacity may be dealt with by considering it as a capacitative shunt C across the input to the line and its effect on line length then calculated.[5]

Let the alternating voltage on the input to the line be represented by

$$e = E_1 \epsilon^{j\omega t} \tag{1}$$

and let the resulting alternating component of electron current be represented by

$$i = I_1 \epsilon \mu^{j\omega(t-\delta)}. \tag{2}$$

Then,

$$z = \frac{e}{i} = \frac{E_1}{I_1} \epsilon^{j\omega\delta} = Z_1 \epsilon^{j\omega\delta} \tag{3}$$

[5] Hund, Bureau of Standards Scientific Paper No. 491, June 23, 1924.

where δ is a phase difference introduced by electron inertia. The real part of z, represented by r, is $Z_1 \cos \omega\delta$.

$$r \text{ is positive when } -\pi/2 < \omega\delta < \pi/2. \tag{4}$$

$$r \text{ is negative when } \pi/2 < \omega\delta < 3\pi/2. \tag{5}$$

The magnitude of δ is determined by tube geometry, space charge, E_g, E_p, and E_1 which enter implicity in the equation

$$\delta = \int_0^{x_0} \frac{dx}{v} \tag{6}$$

where v in turn is defined by the equation

$$\frac{1}{2} mv^2 = eV(x) \tag{7}$$

and must be determined for every point in the tube. x is distance measured from F where $x = O$ to G where $x = x_0$. Graphical methods are most commonly used in evaluating δ.

From the inequality (5) it is evident that one frequency range for which r is negative is given by

$$4\delta > \tau > 1.33\delta \tag{8}$$

where τ is the period of oscillation.

In short, oscillations may be expected in an electrode and lead system tuned to certain frequencies in this range, or in a properly connected transmission line whose fundamental or one of whose overtones lies in this frequency range. In fact, a critically tuned line either oscillates in two modes simultaneously or these modes are separately excited in extremely rapid succession as is shown by photographs of the discharge taking place when such a line is placed in a partial vacuum.[6]

APPARATUS

The generator of short-wave oscillations here used is of the Barkhausen-Kurz type and has been described previously by Carrara[7] and Kozanowski.[8] Two type 552 tubes are used in a push-pull circuit. The grids are held at a potential 500 volts above that of the filaments while the plates are held 90 volts negative with respect to filament. With this circuit it is possible to set up radio-frequency currents of 2.5 amperes in the associated Lecher wires, and potential differences in excess of

[6] Hershberger, Zahl, and Golay, *Physics*, vol. 4, p. 291, (1933).
[7] Carrara, *Elettrotecnia*, vol. 18, p. 874, (1931).
[8] Kozanowski, PROC. I.R.E., vol. 20, p. 957; June, (1932).

several hundreds of volts across the voltage loops while five watts of power are thereby available.

The curves of Fig. 2 show the effect of varying the plate bias on os-
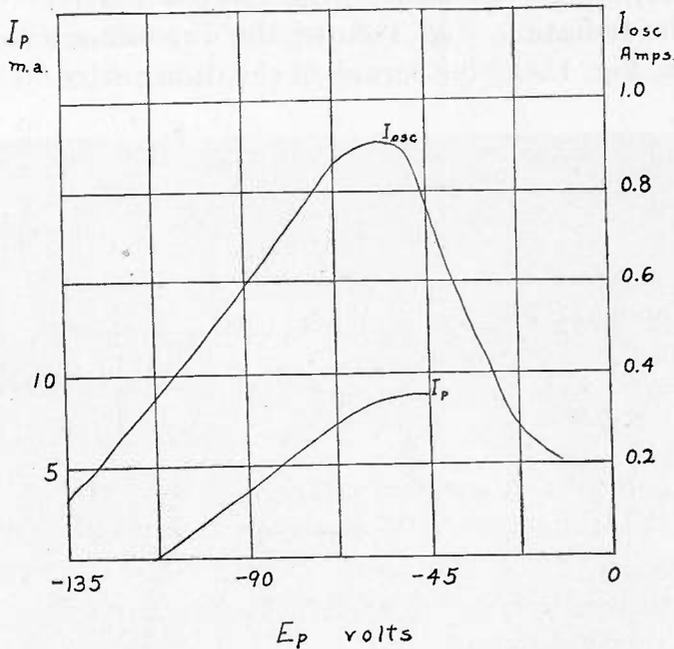
Fig. 2—Oscillating current and plate current shown as a function of the plate bias.

cillating current and on plate current when the constant grid voltage was 500 volts and the grids together drew a current of 450 milliamperes. These curves were used as a basis for calculations on the design of a
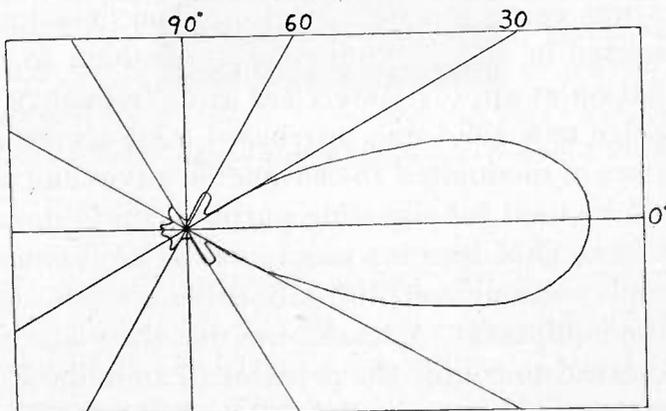
Fig. 3—Directional characteristics of the Yagi antenna.

modulating system. The negative bias actually used was 90 volts while the amplitude of the modulating voltage which was impressed on the plate only was 25 volts. A steady 1000-cycle note was used in making

all measurements on received signal strength although provisions were also made for voice modulation.

A Yagi beam antenna employing one dipole connected directly to the filament Lecher wires, six directors, and one reflector was used to concentrate the radiation. Fig. 3 shows the directional characteristics of this antenna. Fig. 4 is a photograph of the transmitter.
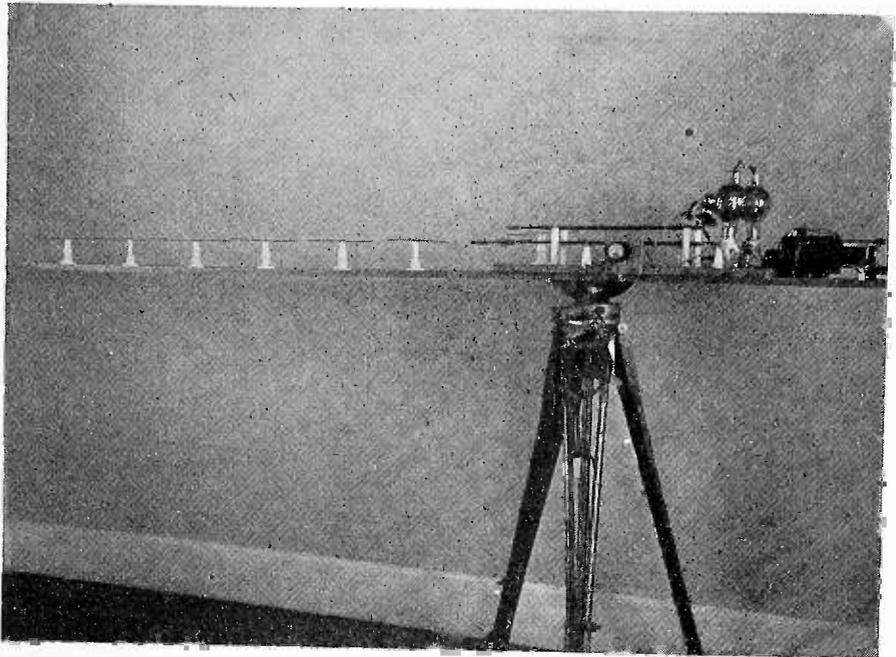
Fig. 4—The transmitter.

An antenna similar to that used with the transmitter but with only two directors was employed for reception. The detector tube was a UV-199 connected in the Barkhausen-Kurz fashion so it was on the verge of oscillation at approximately the same frequency as the transmitter. Of twelve new 199 tubes purchased, nine served satisfactorily both as detectors of modulated 75-centimeter waves and as generators, and they could be used for signaling purposes up to distances of several hundred feet. This detector was found to be far superior to any other type employed and was primarily responsible for the ranges attained with the equipment. A transformer in the plate circuit of this rectifier tube served to couple the detector to an audio-frequency amplifier having a gain of approximately 80 decibels. A 4000-ohm output meter calibrated in volts was used to measure the circuit noise as well as the signal strength. The noise background in the receiving apparatus was of two varieties, a steady hissing sound accompanying detector oscillation, and microphonic disturbances arising because the detector tube was mounted at the mid-point of the receiving dipole where it was

subjected to mechanical shock and wind. The noise level under average operating conditions with no impressed signal gave a reading of the order of 0.2 volt so further audio-frequency amplification was without advantage. Fig. 5 is a photograph of the receiving equipment.

It is considered likely that the frequency changes which accompany the amplitude changes referred to in Fig. 2 are actually of primary
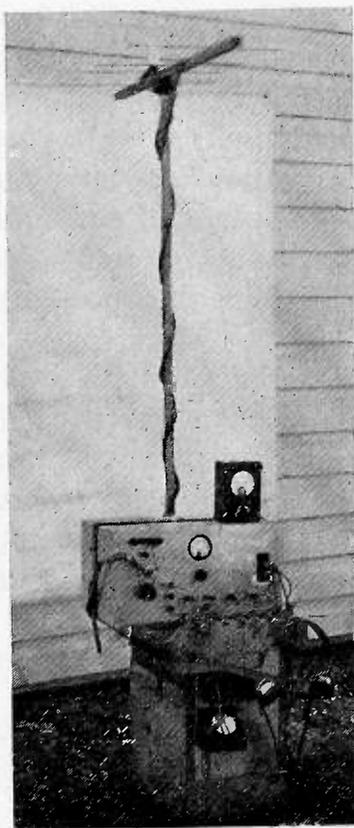
Fig. 5—The receiver.

importance in successful detector operation. It is to be noted that a wavelength change from 75.000 to 75.001 centimeters represents a frequency change in excess of five kilocycles. The 199 detector when on the verge of oscillation is relatively sharply tuned and hence it lends itself well to detection of a frequency modulated carrier. The adjustments on this detector in the matter of plate voltage and filament temperature are rather critical for weak signals, hence the numerical data obtained on such signals are not particularly significant. Also the adjustments for greatest detector sensitivity resulted in some signal distortion thus impairing the intelligibility of speech.

Results

One series of tests with this equipment was conducted over water, and the results are presented at this time as indicative of the utility of the equipment for communication purposes. The transmitter was placed near the Navesink lighthouse at Atlantic Highlands, New Jersey, at an elevation 200 feet above sea level while the receiver was placed on the bridge of a small Army boat and thus had an elevation 20 feet above the surface of the water. This boat maneuvered about in an area south of Long Island. Up to a distance of four land miles the signal saturated the receiving equipment. The signal, measured as volts above noise, dropped rapidly from 10 volts at 5 miles to 2.0 volts at 20 miles in approximately an exponential fashion. Telephonic communication using the 75-centimeter waves was excellent over this entire range of distances. Readings on signal strength were taken every five miles throughout these experiments. Ranges from 20 to 30 miles were characterized by relatively weak signals as well as by unsteadiness. For ranges from 30 to 85 miles the received signal varied from 0.4 to 0.2 volt and showed but slight dependence on distance. The smallness of the boat and inclement weather forced the tests to be discontinued but the greatest range here reported, namely, 88 miles, probably does not mark the limit of the equipment for the received signal was still 0.2 volt and was admirably suited for telegraphic communication at that distance.

The segment of a straight line between Navesink and the horizon drawn through the transmitter and tangent to the surface of the water is 17 miles long. This tangent is 1000 feet above the position occupied by the receiving antenna at the range of 88 miles. That is, the range obtained exceeds the distance from the transmitter to the horizon by a factor of five. This conclusion confirms the observations of other experimenters and the fading or unsteadiness here noted agrees qualitatively at least with previously reported work.[9]

Addendum

With regard to the discussion of the mechanism of Barkhausen-Kurz oscillations in this paper it is to be emphasized that a more complete theory must include a treatment of the conditions which are to be met for the existence of a virtual cathode, in particular, the parts played by space charge and the steady potential differences applied between electrodes. Further, if we are interested in deriving an expression for the phase difference between the alternating components of

[9] *Electrician*, vol. 110, p. 3, (1933).

voltage and current with a degree of rigor greater than is here attempted it is necessary to justify formally the separation of alternating from steady components. Also the starting point would be current densities in the interelectrode spaces instead of total currents which have here been assumed directly. For a more rigorous treatment the papers by Benham are particularly to be recommended as well as a recent paper by Llewellyn, PROC. I.R.E., vol. 21, p. 1532; November, (1933) which appeared after the preparation of this paper.

TRANSMISSION LINES AS FREQUENCY MODULATORS*

By

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Summary—A method of producing frequency modulation is described wherein an eighth-wave radio-frequency transmission line is used as a modulation device. One end of the transmission line serves as a part of the tank circuit of the oscillator while at the other end is placed a variable resistance, such as a vacuum tube. Absolute linearity may be obtained with negligible amplitude modulation. A brief description of different types of transmission lines used is also given.

Introduction

THE most common method of producing frequency modulation at the present time is that of varying one or the other of the oscillatory-circuit constants, L or C, which determine the oscillation frequency. Capacitance variation has been obtained by means of moving diaphragm devices but has not proved to be entirely satisfactory. Examination of the response characteristics of moving diaphragm devices in general shows that, with the exception of the condenser microphone, frequency discrimination and amplitude distortion are present in undesirable amounts. The excellent characteristics of the condenser microphone are produced largely by limiting the diaphragm motion to a very small value and increasing the minute alternating voltage generated with vacuum tube amplifiers. The amount of capacitance variation is much too small in proportion to the fixed capacity of the unit to produce satisfactory frequency modulation.

It is the purpose of this paper to present a method of modulation which is at once simple and linear.

Theory of Transmission Line Modulator

A transmission line will present an impedance at its sending end which is quite different from that connected across its receiving end terminals both in magnitude and in phase. If a line is selected of such a length that a change in resistance only at the receiving end will produce a change in reactance only at its sending end, it may be used to produce frequency modulation. The sending end should be a component part of the oscillatory circuit governing the frequency of transmission, as in Fig. 1, while the receiving end resistance may consist of

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any suitable device, such as a vacuum tube operating over a reasonably linear portion of its plate-resistance—grid-voltage curve. The line itself may consist of a pair of parallel or twisted wires or of a network, depending largely upon the frequency of the modulated wave. For the higher frequencies, the pair of wires would be more economical, but for lower frequencies their length would be prohibitive.

The proper length of line to secure such a condition is determined as follows:

The sending end impedance of a transmission line may be derived from the general transmission line equations as given in any book on telephone or power transmission.

$$Z_s = Z_0 \frac{Z_r \cosh pS + Z_0 \sinh pS}{Z_0 \cosh pS + Z_r \sinh pS} \tag{1}$$
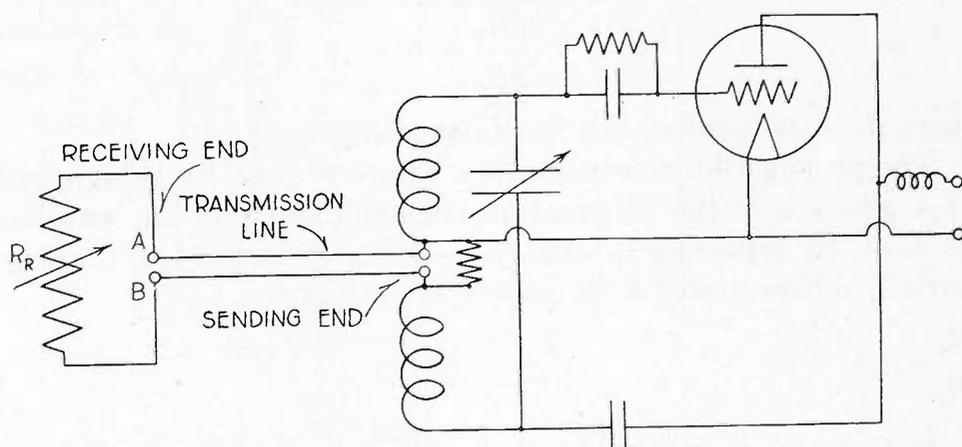
Fig. 1—Circuit diagram of Hartley oscillator with transmission-line modulator.

where,

Z_0 is the characteristic line impedance
Z_r is the receiving end impedance
Z_s is the sending end impedance
p is the propagation constant
S is the length of line in the same units as used for p.

The receiving end impedance is to be purely resistive and we may therefore use R_r instead of Z_r. Furthermore Z_0 in a radio-frequency transmission line has a phase angle so nearly zero that we may also write $Z_0 = R_0$ without appreciable error. Equation (1) may then be written

$$Z_s = R_0 \frac{R_r \cosh pS + R_0 \sinh pS}{R_0 \cosh pS + R_r \sinh pS}. \tag{2}$$

Linear frequency modulation is one of the conditions to be met by this modulation device and will be obtained if the reactive component of the sending end impedance varies linearly with the receiving end resistance. This condition may be expressed mathematically as

$$\frac{\partial X_s}{\partial R_r} = K \tag{3}$$

where X_s is the sending end reactance component and K is any constant, preferably as large as possible in order to produce a high percentage of modulation.

Any amplitude modulation which may be produced by this modulation device is obviously detrimental. It may be avoided by keeping the resistive component of the sending end line impedance constant as the receiver end resistance is varied or

$$\frac{\partial R_s}{\partial R_r} = 0 \tag{4}$$

where R_s is the sending end resistance component.

The propagation constant p is a complex quantity being equal to $a+jv$ where a is the attenuation constant and v is the wavelength constant. By replacing Z_s with R_s+jX_s and p with $a+jv$, (2) may be partially differentiated with respect to R_r to give

$$\frac{\partial R_s}{\partial R_r} = R_0^2 \tag{5}$$

$$\frac{(R_0\cosh aS+R_r\sinh aS)^2\cos^2 vS-(R_0\sinh aS+R_r\cosh aS)^2\sin^2 vS}{(R_0\cosh aS+R_r\sinh aS)^2\cos^2 vS+(R_0\sinh aS+R_r\cosh aS)^2\sin^2 vS}$$

and,

$$\frac{\partial X_s}{\partial R_r} = -2R_0^2 \tag{6}$$

$$\frac{(R_0\cosh aS+R_r\sinh aS)(R_0\sinh aS+R_r\cosh aS)\sin vS\cos vS}{(R_0\cosh aS+R_r\sinh aS)^2\cos^2 vS+(R_0\sinh aS+R_r\cosh aS)^2\sin^2 vS}.$$

But (4) shows that (5) must equal zero or

$$(R_0\cosh aS+R_r\sinh aS)\cos vS = \pm (R_0\sinh aS+R_r\cosh aS)\sin vS. \tag{7}$$

A transmission line must have low losses in order to be a satisfactory modulation device. Therefore, aS must be a very small number and we may write without appreciable error, $\cosh aS=1$ and $\sinh aS=0$. Equation (7) then becomes

$$R_0\cos vS = \pm R_r\sin vS \tag{8}$$

or,

$$vS = \pm \cot^{-1}\frac{R_r}{R_0}.$$ (9)

From (9) it is possible to determine, for any given values of R_r and R_0, the exact length of line required to produce zero change in sending-end resistance as receiver-end resistance is varied. Obviously this length will vary with R_r so that, since R_r must be varied periodically to produce modulation, only an average value of line length, S, may
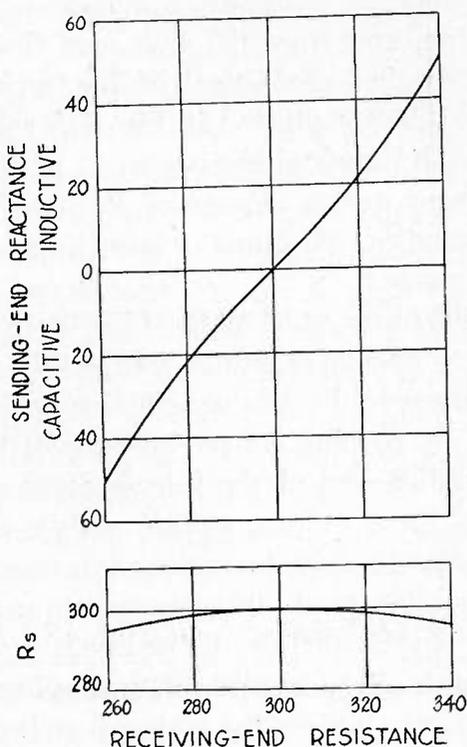
Fig. 2—Curves showing variation of sending end reactance and resistance with receiving end resistance.

be used. However, if R_r does not vary more than 10 to 15 per cent above or below its average value, the change in R_s will be negligibly small. This is clearly shown by the lower curve in Fig. 2 where a change in R_r of 16 per cent results in a change in R_s of only 1.6 per cent.

The rate at which X_s varies with R_r may now be determined by substituting (9) into (6) and letting the values of cosh aS and sinh aS be respectively 1 and 0 as before. The result is

$$\frac{\partial X_s}{\partial R_r} = \mp \frac{1}{2}\sec^2 vS.$$ (10)

The absolute values of R_s and X_s may be determined directly from

(2). By the same process of simplification as used in securing the partial derivatives, the following may be obtained

$$R_s = R_0 \csc 2vS \tag{11}$$

$$X_s = -R_0 \cot 2vS. \tag{12}$$

It is evident from (11) that the lowest possible value of sending end resistance is R_0 and will be obtained when vS is $\pi/4$ or when the line is one eighth of a wavelength long. The sending end reactance for a line of such length is seen to be zero. From (10) it is evident that the rate of change of reactance, while not a maximum, is reasonably large for this length of line, and from (9) it is seen that R_r must be made equal to R_0. Obviously the exact length of line to be used is not critical.

Curves of X_s and R_s are plotted in Fig. 2 for an assumed line one eighth of a wavelength long and having an R_0 of 300 ohms. The reactance variation is linear over a change of R_r of 10 per cent above and below its normal value of 300 ohms, while the resistance variation is negligible over this range.

A resistance of 300 ohms, such as R_s of Fig. 2, inserted into the tank circuit of an ordinary oscillator would completely stop oscillations. It may be reduced, however, by building a line with a lower R_0, or, if this is not feasible, by placing a fixed resistor of the required size in parallel with the sending end of the line. For example, an oscillator, to be stable, should have a tank circuit with a Q of not much less than 12. Let it be assumed that the oscillator is to operate at 1000 kilocycles with a tank condenser of 200 micromicrofarads, having a reactance of 800 ohms at this frequency. If the line is one eighth of a wavelength long and has an R_0 of 300 ohms, a 60-ohm resistor placed in parallel with the line at the sending end will reduce the resistance to 50 ohms, giving a Q of 16. This allows for some additional resistance in the tank inductance without reducing Q below 12.

Such a shunting resistance will necessarily reduce the reactance variation obtained. It is therefore necessary to determine whether sufficient change can be produced. Scott and Woodyard[4] have shown that at a frequency of 1000 kilocycles the maximum change in frequency should be 2.4 kilocycles if the band width is not to exceed 5 kilocycles each side of the carrier. They have also shown that the percentage of frequency variation in the oscillatory circuit is one half the percentage of reactance variation. The maximum reactance variation is therefore 0.48 per cent or a maximum reactance change of 3.84 ohms each side of the normal value of 800 ohms. Fig. 2 shows that the maximum linear change in reactance is about 25 ohms. The shunting resistance of 60 ohms will reduce this effect virtually in the same pro-

portion as it reduces the resistance component, since the actual reactance is very small compared to the resistance. The effective reactance variation will therefore be about 4 ohms which is sufficient. Careful design should give a greater percentage if desired.

Experimental Verification

An experimental 1000-kilocycle oscillator was set up as in Fig. 1 with a noninductive radio-frequency resistance-box at the receiving end of the line. The value of this resistance was varied in steps and the variation in frequency and amplitude of the oscillator was determined. The amount of amplitude variation of the oscillatory current was measured by coupling a low-resistance thermocouple milliammeter through a single turn to the tank circuit inductance and noting the change in current reading as the value of R_r was varied. The frequency variation was measured by determining the change in frequency of the heterodyned signal heard in an oscillating receiver adjusted to a frequency close to that of the tank circuit. A small vernier condenser was connected across the receiver tuning condenser and calibrated so that, with a 500-cycle heterodyned note matching that of a tuning fork and the tuning condenser dial set to an index number, frequency could be obtained from the dial readings. Tests made by means of this method indicated that linear frequency variations of 15 to 18 kilocycles were obtainable with a change in amplitude of the oscillations of less than 0.5 per cent, and that variations of 5 kilocycles were obtainable with a barely perceptible amplitude change, probably in the neighborhood of 0.1 per cent.

The transmission lines used in this experimental work were of the artificial type consisting of cylindrical coils of wire one-half inch in diameter, and seven to eight inches long. The wire was space-wound over a metal coating to make the capacity to ground large in comparison to the turn-to-turn capacity. In construction, aluminum foil was cemented over a half-inch wooden dowl. A longitudinal strip of the foil 0.03 inch wide was removed along an element of the cylinder thus formed so as to avoid a short-circuited turn. Over this foil, waxed paper was wrapped, and the wire was wound over the paper. By varying the turn spacings and insulation thickness in the different designs, the effect of varying the spacing of the wires of a transmission line was simulated. Low values of R_0 were obtained by using thin waxed paper dielectric and by spacing the turns of wire comparatively wide apart. Conversely, high values of R_0 were obtained by including a thicker layer of dielectric between the wire and foil, and by winding the turns closer together.

The maximum change in receiving end resistance required to produce satisfactory frequency modulation has been shown to be not more than 10 per cent. It is therefore easily possible to use a vacuum tube as the receiving end resistance (Fig. 3), operating it over the variable portion of its plate-resistance—grid-voltage curve, without introducing appreciable distortion. It should be pointed out however, that the full voltage of the oscillator is impressed across the line and, therefore, on to the plate of such a tube. If the tube has an impedance higher than that of the line, as most tubes do, and is matched to the line by means of a transformer, the voltage applied to its plate is correspondingly increased. It is therefore desirable to use a tube having a resistance as nearly that of the line as possible.
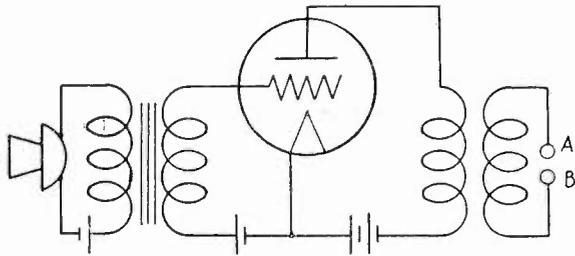
Fig. 3—Circuit diagram of vacuum tube to be used in place of resistance R_R of Fig. 1 to produce frequency modulation.

Undoubtedly there are many other similar applications for a radio-frequency transmission line. It is possible either to magnify or decrease reactance variations by a line of suitable length or to cause a change in receiving end *reactance* to effect a change in sending end *resistance* only.

It might also be stated here that experiments have been run on radio-frequency transmission lines consisting of shielded twisted pairs at frequencies of 8000 kilocycles and higher. These experiments have shown that such lines may be coiled up into a small space without appreciably altering their characteristics. The use of such a line should prove extremely economical. Furthermore these lines had a value of Z_0 as low as 70 ohms. This type of line was not used in the experimental work just described owing to the extreme length required at the frequency used.

ACKNOWLEDGMENT

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THE ACTION OF A HIGH-FREQUENCY ALTERNATING MAGNETIC FIELD ON SUSPENDED METALLIC RINGS AND DISKS*

By

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Summary—An experimental study of the torque on suspended metallic rings and solid disks caused by a high-frequency alternating magnetic field indicates (1) that G. W. Pierce's expression for the torque on a ring is valid over a wide range of frequencies, and (2) that, as a function of frequency, intensity of field, thickness, and conductivity, the torque on a disk is sufficiently similar to that for a ring to suggest that it varies as $\tau p^2 H_v^2 / R(1 + \tau^2 p^2)$ in which τ is the time constant of the disk, R, its effective resistance, and p is 2π times the frequency of the applied magnetic field whose virtual intensity is H_v. Measurement of this torque on a ring or disk apparently constitutes a convenient and practicable method of measuring the intensity of a high-frequency magnetic field.

Comparative measurements on a ring and disk indicate that the greater part of the induced circular current in a disk at high frequencies flows near its periphery, primarily because of the shielding action of the peripheral currents, a result in agreement with experiments made previously by Zenneck.

A reproducible depression in the torque-frequency curve of a thin silver disk was observed at 700 kilocycles.

INTRODUCTION

A METALLIC ring or disk, when placed before a coil which is producing an alternating magnetic field, experiences, in general, a repelling force and also a torque which tends to increase the angle between the normal to the disk and the direction of the field. This effect was discovered by Elihu Thomson[1] in 1884 and used three years later by J. A. Fleming[2] in his "alternate-current disk galvanometer." The repulsion is a differential effect owing its existence to the self-inductance of the movable element which causes a lag in the changes of the induced currents behind the changes in the primary currents. The mathematical analysis of the action has been given both by G. T. Walker[3] and G. W. Pierce.[4] They based the theory upon the simplifying assumption that the movable element is a linear circuit such as a metal ring or short-circuited coil rather than a solid metal disk.

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[1] Thomson, *Electrical World*, May 28, p. 258, (1887).
[2] Fleming, *Electrician*, vol. 18, p. 561, (1887).
[3] Walker, *Phil. Trans.* A, vol. 183, p. 290, (1892).
[4] Pierce, *Phys. Rev.*, vol. 19, p. 202, (1904), and vol. 20, p. 224, (1905).

The theoretical expression for the magnitude of the "electro-inductive repulsion" has been used by F. B. Pidduck[5] for determining the coefficients of self-induction of small suspended loops of wire in terms of their areas and electrical resistances, the strength and frequency of the alternating magnetic field, and the torques experienced. According to his paper, however, Pidduck confined his high-frequency measurements to currents of a single frequency of the order of 100 kilocycles.

Within the knowledge of the writer, no one has experimentally verified, over a wide *range* of frequencies, the theoretical expression for this electro-inductive torque on a linear circuit as developed by Walker and Pierce. As far as the writer is aware no theoretical expression for the variation with frequency of the torque on suspended *solid disks of metal* has yet been developed.

The purpose of the research herein described was twofold: First, to verify experimentally the theoretical expression for the torque on a metal *ring* due to an alternating magnetic field by determining the variation of the torque with frequency and intensity of the field, and second, to investigate experimentally the torque on suspended solid metal *disks* as a function of the frequency and intensity of the magnetic field, the conductivity and thickness of the disk.

Theory

In deriving quantitative expressions for the torque experienced by a circular ring of metal suspended with its center on the axis of a fixed circle which carries an alternating current, both Walker[3] and Pierce[4] based their derivations on the fact that the instantaneous torque is equal to the product of the instantaneous values of the currents in the ring and coil multiplied by the rate of the change of their mutual inductance with respect to the angle between their planes. In the present research, for the fixed element, a pair of Helmholtz coils was used which furnished an alternating magnetic field whose instantaneous intensity was uniform over the region occupied by the movable element, a metal ring, suspended midway between the coils with its center on their common axis. As adapted to fit the above arrangement with the plane of the suspended ring making an angle of 45 degrees with the common axis of the Helmholtz coils, the mathematical expression is

$$T = \frac{Lp^2S^2H_v{}^2}{2(R^2 + L^2p^2)} \cdot \tag{1}$$

Here T is the average torque experienced by the movable ring whose area is S and whose self-inductance and resistance are respectively L

[5] Pidduck, *Phil. Mag.*, ser. 6, vol. 42, p. 220, (1921).

and R; H_v is the virtual value of the intensity of the magnetic field, established at the surface of the movable ring by the current in the Helmholtz coils, the frequency of which is $p/2\pi$.

DESCRIPTION OF APPARATUS AND PROCEDURE

The electrodynamometer was constructed as follows: Two coils, whose common radius was 7.35 centimeters and whose median planes were also separated by this distance, were wound on bakelite forms
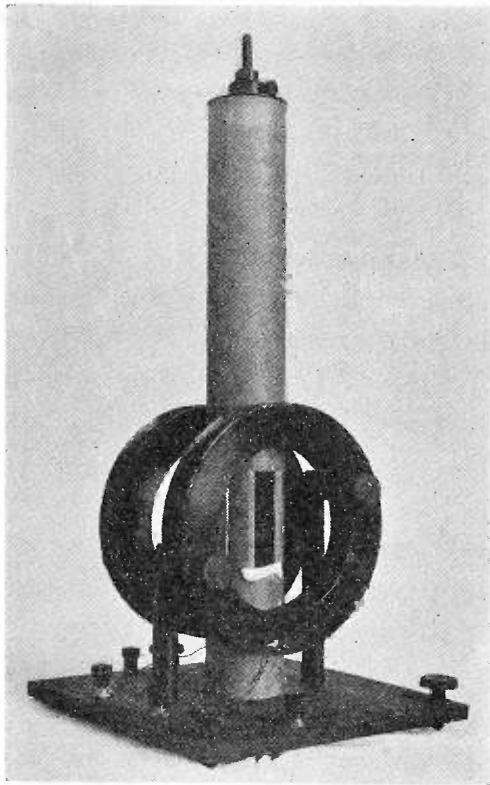
Fig. 1—Photograph of electrodynamometer.

with the turns of each section equally separated in a winding space 1.6 centimeter wide. The forms were mounted on a bakelite base by means of hard rubber bolts and nuts in order to leave the instrument as free as possible from metal parts. That these coils closely approximated an exact Helmholtz pair was tested both by computation from individual pairs of turns and by experiment. Pressed tightly between the coils and set into a groove in the base of the instrument, the tubular formica housing with conventional adjustor for the suspended element was mounted. A photograph of the instrument appears in Fig. 1. Although no electrical connection to the disk was required, it was found convenient to use as suspensions, gold, silver, and phosphor-bronze

ribbons from 12 to 20 centimeters in length which had been rolled from wires 0.0007 to 0.002 inch in diameter. That the suspension might be shielded from sudden temperature changes and vibrations, respectively, the electrodynamometer was placed inside a corrugated cardboard box having double walls and mounted on a cement pier free from the floor. In order to read deflections of the disk from its zero position, a small mirror was fastened on the lower copper stem of the suspension with its plane parallel to that of the disk. A scale and telescope so arranged that a beam of light from the scale was turned through 90 degrees by the suspended mirror before entering the telescope made it easily possible to tell when the plane of the disk made an angle of 45 degrees with the axis of the coils.
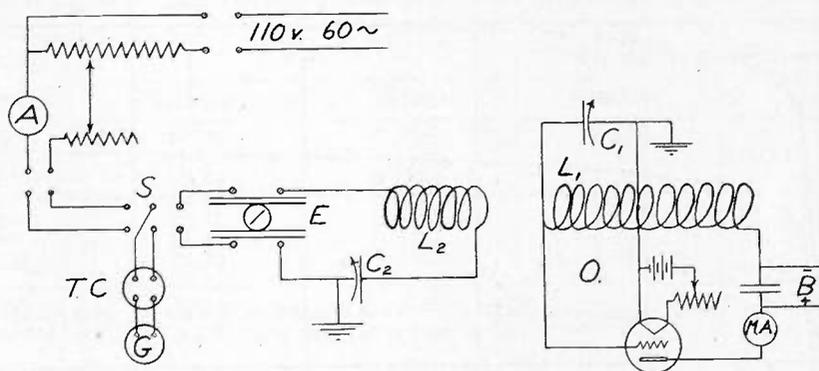
Fig. 2—Circuit diagram of apparatus.

For most of the observations the electrodynamometer was located in a tuned circuit which was inductively coupled to the circuit of a low power shielded vacuum tube oscillator O, as shown in Fig. 2. Directly between the two Helmholtz coils of the electrodynamometer E in the same diagram is shown a double pole, double throw switch S, which was used to connect a Western Electric vacuum thermocouple $T.C.$, (Type 20-R or 20-C) and accompanying direct-current galvanometer G, between the Helmholtz coils during the high-frequency measurements or to a 60-cycle circuit for calibration immediately before and after them. This low-frequency circuit, as shown, contained a potentiometer arrangement for supplying a small fraction of the 110-volt lighting supply to an alternating-current milliameter A (Weston, Model 155) and a control resistance in series with the vacuum thermocouple when switch S was thrown to the left.

Measurements of torque on the suspended element were made for a fixed virtual value of current over a range of frequencies from 12 to 5000 kilocycles, frequencies being determined by means of a General Radio wavemeter (type 224). Observations were also made at a

chosen frequency of the variation of torque with intensity of the magnetic field as determined by the current in the coils of the electrodynamometer.

At frequent intervals during the deflection tests, measurements of the period of vibration of the suspended disk were taken. These readings were averaged and used together with the calculated moment of inertia of the disk and mirror to obtain the coefficient of torsion of the suspension. This was used for calculating torques from observed deflections of the disk.
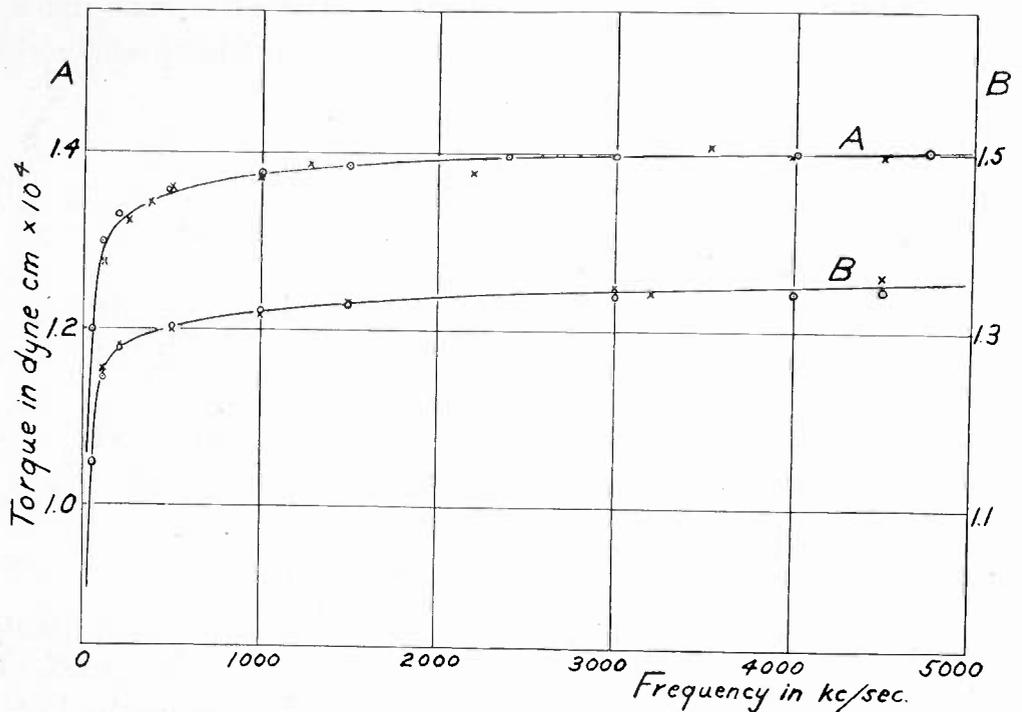
Fig. 3—Torque-frequency curves for silver ring B.
 For curve A, taken with two-turn (per section) electrodynamometer, $H_v = 0.0342$ gauss. For curve B (right scale) taken with five-turn instrument, $H_v = 0.0335$ gauss.

The resistivities of the metals of the various disks were measured by determining the potential difference between two points near the opposite ends of a rectangular plate of the metal when carrying a known direct current introduced by several fine wires connected to saw-tooth projections at each end of the plate. These plates were cut from the same sheet of metal as were the disks.

DISCUSSION OF RESULTS

The experimental results of the deflection tests made on silver ring B are shown in Table I and Fig. 3 together with the values of the torque calculated by the use of (1). In Fig. 3 calculated values are indicated by circles. The values of the resistance R and the self-inductance L of

the rings as used in (1) were calculated from formulas 147 and 207 in Circular 74 of the U. S. Bureau of Standards. It will be noted that there is no systematic discrepancy between experimental and calculated values, and that the agreement is quite close over the range of frequencies studied. The nearly flat portion of the torque-frequency curve was to have been expected since it is to be noted that at frequencies sufficiently high to make R^2 negligible in comparison with L^2p^2, (1) reduces to $T = S^2H_v{}^2/2L$, a form in which T is independent of frequency except for a very slight decrease in L as the frequency is raised.

TABLE I

TORQUES ON SILVER RING B

Ring of 2.135 centimeters diameter made of silver wire 0.0465 centimeter in diameter.

Frequency in Kilocycles	Deflection		Torque in Dyne Centimeters $\times 10^4$	
	Centimeters	Radians	Observed	Calculated
56	19.30	0.0405	1.144	1.147
112	21.17	0.0445	1.254	1.246
200	21.59	0.0454	1.282	1.277
500	21.90	0.0460	1.299	1.303
1000	22.15	0.0465	1.317	1.322
1500	22.49	0.0473	1.333	1.329
3000	22.78	0.0479	1.350	1.340

Current in 5 turn per section Helmholtz coils =55 milliamperes. Scale distance =237.7 centimeters; constant of torsion =0.00282 dyne centimeters per radian.

Measurements taken at a given frequency show very definitely that the torque on either a ring or disk is proportional to the square of the intensity of the magnetic field. Hence in all respects, the experimental results provide, over the frequency range from 50 to 5000 kilocycles, ample verification of the theoretical expression for the electroinductive torque on a suspended ring.

In Table II are to be found the characteristics of the various solid disks together with the torsional coefficients of their respective suspension fibers. It will be observed from this table that an increase in the torsional coefficients of both suspension fibers accompanies an increase in suspended load. This anomaly has been previously observed in the case of fine phosphor-bronze strips by Pealing,[6] Buckley,[7] and Campbell.[8] While interesting in itself, the phenomenon introduces no error in the results of this problem since the torsional coefficients were determined by timing the vibrations of the disks when suspended just as they were used during the deflection tests.

A typical set of measurements showing the variation with frequency of the torque on a solid metal disk is given in Table III wherein are listed the measurements for silver disk No. 3. Results of the measurements on various disks are depicted graphically in Figs. 4 and 5. In

general the shape of the torque-frequency curves for these disks is similar to that for a ring. From ten kilocycles upward, the torque increases rapidly at first but at a rate which decreases as the frequency

TABLE II
CHARACTERISTICS OF DISKS AND THEIR SUSPENSIONS
A. Phosphor-bronze ribbon 20.00 centimeters long rolled from 0.0015-inch round wire used as suspension.

Disk	Conductivity Mho per Centimeter $\times 10^{-5}$	Diameter in Centimeters	Thickness in Centimeters	Mass in Grams	Total Mass Suspended Grams*	Coefficient of Torsion Dyne Centimeters per Radians
Silver No. 1	5.81	2.102	0.0826	2.94	3.22	0.0564
Silver No. 4	5.81	2.102	0.0635	2.28	2.56	0.0541
Silver No. 2	5.81	2.101	0.0422	1.49	1.77	0.0491
Aluminum No. 3	3.33	2.101	0.1548	1.44	1.72	0.0485
Brass No. 1	1.50	2.101	0.0414	1.18	1.46	0.0456
Zinc No. 1	1.68	2.101	0.0411	0.999	1.28	0.0436

B. Gold ribbon 13.00 centimeters long rolled from 0.0007-inch round wire used as suspension

Disk	Conductivity Mho per Centimeters $\times 10^{-5}$	Diameter in Centimeters	Thickness in Centimeters	Mass in Grams	Total Mass Suspended Grams*	Coefficient of Torsion Dyne Centimeters per Radians
Aluminum No. 1	3.33	2.114	0.0989	0.909	1.19	0.00283
Silver No. 3	5.81	2.099	0.0210	0.735	1.02	0.00276
Aluminum No. 2	3.33	2.102	0.0408	0.370	0.653	0.00260

* Includes disk, mirror weighing 0.10 gram and copper stem of suspension weighing 0.18 6ram.

TABLE III
Variation of torque with frequency for silver disk No. 3. Two-turn (per section) Helmholtz coils used. I_v =0.100 ampere. H_v =0.0245 gauss. Scale distance =237.6 centimeters.

Frequency in Kilocycles	Deflection		Torque in Dyne Centimeters $\times 10^4$ =0.00276 $\times$ angle
	Centimeters	Angle in Radians	
8.24	6.90	0.0145	0.400
13.0	9.46	0.0199	0.550
50.9	12.10	0.0255	0.704
253.	13.05	0.0275	0.758
500.	13.12	0.0276	0.761
600.	12.80	0.0269	0.743
750.	12.81	0.0269	0.743
1000.	12.92	0.0272	0.751
2075.	13.07	0.0275	0.758
5820.	13.08	0.0275	0.759

becomes larger, until at frequencies of the order of one thousand kilocycles, the torque is practically independent of frequency.

For the lower range (10 to 100 kilocycles) the torque at a given frequency is larger the greater the thickness of the disk and the conductivity of the metal of which it is made. This is well demonstrated

[6] H. Pealing, *Phil. Mag.*, vol. 25, p. 418, (1913).
[7] J. C. Buckley, *Phil. Mag.*, vol. 28, p. 778, (1914).
[8] A. Campbell, *Proc. Phys. Soc.* (London), vol. 25, p. 203, (1913).

in Figs. 4 and 5. Moreover, the larger the thickness and conductivity, the lower the frequency at which the torque becomes practically independent of frequency.
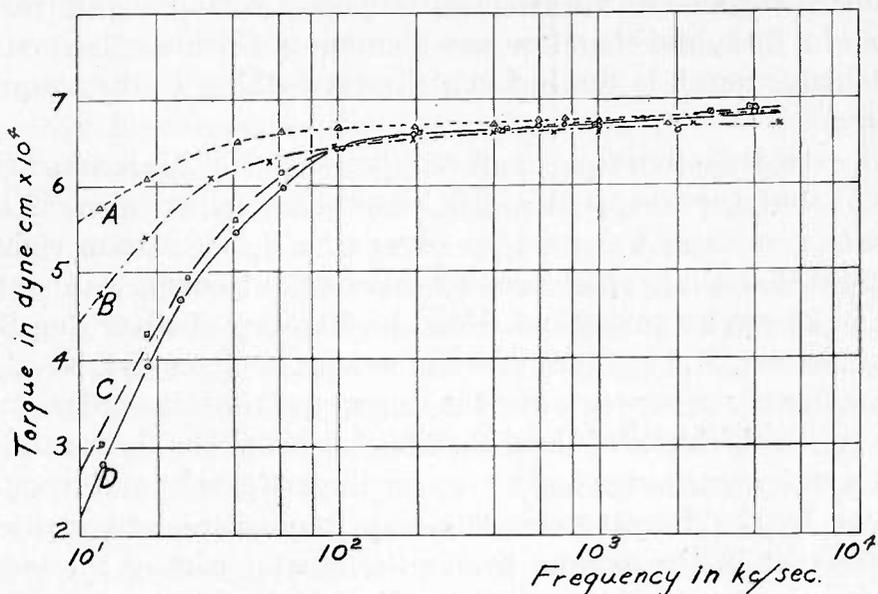
Fig. 4—Torque-frequency curves for disks of different conductances. Curves *A*, *B*, *C*, and *D* are respectively for silver disk No. 2, aluminum disk No. 2, zinc disk No. 1, and brass disk No. 1. $H_v = 0.0731$ gauss.
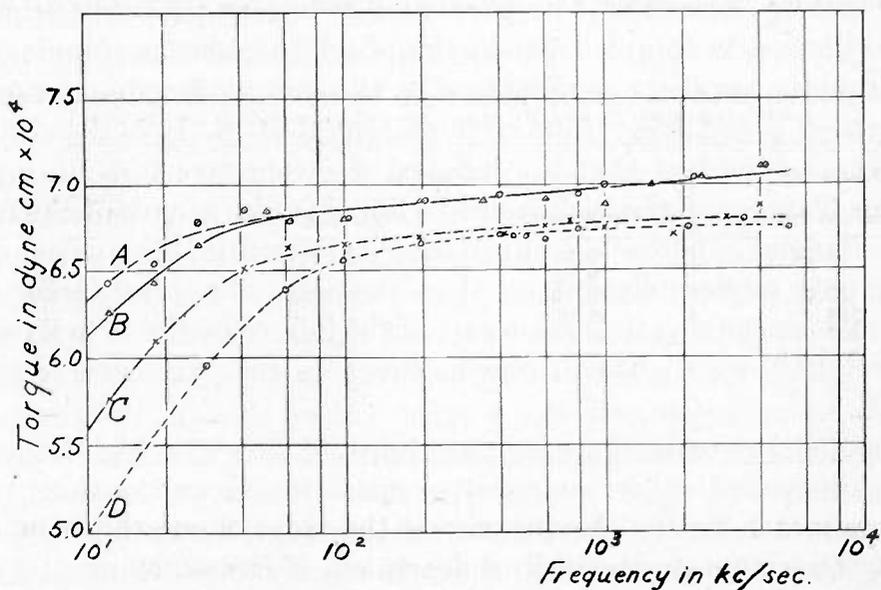
Fig. 5—Torque-frequency curves for silver disks of different thicknesses. Curves *A*, *B*, *C*, and *D* are respectively for silver disks Nos. 1, 4, 2, and 3 which are respectively 0.0826, 0.0635, 0.0422, and 0.0210 centimeter thick. $H_v = 0.0731$ gauss.

The effect of thickness in the range of frequencies above one hundred kilocycles (Fig. 5) is such as to give greater torque for greater

thickness up to a certain limiting thickness, beyond which no greater torque is experienced. This probably indicates that on account of the so-called skin effect which is known to exist at these high frequencies, the induced currents do not penetrate beyond a certain depth from the surface of a disk, and therefore any increase in thickness beyond that at which this depth is reached contributes nothing to the torque experienced.

From the deflection tests made on silver disk No. 2 it is noted (Figs. 3 and 5) that the torque this disk would have experienced with the *same field strength* as was used for silver ring *B* is less than eight per cent larger than that experienced by this ring at frequencies above one thousand kilocycles per second. Now the diameter of silver ring *B* was nearly the same (1.6 per cent greater) as that of silver disk No. 2, and the diameter of the wire forming the ring was of the same order of magnitude as the thickness of the disk. This suggests that the central portion of a disk contributes but a very small part of the total torque experienced by the disk at extremely high frequencies, and apparently indicates that in this region, by far the greater part of the induced circular current in a disk, upon whose value the observed torque depends, flows near the periphery of the disk. This is in agreement with other experiments on the distribution of currents in disks and cylinders performed by Zenneck.[9] The phenomenon is very probably primarily due to the nearly complete shielding at high frequencies of most of the disk from the applied magnetic field by the counter flux due to currents induced in a relatively thin circumferential strip. It is due in part, however, to the fact that the induced electromotive force in a given circular filament of the disk is proportional to the square of the radius of the filament, whereas the impedance is proportional to a power of the radius only moderately greater than the first. The latter factor is responsible for causing the major part of the induced current to flow near the periphery even at low frequencies where the first cause is not effective.

Assuming equal torques on the ring and disk at a high frequency (5000 kilocycles), other comparative measurements show that, as the frequency is lowered, the torque on the ring decreases much more rapidly than that on the disk of nearly equal radius. Noticeable even at the highest frequency considered, this difference in rate of decrease in torque becomes very pronounced in the range from 80 down to 10 kilocycles, the lowest used in this research. In the writer's opinion this is to be explained by the fact that although, as has already been

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[9] Zenneck, "Electromagnetische Schwingungen und Drahtlose Telegraphie," pp. 215 and 480, (1905).

shown, the central portion of the disk contributes very little of the total torque experienced at high frequencies yet, as the frequency is lowered, induced currents of appreciable intensity begin to exist in this portion both on account of the lessened shielding effect due to the counter flux of the peripheral currents and because of lowered impedance. These partially offset the decrease in total induced current in the movable element which occurs purely because of slower rate of change of magnetic field intensity and thus account for the slower rate of decrease of torque with frequency for the disk than for the ring in which, obviously, no such phenomenon can occur.

The torque-frequency curve for the thinnest silver disk used (silver disk No. 3, thickness, 0.0210 centimeter) shows quite definite evidence of a slight depression, the maximum depth of which occurs at 700 kilocycles and amounts to approximately two per cent of the maximum torque experienced by this disk. The depression was observed in four series of measurements each using a different strength of magnetic field and was reproduced on several different days. No appreciable and definitely reproducible similar dip was found with any other disk of silver or other metal. The cause of this phenomenon has not yet been learned but further measurements of the anomaly are contemplated.

One practical use of the electro-inductive torque on a suspended ring is that of the absolute measurement of the intensity of a high-frequency magnetic field. The high-frequency form of (1), previously mentioned, may be written as $H_v = \sqrt{2LT}/S$. For a thin ring, S, the effective area may be determined accurately from dimensions, and L can be computed with a good degree of precision at any frequency by formulas such as those in Bureau of Standards Circular 74 mentioned above. Hence H_v is easily computed as soon as the torque T on the suspended ring has been measured, e.g., by the method used in this research. This method could be used for frequencies above 500 kilocycles with a ring such as silver ring B. Although an *absolute* determination of H_v by this method using a *disk* for the suspended element is impossible since L and S are unknown quantities for the disk, yet the virtual value of magnetic field intensity may be expressed by $H_v = K\sqrt{T}$ where K replaces $\sqrt{2L}/S$ and may be determined once for all for a given disk by observing the torque upon it due to a magnetic field of sufficiently high frequency whose virtual intensity is known. Figs. 4 and 5 show that with disks the measurement of H_v by this method can be extended to appreciably lower frequencies than with thin rings. It may be used throughout the region on which the torque-frequency curve for the disk is flat. This region extends to frequencies approximately as low as 100 kilocycles as shown in Fig. 4.

A satisfactory expression for the torque on a *disk* suspended in an alternating magnetic field has not yet been derived from theoretical considerations. The derivation of such an expression can hardly be made until the distribution of the induced circular currents in the disk is known. As has been demonstrated, this distribution is certainly not uniform. Moreover, it varies with frequency in a manner which, as far as the writer is aware, is as yet unknown. However, in view of the general similarity between the experimentally determined torque-frequency variations for disks and rings, the present work seems, to the author, to suggest that the electro-inductive torque T on a solid metal disk may be expressed as

$$T = C\tau p^2 H_v{}^2 / R(1 + \tau^2 p^2)$$

in which τ replaces L/R and may, perhaps, be called the time constant of the disk, R is its resistance, p is 2π times the frequency of the applied alternating magnetic field whose virtual value or intensity is H_v, and C is a constant of proportionality. This is the same type of expression as that which holds for a ring, placed in what is probably a more convenient form for a disk than (1). This follows from the fact that although there may be some difficulty in conceiving what is meant by the self-inductance, L, of a solid disk and its resistance, R, to the flow of induced circular currents, yet the time constant of a disk[10] and its variation with frequency[11] is already known for very low frequencies. Nothing seems to be known, however, of its value and variation at high frequencies.

It is the writer's intention to extend the deflection tests to very low frequencies with the intent of securing a sufficiently complete set of data to determine whether a satisfactory coefficient, C, for the above expression can be evaluated, and with the added hope of securing information concerning the value and variation of the time constant of a metal disk at high frequencies.

[10] Lamb, *Proc. Roy. Soc.* (London), vol. 42, p. 289, (1887).
[11] Wien. *Ann der Physik*, vol. 49, p. 338, (1893).

DISTRIBUTED CAPACITY OF SINGLE-LAYER COILS*

By

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Summary—The results of previous work done on the distributed capacity of single-layer coils are briefly outlined. Previous theory and experiment do not agree. The reasons for this disagreement are discussed. Two important parameters were omitted from previous theory. The theoretical part of the present paper yields a formula which includes these two additional parameters. This formula is substantiated by experimental evidence and gives the internal capacity of short single-layer coils. By a short single-layer coil, the author means one whose length of winding is, at most, of about the order of the diameter of winding. Any capacity due to leads or terminals should not be considered as part of the internal coil capacity but should be treated separately.

INTRODUCTION

CONSIDERABLE work has been devoted to both the theoretical and the experimental treatment of the distributed internal capacity of single-layer coils. Appended to the present paper will be found a bibliography on the subject. This bibliography, while extensive, is by no means complete.
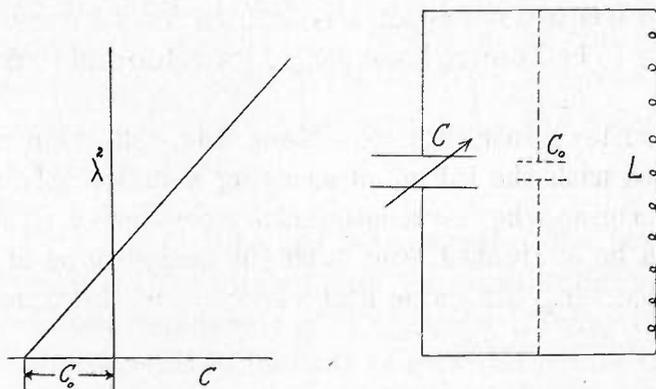
Fig. 1

Lenz,[1] in his theoretical treatment of single-layer coils, determined the total charge of a single ring of an idealized coil. He took the coil as a capacity-free system to which were shunted such capacitances as would have the same effect as the charges actually on the coil. He found that the total coil capacity was one third as much as the ca-

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* Decimal classification: R225. Original manuscript received by the Institute, December 21, 1933.

[1] W. Lenz, "Capacity, inductance and resistance of coils," *Ann. der Phys.*, vol. 37, p. 923, (1912).

pacity between two rings alone. He combined the assumption of an indefinitely long coil with that of negligible curvature. These two assumptions are contradictory; thus Lenz's results would not be expected to apply to actual coils.

Howe,[2] in his work on wave meters, found it necessary to know accurately the internal coil capacity. His well-known method for measuring the coil capacity is shown in Fig. 1. The square of the wavelength is plotted against C, the variable capacity in parallel with the coil necessary to establish resonance. The coil capacity, C_0, is the negative intercept of the straight line on the capacity axis, as shown in Fig. 1. Howe assumes that the coil capacity is independent of the length of winding.

Breit[3] treats the subject theoretically and divides the coil up into small sections perpendicular to the coil axis. He assumes the coil resistance negligible compared to the coil inductance. From the inductance that each section has with respect to the rest of the coil, the electromotive force is found. This electromotive force gives rise to varying amounts of charge along the wire. By summation, an expression for the total electromotive force over the coil is found in terms of charge and capacity. By comparing this expression with the generally accepted procedure of placing the lumped coil capacity in parallel with the coil inductance, Breit's theory yields the following expression for the distributed coil capacity, C_0:

$$C_0 = \int_{x_1}^{x_2} \frac{M(x)}{L} \int_{x_1}^{x} \frac{\alpha(x)}{L} \, dx \, dx. \tag{1}$$

In expression (1):

L is the total self-inductance of the coil
$M(x)$ is the inductance per unit length
$\alpha(x)$ is $Q(x)/(di_1/dt)$ where $Q(x)$ is the charge per unit length
i_1 is the current in the wire at the end of the coil.

For a short single-layer coil having its middle grounded and undisturbed by surrounding objects, (1) yields a capacity in micromicrofarads equal to $0.44r$, where r is the coil radius in centimeters. In this case Breit assumed that the length of winding was small compared

[2] G. W. O. Howe, "Calibration of wave meters," *Proc. Phys. Soc.* (London), vol. 24, pp. 251–259, August, (1919).

[3] G. Breit, "Distributed capacity of inductance coils," *Phys. Rev.*, vol. 17, p. 649; June, (1921); "Distributed capacity of inductance coils," *Phys. Rev.*, vol. 18, p. 133; August, (1921); "Some effects on the distributed capacity between inductance coils and ground," Bureau of Standards Scientific Paper No. 427.

with the diameter of winding and that the inductance per unit length was constant over the coil.

Morecroft[4] obtains some experimental results that do not agree with theory by Breit. No clear conclusions can be drawn from these experimental results because the data given do not include the necessary information concerning the two new parameters of the present paper.

Hubbard[5] shows that Howe's method, mentioned above, can be carried to a good degree of precision. He carries out quite a few measurements which show that the capacity for a short single-layer coil is a minimum and practically equal to the radius of the coil when the length of winding is equal to the diameter of winding. In addition, he finds that the capacity increases by 25 per cent when the length of the coil is four times the diameter of the coil.

Hiecke[6] considers that the capacity between turns of a single-layer coil is negligible and ascribes the measured values to what he terms "external capacity." This "external capacity" is said to be proportional to the radius of the coil.

The results of the previous work, pertinent to the present paper, will now be briefly summarized. Lenz's work cannot be considered because of his inadmissible assumptions. Howe assumes that the coil capacity is independent of the length of winding, namely, independent of the number of turns. Hubbard's measurements show that the capacity is numerically equal to the radius of the coil for short coils. In general, previous work regards the capacity as proportional to the radius of the coil but does not take account of the size of wire or the spacing of the wire on the coil.

THEORY

The present paper has for its purpose the development of a simple formula for the approximate calculation of the internal distributed capacity of single-layer coils when the length of winding is, at most, of about the order of the diameter of winding. A coil whose length of winding is small compared to its diameter of winding, will have greater inductance, for a given total length of wire, than a longer coil would have. Such a coil offers less capacity to surrounding objects than longer coils do. Since the number of turns does not appear as a param-

[4] J. H. Morecroft, "Resistance and capacity of coils at radio frequencies," Proc. I.R.E., vol. 10, p. 261; August, (1922).

[5] J. C. Hubbard, "Effect of distributed capacity in single-layer solenoids," *Phys. Rev.*, vol. 9, pp. 529–541; June, (1917).

[6] R. Hiecke, "On the capacity of coils," *Electrotechnik und Maschinenbau*, vol. 42, pp. 541–545; July, (1925).

eter, the formula may be used with sufficient accuracy for coils where the length is of the order of the diameter of winding. Experimental evidence is presented which indicates the sufficiency of the formula for practical purposes. The disagreement between previous theoretical work and experimental results is largely due to the fact that previous theory neglected to consider two important parameters, namely, the diameter of bare wire and the pitch of winding, that is to say, the distance between centers of adjacent turns. The ratio of these two parameters has an important effect on the coil capacity. These parameters enter in the formula developed in the present paper. It should be noted that the term, internal distributed coil capacity, means only the capacity of the coil itself. Any capacity due to leads and terminals should be treated separately.
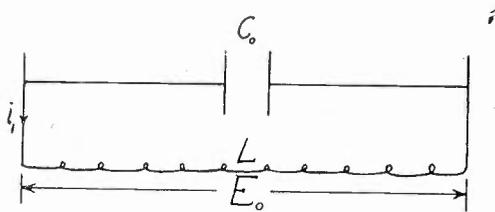
Fig. 2

The generally accepted representation of a coil having inductance and distributed capacity is shown in Fig. 2. The resistance is neglected and the distributed capacity is taken into account by a lumped capacity in parallel with the coil. Thus the current is i_1 and is regarded as the same throughout the winding and the charging currents are considered to flow in the parallel circuit only. Hence the total charge on the coil is summed up and confined to path C_0. The total charge, Q_0, divided by the total voltage, E_0, will give the distributed coil capacity, C_0.

As a matter of fact the amount of the current in the wire varies throughout the coil; it increases in going from the end toward the center of the coil and has its greatest value at the central part. By inserting a noninductive resistance in the wire at various turns of a single-layer coil, Breit measured the current at various distances from the end of the coil. He found that the current in the wire at the central part could be one and a half times as great as that at the end. This effect is explainable when one considers that the turns of the coil have the greatest inductance per turn at the central part. The end turns have the least inductance. A partial trap effect takes place and the central part of the coil is nearer to resonance than the other parts of the coil, for any definite frequency below the natural period of the coil.

Considered from the point of view of circulating current in the path made up of inductance and capacity between two adjacent turns, the circulating current becomes greater toward the center of the coil.

Two straight, coplanar, cylindrical conductors, each of diameter d, having their centers s distance apart, will have a capacity in micro-microfarads per centimeter along the wires, given by the well-known formula[7]

$$\frac{1}{3.6 \cosh^{-1} s/d} \text{ micromicrofarads.} \tag{2}$$

Consider a small distance, dx, on two turns having a diameter of bare wire equal to d centimeters and a distance between their centers equal to s centimeters, as shown in Fig. 3. The curvature of the turns is
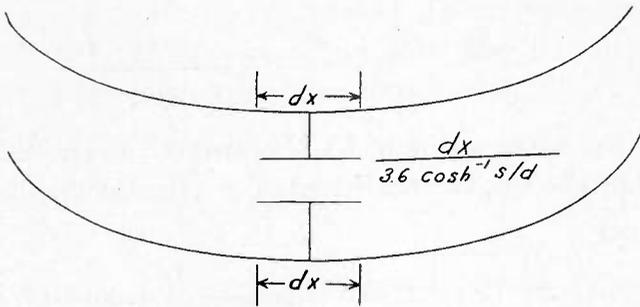
Fig. 3

considered negligible and only those elements, dx, that are s distance apart will be considered. The infinitesimal capacity will be

$$\frac{dx}{3.6 \cosh^{-1} s/d}. \tag{3}$$

If the total voltage of a coil is E_0 and the number of turns is N then the average voltage per turn will be E_0/N. As stated above the charge on the wires is to be summed up. The total charge, Q_0 divided by the total voltage, E_0, will yield the distributed capacity for the whole coil. Diameter of coil is D centimeters.

$$C_0 = Q_0/E_0 = \frac{1}{E_0} \int_0^{N\pi D} \frac{E_0}{N} \frac{dx}{3.6 \cosh^{-1} s/d}$$

$$= \int_0^{N\pi D} \frac{dx}{3.6 \, N \cosh^{-1} s/d} = \frac{\pi D}{3.6 \cosh^{-1} s/d}. \tag{4}$$

Expression (4) will give the distributed coil capacity in micromicro-farads for a short single-layer coil having pitch, s, diameter of bare

[7] Franklin and Terman, "Transmission Line Theory," p. 249.

wire, d, and diameter of winding, D. Formula (4) also holds for non-circular coils; (4) will take the form $L/3.6 \cosh^{-1} s/d$ where L is the length of a single turn. All dimensions are in centimeters. In calculating the charge on the element, dx, the effect of all elements other than those directly below on the next turn, will be neglected. The effect of more distant turns will also be neglected since the intervening turns cause the field between more distant turns to be negligible. The sufficiency of this treatment is to be judged from the agreement of the measured values with those calculated by means of (4).

The same result as (4) can be obtained from the point of view of energy.

$$\frac{1}{2} C_0 E_0{}^2 = \frac{1}{2} Q_0 E_0 = \frac{E_0}{2} \int_0^{N\pi D} \frac{E_0}{N} \frac{dx}{3.6 \cosh^{-1} s/d}$$

$$= \frac{E_0{}^2}{2} \frac{\pi D}{3.6 \cosh^{-1} s/d}. \qquad (5)$$

From (5) the same value of C_0 is obtained. It should be noted that in either one of the above treatments for calculating the coil capacity,
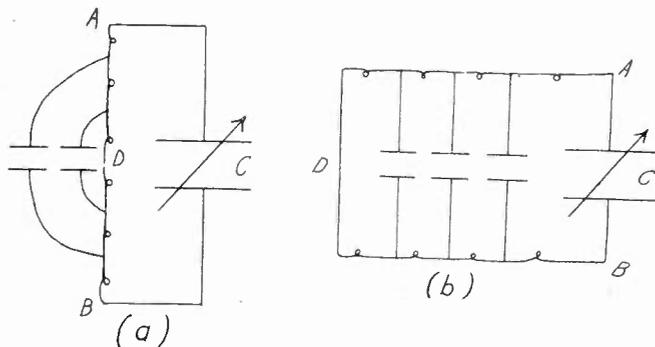
Fig. 4

the charge of the coil should be summed up. To sum up the infinitesimal capacitances as such would require one to consider the various capacitances along the coil as being either in series or in parallel. Contrary to the general treatment of the past, these capacitances cannot be taken as being either in series with one another or in parallel with one another. From the discussion given above, it was seen that the charge on the coil at the central part was greatest and became less at the parts of the coil toward the ends. Thus it can be seen that these capacitances between turns of the coil cannot be taken as being in series because series capacitances must have the same charge throughout the circuit. They cannot be taken as being in parallel because each capacitance should then have the same voltage across it.

For example, Fig. 4 shows an inductance coil represented as a line with uniformly distributed constants as given by Miller.[8] The capacitances between distant turns are considered in parallel with those of turns nearer together. The turns that are far apart will have intervening turns which alter the electric field completely from what it would be without any intervening turns. Furthermore the capacitances would all be unequal and would not permit of treatment as uniformly distributed constants; the inductance of the separate turns also varies. The central turn may have an inductance of thirty per cent more than that of an end turn. The way in which these quantities vary would entail prohibitively complicated mathematics.

The author will show that the same result given above in (4) would follow from the theory given by Breit had the two additional parameters of the present paper been considered. In carrying out his work for a coil whose length is small compared with the diameter of winding, Breit assumes that the inductance per unit length, $M(x)$, is a constant. Expression (1), above, becomes

$$C_0 = \int_{x_1}^{x_2} \frac{M(x)}{L} \int_{x_1}^{x} \frac{\alpha(x)}{L} \, dx \, dx = \int_0^{N\pi D} \frac{M(x)}{L} \int_0^{N\pi D} \frac{Q(x)}{L \, di_1/dt} \, dx \, dx$$

$$= \int_0^{N\pi D} \frac{M(x)}{L} \int_0^{N\pi D} \frac{Q(x)}{E_0} \, dx \, dx = \int_0^{N\pi D} \frac{M(x)}{L} \int_0^{N\pi D} \frac{1}{E_0} \frac{E_0}{N} \frac{dx}{3.6 \cosh^{-1} s/d} \, dx$$

$$= \int_0^{N\pi D} \frac{M(x)}{L} \frac{\pi D}{3.6 \cosh^{-1} s/d} \, dx = \frac{L}{L} \frac{\pi D}{3.6 \cosh^{-1} s/d}$$

$$= \frac{\pi D}{3.6 \cosh^{-1} s/d}. \tag{6}$$

Expression (6) is the same as (4) above. Since $M(x)$ is a constant in the integration carried out in (6), the upper limit for each integral becomes $N\pi D$ while the lower limit is zero. In his theoretical treatment, Breit neglects to consider the size of wire and the pitch of winding; consequently he renders the coil a cylindrical current sheet mathematically.

EXPERIMENTAL RESULTS

In Table I will be found the experimental results of the present paper. All linear dimensions are in centimeters. The average diameter of winding is D; the diameter of bare wire, d; the true inductance of the coil, L, in microhenrys; the pitch of winding, s; and the distributed coil capacity is C_0, in micromicrofarads.

[8] J. M. Miller, "Electrical oscillations in antennas and inductance coils," PROC. I.R.E., vol. 7, p. 299; June, (1919).

Those coils whose self-inductances were very small, say of the order of 15 microhenrys or less, were tuned to a known high frequency with a known value of tuning capacity, C, in parallel with the coil. The value of the inductance was measured and also checked by means of the Grover[9] formulas. The value of C_0 was computed from the following expression (7):

$$\lambda^2 = 3.553 \, L(C + C_0). \tag{7}$$

In the case of the coils having high enough self-inductance, the coil capacity was measured by Howe's method, described above. A

TABLE I

Coil No.	Turns	D	d	L	s	s/d	C_0 (measured)	C_0 (computed)
1	2	7.47	0.624	0.73	1.67	2.67	3.2	3.9
2	10	7.60	0.477	4.33	0.95	1.99	4.4	5.0
3	25	7.80	0.165	33	0.315	1.90	5.7	5.4
4	20	15.20	0.129	96	0.15	1.16	21.0	20.5
5	29	15.20	0.129	174	0.15	1.16	20.5	20.5
6	31	8.60	0.129	82	0.15	1.16	14.0	14.3
7	31	8.60	0.129	45	0.357	2.76	4.5	4.5
8	50	7.95	0.100	157	0.126	1.26	9.0	9.9
9	28	10.40	0.326	58	0.345	1.06	20.0	20.5
10	26	12.75	0.163	79	0.305	1.87	9.5	9.0
11	21	8.05	0.103	18	0.568	5.50	4.0	3.0
12	35	12.87	0.081	193	0.130	1.60	15.2	12.5
13	18	16.50	0.257	62	0.384	1.49	12.0	13.8
14	15	12.70	0.163	37	0.286	1.75	9.5	9.1
15	12	12.70	0.257	21	0.475	1.85	10.0	8.9
16	9	12.70	0.317	12	0.568	1.79	9.8	8.2
17	6	12.70	0.317	7	0.568	1.79	9.0	7.8
18	5	12.70	0.317	4	1.10	3.47	8.6	5.5
*19	112	22.20	0.070	2190	0.089	1.26	12.8	12.9

* Coil 19 was wound with 48–38 litz and the value of d given in the data is that for an equivalent solid round wire.

very severe test of the results of the present paper is to be had in the case of coil 1. This coil had but two turns and a self-inductance of 0.73 microhenry. The self-inductance of the wiring in the tuning circuit was appreciable. Still the comparison of the measured and computed values of the capacity is quite favorable. Coil 1 was tuned to a frequency of 30,000 kilocycles with a known capacity of 35.5 micromicrofarads in parallel with it. Coil 2 was tuned to a frequency of 12,700 kilocycles with a known tuning capacity of 32 micromicrofarads in parallel with it.

Through the courtesy of the General Electric Company, permission was granted to the author to carry on measurements at the General

[9] F. W. Grover, "Formulas and tables for the calculation and design of single-layer coils," Proc. I.R.E., vol. 12, pp. 193–208; April, (1924); M. Rietz, "On the capacity of coils," Ann. der Phys., vol. 41, p. 543, (1913); Bureau of Standards Circular No. 74; R. R. Batcher, "Rapid determination of the distributed capacity of coils," Proc. I.R.E., vol. 9, p. 300; August, (1921); P. Drude, "On the capacity of coils," Ann. der Phys., vol. 9, pp. 293–339, (1902); W. D. Oliphant, "High frequency coil measurements," Jour. Sci. Instr., vol. 9, p. 121; April, (1932).

Engineering Laboratory in Schenectady for coils 1 to 12, inclusive. Measurements for coils 13 to 19, inclusive, were communicated to the author from the Radio Laboratory of the Bureau of Standards in Washington, D. C.

The effects of the various parameters will now be studied. In coils 4 and 5, the only quantity that is different is the number of turns. The capacity of 4 was 21 while that of 5 was 20.5, measured values. This shows that the number of turns has negligible effect on the capacity. Coils 2 and 3 have about the same values for D and s/d but their number of turns are 10 and 25, respectively. Still their capacities are not widely different. This shows again that the number of turns needs not be considered as a parameter.

Coils 14 to 17, inclusive, have about the same s/d and D while their values of s and d vary. Their capacities are not widely different. Only the pitches vary in the case of coils 6 and 7. The capacity of 6 is more than 3 times the capacity of 7. Coil 7 had a value of 2.76 for s, a relatively large value. This means that the conductors were spaced quite far apart. The good agreement between the measured and computed values of the capacity for coil 7 shows that even at this high value of pitch the effect of distant turns of the coil on one another is negligible. It is quite possible that if the turns were to be spaced extremely far apart on the coil, the effect of distant turns would have to be considered.

In many cases the measured value of capacity was somewhat larger than the computed value. For example, the measured capacity for coil 12 was 15.2 while the computed value was 12.5. This particular coil had large terminals on it, hence the measured value of the capacity would have to be corrected for the effect due to the terminals. Correction should be made for terminals or leads where it is found to be necessary.

Conclusion

From the results given above, it is evident that the important parameters in the coil capacity are the diameter of winding and the ratio of pitch of winding to the diameter of bare wire. The capacity is practically independent of the number of turns. The distributed coil capacity can be obtained with accuracy sufficient for practical purposes by means of (4) of the present paper.

Acknowledgment

CONTINUOUS RECORDING OF RETARDATION AND INTENSITY OF ECHOES FROM THE IONOSPHERE*

By

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Summary—Pulse retardation method of Breit and Tuve has been modified to record continuously the equivalent height as well as the intensity of reflections from the ionosphere. Synchronized pulses are transmitted, and the received ground pulse and the reflected pulses, after amplification and suitable distortion, are applied to the focusing cylinder of a cathode ray tube the horizontal deflecting plates of which are connected to a synchronized linear time base circuit. The pattern on the screen is composed of a bright straight line corresponding to the time base with dark gaps corresponding to the received pulses. The distance between the initial points of the gaps represents retardation while the widths of the gaps correspond to the intensity of the pulses. The pattern is photographed on a vertically moving film.

One of the first few records taken at Bangalore on 4 megacycles is reproduced. It shows, among other things, that the less retarded component of magneto-ionic splitting from the F layer is present most of the time. Whenever the longer retardation component does occur, it has stronger intensity than the former. Towards the late evening hours, just before disappearing, when the F layer rises and exhibits magneto-ionic splitting, the intensity of the less retarded component is extremely low compared with the other component.

Loss of resolving power at high intensities could be avoided by the use of a television type of cathode ray oscillograph and undistorted pulses applied to the focusing cylinder, thus recording the intensity of the pulses in terms of the photographic density of the record.

I. INTRODUCTION

THE study of the ionized regions of the upper atmosphere have generally involved the observations of the equivalent height at which the reflections of the radio waves apparently seem to take place. Also, Appleton has shown how, from measurements of equivalent height over a range of frequencies, critical frequencies can be determined at which the waves penetrate the ionized regions. From such critical frequencies one can compute the approximate ionization maxima of the various regions. Recently records have been made which distinguish between the right- and left-handed polarized components in the reflected signals.[1,2] Comparatively little attention, how-

[1] J. A. Ratcliffe and E. L. C. White, "Automatic recording method for wireless investigations of the ionosphere," *Proc. Phys. Soc.*, vol. 45, p. 399; May, (1933).

[2] J. A. Ratcliffe and E. L. C. White, "Fine structure of the ionosphere," *Nature*, vol. 131, p. 873; June 17, (1933).

ever, has been paid to the study of the variations of the intensity of reflections from the various regions. Hollmann and Kreielsheimer[3] have described a system for intensity and height recording which does not seem to be adequate enough to record complicated reflection patterns when a large number of echoes are present and some of them all very close to each other. Such a study may yield additional information with regard to the variations in the ionization content, mechanical movements such as turbulence and winds in the ionized regions, comparative attenuation of the various components of reflection, etc. Perhaps the main hindrance in the way of intensity observations has been the tediousness of the methods used for determining the equivalent height. Gilliland and Kenrick[4] have briefly summarized the situation in one of their papers, in which they have described one of the first attempts to overcome the tediousness and expense involved in the original pulse retardation method of Breit and Tuve.[5] Since then there have appeared a number of other papers describing improved recorders.[2,6,7,8,9] These developments considerably simplify the study of the equivalent height, but it would still further increase the value of these records if simultaneous recording of intensity were incorporated in them. This the authors have attempted to effect by the method described in this paper.

In brief, the method is another modification of the pulse retardation method of Breit and Tuve.[5] Radio-frequency groups or pulses of about 0.1 millisecond duration are sent out from the transmitter at the rate of 125 pulses per second synchronized with the power line frequency of 62.5 cycles per second. This choice of pulse frequency is made by the consideration that the maximum equivalent height so far observed, that is, about 1000 kilometers, should lie within the range of the record, and at the same time sufficient separation should be obtained between the ground ray and the lowest reflection from the E

[3] H. E. Hollmann and K. Kreielsheimer, "Selbsttätige Registierung der Heavisideschichf," *E. N. T.*, vol. 10, p. 392; October, (1933).

[4] T. R. Gilliland and G. W. Kenrick, "Preliminary note on an automatic recorder giving a continuous height record of the Kennelly-Heaviside layer," *B.S.J.R.*, vol. 7, p. 783; November, (1931); Proc. I.R.E., vol. 20, p. 540; March, (1932).

[5] G. Breit and M. A. Tuve, "A test of the existence of the conducting layer," *Phys. Rev.*, vol. 28, p. 554, (1926).

[6] H. Rukop and P. Wolf, "Ein leistungsfähige Einrichtung für Messungen an den Heavisideschichten," *Zeit. für tech. Physik*, vol. 13, p. 132, (1932).

[7] E. L. C. White, "Automatic recording of Heaviside layer heights," *Nature*, vol. 129, p. 579; April 16, (1932).

[8] H. R. Mimmo and P. H. Wang, "Continuous Kennelly-Heaviside layer records of a solar eclipse," Proc. I.R.E., vol. 21, p. 529; April, (1933).

[9] G. W. Kenrick and G. W. Pickard, "Observations of the effective height of the Kennelly-Heaviside layer and field intensity during the solar eclipse of August 31, 1932," Proc. I.R.E., vol. 21, p. 546; April, (1933).

layer, that is, about 100 kilometers. At the receiving station, about 5 kilometers away from the transmitter, a linear time base circuit synchronized with the same power line is made to sweep the spot across the screen of a cathode ray oscillograph tube 125 times per second. The received pulses are amplified, rectified, and introduced in the Wehnelt cylinder used to focus the electron beam in the cathode ray tube. Each received pulse thus produces a dark gap in the time base line by de-
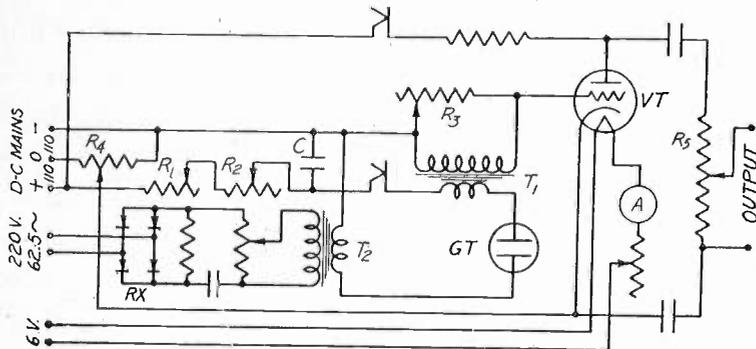
Fig. 1—Pulse generator.

focusing the cathode beam at the time of its occurrence. The circuit is so arranged that the width of the gap corresponds to the intensity of the pulse. The pattern thus produced is photographed on a film moving in a direction at right angles to the direction of the time base line, thus producing a record of equivalent height and intensity plotted as a function of time.

II. Apparatus

1. Pulse Generator

A number of schemes have been devised for the generation of short duration pulses by the various workers in this field.[1,4,5,10,11,12,13,14,15] The schematic diagram of the one developed by the authors is given in Fig. 1. This circuit has been evolved out of the linear time base circuit commonly used in connection with cathode ray oscillograph tubes. Con-

[10] M. A. Tuve and O. H. Dahl, "A transmitter modulating device for the study of the Kennelly-Heaviside layer by the echo method," Proc. I.R.E., vol. 16, p. 794; June, (1928).

[11] George Goubau, "Eine Methode zur Untersuchung von Echos bie der Ausbreitung electromagnetischer Wellen in der Atmosphäre," *Physik. Zeit.*, vol. 31, p. 333, (1930).

[12] E. L. C. White, "A method of continuous observation of the equivalent height of the Kennelly-Heaviside layer," *Proc. Cambridge Phil. Soc.*, vol. 27, p. 445, (1931).

[13] J. P. Schafer and W. M. Goodall, "Radio transmission studies of the upper atmosphere," Proc. I.R.E., vol. 19, p. 1434, (1931).

[14] E. V. Appleton and G. Builder, "Wireless echoes of short delay," *Proc. Phys. Soc.*, vol. 44, p. 76, (1932).

[15] J. P. Schafer and W. M. Goodall, "Kennelly-Heaviside layer studies employing a rapid method of virtual height determination," Proc. I.R.E., vol. 20, p. 1131; July, (1932).

denser C is charged through rheostats R_1 and R_2 (rough and fine frequency controls) from the 220-volt direct-current mains. Across this condenser is connected a neon glow tube GT. When the potential across this condenser reaches the breakdown voltage of GT, it discharges through the latter, the discharge current passing through the windings of the transformers T_1 and T_2. The discharge ceases when the potential across C has reached the value of the extinction potential of GT and C starts to charge again. Thus continuous "relaxation oscillations" are set up. The momentary discharge current passing through
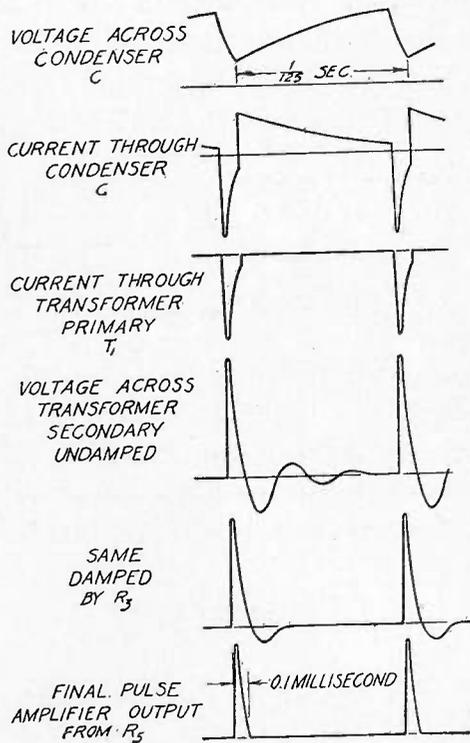
Fig. 2—Voltages and currents in various parts of the pulse generator.

the low tension windings of the transformer T_1 causes a high voltage pulse to be induced in the secondary. Due to the distributed capacity and the inherent inductance in the secondary windings, each of these short duration pulses sets up damped oscillations of a comparatively longer period than itself. Resistance R_3 across the secondary helps to damp these out so that they die out within about one cycle. The pulse is further improved by amplifying it through a highly biased class C amplifier VT. By proper adjustment of the suppressor rheostat R_4, which controls the grid bias of VT, the output of the latter across R_5 can be made to consist of a unidirectional pulse free from any oscillations and of a somewhat shorter duration than the original discharge from the condenser C. The estimated duration of the pulse used is about 0.1

millisecond. The function of T_2 is to introduce into the discharge circuit a small amount of 125-cycle ripple voltage obtained from the rectifier RX supplied by the 62.5-cycle power line. Fig. 2 shows qualitatively the functioning of the circuit,[16] the reference letters corresponding to those in Fig. 1.

The fundamental difference between this circuit and that of Schafer and Goodall[14] is that the coupling between the glow discharge tube and the output circuit is effected by means of a transformer instead of by resistance and capacity. The latter scheme necessitates the introduction of a considerably high resistance in the discharge circuit, which naturally increases the discharge time and hence the pulse duration. In the present case, however, only low tension transformer windings are introduced, the effect of these on the pulse duration being compara-
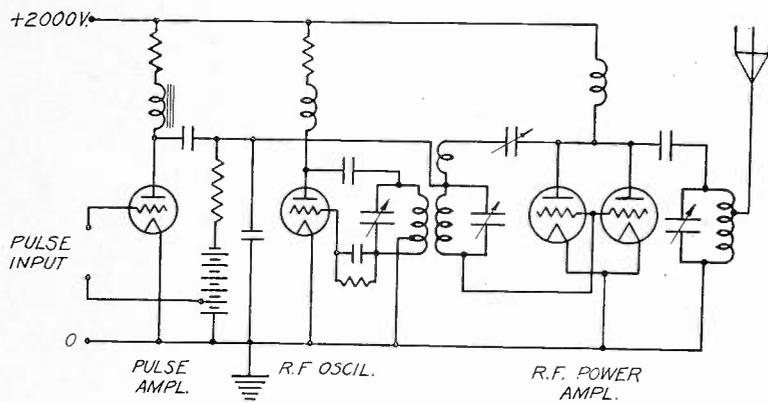
Fig. 3—Transmitter.

tively negligible. Besides, the use of a highly biased amplifier makes it possible to reduce the pulse duration still further if necessary by merely increasing the grid bias.

An observation may be mentioned here in passing, which might be of interest to workers in this field as well as to others interested in multivibrator circuits. When an attempt is made to stabilize directly a pulse frequency by means of a standard frequency which is a submultiple of the former, it is observed that the successive pulses are not evenly spaced but that there exist groups of pulses, the frequency of the groups being that of the standard control voltage. The explanation of this effect becomes clear when one considers in detail the mechanism of stabilizing in such circuits. The stabilization is brought about by

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[16] The whole circuit can be run from the alternating-current mains by simply using the unfiltered rectifier output to charge the condenser C, stabilization being caused by the ripple in it and the filtered direct-current output to supply the necessary voltages for VT. But since both mains are available, the circuit is left in its present simple form.

means of a voltage impulse imparted to the discharge circuit by the control voltage, once every cycle near the peak of the voltage wave. Now, if the stabilized circuit is discharging at a higher frequency than that of the control voltage, the discharges that occur between the time interval of the successive peaks of the control voltage wave are assisted only to a lesser extent by the latter and are even retarded during the opposite half of the cycle. This sort of fine-grain discrepancy must also occur in the multivibrator circuits so commonly used for multiplying frequency.
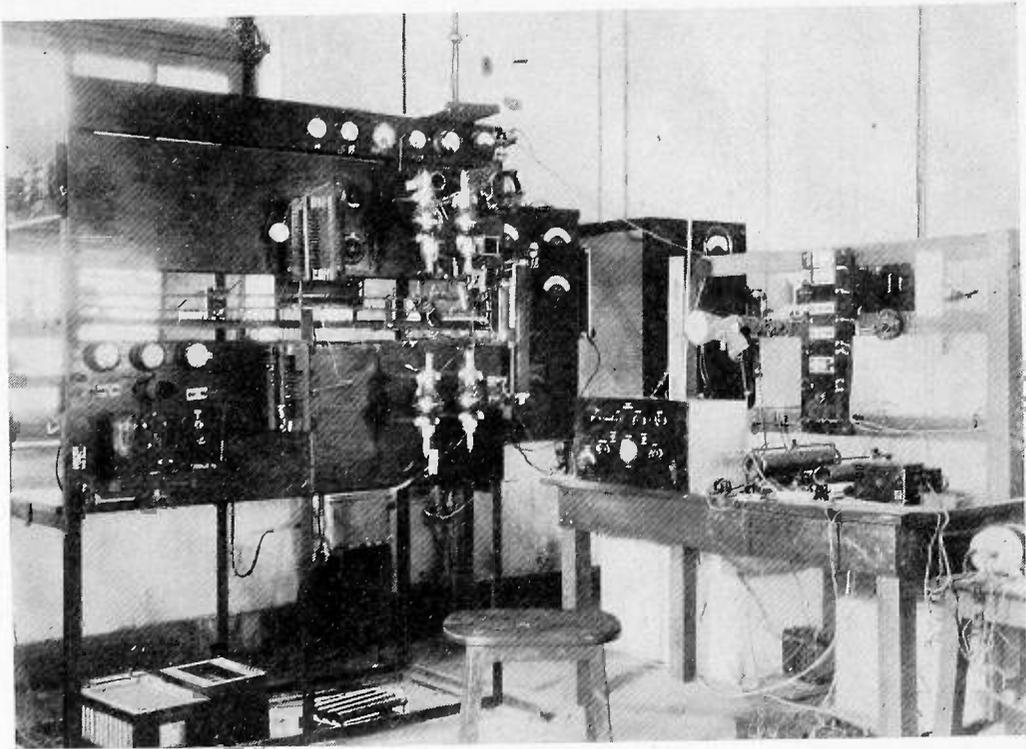
Fig. 4—Transmitter and pulse generator with
cathode ray oscillograph monitor.

2. *Transmitter*

The circuit diagram of the transmitter is given in Fig. 3. It consists of an oscillator, a low-frequency amplifier for pulses, and two neutralized radio-frequency power amplifiers in parallel. The radio-frequency power amplifiers are highly biased, so that the radio-frequency output from them is negligible when there is no pulse. The effect of the pulse is to decrease momentarily the bias and cause the radio-frequency power to be radiated during the time it lasts. Fig. 4 shows the pulse generator and transmitter set-up.

3. Receiver

The receiver consists of two stages of tuned radio-frequency screen-grid amplifiers, a power detector, and a three-stage resistance coupled audio-frequency amplifier.

4. Pulse Indicator

The signals from the receiver are passed on to what is termed the pulse indicator circuit shown in Fig. 5. It is composed of a cathode ray oscillograph tube, a linear time base circuit of conventionl type, the switching arrangement for pulses and that for time marking, and a
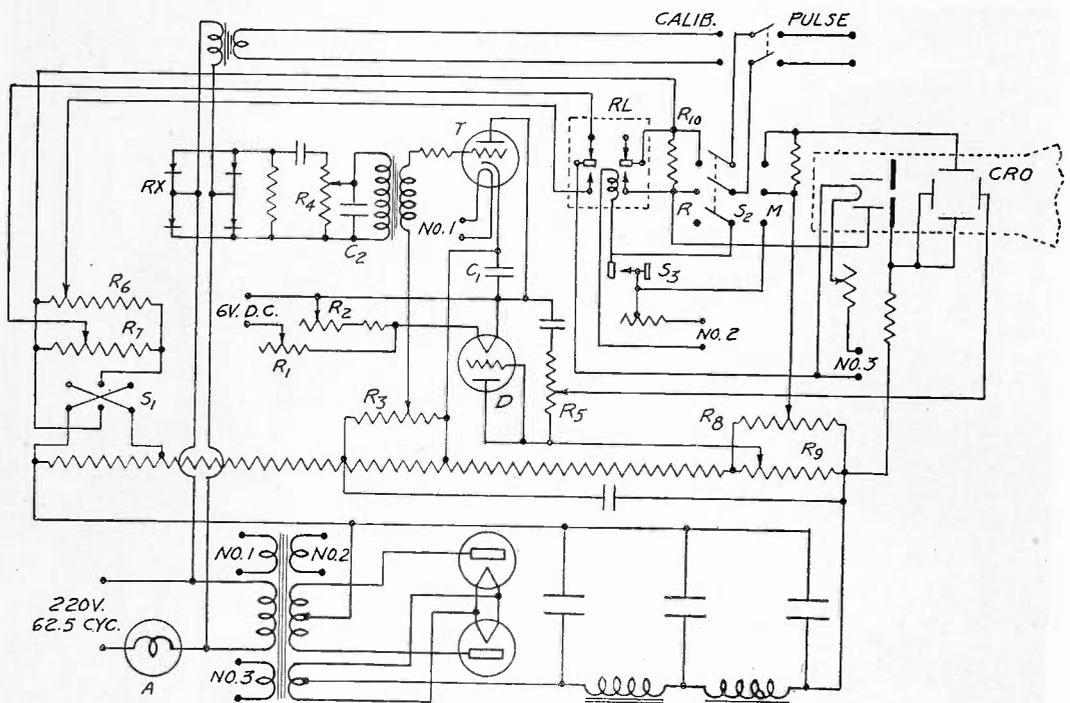
Fig. 5—Pulse indicator.

power supply rectifier and filter circuit to supply the necessary voltages to all these circuits.

The time base circuit contains a saturated diode D, condenser C_1, and a General Radio Thyratron T. R_1 and R_2 are the frequency controls; R_3 controls the grid bias of T and thus the total available amplitude of the time base voltage; R_4 is the stabilizer controlling the 125-cycle input to the Thyratron grid; and R_5 controls the output of this circuit and thus the length of the time base on the cathode ray oscillograph screen. The cathode ray oscillograph tube CRO is supplied from the same rectifier-filter source; R_6 and R_7 controlling the focusing cylinder bias; S_1 being used to make the polarity of the focusing cylinder positive or negative thus making the circuit useful for all types of

cathode ray tubes; and R_8 and R_9 furnishing the necessary biases to the deflecting plates for centering the pattern on the screen. In order to bring the ground ray pulse on the appropriate part of the time base, a fixed condenser C_2 of appropriate value is inserted to adjust the phase of the time base with respect to the ground ray pulse. A continuous phase shifting device seems unnecessary. Switch S_2 is arranged so that, when thrown towards M, the received pulses are applied to the vertical deflecting plates, or when thrown towards R they are introduced
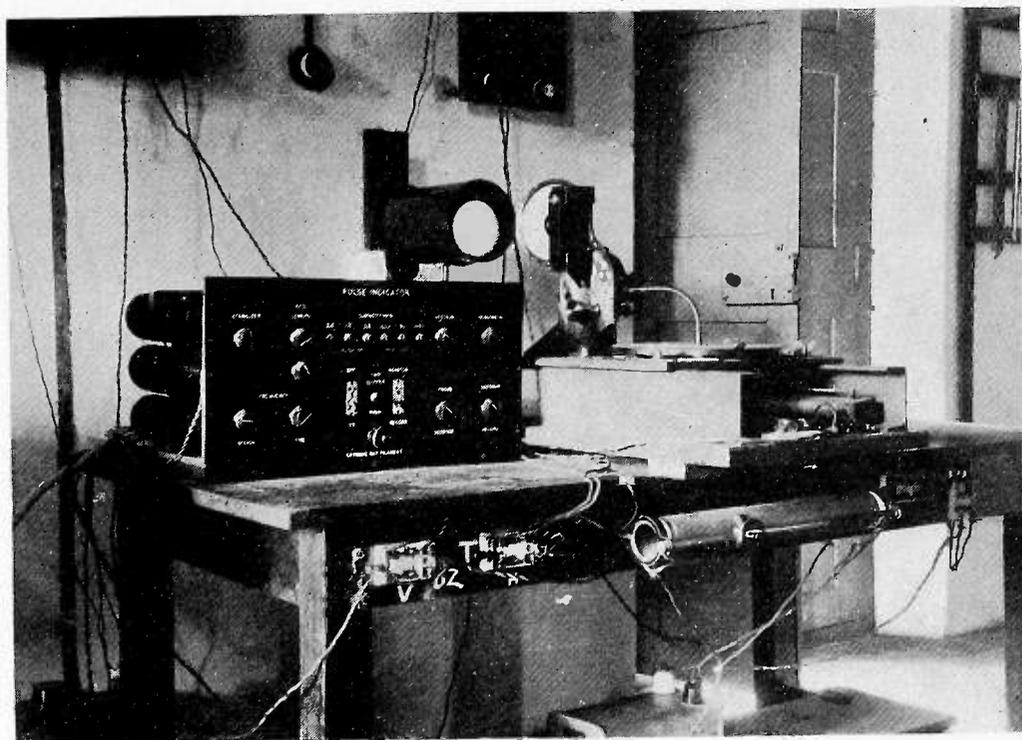
Fig. 6—Pulse indicator and recorder.

in the focusing cylinder circuit of the cathode ray oscillograph tube. In the former case the pulses appear plotted against time in the Cartesian coördinates, this arrangement being used for monitoring. In the latter case the pulse voltage modulates the focusing cylinder bias. This arrangement may be used in two different ways for recording.

1. The rheostat R_7 may be so adjusted as to keep the time base line in sharp focus. The effect of the pulse, if applied in the correct sense, is to produce a sharp break in this line due to the change in the biasing voltage produced by it. Thus the pattern is composed of bright base line with dark gaps corresponding to the ground ray and reflected ray pulses, spaced according to the time of their reception. This arrangement may be termed the defocusing method of recording.

2. The rheostat R_7 alternatively may be so adjusted as to keep the time base itself out of focus. The pulse voltage then modifies the bias so that the electron beam is momentarily brought to focus during the interval that the pulse lasts, that is if the pulse voltage is introduced in the correct sense. Thus, the pattern produced would have a dark background with bright spots corresponding to pulses, correctly spaced in time. This may be called the refocusing method.

It may be mentioned that not all types of cathode ray oscillograph tubes can be operated under either of the two conditions. For example, of the two types of tubes tried, the Standard Telephone tubes give
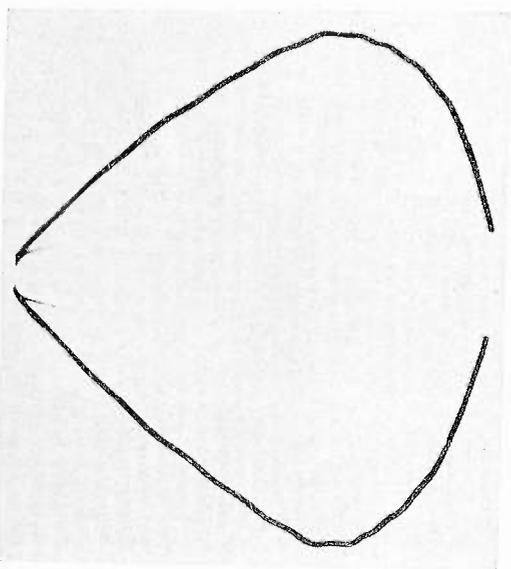
Fig. 7—Time base calibration pattern.

satisfactory results only under defocusing conditions while the von Ardenne type of tubes can be utilized in either of the two ways, defocusing or refocusing. For the results described in this paper the former type of tube has been used.

To introduce the intensity recording feature, the time constant of the low-frequency pulse amplifier of the receiver is so adjusted that the wave form of the output is composed of a sharp positive pulse followed by a trailing negative dip, the duration and intensity of which is roughly proportional to the intensity of the positive portion. Thus the widths of the gaps produced in the defocusing arrangement are increased or shortened corresponding to the intensity of the received pulses and the distances between the initial edges of the gaps give a measure of the equivalent height. What holds for the gaps in the defocusing arrangement would be true for the spots in the refocusing

scheme. The broadening of the defocused gaps or the refocused spots evidently prevents the system from recording distinctly the pulses that may occur very close to each other and have rather high intensity, which may well be the case when magneto-ionic splitting occurs. To avoid this trouble a further refinement is proposed which will be mentioned later; suffice it to say here that, in spite of this difficulty, the record reproduced in Fig. 10 shows several magneto-ionic splits quite distinctly (see Fig. 11 also).
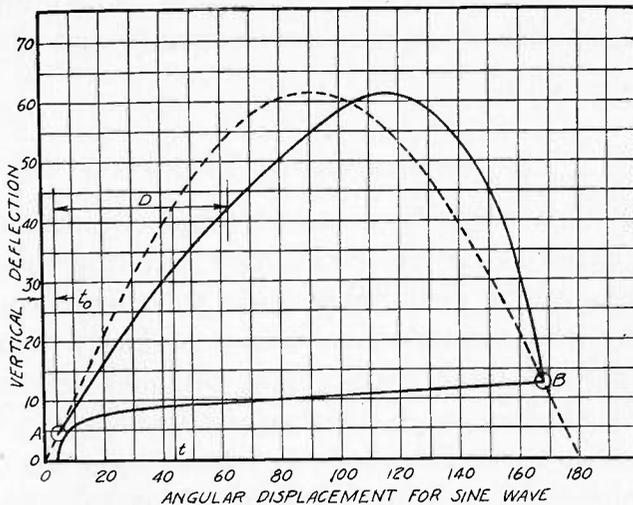
Fig. 8—Time base calibration pattern and sine wave curve.

To provide for making time marks on the record, key S_3, double contact relay RL, and potentiometer R_6 are provided, Fig. 5. At regular intervals, say every hour, S_3 is depressed or short-circuited by means of an automatic clock device, thus operating the relay RL. One set of contacts on this relay short-circuits the resistance R_{10}, thus cutting out the pulse input to the cathode ray tube, while the other set of contacts introduces R_6 in place of R_7 as the focusing cylinder voltage control. R_6 may be preset (a) to keep the time base in focus, thus giving a bright line as a time mark, or (b) to defocus the time base, thus giving a continuous dark line for time mark. It is clear that either of the two schemes may be used for time marking with either of the arrangements for recording. It will be noticed, however, that, when (a) is used for time marking, as has been done for the record in Fig. 10, monitoring by means of switch S_2 becomes a simple matter since it simultaneously operates the relay just like the key S_3, thereby short-circuiting the pulse input to the focusing cylinder and introducing R_6 in place of R_7 for cylinder biasing.

5. Recorder

Fig. 6 shows the pulse indicator and the recording camera with its associated drive. A Leica camera is employed which uses the standard 35-millimeter motion picture film. It is driven by a two-speed gramophone induction motor through the necessary reduction gears to obtain film speeds of about 2.0 and 4.7 centimeters per hour, the intention being to use the higher speed for the sunset and sunrise hours and the other for the rest of the day. In the drive mechanism, a friction clutch of the type used in micrometer gauges is provided to limit the torque transmitted to a value below a predetermined maximum, in order to avoid injury to the delicate camera mechanism in case of film jamming, etc.

III. CALIBRATION OF TIME BASE

A perfectly linear time base would require no elaborate calibration, except the determination of the time taken by the charging and discharging parts of the cycle. But, since the diodes commonly used for charging the condenser do not yield a perfectly constant current over large range of voltage variations it becomes necessary either to replace them by another more satisfactory device or to resort to calibration of the time base. The plate current of some pentodes remains quite constant over a large range of plate voltage variations, and some preliminary work has been done by the authors to make use of this characteristic for charging the condenser in the time base circuits.[17] In the meantime, however, a method is used for time base calibration which does away with the necessity of employing a separate oscillator having an exact multiple frequency. At the end of a run a photograph is made of a pattern obtained by connecting the vertical pair of deflecting plates to a 62.5-cycle voltage wave and leaving the 125-cycle base on the horizontal pair of plates (see Figs. 5 and 7). Fig. 7 is carefully measured and replotted on a graph paper along with a half cycle of a sinusoidal wave of the same amplitude, Fig. 8. From these curves the calibration of time base can easily be deduced. A time base calibration pattern and a half sine wave curve are drawn to coincide at the end points of the former A and B. Then a point on the time base whose displacement is D, corresponds to an instant of time $t - t_0$ reckoned from the starting point A. A calibration curve obtained in connection with the record of Fig. 10 is given in Fig. 9.

[17] A new pentode tube has recently been developed specially for this purpose; *see* Cecil E. Haller, "Linear timing axis for cathode ray oscillographs," *Rev. Sci. Instr.*, vol. 4, p. 385, (1933).

IV. RESULTS

In Fig. 10 is reproduced one of the first few records obtained at Bangalore, South India, latitude 13°N. This particular record was taken during the evening hours of June 8, 1933, on a frequency of 4 megacycles. This, by the way, is probably the first time[18] that radio echo records have been made in such low latitudes. Work along these lines is intended to be continued at the Indian Institute of Science
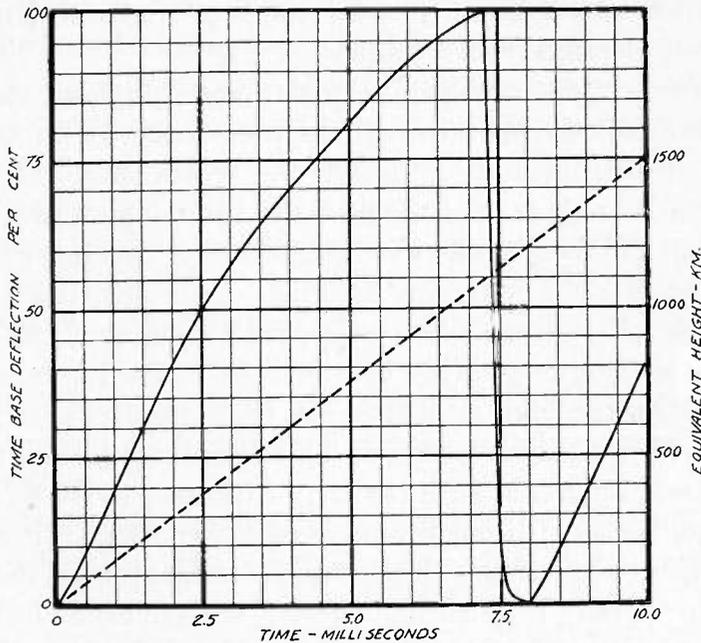
Fig. 9—Calibration curve of time base. (Dotted line in connection with right-hand scale gives equivalent height.)

and will, no doubt, yield valuable information with regard to the behavior of the ionosphere in low latitudes.

It will be observed from Fig. 10 that the intensity of the reflections from the ionosphere fluctuates quite rapidly and erratically especially during the last two hours of recording. With the few records on hand it is not yet possible to draw any general conclusions with regard to the intensity fluctuations. For clearly bringing out the various features with regard to the variations in equivalent height, this record has been carefully measured at intervals of 2.5 minutes and, by making use of the calibration curve of Fig. 9, replotted to an enlarged linear scale, Fig. 11. It will be observed that both E and F reflections are

[18] Since this paper was written the author's attention has been called to the work of The Carnegie Institution of Washington in Peru (latitude 12°S) started in May, 1933, now awaiting publication. Also refer to Ivo Ranzi, "Ionospheric investigations in low latitudes," *Nature*, vol. 133, p. 29; January 6, (1934).

present with magneto-ionic splitting appearing here and there. The equivalent height of the E region varies but slightly around 120 kilometers. It disappears just before sunset. Second order reflection from E is in evidence between 1803 and 1833 hours. A number of long retardation signals appear just before sunset for short intervals which are somewhat difficult to interpret. Those marked E_6 have a retardation value roughly 6 times that of the first order reflection from E, and the doublet marked E_4 has a retardation roughly four times E. The doublet just above it at 1830 hours is very likely a magneto-ionically split second order reflection from F region.

F region reflections come in very strong throughout the period of recording, with some discontinuities before sunset. From 1827 to 2158 hours they are continuous, although the second and third order reflections from F are weak for most of the time and show a number of discontinuities. The equivalent height of the F region varies roughly between 320 and 220 kilometers. Magneto-ionic splitting is evident in the third order F reflections on two occasions about 1910 and 1950 hours. This splitting is possibly also present in the lower order reflections, but due to the high intensity, the resolving power of the instrument is rather low at these points and hence splitting is not recorded in lower order reflections. It is, however, of interest to note that, in the third order reflections, the component of lesser retardation is present most of the time while that of longer retardation appears only on these two occasions. It may be noticed also that, whenever the latter component does come in, it has a much greater intensity than the former. Furthermore, it will be observed that, throughout the record, the third order reflection has somewhat less retardation than three times that of the first order F reflection. The discrepancy has been calculated to be about 52 kilometers on the average, which is far beyond the experimental error of the apparatus. Again, at about 2140 hours there begins to appear a magneto-ionic splitting in the first order F reflections before they disappear. The higher component rises rapidly and disappears about five minutes before the lower one, which rises rather slowly. The point worth noting is that the intensity of the lower component is very much less than that of the upper one; so much so that it is only faintly observable on the original record.

Appleton, his coworkers and others[19,20,21,22,23] have published a good deal of experimental data concerning magneto-ionic splitting, along with some theoretical explanations of the same. It has been shown that

[19] E. V. Appleton, "Present knowledge of the upper atmosphere," *Wireless Engineer*, vol. 8, p. 513; September, (1932). (A résumé of a lecture to the I.E.E. (London), May 25, 1932).

the two components have different polarizations—one has right-handed and the other left-handed circular polarization. Either one or the other of these may have longer retardation depending on the conditions of ionization and the frequency. Normally, however, at this frequency the longer retardation component is expected to be left-handedly polarized and stronger in intensity than the other.

In view of this knowledge one is tempted to draw some conclusions about the conditions recorded in Fig. 10. Inasmuch as they are based on a single record, it must be clearly borne in mind that their value only lies in the fact that they raise certain questions that point the way for further investigation. Since the third order reflections of the F region show a retardation some 50 kilometers less than three times that of the first order and on two occasions somewhat longer retardation components appear for short intervals of time, it may be assumed that the former has a left-handed polarization and the latter a right-handed one. According to Appleton's theory, such a state of affairs can exist under either of the following two conditions, namely (using his nomenclature):

1. A prominent ledge below the F region may be present, which at certain frequencies may cause the right-handed component to suffer longer retardation.

2. The splitting may not be due to stratification effect, but to group retardation effect brought about by the residual ionization of the E region.

Evidently, the second possibility does not seem likely. In any case one question still remains open which concerns the reason for the greater intensity of the longer retardation component which appears at the two relatively short intervals. The splitting that appears at 2140 hours is, however, quite a normal phenomenon, but it may be pointed out that the relative intensities of the two components is here recorded for the first time, though they have been otherwise observed before. It would be evident that if polarized aerials be incorporated in this system, somewhat in the manner of Ratcliffe and White,[1,2] it would be possible to study the relative attenuation of the variously polarized components under different conditions.

[20] T. L. Eckersley, "Polarization of echoes from the Heaviside layer," *Nature*, vol. 130, p. 398; September 10, (1932).

[21] E. V. Appleton and J. A. Ratcliffe, "Polarization of wireless echoes," *Nature*, vol. 130, p. 472; September 24, (1932).

[22] E. V. Appleton and G. Builder, "The ionosphere as a doubly refracting medium," *Proc. Phys. Soc.*, vol. 45, p. 208; March, (1933).

[23] E. V. Appleton, "On two methods of ionospheric investigations," *Proc. Phys. Soc.*, vol. 45, p. 673; September, (1933).
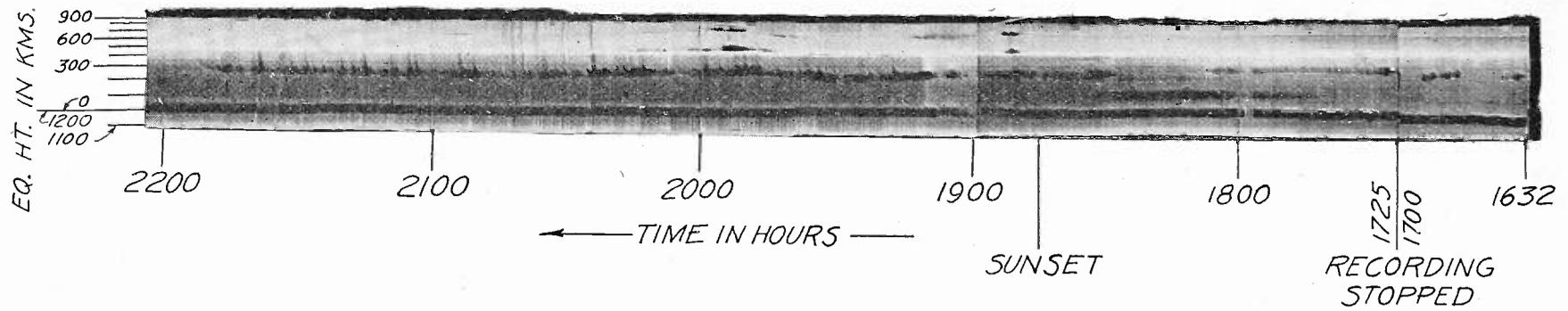
Fig. 10—Retardation and intensity record taken at Bangalore, India, on June 8, 1933.

Finally, attention may be drawn to a rather long time retardation signal received at about 1953 hours for about three minutes, which corresponds to an equivalent height of 1105 kilometers. On Fig. 10 it appears below the ground-wave line. This may probably be a fourth order F reflection.

V. Further Possible Improvements

In spite of the loss of resolving power at high intensities, it has been shown in the above discussion that the system as it stands is capable of gathering quite valuable information with regard to the intensity of the reflections. The objection of low resolving power can, however, be overcome by using a television type of cathode ray oscillo-
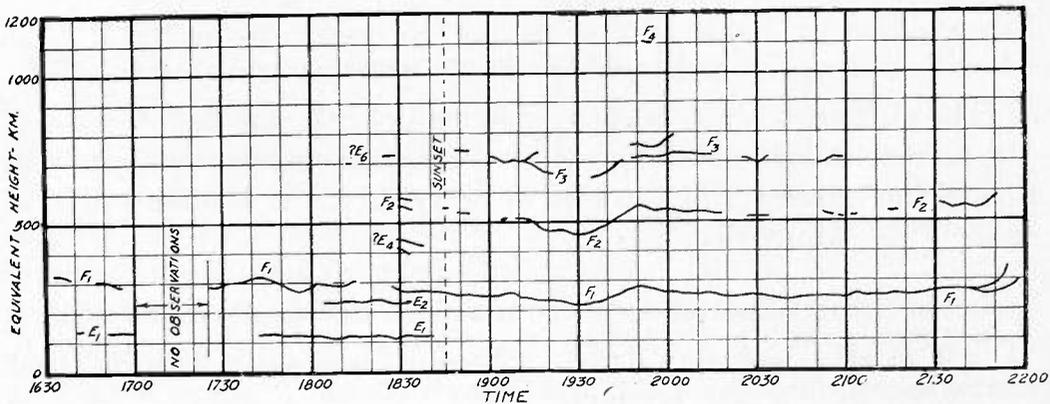
Fig. 11—Equivalent height as measured from Fig. 10.

graph tube instead of the ordinary laboratory model. With the former type of tube the *intensity* of the cathode ray beam can be modulated by the variations of the voltage on the focusing cylinder. Thus the records obtained by either the defocusing or the refocusing scheme would consist of narrow traces of lines the photographic density of which would bear a definite relation (which can be easily made approximately linear) to the intensity of the reflections from the ionosphere. It would, no doubt, be necessary to reduce the time constant of the amplifier in this case to a value less than the duration of a single pulse. The resolving power of the system would then be a function of the pulse duration only. With pulses of 0.1 millisecond duration, it would be possible to distinguish two reflections whose equivalent heights differ by 15 kilometers, or even less.

An accurately linear time base obtained by using the newly developed constant current tube[17] coupled with the ingenious time marking calibration device of the type developed by Ratcliffe and White[1] would be valuable additions to the system. But in order to take full

advantage of this automatic calibration and obtain accurate results thereby, it seems necessary to have some kind of automatic phase control of the synchronizing voltage. Slight variations in the phase cause the ground ray base line to shift about. This shift, though small, is noticeable not only in records taken by the authors but also in those published by other observers. The authors have eliminated the error due to these phase shifts by measuring the record point by point and replotting it. However, with automatic calibration, it would be desirable to get rid of the phase variations entirely. The easiest and quickest way out seems to be to use an independent alternating-current source accurately regulated and to reduce the distance between the transmitter and the receiver. The latest developments indicate that the distance can be reduced to almost zero.[24,25,26,27]

VI. ACKNOWLEDGMENT

[24] Ivo Ranzi, "Observations of the stratification of the Heaviside Region," *Nuovo Cimento*, vol. 8, p. 258, (1931).

[25] S. K. Mitra and H. R. Rakshit, "Recording wireless echoes at the transmitting station," *Nature*, vol. 131, p. 657; May 6, (1933).

[26] R. A. Watson-Watt and L. Bainbridge-Bell, "Recording wireless echoes at the transmitting station," *Nature*, vol. 131, p. 657; May 6, (1933).

[27] Ivo Ranzi, "Recording wireless echoes at the transmitting station," *Nature*, vol. 132, p. 174; July 29, (1933).

ECHOES OF RADIO WAVES*

By

N. Janco

(New York University, New York City)

Summary—Following the magneto-ionic theory of Appleton-Hartree, an explanation for the existence of echoes of long delay has been formulated. Realizing that the ordinary and extraordinary rays have opposite senses of polarization, it may be seen that one of these rays will penetrate the E layer while the other may not. The wave which penetrates the E layer will be reflected back from the F layer to the E layer and being repeatedly reflected between the two layers will travel around the earth and finally back to the ground. This time delay will account for the existence of echoes of long delay.

WHILE the subject of "echoes of long delay" is, perhaps, still a matter of controversy, enough experimental evidence for the existence of such "echoes" has been accumulated to justify speculation as to their origin. In addition to the suggestions already offered, it has seemed to the author that the magneto-ionic theory of Appleton-Hartree as discussed by M. Taylor[1] presents an avenue of investigation of some promise. The argument follows:

According to the above theory, as may be noticed from the values of the refractive index, two different waves are in general present in an ionized region. These two waves are called the "ordinary" and the "extraordinary" waves. The ordinary wave has a left-handed sense of rotation while the extraordinary wave is polarized in a right-handed sense. This polarization will be elliptical, and with the directions of the polarizations above mentioned, when the direction of propagation makes an acute angle with the positive direction of the earth's magnetic field.[2]

The theory indicates and experiment proves[3] that ordinarily two waves will be reflected from an ionized region with different equivalent path lengths. This indicates that, indeed, two different waves are propagated and that one is able to penetrate the ionized region with greater facility than the other. In fact, for short waves, under a favorable density of ionization in the E layer, the ordinary ray may penetrate the layer while the extraordinary ray will be reflected.

The reason for this is that there is a "characteristic gyroscopic frequency" and direction of rotation of electrons in the ionized layer

(revolving about the earth's magnetic lines of force). This seems to indicate that a certain radio frequency would be most favorably transmitted through this layer, and that a wave with a polarization in the same direction of rotation as the electrons would also be most favorably transmitted.

From the preceding, we can see that under certain circumstances the ordinary ray can penetrate the E layer. Now, the ionization density of the F layer is much higher than that in the E layer so that the ray would be reflected (or refracted) from the F layer. Let us now consider that this reflection changes the direction of the polarization or rotation of the polarized wave. Then the wave, now having a right-handed sense of rotation, will be reflected (as the extraordinary ray was reflected initially) back from the E to the F layer with another similar change in polarization. This process would then repeat itself, the wave being repeatedly reflected back and forth between the two layers in a region of low attenuation. However if the ray initially was inclined at even a small angle from the vertical it will tend to travel around the earth while being repeatedly reflected. After a partial or complete circuit of the earth it is possible for the ray to find a hole (that is, a part of the E layer which has a low density of ionization) which would allow the ray to penetrate the layer and return to the earth to a receiving station.

It has been noticed experimentally by Stormer[4] and van der Pol[5] that within a short period echoes may be heard with delays from three to thirty seconds. This would be expected from the above explanation as it is known that the density of ionization and therefore the equivalent path length varies continuously. This would, because of the many reflections, increase or decrease the path length by a sufficient amount to account for the delay of the echoes in question. Pedersen[6] has shown that the attenuation losses accompanying reflection from an ionized layer can be small, under proper conditions. This would, of course, be essential to the above explanation.

The preceding discussion can only be true if the density of ionization is such that the ordinary ray penetrates the E layer while the extraordinary ray is reflected. It is also necessary for the ray to find a hole to return to the earth; and an observer must be located, if the echo is to be observed, where it does return to earth (a location which cannot be predetermined). Thus it may be noted that although the

[4] Stormer, *Nature*, November, (1928).
  Stormer, *Proc. Roy. Soc.* (Edinburgh), vol. 50, pt. 2, (1930).
[5] Balth. van der Pol, *Nature*, December, (1928).
[6] Pedersen, "Wireless echoes of long delay," PROC. I.R.E., vol. 17, pp. 1750–1785; October, (1929).

phenomenon of "long delay echoes" might occur quite frequently, it would be recorded only on very infrequent occasions.

For waves in the broadcast band the conditions are reversed. The extraordinary ray tends to penetrate—not the ordinary ray. (As stated previously, a particular frequency will be favored and this frequency is in the neighborhood of 30 meters.) However, since the attenuation increases with the wavelength, we should not expect to find echoes for such long waves. Experiments which would determine the type of polarization of the echo could throw much light on this subject.

ACKNOWLEDGMENT

DISCUSSION ON "AN OUTLINE OF THE ACTION OF A TONE CORRECTED HIGHLY SELECTIVE RECEIVER"*

E. B. MOULLIN

C. B. Fisher:[1] In the above paper it is pointed out that if a modulated voltage wave is applied to a simple resonant circuit with a power factor less than 0.05 per cent, then the response is sensibly independent of the power factor for all modulation frequencies greater than about 500 cycles per million cycles of carrier frequency. Moullin hence shows that for a given power factor, the ratio of the depth of modulation of the current in the resonant circuit to the depth of modulation of the voltage wave, is inversely proportional to the modulation frequency. In many important instances, however, we are interested in the cases which lie outside this proportional limit, either because of a larger power factor of the tuned circuit, or a lower modulation frequency. Accordingly the curves of Fig. 1
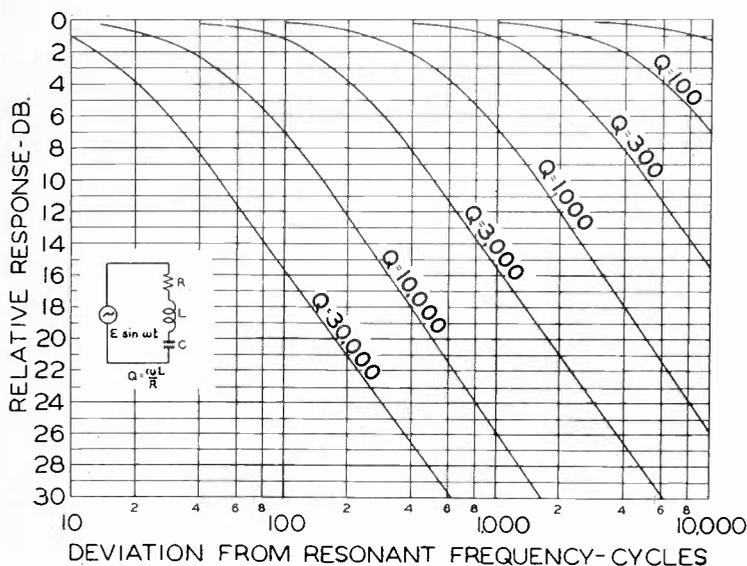
Fig. 1—Response of simple resonant circuits to frequencies removed from resonance. Resonant frequency 1000 kilocycles.

have been drawn to show the relative response of a simple resonant circuit to a voltage wave not of the resonant frequency of 1000 kilocycles. These calculations are based on data given by B. de F. Bayly before the Toronto section of the I.R.E., May 8, 1929. Moullin's symbol F, the power factor, has been replaced by Bayly's symbol Q, the reciprocal of F.

Moullin discusses cases where the power factor is of the order of 0.01 per cent; i.e., Q is 10,000. Such circuits can be obtained only by the use of crystals, or the introduction of a considerable amount of regeneration, and against both these solutions serious objections may be raised. There appear to be better possibilities in the use of several resonant circuits with easily obtainable values of power factor, connected in tandem through unilateral circuits, that is, neutralized or screened vacuum tubes. In Fig. 2 the curve of Fig. 1 for $Q = 300$ has been

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redrawn and additional curves drawn to show the response of two and three similar circuits, respectively. The response of three circuits shows a reduction at frequencies near 5000 cycles equal to that of a single circuit with $Q = 3000$, and at the same time the frequency for a reduction in amplitude of one decibel is more than six times the corresponding value for the single circuit with $Q = 3000$. When it is considered how difficult it is to obtain stable circuits with high values of Q, and the very real difficulty in maintaining accurate tuning of highly selective receivers, the essential advantages of the multiple circuits are obvious.

E. B. Moullin:[2] I agree with Mr. Fisher that there is much to be said for the use of several resonant circuits connected in tandem through amplifiers. But such an arrangement involves several more triodes and also carefully ganged
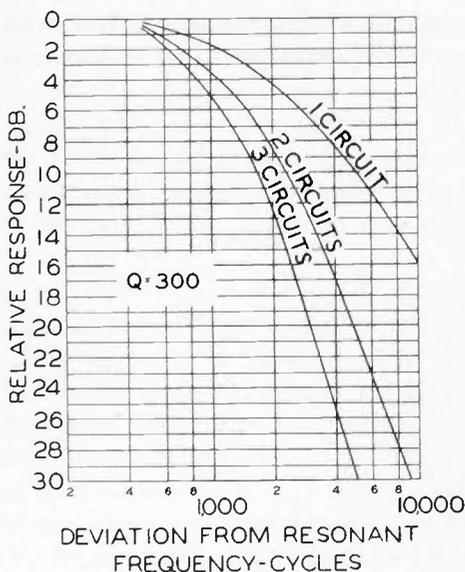
Fig. 2—Response of 1, 2, and 3 isolated similar simple resonant circuits, respectively, to frequencies removed from resonance. Resonant frequency 1000 kilocycles.

circuits. It may be that such procedure is necessary, but if so I think much of the simplicity has been lost. It appears to me there are two ways of obtaining selectivity; one which involves an approximately square-shaped resonance curve and one which involves the simple resonance curve having abnormal sharpness. The first method required band-pass filters and a superheterodyne system if the receiver is to be practicable over a band of wavelengths. The second system aims at obtaining adequate selectivity without the use of a chain of circuits or a superheterodyne. If practical difficulties of stability demand the use of Mr. Fisher's system, then I should have supposed a better solution was to use a band-pass filter and so avoid the necessity of tone correction.

I have no first-hand experience of the power factors which can be obtained in stable operation with valve retroaction. But information concerning this is obtainable on p. 46 of Special Report No. 12 of the Radio Research Board of Great Britain. In this report an experimental resonance curve is shown having a power factor of 1.18×10^{-4} and it is stated that the system remained sufficiently

stable for making precision measurements over a period of an hour or so. Other cases are cited as follows: $F = 2.7 \times 10^{-4}$ for a carrier frequency of 1000 kilocycles, and 1.5×10^{-3} for a carrier frequency of 100 kilocycles. These results were obtained by the use of a valve whose sole function was retroaction and doubtless it would be essential to have a separate valve for this purpose. Whether figures such as I have quoted are practicable in laboratory conditions only, I am unable to say. But it was knowledge of the figures in this report which led me to discuss power factors of the order of 10^{-4}.

My article was a discussion of the action of a receiver in which the desired sensitivity is obtained by the use of a single receiving circuit. Such a system combined with tone correction is an alternative solution to the use of band-pass filters with or without superheterodyne. Its claim to consideration seems to me solely that of simplicity.

If experience shows that several ganged tuned circuits are necessary, then much of its claim to consideration seems to me to have been lost.

BOOK REVIEWS

Elements of Radio Communication, Second Edition, by John H. Morecroft, John Wiley and Sons, Inc., 1934. 286 pages, 241 figures. Price $3.00.

The second edition of *Elements of Radio Communication*, by the late Professor Morecroft, has been entirely reset and printed with smaller margins, making perhaps about one hundred more words per page than in the first edition, type size being the same.

A small amount of material giving obsolete practice has been deleted and a large amount of new material has been added. The new material deals with the new tubes and advances made in radio since the first, 1929, edition was printed. The first edition contained a paragraph or two on the screen-grid tube, but in 1929 it will be remembered that the screen-grid tube was, to a large extent, a talking point used by salesmen to sell broadcast receivers. This edition, like the first edition, deals for the most part with the broadcast field of radio. Short waves, the transatlantic radiotelephone, radio beacons and kindred subjects are discussed briefly.

One noticeable change of this edition in the first chapters is the large number of illustrative problems in the text. These problems are stated and detailed solutions are given illustrating the point in question. The mathematics involved in these problems is quite elementary. "The Elements" is written for those readers who can not master Professor Morecroft's standard text, *Principles of Radio Communication*. Besides the solved problems in the text, a large list of unsolved problems is given at the back of the book.

When one remembers Professor Morecroft's reputation as a teacher of and a writer on radio subjects, no higher praise can be given the book than to say it is a "Morecroft book." Perhaps this book represents his last work. The preface is dated October first, 1933.

<div align="right">*R. R. RAMSEY</div>

Signals and Speech in Electrical Communication, by John Mills. Harcourt, Brace and Company, 281 pages. Price $2.00.

By his extensive scientific training, wide engineering experience, and previous success in presenting technical subjects in understandable language, Dr. Mills is well qualified for the task he has undertaken. He has elected to convey his message by simple and effective words rather than by the usual mathematical and vector language of the engineer and as a result the quantitative element is necessarily lacking, however, he has succeeded well with the qualitative treatment. Although the historical aspects are not emphasized one obtains a fairly complete knowledge of the various milestones. Some of the chapter headings could be considerably improved, for example, "Vivisection of Speech" may be colorful but certainly not precise. It is more suggestive of the scalpel than a scientific analysis of speech. "Modulation, a Marriage of Currents" also seems inappropriate and if the analogy were carried too far it would probably lead to

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more confusion than clarity. These, of course, are minor criticisms. I do not hesitate to recommend the book which will be of interest both to engineers and the general public.

*H. M. Turner

Federal Radio Commission's Rules and Regulations (1934 Revision). Published by U.S. Superintendent of Documents, Washington D.C. 186 pages, loose leaf. Price 30 cents.

The newly published Rules and Regulations of the Federal Radio Commission are, in general, not greatly different from the first edition, effective February, 1932. They have been broadened to include regulations covering certain phases of radio regulation previously under the jurisdiction of the Radio Division of the Department of Commerce which was transferred to the Federal Radio Commission by executive order on July 20, 1932. The added regulations consist mainly of matters relating to the issuance of amateur and commercial operators' certificates although certain other additions are scattered here and there throughout the text. In addition the new regulations contain certain revisions made from time to time during the past two years to conform to new practices which experience has indicated as desirable and to meet the terms of the international agreement made at the North American Regional Conference held in Mexico City, July 10 to August 9, 1933, and the United States-Canadian broadcast agreement.[1]

The volume is divided into five main parts covering various phases of matters subject to government regulation.

Part I, General Rules and Regulations, covers the filing and amending of applications, certain general conditions under which applications will not be granted, regulations governing the use of radio facilities in emergencies, a table showing the normal license period of the various classes of radio stations, and a list of the radio field districts in the United States.

Part II, Practice and Procedure, deals with the routine established by the Commission for acting on applications, the rights of parties whose applications may be denied in whole or in part, the routine procedure in cases set for hearing or in revocation and suspension proceedings and matters relating to appeals from decisions of the Commission.

Part III, Broadcast Service, contains a compilation of the regulations dealing exclusively with stations engaging in this form of radio communication. These regulations are subdivided into the following sections:

Classes of Broadcast Stations	Equipment
Definitions	Technical Operation
Quotas of Facilities	Operation
Allocation of Facilities	Log Records

Part IV, Services Other than Broadcast, contains the detail regulations governing the operation of all classes of nongovernment radio stations except broadcasting. The first chapter of Part IV covers the general regulations relating to all of these stations and subsequent chapters cover regulations relating specifically to stations engaged in particular services. The chapter headings of Part IV are as follows:

General Regulations   Mobile Press Service
Fixed Services   Experimental Services
Aviation Service   Emergency Service
Coastal Services   Geophysical Service
Marine Relay Service   Temporary Services
Ship Service   Amateur Service

Alaska

Part V is a compilation of the regulations governing operators' licenses and outlines the conditions for securing and renewing such licenses. Following it there is included a copy of the Radio Act of 1927, as amended and annotated.

*R. D. CAMPBELL

* American Telephone and Telegraph Company, New York City.
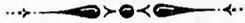
BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Copies of the publications listed on this page may be obtained gratis by addressing a request to the publisher or manufacturer.

Cunningham-Radiotron, 415 S. 5th Street, Harrison, N.J., has issued Application Note No. 37 on 100-volt operation of 6C6 and 6D6 tubes, No. 38 on a simple method for converting pentode characteristics, and No. 39 on the design of voltage supply for the 905 and 906 cathode ray tubes. A bulletin has also been issued on the type 6C6 triple-grid detector and amplifier tube.

A shadow tuning instrument is described in catalog Section 43-345 of the Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pa. In addition, a booklet on adjustment and operation of modulators with rectox and thermal instruments is available.

A socket operated sound system is described in a leaflet published by the Miles Socket Mike Company of 244 West 23rd Street, New York City.

The Acheson Oildag Company of Port Huron, Mich., has issued technical Bulletin No. 191.1 on colloidal graphite—an ideal ray focusing anode material for cathode ray tubes.

Measuring instruments suitable for radio purposes are described in a leaflet issued by the Triplett Electrical Instrument Company of Bluffton, Ohio.

A set of engineering news letters issued by Hygrade Sylvania Corporation of Emporium, Pa., which was started in November, 1932, has just come to our attention. The contents are as follows: (1) November, 1932, an improved 41 output tube, class A prime with pentodes, the development of automobile radio; (2) December, 1932, experimental data on the 79, rectifier tubes for special service; (3) January, 1933, new tubes for new receivers; (4) April, 1933, filter circuits for the 25Z5 voltage doubling service, application of types 2B7 and 6B7, amplified delayed AVC with double diode triodes; (5) September, 1933, some problems in ac-dc radio receivers, phase inversion with the 79; (6) October, 1933, automatic volume control circuits; (7) January, 1934, oscillator performance with pentagrid converters; and (8) February, 1934, resistor data for self-biased tube operation. Technical information booklets are also available for the following types of tubes: 1V, 25Z5, 2A3, 6C6, 6D6, 43, 5Z3, 2A5, 18, 2A7, 6A7, 2B7, 6B7, 1A6, 6A4, 6F7, 53, and 76.

The Earl Webber Company, Daily News Building, Chicago, Ill., has issued several leaflets describing its radio tube testers and measuring equipment.

Microphones and piezo-electrical crystal holders are described in a leaflet issued by Eastern Coil Company of 56 Christopher Avenue, Brooklyn, N.Y.

The RCA Victor Company of Camden, N.J., has available for distribution booklets on the following equipment: Model AVT-1 airport radio traffic control transmitter, model AVR-3 aircraft communication receiver, TMV-52-E beat frequency oscillator having a range from 20 to 17,000 cycles, TMV-18-D standard signal generator with range from 25 kilocycles to 25 megacycles, TMV-36 universal curve recorder suitable for over-all loud speaker measurements, TMV-69-A wire testing equipment, TMV-107 high frequency modulated-signal system for the distribution of signals from two to twenty megacycles

over a common circuit for test purposes, TMV-75-B field intensity meter with range from 500 to 20,000 kilocycles, and TMV-97 portable test oscillator covering from 150 to 25,000 kilocycles.

Magnetic and dynamic speakers are described in Bulletin S2343 issued by the Arlab Manufacturing Company of 1250 N. Paulina Street, Chicago, Ill.

The J. W. Miller Company, 5917 S. Main Street, Los Angeles, Calif., has issued leaflets describing its service test oscillator, other test equipment, and superheterodyne coil kit.

Part 2 of Catalog G of the General Radio Company, 30 State Street, Cambridge, Mass., describes components and measuring equipment.

Radio replacement products, chiefly transformers, are described in Catalog 341R published by the Jefferson Electric Company of Bellwood, Ill.

Leaflet No. 2167 of the Allis-Chalmers Manufacturing Company, Milwaukee, Wis., describes a metal-clad grid-control mercury arc power rectifier suitable for high voltage direct-current power supplies for radio transmitting stations.

Premier Crystal Laboratories, 53 Park Row, New York City, has issued Bulletins No. 100 on inductance and capacitance measurements, 101 on constant frequency control equipment, and 102 on a visual capacity meter.

Insulators made from lava, alsimag, alumina, beryllin, and magnesia are described in Bulletin 34 issued by the America Lava Corporation, Chattanooga, Tenn.

Ward Leonard Electric Company of Mt. Vernon, N.Y., describes in its Bulletin 2501 a line of battery charging rheostats and resistors.

The Kathetron out-voltage regulator is described in Catalog 10 of the Roller-Smith Company, 233 Broadway, New York City.

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Janco, Nathan: Received B.S. degree in mechanical engineering, New York University, 1932; M.S. degree, 1934. Graduate assistant in mechanical engineering, New York University, 1932–1934. Nonmember, Institute of Radio Engineers.

Kear, Frank G.: Born 1903 at Minersville, Pennsylvania. Received E.E. degree, Lehigh University, 1926; S.M. degree, Massachusetts Institute of Technology, 1928; Sc.D. degree, 1933. Instructor and member of research staff, Massachusetts Institute of Technology, 1926–1928. Assistant physicist, United States Bureau of Standards, 1928–1930; associate physicist working on radio aids to air navigation, 1930–1933. Instructor and graduate student, Massachusetts Institute of Technology, 1931–1932. Physicist, Washington Institute of Technology, 1933 to date. Associate member, Institute of Radio Engineers, 1924; Member, 1931.

Mohammed, Aijaz: Born 1904 at Jaipur, India. Graduated, Aligarh University, 1926. Studied electron theory, X-rays, and wireless telegraphy at Lucknow. Received M.S. degree, 1929; research work, communication engineering department, Indian Institute of Science, Bangalore, India, 1930 to date. Associate member, Institute of Radio Engineers, 1931.

Palermo, A. J.: Born May 16, 1899, at Schenectady, New York. Received B.S. degree in electrical engineering, Union College, 1922; M.S. degree, 1924, PhD. degree, 1931. Instructor in electrical engineering and mechanics, Union

College, 1922 to date. Consultant in Italian and French patents for International General Electric Company. Member, Sigma Xi. Associate member, Institute of Radio Engineers, 1930.

Scott, Earl D.: Born 1906 at Ellensburg, Washington. Received B.S. degree, University of Washington, 1932; M.S. degree, 1933. Marine radio, 1925–1932. Teaching assistant, University of Washington, 1932–1933. Student member, Institute of Radio Engineers, 1932.

Taylor, Paul K.: Born September 27, 1902, at Buffalo, New York. Received B.A. degree, University of Rochester, 1924; M.S. degree, Northwestern University, 1926; Ph.D. degree, 1931. Instructor in physics, Northwestern University, 1924–1928, and 1930–1931; assistant professor of physics, Furman University, 1928–1930; professor of physics, DePaul University, summer of 1931; professor of physics and instructor in mathematics, State Teachers College, Minot, North Dakota, 1931 to date. Member, American Physical Society and Sigma Xi. Nonmember, Institute of Radio Engineers.

Verman, Lal C.: Born October 2, 1902, at Amritsar, India. Received B.S. degree in electrical engineering, University of Michigan, 1927; M.S. degree, Cornell University, 1928; Ph.D. degree in physics, 1930. Research assistant in engineering research department, sound division, University of Michigan, 1925–1927; Donnovan Scholar, University of Michigan, 1925–1927; Charles A. Coffin Fellow (General Electric Company), Cornell University, 1927–1930. Heckscher Research Assistant, Cornell University, 1930–1931; Research Fellow, Indian Institute of Science, Bangalore, 1932–1933; Lac research physicist, London Shellac Research Bureaux, London, 1933 to date. Associate member, American Institute of Electrical Engineers. Associate member, Institute of Radio Engineers, 1928.

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XI

When writing to advertisers mention of the PROCEEDINGS *will be mutually helpful*

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)

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(Address for mail)

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(Date)   (City and State)

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Address ....   Address ....

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Mr. ....

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The following extracts from the Constitution govern applications for admission to the Institute in the Associate grade:

ARTICLE II—MEMBERSHIP

Sec. 1: The membership of the Institute shall consist of: * * * (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

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..........

Alphabetical Index to Advertisements

INSTITUTE SUPPLIES

EMBLEMS

Three styles of emblems are available to members of the Institute only. They are all of 14k gold with gold lettering on an enameled background, the color of which indicates the grade of membership. The approximate size of each emblem is as illustrated.

The lapel button is supplied with a screw back having jaws which fasten it securely to the coat. The price is $2.75, postpaid, for any grade.

The pin which is provided with a safety catch may be obtained for any grade for $3.00, postpaid.

The watch charm is handsomely finished on both sides and is equipped with a suspension ring for attaching to a watch chain or fob. Price for any grade, $5.00, postpaid.

BINDERS
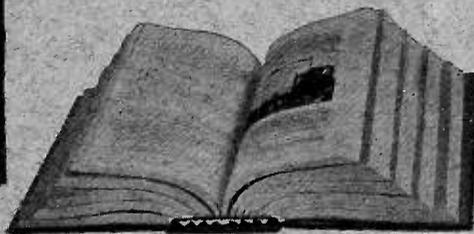
Binders are available for those who desire to protect their copies of the PROCEEDINGS with stiff covers. Each binder will accommodate the twelve monthly issues published during the year. These binders are of blue Spanish Grain Fabricoid with gold lettering, and will serve either as temporary transfers or as permanent binders. They are so constructed that each individual copy of the PROCEEDINGS will lie flat when the pages are turned. Copies can be removed from the binder in a few seconds and are not damaged by their insertion. They are priced at $1.50 each, and the member's name or PROCEEDINGS Volume Number will be stamped upon the binder in gold for $0.50 additional.

BACK ISSUES OF THE PROCEEDINGS

Complete volumes of back issues of the PROCEEDINGS are available in unbound form for the years 1918, 1920, 1921, 1922, 1923, and 1926 at $6.75 per year (six issues). Also complete unbound for 1931, 1932 and 1933 at $7.50 per year. A list of the available miscellaneous copies published from 1913 on may be obtained from the Secretary. Single copies published prior to 1927 are sold to members at $1.13 each, and those issued during and after 1927 are $0.75 each. Foreign postage on single copies is $0.10 per copy, making $0.60 on unbound volumes before 1927.

Issues for 1922 may be obtained bound in blue buckram at $8.75 per volume. Similar bound volumes for 1930, 1931, 1932, and 1933 are priced at $9.50 each. Foreign postage is $1.00 additional per volume. Volumes for 1922 and 1926 bound in Morocco leather are available at $10.25 each; volumes for 1930 and 1931 are $11.00 each.

TWENTY-TWO-YEAR INDEX

An Index of the PROCEEDINGS for the years 1909-1930, inclusive, is available at $1.00 per copy. It is extensively cross indexed. The earlier references are to the PROCEEDINGS of the Wireless Institute; the PROCEEDINGS of the Institute of Radio Engineers has been published since 1913.

When ordering any of the above, send remittance with order to the Secretary, Institute of Radio Engineers, 330 West 42nd Street, New York City.

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General Radio frequency-measuring equipment encompasses all requirements, from national standards of frequency to amateur station monitors.

STANDARDS

Primary standards for direct standardization of frequency against time signals or star transits. The primary standard will produce frequencies (or time intervals) having a precision of one part in five million. The secondary standard, calibrated against a primary standard, produces frequencies having a precision of twenty parts per million.

COMPARISON EQUIPMENT

Comparison equipment is available for a wide range of requirements. It includes interpolation oscillator and heterodyne frequency meters for use with the standards, and wavemeters of suitable grades of precision for all laboratory and commercial requirements.

MONITORS

Equipment for continuous monitoring of frequency is also available. A direct-reading precise monitor for broadcast stations has recently been improved and is now entirely a.c. operated. A new monitor for high-frequency channels will be announced shortly.

Bulletin 10-R describes frequency-measuring methods and equipment. A copy will be sent on request.

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