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PROCEEDINGS
of
The Institute of Radio Engineers

1930 CONVENTION
Toronto, Ont., Canada
August 18, 19, 20, and 21

Form for Change of Mailing Address or Business Title on Page XLIX

Fifth Annual Convention

of the

Institute of Radio Engineers

Toronto, Ontario, Canada

August 18, 19, 20 *and* 21

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 18 — June, 1930 — Number 6

Contents

Part I

Page

Frontispiece, Donald McNicol, President of the Institute, 1926 904
Institute News and Radio Notes 905
  May Meeting of Board of Direction 905
  Radio Signal Transmissions of Standard Frequency 905
  Committee Work 906
  Institute Meetings 907
  Personal Mention 910

Part II

Technical Papers

Some Problems in Short-Wave Telephone Transmission .. J. C. SCHELLENG 913
A 12-Course Radio Range for Guiding Aircraft with Tuned-Reed Visual Indication .... H. DIAMOND and F. G. KEAR 939
A Tuned-Reed Course Indicator for the Four- and Twelve-Course Aircraft Radio Range .... F. W. DUNMORE 963
Single- and Coupled-Circuit Systems .... E. S. PURINGTON 983
The Establishment of the Japanese Radio-Frequency Standard. Y. NAMBA 1017
The Amplification and Detection of Extra-Short Electric Waves .... KINJIRO OKABE 1028
The Radio Engineer and the Law .... PAUL M. SEGAL 1038
Note on Variations in the Amplification Factor of Triodes .... F. E. TERMAN and A. L. COOK 1044
Radiotelegraphy and Radiotelephony on Half-Meter Waves .... SHINTARO UDA 1047
The Effect of Rain and Fog on the Propagation of Very Short Radio Waves .... J. A. STRATTON 1064
Meteorological Influences on Long-Distance, Long-Wave Reception .... E. YOKOYAMA and TOMOZO NAKAI 1075
Book Review, "The Radio Manual" .... S. S. KIRBY 1084
Booklets, Catalogs, and Pamphlets Received 1085
Monthly List of References to Current Radio Literature 1088
Contributors to this Issue 1094

The Institute of Radio Engineers

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	Maplewood, 99 Tuscan Road	Poppele, J. R.
	Newark, 108 Schofield St.	Meyer, H. E.
	New Brunswick, c/o National Air Transport, Inc., Hadley Field	Barker, F. C.
	Orange, 651 Lincoln Ave., Apt. 305	Maul, G. E.
	Orange, 674 Scotland Rd.	Reifel, Harry
	Passaic, 50 Park Ave.	Webster, R. G.
New York	Brooklyn, 25 Billings Pl.	Sobel, A. D.
	Brooklyn, U. S. Naval Radio Station, Navy Yard	Tallman, Clare Le Roy
	Buffalo, 345 Shirley Ave.	Cameron, Richard.
	Buffalo, 401 Forest Ave.	Meek, D. A..
	Elmhurst, L. I., 8642-57th Road	Content, E. J.
	Kenmore, 171 W. Girard Ave.	Sheets, H. M.
	New Rochelle, 29 Burling Lane	Cantor, A. B.
	New York City, 138 E. 38th St.	Bylander, J. C.
	New York City, Century Hotel, 111 W. 46th St.	Conley, S. D.
	New York City, Bell Tel. Labs., 463 West St.	Custer, C. J.
	New York City, 17 W. 60th St., Room 1101	Haldane-Duncan, B.
	New York City, USS Saratoga, c/o Postmaster	Honeycutt, V. C.
	New York City, 25 South St., Box No. 1080	Jordan, C. H.
	New York City, 26 Broadway, SS A. S. Bradford, c/o Standard Shipping Co.	McCarroll, G. M.
	New York City, 67 Broad St., c/o Mackay Radio Telegraph Co., Inc.	Miles, P. D.
	New York City, 152 West 42nd St., c/o S. S. White Dental Mfg. Co.	Smack, J. C.
	New York City, 326 East 155th St.	Speed, R. B.

New York (cont.)	New York City, 865 East 167th St.	Weintraub, D. H.
	New York City, 18 East 3rd St.	Wood, R. C.
	Pelham Manor, 1333 Manor Circle	Lewis, F. W., Jr.
	Rochester, 822 University Ave.	Proctor, Benj., 3rd.
	Rocky Point, Radio Corporation of America	Wooldridge, A. C.
	Scarsdale, 85 Alkamont Ave.	Weis, E. M.
	Schenectady, 6 Jackson Place	Rohner, A. J.
Ohio	Akron, 59 Hawthorne Ave.	Myers, W. H.
	Cleveland, National Air Transport, Municipal Airport	Clemans, H. L.
	Cleveland, National Air Transport	Costa, Humbert
	Columbus, 111 West 11th Ave.	Newhouse, R. C.
	East Cleveland, 15987 Nelamere Road	Williams, H. R.
	Medina	Kohli, H. J.
Oklahoma	Oklahoma City, 1110 West 2nd St.	Spooner, A. J.
Oregon	Eugene, 492 West Broadway	Du Sair, P. E.
Pennsylvania	Philadelphia, 1817 Fulmer St.	Berrien, P. H.
	Philadelphia, 1624 E. Eyre St.	Bloom, Walter
	Philadelphia, 128 S. Salford St.	Podolsky, Leon
	Philadelphia, 4910 Stenton Ave.	Williams, A. J., Jr.
	Pittsburgh, 929 Penn Ave.	Clary, Howard L.
	Upper Darby, 264 Wembley Road	Lenhert, G. R., Jr.
	Wilmerding, Faller Bldg., Apt. 25	Irlam, William
Rhode Island	Providence, 204 Bowen St.	Huddy, Franklin S.
Texas	Brownsville, Cia. Mexicana de Aviacion, S. A.	Fernandez, Manuel
	Dallas, Box No. 1151	Eilert, E. F.
	Dallas, Box No. 181, S. M. U.	Tucker, D. J.
Washington	Everett, 2511 East Grand	Malmstrom, W. H.
Wisconsin	Milwaukee, 672 Van Buren St.	Gebhart, B. R.
	Platteville, R. R. No. 1	Bentz, C. F.
	Sauk City	Schmitz, A. J.
Canada	Kenora, Ont., Box No. 135	Doan, G. F.
	Montreal, Que., 173 William St.	Bailey, F. A. A.
	Montreal, Que., 637 Craig St., West	Higgins, H. H.
	Montreal, Que., 4922 Sherbrooke St., West	White, W. E.
	Ste. Therese, Que., Box No. 111	Farmer, E. W.
Canal Zone	Coco Solo, U. S. Naval Air Station	Pritchard, C. A.
Colombia	Cúcuta, Estación Inalambrica	Caicedo, A. D.
England	Colchester, Essex, 13 Salisbury Ave.	Wyborn, R. B.
	Liverpool, 86 Canning St.	Worall, E. J. B.
	London, 599 High Ro Leyton	Summers, R. H. J.
	Sale, Nr. Manchester, 17 Hulme Road	Critchley, W. H.
	Thornton Heath, Surrey, 89 Parchmore Road	Hall, R. C.
	Thornton Heath, Surrey, 66 Ingram Road	Thomas, L. H.
Philippine Islands	Manila, c/o Radio Division, Bureau of Posts	Soriano, Alfonso
Scotland	Edinburgh, 22 Walker St.	Kerr, John
South Africa	Umbilio, Durban, Natal, 23 Francois Road	Thomson, H. J.

For Election to the Junior grade

Illinois	Chicago, 801 S. Marshfield Ave.	Nichols, L. R.
Michigan	Flint, 2208 Cadillac St.	Blanford, E. K.
Ohio	Akron, 188 Gale St.	Branch, G. E.
Canada	Toronto, Ont., 345 Adelaide St., West	Blackburn, Wesley
	Winnipeg, Man., 205 Cambridge St.	Sinclair, D. B.

OFFICERS AND BOARD OF DIRECTION, 1930

(Terms expire January 1, 1931, except as otherwise noted)

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Donald McNicol

President of the Institute, 1926

Donald McNicol was born in Hopetown, Ontario, Canada, July, 1875.

He was a pioneer in radio research and experimentation, and had a transmitting and receiving system in operation in Minneapolis as early as 1900. His first important published work was an illustrated serial on "Wireless or Radio Telegraphy" which was published in the *Western Electrician* during 1906 and 1907.

For twenty years he was in the service of the land-line telegraph companies and wrote three books on telegraph engineering which have been used very widely. One of these was translated into French.

In 1922 he became assistant to the president of the Radio Corporation of America, and a year later engaged in consulting work in communication engineering, serving a number of engineering and manufacturing companies. In addition, he became vice president of a company manufacturing terminal-office equipment.

He was recently appointed editorial director of the publications *Radio Engineering* and *Projection Engineering*, and his "The Engineering Rise in Radio," which was published in *Radio Engineering*, will undoubtedly be remembered by many.

Mr. McNicol became an Associate member of the Institute in 1914, advancing to the Member grade during the same year. In 1924 he became a Fellow. From 1919 to 1924 he was a manager of the Institute, in 1925, vice president, and President in 1926. He was chairman of the Standardization Committee in 1921–1924.

He is a Fellow of the American Institute of Electrical Engineers, and has been a member of the Publications Committee, and for five years chairman of the Communications Committee of that Institute.

INSTITUTE NEWS AND RADIO NOTES

May Meeting of Board of Direction

A meeting of the Board of Direction was held at 4 P. M., May 7, 1930, the following being present: Alfred N. Goldsmith, acting chairman; J. H. Dellinger, R. A. Heising, J. V. L. Hogan, C. M. Jansky, Jr., R. H. Manson, R. H. Marriott, A. F. Van Dyck, and H. P. Westman, secretary.

The following were elected to the Member grade: F. A. Cowan, Eugene Peterson, W. G. Eaton, Ernst Lübcke, and W. D. L. Starbuck. The following were transferred to the Member grade: A. P. Bock, H. M. Booth, E. J. H. Buzzard, J. A. Chambers, R. C. Hitchcock, D. D. Israel, C. E. Kilgour, Charles W. Peterson, and P. S. Scofield.

One hundred and seventy Associate and nine Junior members were elected.

Board of Editors

A meeting of the Board of Editors was held at 2 P. M., May 2, 1930, at the offices of the Institute. Those present were Alfred N. Goldsmith, editor; J. W. Horton, L. E. Whittemore, W. Wilson, and H. P. Westman, secretary.

Associate Application Form

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

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

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

Radio Signal Transmissions of Standard Frequency July to December, 1930

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

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

Eastern Standard Time	July 21	Aug. 20	Sept. 22	Oct. 20	Nov. 20	Dec. 22
10:00 P.M.	1600	4000	550	1600	4000	550
10:12	1800	4400	600	1800	4400	600
10:24	2000	4800	700	2000	4800	700
10:36	2400	5200	800	2400	5200	800
10:48	2800	5800	1000	2800	5800	1000
11:00	3200	6400	1200	3200	6400	1200
11:12	3600	7000	1400	3600	7000	1400
11:24	4000	7600	1500	4000	7600	1500

Committee Work

Committee on Admissions

A meeting of the Committee on Admissions was held at 1 P. M., May 7, 1930, at the offices of the Institute. The following were present: R. A. Heising, chairman, C. M. Jansky, Jr., A. V. Loughren, R. H. Marriott, and A. F. Van Dyck.

The committee considered fifteen applications for transfer or election to the higher grades of membership in the Institute.

Committee on Broadcasting

The following were present at the meeting of the Committee on Broadcasting held at 10 A. M., May 7, at the offices of the Institute: L. M. Hull, chairman, P. A. Greene, J. V. L. Hogan, C. W. Horn and C. M. Jansky, Jr.

Committee on Constitution and Laws

A meeting of the Committee on Constitution and Laws, held at 10 A. M., May 7, at the offices of the Institute, was attended by R. H. Marriott, chairman, H. E. Hallborg, R. A. Heising, and G. W. Pickard.

Committee on Membership

A meeting of the Committee on Membership was held at 7 P. M., April 16, with the following present: I. S. Coggeshall, chairman; F. R. Brick, B. Dudley, assistant secretary, H. P. Gawler, C. R. Rowe, A. M. Trogner and H. P. Westman, secretary.

Another meeting of this Committee was held at 5:30 P. M., May 7, being attended by I. S. Coggeshall, chairman; F. R. Brick, H. C. Gawler, S. R. Montcalm, A. F. Murray, C. R. Rowe, J. E. Smith and A. M. Trogner.

Committee on Nominations

On May 2, at 4 P. M., a meeting of the Committee on Nominations was held at which Dr. Alfred N. Goldsmith, chairman; W. R. G. Baker, Melville Eastham, Donald McNicol and H. P. Westman, secretary, were present.

Committee on Sections

A meeting of the Committee on Sections was held at 12:30 P. M., April 16, at 195 Broadway, Room 1531, New York City. The following were present: Austin Bailey, chairman; C. W. Horn, B. E. Shackelford, B. Dudley, assistant secretary, and H. P. Westman, secretary.

Committee on Standardization

A meeting of the Executive Committee of the Sectional Committee on Radio, operating under the American Standards Association procedure, and the Executive Committee of the Committee on Standardization of the Institute, was held at the offices of the Institute at 2 P. M., May 7, 1930. Those present were: J. H. Dellinger, chairman of the I. R. E. Committee on Standardization, acting chairman, Alfred N. Goldsmith, chairman of the Sectional Committee on Radio, E. T. Dickey H. E. Farrar, H. A. Frederick, R. H. Manson, Haraden Pratt, C. H. Sharp, J. C. Warner, L. E. Whittemore, Irving Wolff, B. Dudley, secretary, committee on Standardization, and H. P. Westman, secretary, Institute of Radio Engineers.

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

A meeting of the Technical Committee on Radio Receivers of the Institute was held at the offices of the Institute at 10 A. M., May 8, 1930. In addition to the chairman, E. T. Dickey, the following were present: C. M. Burrill, M. Ferris, E. J. T. Moore (representing V. M. Graham), F. X. Rettenmeyer, R. S. Shankland, (representing H. Diamond), and B. Dudley, secretary.

Institute Meetings

Buffalo-Niagara Section

Thirty-one members and visitors attended the April 16 meeting of the Buffalo-Niagara Section which was held at the University of Buffalo, Dr. L. Grant Hector, chairman, presiding.

A paper on "The Suppression of Interference" was presented by H. J. Klumb, Director of the Electrical Laboratory of the Rochester Gas and Electric Corporation.

An interesting discussion on the general subject of interference and its suppression ensued.

New York Meeting

The regular monthly meeting of the Institute was held in the Engineering Societies Building, 33 West 39th Street, New York City, on Wednesday, May 7, 1930. R. H. Marriott, member of the Board of Direction, presided.

The paper of the evening was presented by O. H. Caldwell, editor of *Electronics*, published by McGraw-Hill Publishing Company, and was entitled "Radio's Contribution to Modern Civilization." It was summarized as follows:

> This paper undertakes to outline and summarize the almost countless ways in which the development of radio has affected human progress and civilization, during the past quarter century—both directly as space-radio and ether-wave communication, and also through the many non-radio applications of the electronic tube and associated circuits in their amazing variety of uses, throughout industry, power distribution, medicine and therapeutics, and the whole gamut of the arts and sciences.
>
> Applications of the vacuum tube to human safety, in both war and peace, comprise one of the most significant human uses of this new electronic tool. These safety uses of electronic devices are already found in many forms—in navigation, aviation, street traffic, railroad signals, protection of industrial machinery, warnings against charged electrical apparatus, etc. The Army and Navy make numerous uses of electron-tube devices for safety purposes, throughout all arms of the services. The author offers some speculation as to the important war-time value of our existing system of broadcast stations for general emergency communication throughout the nation in the event of another war. These stations are already pressed into service during regional emergencies, and this use might be extended, even in peace time.
>
> The paper mentions many novel uses of devices derived from radio research, for example,—new musical instruments; alarms; detection of larvae in fruit; diagnosis of heart murmurs; talking pictures as wills and legal documents; the "talking book"; aiding blind to read ordinary print; sound pictures with acoustic depth; structure analysis; food preservation; emotion recorders; "lie detectors"; etc.
>
> The paper closes with a listing of nearly two hundred principal classes of applications of radio and electronic tubes, nearly all of which are subject to further development and wide subdivision.
>
> In short, it is declared that both the radio and non-radio applications of the vacuum tube are just beginning to find uses, and the prediction is made that in a few years there will be nothing that the average man hears, sees, or buys that will not be controlled, regulated, or affected in some important respect by an electronic tube.

Approximately two hundred members and guests of the Institute attended.

Rochester Section

A joint meeting of the Rochester Sections of the Institute of Radio Engineers, American Institute of Electrical Engineers, American Society of Mechanical Engineers, American Society of Civil Engineers, American Chemical Society, and the Optical Society of America was held at the Rochester Chamber of Commerce on April 14th, M. Herbert Eisenhart of the Rochester Engineering Society, presiding.

R. A. Millikan spoke on "Fire or Relation of Science to Mankind." He treated this subject in a popular manner so as to interest the many guests who were present, among whom were many ladies. On the same evening the Eighth Annual Banquet of the combined Engineering Societies of Rochester was held.

The regular April meeting of the Section was held at the Sagamore Hotel on the 17th, H. J. Klumb, presiding. The seventy members and guests in attendance heard Professor H. B. Smith of Worcester Polytechnic Institute deliver an illustrated lecture on "The Quest of the Unknown."

Professor Smith's paper concerned itself primarily with his experience in the design of very high voltage transformers and insulators.

Toronto Section

A paper by E. B. Ferrell of the Bell Telephone Laboratories on "High Frequency Telephone and Telegraph Communication" was presented at the April 9th meeting of the Toronto Section.

This paper described the New York-London and New York-Buenos Ayres communicating system and was illustrated with slides covering many important details of the work. The discussion that followed was participated in by Messrs. Carruthers, Fox, Hackbusch, Leslie, Oxley, Pipe, and Shane.

Sixty-five members and guests were in attendance.

Washington Section

Dr. L. P. Wheeler of the U. S. Naval Research Laboratory delivered a paper on "Master Circuit of Crystal-Control Transmitters" at the April 10th meeting held at the Hotel Continental, T. McL. Davis, presiding.

The paper was illustrated and considered the fundamental crystal-controlled circuits used for high-frequency transmitters. Provisions for changing connections to any one of a number of crystals to allow the frequency of the transmitted wave to be changed was described. The important considerations for obtaining frequency stability were analyzed and methods for obtaining best results were discussed.

One hundred and two members and guests attended the meeting.

Personal Mention

Keith Henney, previously associated with *Radio Broadcasting*, has become associate editor of *Electronics*, the new publication issued by McGraw-Hill Publishing company.

R. K. Bonell is now an engineer at the Bell Telephone Laboratories. He was formerly with the American Telephone and Telegraph Company.

R. O. Brooke has recently been appointed chief engineer of WRHN at Minneapolis.

J. R. Donovan is now broadcast engineer at WOAN, Lawrenceburg, Tenn., previously being chief engineer of WCOC at Meridian, Miss.

R. E. Downing has been shifted from the Elmira, N. Y. branch of the American Telephone and Telegraph Co. to the Department of Development and Research, in New York City.

Previously chief engineer of the Sonora Phonograph Company, R. M. Dunning has joined the Research Engineering Department of the Thomas A. Edison Company, at West Orange, N. J.

C. A. Ellert is now U. S. Radio Inspector at Fort McHenry, Baltimore. He was formerly an engineer at the East Pittsburgh plant of the Westinghouse Electric and Manufacturing Company.

Henry W. Baukat, who was technical editor of *Radio Retailing*, is now with the RCA Radiotron Company at Harrison, N. J.

Manuel Escolano, formerly with the Compañía Nacional de T. S. H. of Madrid, is now director general of the Compañía Telmar of Madrid.

F. Clifford Estey is now sales engineer in the Radio Division of the Aluminum Company of America. He was previously assistant to the President of the United Reproducers Company at Springfield, Ohio.

J. T. Hallam is now an engineer at WOAI in San Antonio, Texas, and was formerly assistant engineer at the West Virginia Broadcasting Corporation at Wheeling.

R. K. Hansen is now engineer for RCA Radiotron Corporation at Harrison, N. J. He leaves the Westinghouse Lamp Company of Bloomfield, N. J.

N. P. Hinton, who was in the engineering department of the Metropolitan Vickers Company of Manchester, England, is now with the British Thomson-Houston Co. at Coventry.

At the Constant Frequency Monitoring Station at Grand Island, Nebraska, we find G. L. Jensen who was previously a junior radio inspector of the Department of Commerce, New York.

E. V. Keshimer, who was formerly engineer for the United Reproducers Corporation, is now an inspection engineer with the Crosley Radio Corporation at Cincinnati.

W. W. Lindsay, Jr., who was a research engineer with the Movietone Department, William Fox Studios, is now recording engineer for the Fox Film Corporation, at the Fox Hills Studio, Beverly Hills, Cal.

D. S. Little has been made superintendent of the Great Lakes Division of the Radiomarine Corporation and is located at Cleveland, Ohio.

E. C. Manderfeld has joined Electrical Research Products, Inc., as engineer, Sound Recording Department, having previously been a research engineer at the Bell Telephone Laboratories.

H. G. Miller, formerly radio engineer for the Westinghouse Electric and Manufacturing Co. at East Pittsburgh, is now a receiving engineer for the Jenkins Television Corporation of Jersey City, N. J.

J. N. Mogridge, who was manager of the W. J. Crogsdill Sales Company, has recently been made chief engineer of CKPC at Preston, Ont., Canada.

H. P. Morris has been transferred from the Marion station of RCA Communications to the Rocky Point Station.

W. R. G. Baker, formerly managing engineer of the General Electric Company, is now vice president of the RCA-Victor Company at Camden, N. J.

B. Ray Cummings, formerly of the General Electric Company, is now in charge of the Transmitter Division of the Engineering Department of the RCA-Victor Company at Camden.

Previously with the Victor Talking Machine Company at Langhorne, Pa., W. H. Newbold is now in the Development Department of the RCA-Victor Company at Camden.

H. J. Nichols has been made manager of the Radio Engineering Department of the Chicopee Falls plant of the Westinghouse Electric and Manufacturing Company.

N. J. Oman is now a member of the technical staff of the Bell Telephone Laboratories having left the Brandes Laboratories.

Previously chief engineer of the Radio Corporation of the Philippines at Manila, D. S. Rau is now engineer in charge, RCA Communications, Inc., Marion, Mass.

E. W. Ritter, who was formerly at the Cleveland Vacuum Tube Works of the General Electric Company, is now with the RCA Radiotron Company at Harrison, N. J.

H. L. Roosevelt, formerly European manager of the Radio Corporation of America at Paris, is now located at the 233 Broadway office in New York City.

George Rodwin, formerly an engineer for the Radio Corporation of America, is now a member of the technical staff of the Bell Telephone Laboratories.

Previously a radio engineer with the Westinghouse Lamp Company Bloomfield, N. J., R. L. Schoene has become radio tube engineer for the U. S. Radio and Television Corporation., Marion, Ind.

From the Television Division of the General Electric Company at Schenectady, D. W. Short has come to the Jenkins Television Corporation.

Zangwill Ward has left the Radio Corporation of American to become a radio engineer for the Radio Administration, Board of Communications, Shanghai, China.

R. R. Winans, previously a radio engineer for the Gold Seal Manufacturing Company, has joined the staff on the American Bosch Magneto Corporation at Springfield, Mass.

J. Wienberger, formerly Division Engineer at the Van Cortlandt Park Laboratory of the Radio Corporation, is now engineer in charge of research of RCA Photophone, in New York City.

Sidney Bloomenthal, formerly of the Research Department at the Van Cortlandt Park laboratory of the Radio Corporation of America, is now a physicist for the RCA-Victor Company.

H. E. Reeber is now with the RCA-Victor Company, previously being at the Van Cortlandt Laboratory of the Radio Corporation.

The following engineers who were formerly with the General Electric Company are now in the engineering department of the RCA-Victor Company at Camden: H. R. Batchelor, C. F. Coombs, A. H. Demmer, E. W. Enstrom, C. W. Frank, W. J. Poch, Robert Serrell, J. A. Terrell, and G. G. Thomas.

This list gives those engineers who were formerly with the Westinghouse Electric and Manufacturing Company at their East Pittsburgh, Pa., and Chicopee Falls, Mass., plants, who are now in the engineering department of the RCA-Victor Company: R. C. Ballard, G. L. Beers, R. W. Carlisle, M. E. Karns, D. Stayer, E. F. Sutherland, J. Q. Tiedje, and M. C. Walker.

P. R. Cunliffe, formerly at the Bloomfield plant of the Westinghouse Electric and Manufacturing Company, is now at the Harrison plant of the RCA Radiotron Company.

R. D. Cunningham, formerly secretary of the Pittsburgh Section, has joined the Photophone and Applications Division of the RCA-Victor Company, going from the Westinghouse Electric and Manufacturing Co. plant at East Pittsburgh, Pa.

PART II

TECHNICAL PAPERS

Proceedings of the Institute of Radio Engineers
Volume 18, Number 6 *June, 1930*

SOME PROBLEMS IN SHORT-WAVE TELEPHONE TRANSMISSION*

By
J. C. Schelleng
(Bell Telephone Laboratories, New York City)

Summary—*In this paper are discussed certain phases of short-wave telephony, primarily, though not entirely, from the point of view of the transmitter. The field strengths which the transmitting station must provide at the receiver are considered. Typical data are given showing results obtained in transmission from Deal, New Jersey, to England. This is followed by a discussion of requirements and limitations of the transmitting antenna. The gains which arrays may reasonably be expected to provide are considered. The phenomenon of non-synchronous fading at nearby points is examined as to its bearing on the dimensions and performance of directive arrays. Other directional properties of the transmitting medium are also considered. Attention is then directed to the transmitting equipment, particular attention being given to the high-power part of it. Requirements, rather than circuit details, are emphasized. These include stability of operation, flexibility, and freedom from amplitude distortion, and phase and frequency modulation. The results of tests in which some of these matters were considered quantitatively are given.*

Introduction

Since May, 1927, short-wave telephony has been one of the commercial means of communication between this country and Europe. Beginning as a support for the West-to-East long-wave circuit during the difficult summer months when experimental equipment was called upon to carry traffic, short waves have been brought into use by the Bell System and the British General Post Office to provide three independent two-way channels.

Undertakings of this sort necessarily involve several groups working on different aspects of the problem. It is the purpose of this paper to touch upon a few of the questions which have arisen in the course of the work of only one of these groups. We shall consider certain aspects of short-wave transmission, the efficiency of transmitting antennas and the requirements which transmitters must meet. It is not the purpose to enter here into a detailed discussion of any of these matters, since this will be done in other papers now being prepared by various authors.

Transmission

General

The electric field strength which the short-wave transmitting station should be called upon to lay down at the receiver depends on

* Dewey decimal classification: R412. Presented at New York meeting of the Institute, November 4, 1929.

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Transmission

General

The electric field strength which the short-wave transmitting station should be called upon to lay down at the receiver depends on

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so many variables that a simple statement specifying it cannot be made. However, it has been found that a value in the general region of ten μv per m has given useful results in transatlantic work. Telephone communication can sometimes be established on days when radio noise is low with considerably less than 1 μv per m while on days of extremely bad noise and fading 100 μv per meter may not be sufficient.

Low signal level, or rather low signal-to-noise ratio, is naturally only one of the factors limiting the usefulness of a telephone circuit. Fading, with associated bad quality, is another. It is difficult to separate these two factors since they occur so often at the same time. Low signal-to-noise ratio occurs alone more than does fading.

In this connection it may be said that probably the most important and difficult problem is the maintenance of service on days of disturbed solar conditions (magnetic storms). It appears that these difficulties are greater in high latitudes than in the case of communication across the equator. However this may be, transatlantic signals are sometimes so attenuated that the carrier cannot be detected even as a beat note. To overcome such extreme conditions would require an increase in signal-to-noise ratio of more than 30 db over that given by the system now used. It is not absurd to suppose that an improvement of 30 db can be effected. There is no doubt, however, that it will be very difficult, and there is no assurance that the gain needed is not very much greater than this figure. At any rate lesser improvements should materially reduce the time during which the circuits are made "uncommercial" by these unusually weak signals.

Wavelength Selection

It is well known that for best results, long distance short-wave systems require several wavelengths for 24-hour operation throughout the year.

For transmission through the daylight period, transatlantic work is carried on at frequencies in the region from 17 to 19 megacycles. On winter days, somewhat better results may be obtained from lower frequencies than in the summer, though different years differ with respect to the best daylight frequency. Thus in the winter of 1926–1927, a frequency of 13 megacycles was good over a longer period than in either winter since.

Transmission during winter nights presents the other extreme. A frequency as low as 6 or 7 megacycles is then desirable.

Between these limits represented by the summer day and the winter night lies a range of conditions for which intermediate frequencies are preferred. The number of these frequencies needed depends upon the importance of service during the transition periods when part of the path is illuminated and the rest is in darkness.

One important time, probably the most difficult that is encountered during the year, is the winter afternoon. The sun has set in England though it is still daylight in America. In passing through this period the optimum frequency changes gradually from the morning value of 18 megacycles to half this value at 6 P.M., E.S.T.

The fact that conditions change considerably from day to day makes selection of operating wavelengths more difficult. A system involving several channels has a distinct advantage. Essential information peculiar to the day can be obtained by making a frequency change in one of them. The result will be a guide in deciding the proper time to shift wavelengths in the remaining channels.

Each of the short-wave transatlantic telephone channels is equipped for operation in three frequency ranges, namely 19, 14, and 9 megacycles, (16, 21, and 31 m). A still lower frequency is provided for one of them, the purpose being to cover the winter night satisfactorily. This frequency is about 6.7 megacycles (45 m).

During the long days of summer very little frequency changing is necessary. In fact, the usefulness of the 18-megacycle band does not disappear with sunset, but continues on into the night. Often it may be employed later than midnight, but it always fails during the few hours preceding sunrise in America. The result is that in summer short-wave transmission is at its best. This, of course, is fortunate in view of the fact that this season is the most difficult for long-wave transmission.

Field-Strength Data

Systematic tests have been made to determine the properties of different frequencies throughout the short-wave range. Intensive tests were made in June, 1926. These have continued as weekly tests with small interruptions to the present time. Throughout this work signal measurements have been made in England by a group under the direction of A. G. Jensen. For about a year similar work has been carried on in Buenos Aires by E. J. Howard. At Deal, E. B. Ferrell, N. F. Schlaack, and A. E. Kerwien have at different times been in charge. Frequencies from 2.7 to 27 megacycles have been used. A small number of measurements has been made at higher frequencies. Unless otherwise stated the results given apply to data taken in England.

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No attempt will be made here to summarize completely the results obtained in these tests but it may be of interest to indicate some of the conclusions.

The field-strength curves are of especial interest. The method of plotting has usually been to draw lines of equal signal on a "map" whose coordinates are time of day and frequency. Such, for example, is Fig. 1, which represents short-wave data obtained in June, 1926.[1] The figure resembles a topographical map, high altitude corresponding to strong signals. "Sea level" is the interference limit. The broad ocean in the lower right corner is due to daylight absorption of the

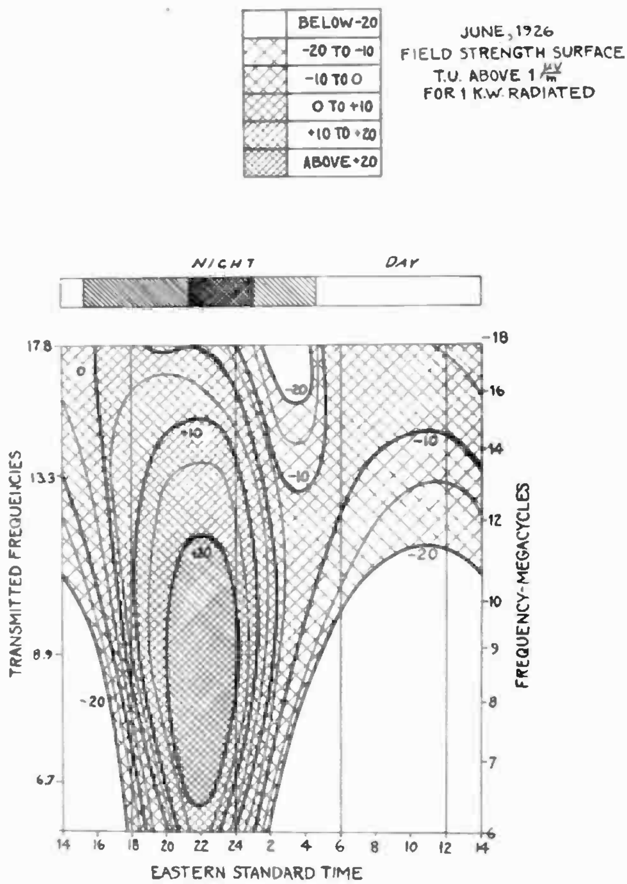

Fig. 1—High-frequency radio transmission, Deal, N. J., to New Southgate, England. Field strength in decibels above one μv per meter, June, 1926.

lower frequencies in the short-wave region. The mountain represents the excellent transmission of these same frequencies at night. The plane in the upper right corner corresponds to the satisfactory transmission of the higher frequencies by day, while the depression at the middle of the top is the skip region.[2]

[1] These curves have been obtained from C. R. Burrows who has made a detailed study of the results.

[2] Whether this region is due to penetration of the layer due to insufficient ionization, as suggested by Taylor and Hulburt, or to absorption in the layer because of an insufficiently steep ionic gradient, as suggested by Appleton and elaborated by T. L. Eckersley, is one of the more interesting unsolved questions

This figure is typical of mid-summer results in these latitudes. The narrowness of the nighttime peak is due to the short time that the whole path is in darkness. Contrasted with this is a typical plot for winter, Fig. 2. The period of satisfactory low-frequency transmission has now increased because of the longer nights, and the main features of the map have moved to frequencies which are some thirty per cent

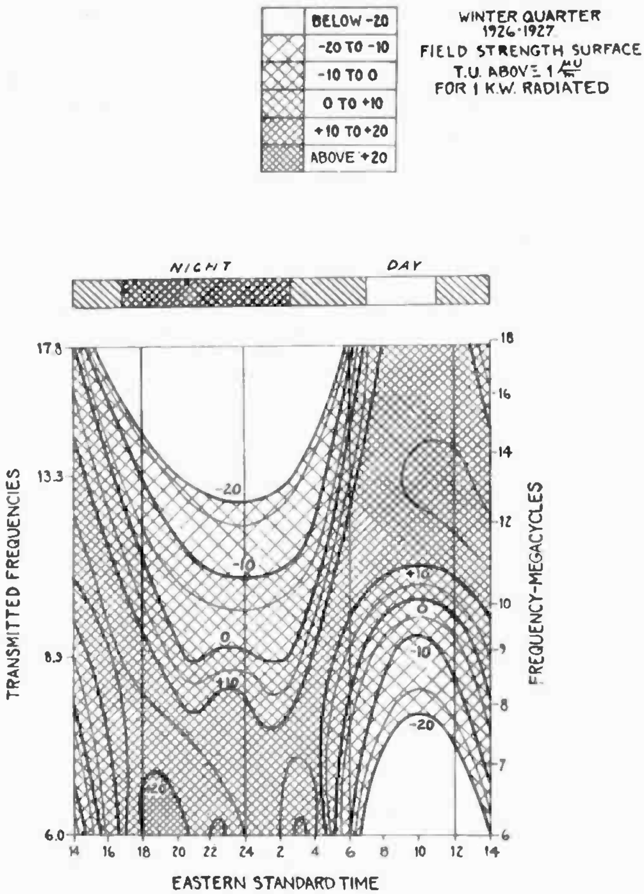

Fig. 2—High-frequency transmission, Deal to New Southgate, winter, 1926–1927.

lower than in the summer curve. The skip zone has increased in area while the absorption region has receded toward the lower frequencies.

It should be observed that such curves depend somewhat on the latitude of the stations. If instead of Deal, a transmitter farther north

in radio transmission. It is difficult to reconcile the Hals-Störmer long delayed echoes with the absorption hypothesis. If these long delays are really due to the waves having travelled *long distances*, it is probable that the paths are such as Störmer assumes. This would require penetration of the dense layers assumed by Eckersley. If the long delays are due to exceedingly *low group velocities*, the assumption that the skip effect is due to absorption becomes difficult and probably unnecessary to support. Therefore, the existence of these echoes supports the hypothesis of insufficient ionization as an explanation of the skip effect. It also seems that the explanation of the echoes given by Störmer is more probable than the low group velocity explanation of van der Pol.

had been used, it is likely that the summer nighttime peak at 9 megacycles would have been reduced or even eliminated, since conditions in the upper atmosphere might then have approximated those of daytime all along the path. Similarly, an opposite change would perhaps have been observed if the receiver had been moved south from England to the latitude of New York, (e.g., to Spain).

The results of tests which have been carried out between the Deal experimental station and Buenos Aires will not be discussed in this paper. Most of the general characteristics are the same as those already given. The frequencies required are higher because the distance is greater, and the seasonal variations are much less since opposite seasons are encountered at the two ends of the path.

Reliability

Along with the field-strength measurements, intelligibility tests have been made. These have involved determinations of the percentages of words understood. On the basis of such results it is possible by extrapolation to obtain some idea of the communication to be expected with facilities which are better or poorer than those actually used. Such estimates, while not thoroughly reliable, are of value in the absence of better and they do bring out certain tendencies which are of interest.

The average percentage of the day that a good transatlantic shortwave circuit is successful is naturally a figure of interest and importance and data on this are being accumulated from commercial operation. It is realized, however, that no dependable estimate of commercial reliability can be made in advance of actual use under commercial conditions. Our present remarks are confined to experimental data and the discussion of methods used in estimating the effects of increasing or decreasing the signal-to-noise ratio.

The method used has its starting point in curves such as those of Fig. 3.[3] The original data disclose for each individual observation whether or not the experimental intelligibility was greater than that assumed to represent the border line between a "commercial" and an "uncommercial" circuit. In this case the understanding of 60 per cent of a list of unconnected words was arbitrarily taken as the criterion. In addition the field strength, (or field-strength-to-noise ratio) is known. An elementary antenna with three "reflectors" was used at the receiving station. The readings are collected into groups according to field strength—for example, the groups may be 5 db wide—and the

[3] The data used in this particular example were obtained in some tests made in cooperation with engineers of the American Telephone and Telegraph Company. They do not involve 24-hour operation.

percentage of each group which is commercial is plotted as in the upper figure. This curve expresses the fact that as the field strength is increased the probability of the circuit's being satisfactory also increases, approaching a value somewhat less than 100 per cent. The second curve in Fig. 3 gives the number of readings in each group as a percentage of the total number; it therefore represents the probability

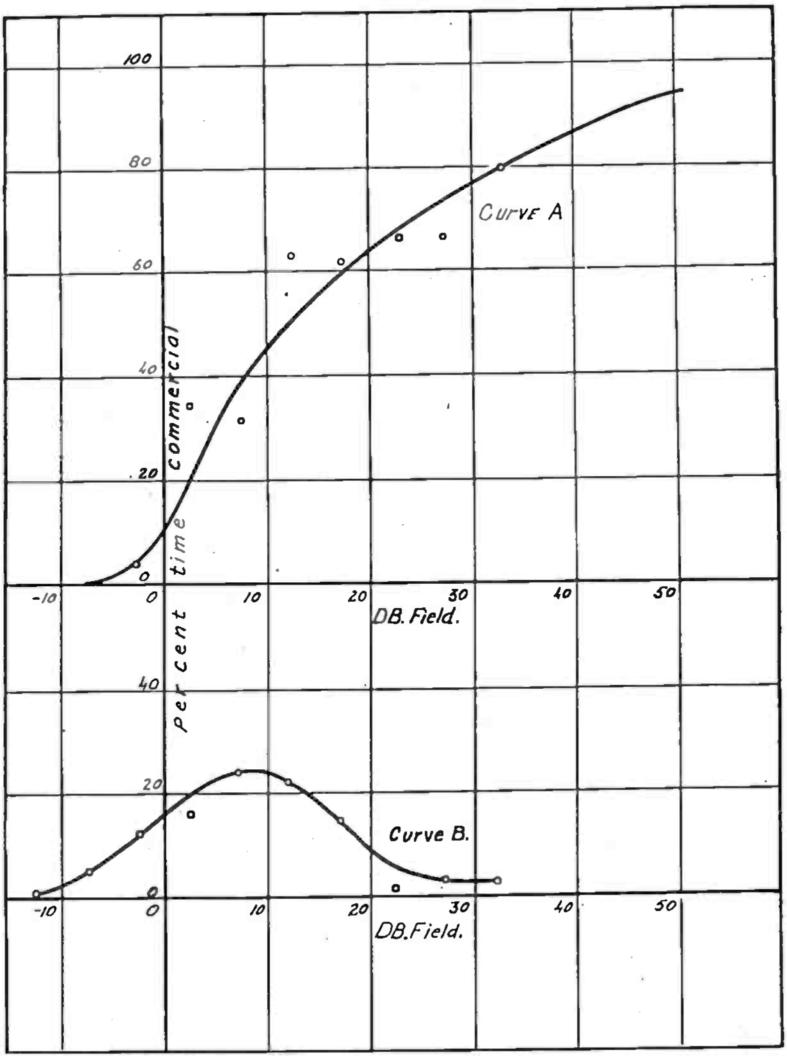

Fig. 3

of occurrence of the various field strengths. The integral of the product of these curves divided by the integral of the second curve gives the weighted average of all the readings. In making extrapolations the assumption is made that while intelligibility is affected by fading, the *differences* in intelligibility at different levels are caused by differences in signal strength only. This postulate is not entirely above question, but it probably is not far wrong. Calculations of the reliability obtained with various assumed improvements in signal value are made by moving the second curve to the right a number of db corresponding to the assumed improvements.

In this way the curve in Fig. 4 is obtained. The curve brings out certain tendencies which are of interest, although these particular numerical data are not applicable in general. In the middle portion,

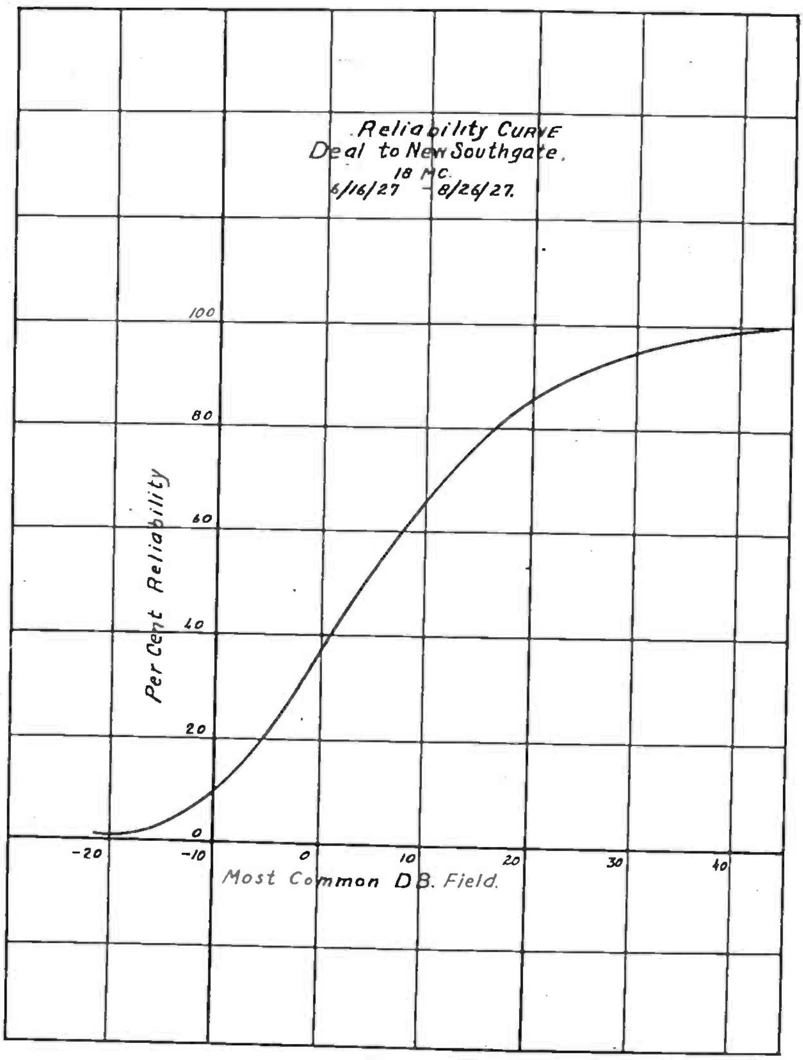

Fig. 4

the reliability increases rapidly with the field strength. At the upper end a large increase is necessary to improve reliability. With a power of 5 kw, with simple transmitting and receiving antennas and with transatlantic distances, we might obtain a reliability in the order of 20 or 30 per cent. In other words, the point would be at the bottom of the steepest part of the curve. A small increase in the gain through the system would therefore result in a considerable improvement. This circumstance, combined with the fact that for short waves, directivity provides a cheap way of obtaining a considerable signal increase, makes it *highly advisable to include at least elementary directive antennas at both ends of point-to-point radio links.* On the other hand, the fact that the curve tends to become horizontal at the upper end where also improvements may be very costly, indicates that an economic balance will have to determine how far to go in this direction. Different answers may be reached depending on the importance attached to the maintenance of continuous communication.

This type of curve also throws light on the striking success which is often obtained with simple apparatus. In the particular case illustrated, there is a spread of 30 to 40 db between the levels required

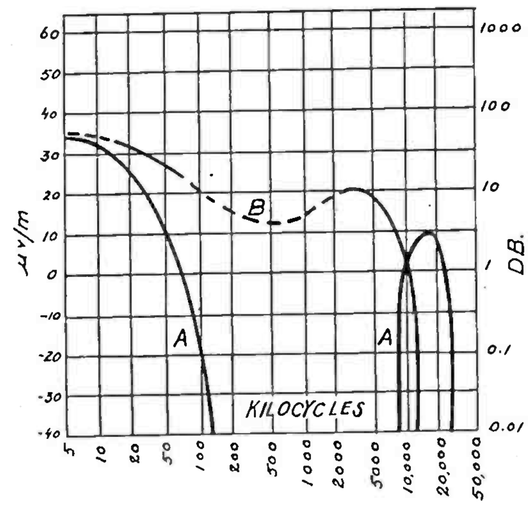

Fig. 5—Approximate transatlantic transmission characteristic through entire radio spectrum; curve A for day, curve B for night. One kw radiated.

for reasonably reliable communication and for occasional communication. This corresponds to a power ratio of 1,000 to 10,000 times. Considering with this the fact that for slow-speed telegraphy, a much weaker signal can be employed successfully than in telephony or in high-speed telegraphy, some idea is obtained as to the reason why commercial services require so much more elaborate facilities than amateur work.

Comparison of High and Low Frequencies

Fig. 5 is an attempt to represent some general characteristics of the entire radio-frequency range in one field-strength diagram. The data are for transatlantic transmission and they assume 1 kw to be radiated. Curve A is for daylight transmission, curve B for nighttime. The low-frequency portion of the former is based on the empirical equation obtained by Espenschied, Anderson, and Bailey,[4] while the high-frequency data were obtained between Deal and New Southgate during the winter of 1926–1927.[5] Curve *A* falls in two distinct parts, separated by a wide region (approximately 125 kc to 7000 kc) in which absorption completely annihilates the signal. For very low frequencies[6] the signal approximates the value obtained on the inverse distance basis. On comparing the two parts of curve *A* the fact is brought out that, for 1 kw radiated, frequencies below 50 kc give stronger signal fields than very high frequencies. The same is true of the night conditions depicted by curve *B*. Here there is uncertainty as to the behavior of intermediate frequencies as indicated by the use of a dotted line. These curves clearly bring out the fact that, solely from the viewpoint of transmission, very long waves (e.g., 20,000 m) are for transatlantic distances somewhat superior to very short (e.g., 15 m). 1 kw gives about 20 μv per m at 20,000 m and only 3 or 4 at 15 m.

It is hardly necessary to add, however, that this is only one phase of the comparison. It is considerably more expensive to radiate 1 kw with long waves than to radiate the same power with short and it is out of the question in a long-wave transmitting antenna to obtain

[4] *Bell Sys. Tech. Jour.*, July, 1925.

[5] This gives the average for a large number of days.

[6] The lowest frequencies in the figure are audible; no signal-strength data are available. This is the region which Watson-Watt has studied oscillographically. It includes "musical X's." T. L. Eckersley's interesting observation that the latter occur more often on magnetic-storm days may possibly be explained as a result of improved long-wave transmission at such times. It has occurred to the writer that the following simple explanation of these "whistlers" may have some merit. The wave set up by a non-oscillatory electric discharge somewhere between the earth and ionized layer would undergo multiple reflection in such a way that the portion of the disturbance travelling nearly vertically would take a longer time to travel, say 10,000 miles around the earth, than that which proceeds horizontally. Furthermore, the multiple reflections of the former would make up a lower musical tone than the latter, thus giving an explanation of the change in pitch. The frequency of the last part would approach $c/2h$, in which h is the effective height of the ionized layer for low frequencies. If for an order of magnitude we identify this with the height found in the radio range on undisturbed days, (e.g., 100 km), and multiply this by a factor of two because of the increase of height due to the magnetic storm, (see Hafstad and Tuve, *Terr. Mag.*, March, 1929; Dahl and Gebhardt, Proc. I.R.E., 16, 290–297; March, 1928; and Maris and Hulburt, Proc. I.R.E., 17, 494–500; March, 1929.), this frequency is found to be 750 cycles. According to Eckersley the frequency observed varies between 200 and 1,000 cycles. (*Phil. Mag.*, 49, 1250; 1925.) This explanation, if valid will furnish a method for determining the "height" of the layer. Is it not also possible that the different types of "musical X's" have fundamentally different explanations, some being caused by dispersion and others by multiple reflection?

The mechanism suggested above bears some resemblance to the cause of similar musical sounds which are often observed while walking past a picket fence. Secondary waves are set up, and return to the observer, when the sound of the foot-step reaches the various pickets. The latter correspond to the multiple images of the assumed discharge. The analogy, however, is not perfect.

the efficiency which is possible in a short-wave directive array. In the case of the receiver, the difference is not definite. Large gains in signal-to-noise ratio can be obtained with both short and long waves by the use of directional reception. Another consideration is that it is not customary at the present time to employ as high a power with short waves as with long. A further item in the comparison is the amount of noise encountered in the two frequency ranges. While the difference has sometimes been over-stated, static is less in the case of short waves. The net result of the comparison is that the greater power at low frequencies, together with the smaller amount of time lost due to magnetic storms and fading, offsets the advantages of high frequencies resulting from the improved transmitting antenna efficiency and reduced static.

Transmitting Antennas

In discussing antennas there are three more or less distinct steps of the problem to consider:

(1)—What are the directional properties of the medium; i.e., what are the requirements of directive antennas and what are the possibilities and limitations from the standpoint of transmission?

(2)—What distribution of currents, both with respect to phase and spacial arrangement, is necessary in order to satisfy the conditions of (1)?

(3)—What circuit means may be employed to provide these currents?

We shall make no attempt at a comprehensive answer to these questions. Since (1) in the past has been by comparison somewhat neglected, most of our attention will be directed to it. It must be admitted at the outset, however, that our information is far from complete. A study of certain phases of (2) and (3) will be given in a paper now being written by E. J. Sterba.

Directional Properties of the Medium

The term "broadside" is applied to antenna arrays whose elements are located along a horizontal line perpendicular to the line of transmission and excited in the same phase. The calculation of the gain will be touched upon later. Various antennas of this sort have been constructed and tested by transmission to England. The length of the arrays has ranged from 1 to 9 wavelengths. At least up to 6 wavelengths, the gain calculated has been actually obtained within experimental error with a wavelength of 16 m and with daylight conditions. Longer antennas also give the calculated gain a large part of the time. Since the antenna, 6 wavelengths long, had a major lobe whose width from minimum to minimum was 19 deg., it appears that usually the range of useful bearing was less than one half of this value. In other words, under the condition of these tests the direction of the more important components of the signal did not deviate from the geographical bearing by more than ± 5 deg., and possibly less.

In one antenna, provision was made for quickly deviating the direction of the signal maximum by about 5 deg. In one experiment two receivers were employed. One was in London, which is on the great circle, normal to the antenna length; another receiver was located in Scotland on a great circle about 3 deg. different. Under these conditions, the relative signal strengths were measured for the two directivities of the system. These readings lasted over a period of several

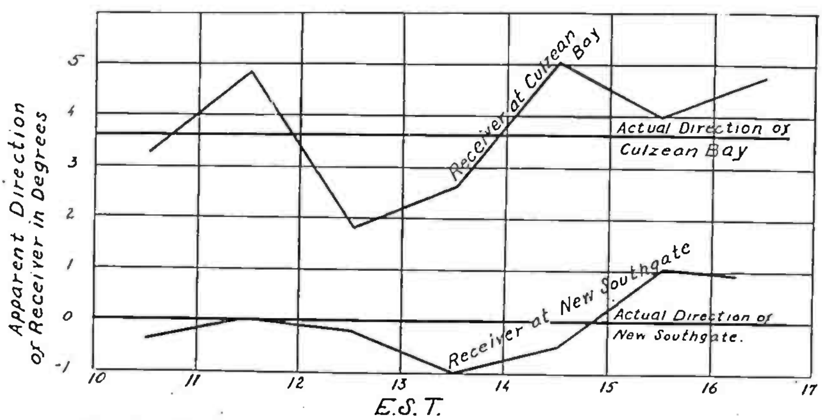

Fig. 6—Eighteen-megacycle antenna used as a beacon.

hours of daylight transmission at 16 m. The antenna, having a major lobe of $9\frac{1}{2}$ deg. width from maximum to first minimum, would be expected to give noticeably different signals under the two conditions. This was found to be the case, and with a more careful analysis of the results, E. B. Ferrell found that on the average the difference numerically checked with the assumption that the maximum signals cluster closely about the great-circle path. The agreement was so good in fact that the measurements might have served with good accuracy to determine the bearing of the receivers. In Fig. 6 are plotted the apparent bearings. From the manner in which they cluster about the true bearing it appears that even these very short wavelengths are not without utility for beacon service. Unfortunately it was not possible to determine the limitations[7] of this method by long-continued tests.

[7] That such limitations exist is indicated by some results obtained by Friis, but it may be mentioned that the method employed here, depending on an average over a period of from several minutes to a half hour, will naturally show

It also is not known how short the measurement period might be made without loss of accuracy.

Additional information is obtained by a consideration of the well-known phenomenon that near a receiver fading differs at two points which are a few wavelengths apart.[8] If there are simultaneously present in the received wave two components from somewhat different directions, and if they add in phase at one of these locations, they will be out of phase at a second point sufficiently distant. Conversely, the existence of this fading phenomenon is always caused by the existence of two or more components travelling in different directions. It therefore becomes desirable to consider directivity of antennas from this point of view.

Consider the following simple case. We shall assume that the two points, A and B, both at the earth's surface are along a line perpendicular to the great-circle path (see Fig. 7).

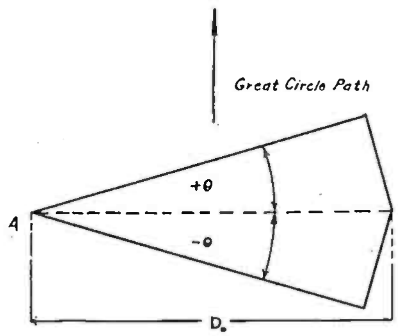

Fig. 7

Two wave components differing in azimuth by a small angle arrive at A with a certain phase difference. In order that the fading at B should be random with respect to that at A, the phase difference of the components at B should be independent of that at A. This condition will be satisfied if the directions of arrival of the components are always between $+\theta$ and $-\theta$, having all values with equal probability.[9] θ may then be determined in terms of D_0, the minimum distance for which fading is random. For, when θ is small the maximum change in phase difference of the components as we pass from A to B is equal to $4\pi D_0\theta/\lambda$ and should equal 2π in order to have substantially random

less deviation than Friis' instantaneous method. H. T. Friis, "Oscillographic observations on the direction of propagation and fading of short waves," Proc. I.R.E., 16, 658; May, 1928.

[8] The reverse of this phenomenon is also true, so that while it is easier to discuss the matter from the point of view of the receiver, any conclusions drawn regarding receiving antennas apply to transmitting arrays as well.

[9] The assumption of equal probability is made in the interest of mathematical simplicity, as is also the assumption that two components of equal amplitude are always present.

conditions. This leads to $D_0/\lambda = 1/2\theta$. If, for example, D_0 by experiment is found to equal 6λ, θ must equal 4.8 deg. From tests made by Friis and Jensen it is known that the values assumed are typical.

It is of interest to carry the computation of this simple case a little further. Suppose, that the detector outputs of the two identical radio receivers were combined differentially, so that for equal signals the reading of the meter would be zero. For simplicity in making the calculation this meter is assumed to have a square-law characteristic. When the receivers are located side by side the reading of the meter E will be zero. As the separation perpendicular to the great circle is increased, the reading (averaged over a time long compared with the

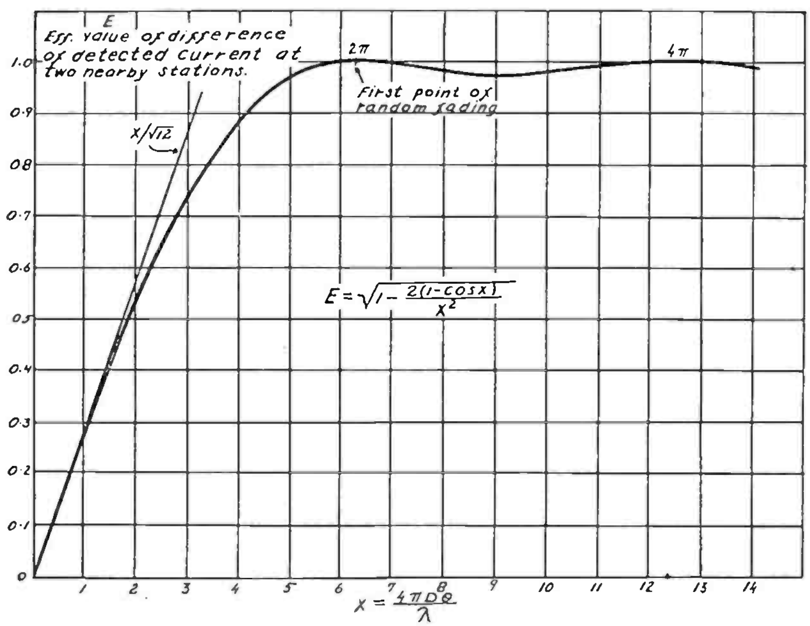

Fig. 8—Average difference in output of two receivers separated D/λ wavelengths on line normal to great circle.

period of fading) will also increase until the fading at the two points becomes random. For greater separations the reading will remain the same.

On this basis it can be shown that the assumption of two rays always lying between $+\theta$ and $-\theta$ leads to the relation

$$E = \sqrt{1 - \frac{2(1 - \cos X)}{X^2}}$$

where $X = 4\pi D\theta/\lambda$. (Fig. 8). It consists of a straight portion near the origin, a knee and then a nearly horizontal portion. It is consistent with the experimental fact that the difference of the currents increases somewhat linearly for small spacings and becomes constant for great separations. The first point of random fading is, of course, $D_0 = \lambda/2\theta$ as was found previously. The average value of the current of one receiver is the same as the effective value of the difference of the two for a spacing of D_0.

We have been led to examine this case theoretically because it is natural to assume that the length of a directive transmitting antenna should be substantially less than the minimum distance of random fading. Fig. 9 gives the results of a calculation of the loss which is

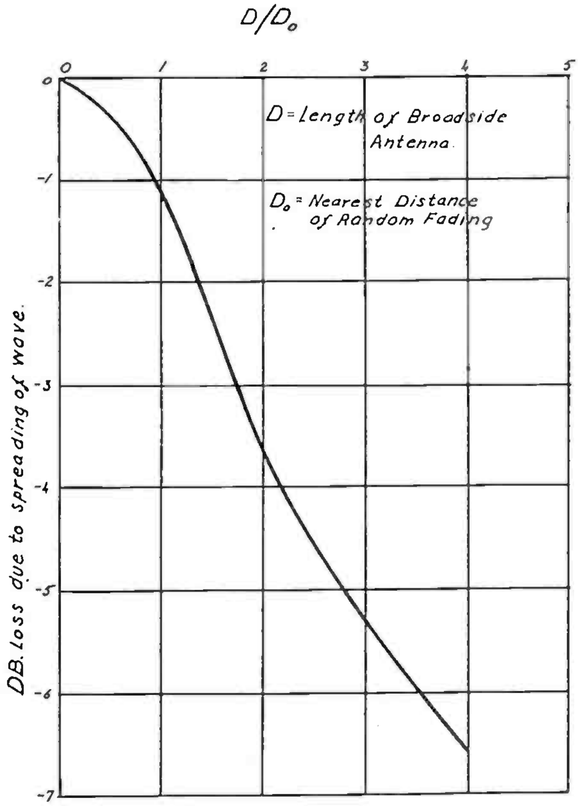

Fig. 9—Loss with broadside antenna due to scattering in horizontal plane.

incurred because of this spreading of the wave in the case of a broadside antenna. When $D/D_0 = 1$, the signal is 1.1 db less than it would have been if the wave had travelled only along the great circle. By increasing the antenna length to $D = 2D_0$, this relative loss would become 3.6 db. Since doubling the length increases the efficiency without scattering by 3.0 db, there would still be a net gain over the case $D = D_0$ equal to 0.5 db, but this would usually be too small an improvement to justify the increased expenditure. Should D/D_0 become equal to 2, the polar diagram of the antenna would have minima along θ and $-\theta$, and the fading of the signal would be affected to a marked degree. The net effect might not be an adverse one.

It is well to remember, of course, that when conditions are such that the wave clusters about a direction different from the great circle bearing it will be necessary, in order to obtain the greatest efficiency, either to shift the direction of maximum signal correspondingly or to increase the angular width of the main lobe. The first alternative is possible in suitably designed arrays and has actually been accomplished experimentally. To avoid operating difficulties, it is simpler to avoid too sharp a "beam."

Each of the transatlantic telephone links employs three wavelengths in the ranges 16 m, 22 m, and 32 m, while one employs a fourth at 45 m. In all cases the antennas at the American transmitting terminal (located at Lawrenceville, New Jersey) are broadside with respect to the great circle and employ "reflectors."

The antenna lengths are roughly 8, 6, 4, and 4 wavelengths, respectively. There are several conditions in addition to those discussed here which influenced this choice. Such, for example, is the desirability in a row of towers of maintaining uniform spacing. These matters, however, are beyond the scope of this paper.

Under the same conditions for which the components of the wave are confined in azimuth to a few degrees, it is common to find a much wider spread in the vertical plane, or plane of incidence. This statement is based in part on experimental observations made by Friis and Jensen, which show that two adjacent receivers located on the same great circle often fade at random when their separation does not exceed 4 or 5 wavelengths. The general considerations are similar to those involved in the discussion above on variations in the horizontal plane and therefore will not be repeated. Fig. 10 gives computations based on various assumed distances of random fading, D_0. θ_1 and θ_2 represent the lower and upper limits of the angle from the horizontal. Thus in order that fading be random with a separation as small as 3 wavelengths (see Fig. 10, $D_0/\lambda=3$) there must be a noteworthy amount of radiation at angles as high as 40 deg. to 50 deg. above the horizontal.[10] This numerical example assumes a minimum angle, θ_1, equal to zero.

While high-angle radiation certainly exists, low-angle waves are more dependable in the 15-m range. This is concluded from the fact that antennas designed to narrow down the vertical-polar char-

[10] These considerations have some theoretical interest as they indicate that high- and low-angle radiation are commonly present *at the same time*. They therefore strongly suggest that in long-distance radio transmission, propagation by multiple reflections is the general rule. It seems, however, that there are so far no completely conclusive arguments either way on the question of "short-step" versus "long-step" propagation of short waves. See Pedersen, "Propagation of Radio Waves," p. 194–200.

acteristics to moderately small angles, experimentally give nearly the gain which is calculated on the assumption that the maximum signal should be obtained in a nearly horizontal direction. On the other hand the danger exists that this may be overdone as we have sometimes found, particularly at the longer waves. The Lawrenceville antennas employ a single "tier"[11] at 45 m, two at 22 and 32 m, and three at 16 m.

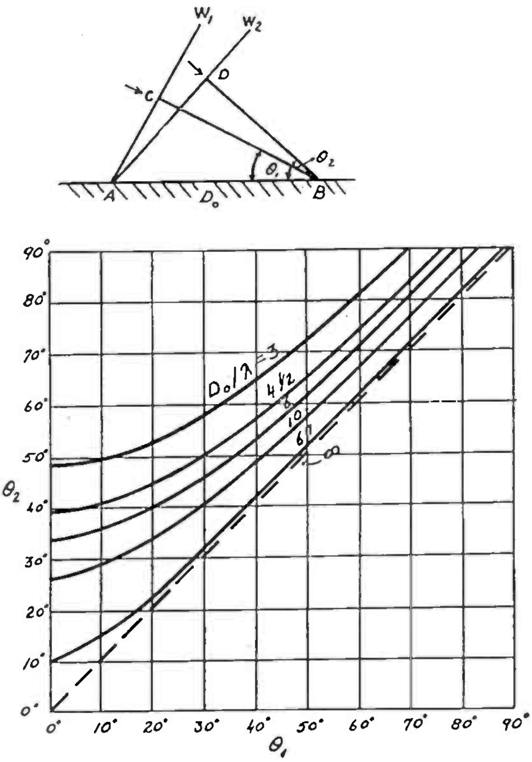

Fig. 10—Spread in angle of elevation necessary to cause random fading.

Gain of Antenna Arrays

As previously mentioned, the arbitrary standard usually employed in the comparison of short-wave antennas is the half-wave vertical antenna. In decibels the gain over the standard is 20 log E/E_s, E and E_s referring to the field strengths at the receiving point for equal power inputs to the two antennas.

Fig. 11 gives the result of calculations of two types of array, namely the broadside and the end-fire type. In both cases it is assumed that doublets 1/4 wavelength above a perfectly conducting ground are employed. These are arranged along a horizontal straight line with a spacing of $\frac{\lambda}{2}$. In the first case the phases are the same. In the second

[11] A "two-tier" antenna array is one having two half-wavelength elements one above the other. The larger the number of tiers the sharper is the directivity in the vertical plane.

there is a progressive phase shift of 180 deg. per element so that the maximum signal is transmitted in the direction of the line of the array. The curves indicate that for a given size of structure the broadside structure is more efficient than the end-fire type.*

The assumption of doublets instead of half-wave wires is made because of mathematical convenience. The result cannot in any event be accurate unless account is taken of the actual properties of the earth which unfortunately, is not a perfect conductor. With this in mind it will probably be agreed that the small error involved in the doublet

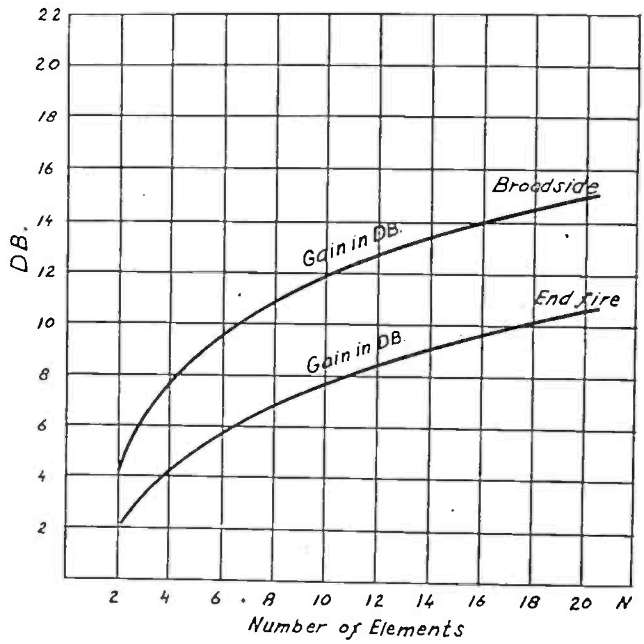

Fig. 11—Gain produced by linear array of vertical doublets $\lambda/4$ above a perfectly conducting earth.

assumption need not concern us. The errors involved, moreover, are to some extent compensating because the calculation of gain is a comparison between two antennas in which similar approximations have been made.

To a good approximation the use of a well-designed reflector improves the gain by 3 db, and for nearly horizontal directions the additional gain obtained by the use of a total of S tiers is somewhat less than $10 \log_{10} S$. Obviously an actual loss may be obtained by the use of more than one tier when high-angle radiation is more effective than that at low angles.

Reference has been made to the fact that the first few decibels obtained with directivity are cheap. However, the cost of each

* See Appendix.

additional improvement increases with the number which have already been acquired. This is apparent from the fact that the gain is roughly proportional to the logarithm of the number of elements while the cost increases more nearly linearly with that number. This diminishing return is accentuated by the fact, already mentioned, that at the higher levels the value in average reliability increase per decibel improvement becomes relatively small.

While the effectiveness of antennas may be estimated in the general way indicated, such calculations recognize only ideal conditions of transmission. The estimated gains are not always realized in practice. In the 15-m range, however, gains of 18 or 20 db are obtained a sufficient portion of the time to justify the use of antennas of this general size. The theoretical gains for the Lawrenceville antennas range approximately from 20 db at 15 m to 13 db at 45 m.

Transmitters

Telephony requires of the short-wave transmitter certain characteristics which are different from those encountered by the designer of telegraph apparatus. To be of greatest utility, development must be planned with the fact in mind that the future will probably require certain types of transmission which will necessarily involve radical departures from present practices in modulation. Single side-band telephony is an example of a system of this sort. Its success depends on satisfactory amplification *after* modulation. Harmonic generators, oscillators, or the like cannot in this case be included at points beyond that where modulation takes place. In order to salvage a large part of existing plant when making such a change, it is necessary to modulate at low-power level and to provide an amplifier which may be employed with almost any system. This is one of the reasons why so much emphasis has been placed in this work on the development of amplifiers.

In the second place, telephone equipment, or more specifically a telephone amplifier, needs to be stable at all radio-input levels from zero to the maximum that is used. It does not suffice to obtain stability with zero output by means of a very high negative grid bias; or with maximum output, by means of very strong input. All levels are equally important. In fact, in the long run, it is found that no set is sufficiently stable for telephony unless it is possible to reduce the grid bias to zero and remove the grid-input voltage without the occurrence of singing at any frequency. This must be true for all positions of the variable tuning elements and when capacity neutralization is used a generous margin of adjustment must be provided.

Another result which must be obtained is the ability to amplify without distorting either the amplitude or the phase. That is, the envelope of the output must be a replica of that of the input, and the phase of the radio frequency in passing through the amplifier must not suffer a change which is a function of amplitude.

It is not intended here to enter into a detailed discussion of the power amplifier which has been developed.[12] It is of the push-pull type, employs two water-cooled tubes in the first stage and six of the same kind in the second. Capacity neutralization is used. The output of the second stage is coupled to the antenna by means of a two-wire 600-ohm transmission line, the impedance of which is matched at the antenna.

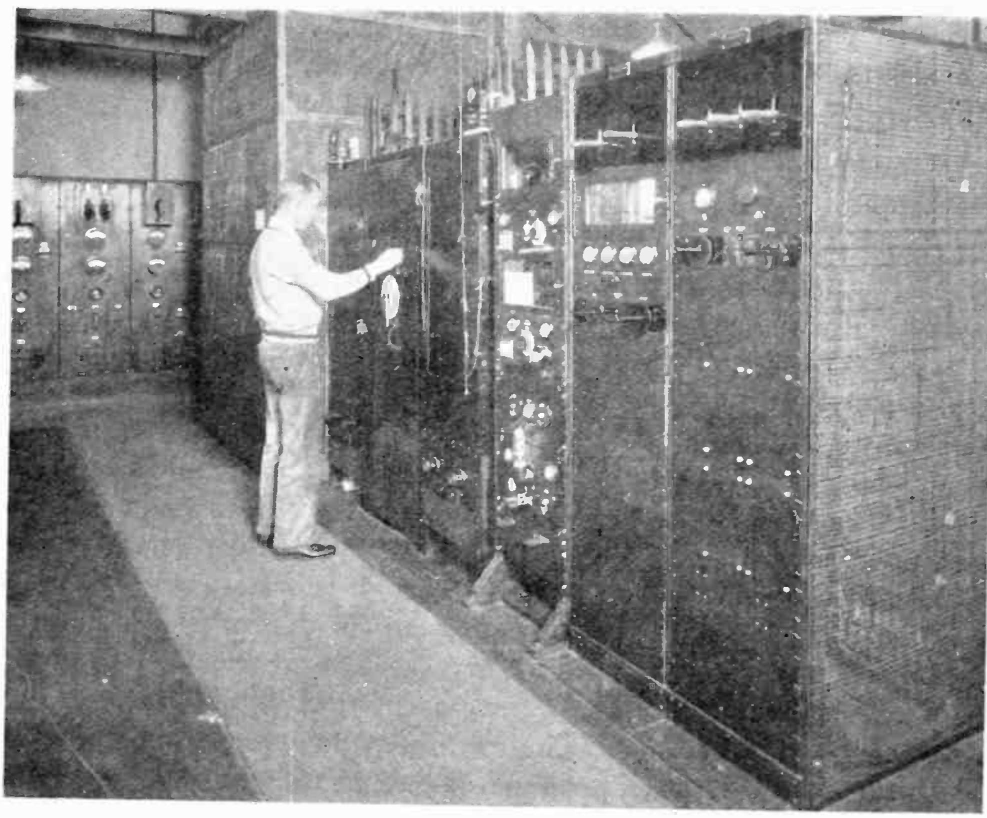

Fig. 12—Experimental transmitter at the Deal laboratory.

The last two stages of the later sets amplify an input of the order of 150 watts to an output of 15 kw. In other words the gain is about 20 db. In speaking of the output as 15 kw, we are adhering to the method of rating used in broadcast transmitters. The figure given applies to the unmodulated carrier. However, the tube power capacity which has necessarily been provided is four times that value. In other words, for telegraphy or single side-band telephony a peak output of 60 kw should be possible.

[12] A complete account of the high-power work is being prepared.

Fig. 12 is a view of the transmitter at the Deal laboratory which has been used for much of the experimental work referred to in this paper. This is an early model. In external appearance it differs greatly from those of later design.

In short-wave transmission it is necessary to be able to change frequencies quickly. For transatlantic work a range of three-to-one in frequency is necessary. This can be accomplished in a single amplifier by the use of plug-in coils together with variable condensers or inductances for finer adjustment. In the amplifiers at the Deal laboratory and at Lawrenceville, frequencies can be changed from one to another of several predetermined values in less than five minutes. The frequency range of the sets is from 22 megacycles (14 m) to 6 megacycles (50 m). These values, however, represent the range required in transatlantic work rather than the extreme limits which are possible.

Distortion

It is well known that when a complicated wave is applied to a non-linear device, many components not present in the input appear in the output. Without entering into extended computations of this kind, certain criteria useful in the avoidance of distortion will be mentioned. In high-frequency signalling the input frequencies are confined to a band which is small as compared with the "mean frequency" which is employed. With this limitation, the usual Fourier representation of any wave may be replaced by the form[13]

$$f(t) \sin \left[\omega t + \alpha(t)\right].$$

This equation represents the wave mathematically in much the same way as that by which the engineer naturally visualizes it. It is characterized by a variable amplitude, $f(t)$. There is a certain mean frequency ω and a variable phase $\alpha(t)$. The high-frequency non-linear device itself "sees" the wave in this way. In translating input into output, it has to work substantially "from hand to mouth," since it is not endowed with that memory and foresight which enables a Fourier integral to form its decision on the basis of all happenings from minus to plus infinity. We may, therefore, say that an amplifier has faithfully performed its function if the envelopes, $f(t)$ of the input and output do not differ except by a constant amplification factor, and if also the phases $\alpha(t)$ of the resultant waves are the same.

We are naturally lead in this way to a consideration of two types of distortion, viz., distortion of amplitude and of phase. It may there-

[13] J. R. Carson, Proc. I.R.E., 16, 967-975; July, 1928.

fore be said that if the amplifier, *however curved its static characteristics may be,* nevertheless has a linear input-output characteristic passing through the origin, and if also the phase in passing through the device is not changed, then a Fourier analysis will not reveal the presence of any new frequencies in the output.[14] The device is then satisfactory as a high-quality amplifier.

If it is not obvious that a change in phase constitutes distortion,[15] it becomes so after a simple analysis. A wave of frequency ω, whose amplitude and phase are both modulated sinusoidally by a frequency p may be represented as follows

$$(1 + k_1 \cos pt) \cdot \cos \left[\omega t + k_2 \cos (pt + \Phi)\right]$$

where k_1 is the percentage of amplitude modulation and k_2 is the maximum angle of phase modulation. If for simplicity we limit the consideration to small angles (10 deg. or less) the expansion of this expression simplifies and we obtain the following

$$\begin{aligned} \cos \omega t &- \frac{k_1 k_2}{2} \cos \Phi \cdot \sin \omega t \\ &+ \frac{k_1}{2} \cos (\omega + p)t - \frac{k_2}{2} \sin \left[(\omega + p)t + \Phi\right] \\ &+ \frac{k_1}{2} \cos (\omega - p)t - \frac{k_2}{2} \sin \left[(\omega - p)t - \Phi\right] \\ &- \frac{k_1 k_2}{4} \sin \left[(\omega + 2p)t + \Phi\right] \\ &- \frac{k_1 k_2}{4} \sin \left[(\omega - 2p)t - \Phi\right]. \end{aligned}$$

Distortion appears

(1)—As an actual change in the amplitude and phase of the carrier ω.

(2)—By the addition of new terms to the expressions for the first-order side bands, $(\omega+p)$ and $(\omega-p)$.

(3)—By the presence of terms $(\omega+2p)$ and $(\omega-2p)$. Had the simplifying assumption of small angles not been made, we should also have had $(\omega \pm np)$, n being any positive integer.

[14] It has also been tacitly assumed as a matter of course that all time constants in the circuit are small compared with the components involved in $f(t)$ and that there are no low-frequency impedances involved which might disturb the assumed constancy of the supply voltages. Filters eliminating from the output all components except those about ω are also necessary. The latter function is adequately fulfilled by the usual tuned circuits in short-wave transmitters.

[15] R. Bown, De L. K. Martin, and R. K. Potter, "Some studies in radio broadcast transmission," Proc. I.R.E., 14, 57; February, 1926. These authors were the first to point out the serious effects of this important type of distortion in radio transmission.

The interesting fact appears that excepting when Φ is zero, the upper first-order side band does not equal the lower in amplitude. When Φ is positive, the amplitude of $(\omega - p)$ is greater than that of $(\omega + p)$. This has been observed experimentally.

The expressions "phase modulation" and "frequency modulation" are sometimes used interchangeably. It is true that they have much in common, in particular from the point of view of distortion. As long as we confine our interest to the *effect* of the distortion, once produced, the distinction is doubtful in value, if not quibbling in nature. The fundamental reason for the distortion is a rate of change of phase or frequency—call it what we will—which results in distortion when two components differently delayed, are recombined at the receiver.

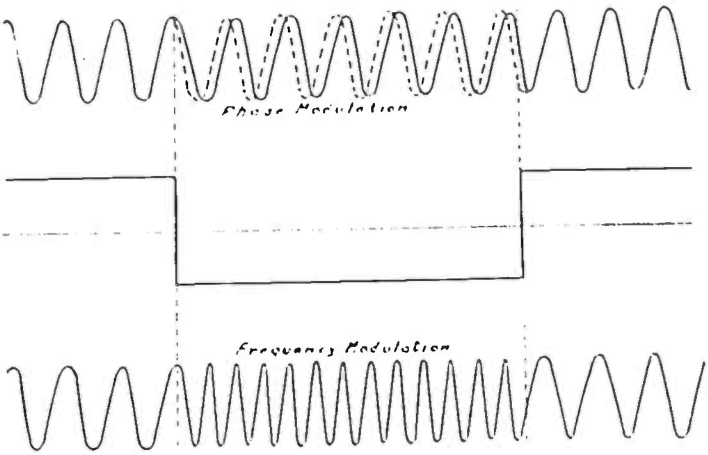

Fig. 13.

When, however, we direct our attention to the *cause* of the distortion, it is often useful to differentiate. In the first place, there are types of transmitter defects, (such as coupling from the output back to frequency-controlling oscillator) in which the frequency may be expected to vary through the modulation cycle and in such a way that the frequency is a function of the amplitude of the output. Here only the instantaneous *amplitude* of the audio wave is a determining factor of frequency. In the second place, there are defects in which the frequency of the control oscillator is held rigidly constant, but with different amplitudes different relative phase shifts are produced between that oscillator and the antenna. The latter might for example be produced by the impedance of one element of the circuit (e.g., an amplifier) changing as a function of input. In other words, the frequency in the first case and the phase in the second, are functions of the instantaneous amplitude of the modulating envelope. The existence of these two more or less distinguishable types of defect seems sufficient justification for the use of the distinguishing names, "frequency modulation" and "phase modulation."

While we are discussing this from the point of view of distortion, the distinction seems particularly fitting in view of the fact that it coincides with definitions made previously in connection with these phenomena as a means for, rather than as a hindrance to, communication. The difference is illustrated in Fig. 13, in which a wave of constant amplitude is modulated first in phase, and second in frequency, by the same square-topped modulating wave.[16]

Neither phase nor frequency modulation can be found by simple detection. The phases and amplitudes of the many components are such that the result with a square-law detector is the same whether or not the phase is modulated. When, however, the wave passes to the receiver along several paths of widely different optical lengths (as measured in wavelengths) this nice balance of detection products is destroyed. The result is that harmonics of pt appear in the output while the amplitude of the pt term may be reduced to small values. It should be possible to detect phase modulation locally if a sufficiently selective frequency analyzer were available.

Returning to the transmitters under discussion, the extent to which the performance of high-power amplifiers conforms to those requirements has been studied by means of cathode-ray oscillograph devices and frequency analyzers.

By applying a low-frequency modulating tone to the set, and by applying this input together with the rectified output to the two pairs of deflecting plates of a cathode-ray tube, the overall characteristic of the set may be visually observed, traced or photographed. In this way, these sets have been adjusted so as to fulfill the above mentioned requirements that this characteristic be linear.

An experimental study of the phase-modulation occurring in the transmitter shown in Fig. 12 was carried out by C. R. Burrows. A laboratory method was developed which made possible the direct measurement of the relative phases in different parts of the audio

[16] When modulation is sinusoidal, however, the distinction is not so obvious. In a phase-modulated wave, the phase may be written $(\omega t + k_2 \cos pt)$. Regarding this as the result of a variable frequency whose value equals the rate of change of the phase, we have

$$\text{Frequency} = 1/2\pi \ d/dt \ (\text{phase}) = 1/2\pi \ (\omega - k_2 p \sin pt).$$

This looks in some ways like frequency modulation but the audio amplitude of the latter contains a factor p not found in the first and also the audio terms of phase and frequency are 90 deg. out of phase. In other words, while the "wave forms" are the same in the sinusoidal case, they become different as soon as that case is departed from, *a thing which is always necessary in order to transmit signals.*

cycle. The method involved application of a voltage, derived from the output of the transmitter, to one of the pairs of deflector plates of a cathode-ray tube, while across the other pair was applied an unmodulated voltage of carrier frequency taken from the quartz crystal. For the latter, auxiliary harmonic generators were required so as to give the same final frequency. A phase-shifting device introduced before the harmonic generator made possible actual measurements of relative phase. To insure satisfactory operation of the oscillograph tube the frequency of the voltages applied to the two deflector pairs was reduced by the means of intermediate detection. From the shapes of the figures on the screen of the oscillograph, conclusions could be drawn regarding the phase. Thus if without tone input, the figure is a straight line, and if with tone no ellipse appears, it follows that there is no phase modulation present. In this way it was found that (as expected) little or no phase modulation is produced by the amplifier. A considerable amount may be expected in parts of the circuit where regeneration is present. Thus phase modulations up to plus and minus 90 deg. were sometimes measured in an improperly adjusted circuit, though the same circuit could be adjusted to very much smaller values. In a separate investigation it was found that the phase variation due to frequency modulation in the crystal itself was only a matter of a few degrees. This being hardly greater than the limit of accuracy of the measurement, it is believed that the improvements made in sets of a later design eliminate frequency modulation in the crystal as a source of distortion. In short, phase or frequency modulation in a properly designed and adjusted set should not be an important source of trouble.

It should be added that phase modulation, as well as amplitude distortion, can be produced only in a system employing non-linear elements.

With respect to the amplitude distortion in the transmitting sets being described, it is possible to obtain from 80 per cent to 100 per cent modulation without the production of noticeable distortion when normal commercial speech is transmitted.

Appendix

Curves A and C, Fig. 11, have been calculated from the formula

$$\text{Gain} = 10 \log_{10} M/N,$$

in which

$$M = (1 + 3/\pi^2)R$$

and

$$N = (1 + 3/\pi^2)nR$$
$$+ 3R \sum_{k=1}^{n-1}(n - k) \cos k\Phi \left(F(kS) + \propto F\sqrt{k^2S^2 + \pi^2} + 2\pi^2 \frac{\sin \sqrt{k^2S^2 + \pi^2}}{(k^2S^2 + \pi^2)^{3/2}} \right)$$

and

$$\propto = \frac{k^2S^2 - 2\pi^2}{k^2S^2 + \pi^2}$$

$$F(x) = \frac{\sin x}{x} + \frac{\cos x}{x^2} - \frac{\sin x}{x^3}.$$

The notation is as follows

R = radiation resistance of doublet in space
n = number of elements in array
Φ = phase shift per element (respectively zero and 180 deg. for cases given in Fig. 11)
$S = 2\pi D/\lambda$
D = spacing of elements.

These equations have been derived by taking account of the power which must be supplied to each element. In this method the e.m.f.'s in each wire necessary to maintain the assumed current distribution are calculated. These e.m.f.'s balance the drop through the radiation resistance of the element, assumed in free space, and the in-phase component of the e.m.f.s introduced by neighboring wires. The same equations have also been obtained by E. J. Sterba by integrations based on Poynting's theorem.

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A 12-COURSE RADIO RANGE FOR GUIDING AIRCRAFT WITH TUNED-REED VISUAL INDICATION*

By

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Summary—*This paper describes a directive radiobeacon (or radio range) of the visual indicating type, which has been developed by the Bureau of Standards to provide radio-marked courses at air terminals where more than four airways converge.*

The radio range is similar in design to the double-modulation type, with the addition of a third amplifier train and accompanying modulation frequency. It provides twelve equisignal zones which may be oriented within rather wide limits and made to coincide with the converging airways.

In order to prevent coupling between amplifier branches special means are employed to supply them successively rather than simultaneously. These methods are described in detail.

Three-phase radio-frequency supply is used to excite the amplifier trains, which is received from a single-phase oscillator of conventional type by means of a phase divider. This prevents possible trouble due to the employment of a three-phase oscillator with three tuned oscillating circuits.

To transfer the energy to the antenna system a goniometer with three primary coils is employed. These are displaced 120 deg. from each other and coupled to the two crossed loop antennas through two secondary coils whose displacement is 90 deg. This permits the use of two antennas to establish a space pattern from three amplifier branches.

Means for aligning this resultant space pattern with the airways are also discussed and several examples are given. Satisfactory tests have been made at distances of more than 100 miles.

I. Introduction

THE RAPID increase in the number of airways emanating from the more important airports of the United States has created a need for a directive radiobeacon capable of marking out a greater number of courses than has hitherto been possible. This paper describes a radio range (i.e., directive radiobeacon) which provides twelve beacon courses normally disposed 30 deg. from each other. By means of simple adjustments at the transmitter the angles between courses can be set as desired. This range fulfills all the present requirements for guiding aircraft along the civil airways.

The 12-course radio range is similar in operation to the 2- and 4-

* Dewey decimal classification: R526.1. Published by permission of the Director of the Bureau of Standards of the U. S. Department of Commerce.

course radio ranges previously described.[1] The increase in apparatus is not great. The same crossed-coil antenna system and the same circuit arrangements are employed, (see Fig. 1) except that three amplifier branches modulated to three different low frequencies are necessary. The modulation frequencies used are 65, 86.7 and 108.3 cycles, respectively. In addition a special goniometer is required. The rotor system of this goniometer comprises two coils crossed at 90 deg., and each connected in series with one loop antenna. For convenience of goniometer design each rotor coil is made up of three sections. Three stator coils, normally placed at 120 deg. with each other, are employed, one stator coil being connected to each power amplifier tube of the transmitting system.

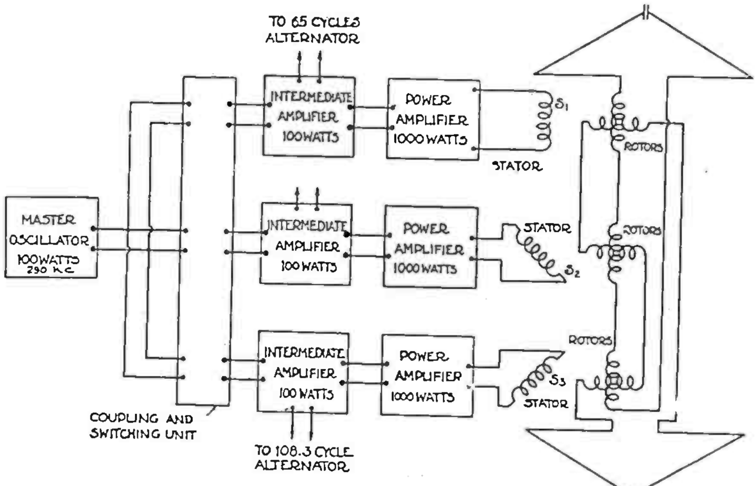

Fig. 1—Schematic diagram of 12-course radio range.

The 12-course beacon transmitter installed at College Park, Maryland, is shown in Fig. 2, and the special goniometer employed is pictured in Fig. 3.

Referring to Fig. 1, it will be observed that stator coil S_1 carries a radio-frequency current modulated to 65 cycles, stator coil S_2 carries a radio-frequency current modulated to 86.7 cycles and stator coil S_3 a radio-frequency current modulated to 108.3 cycles. Each stator coil, acting in conjunction with the two crossed rotor coils and the two crossed loop antennas, sets up a system which is electrically equivalent to a single loop antenna. The plane of this phantom antenna coincides with the plane of the stator winding for zero setting of the rotor, but rotates in space as the rotor system is rotated. Since there are three stator coils, placed at 120 deg. with each other, three such phantom antennas

[1] J. H. Dellinger and H. Pratt, "Radio aids to air navigation," Proc. I. R. E., 16, 890–920; July, 1928.

J. H. Dellinger and H. Diamond, "Radio developments applied to aircraft," *Mech. Eng.*, 51, 509–514; July, 1929.

(crossed at 120 deg.) exist. When special precautions are taken in circuit design the combined space pattern consists of a circular carrier with three figures-of-eight side bands crossed at 120 deg. See Fig. 4. The corresponding polar pattern as received on the reeds is shown in Fig. 5.

Now assume that a pilot is equipped with three vibrating-reed course indicators of the two-reed type,[2] the first tuned to 65 and 86.7 cycles, the second to 86.7 and 108.3 cycles and the third to 65 and 108.3 cycles,

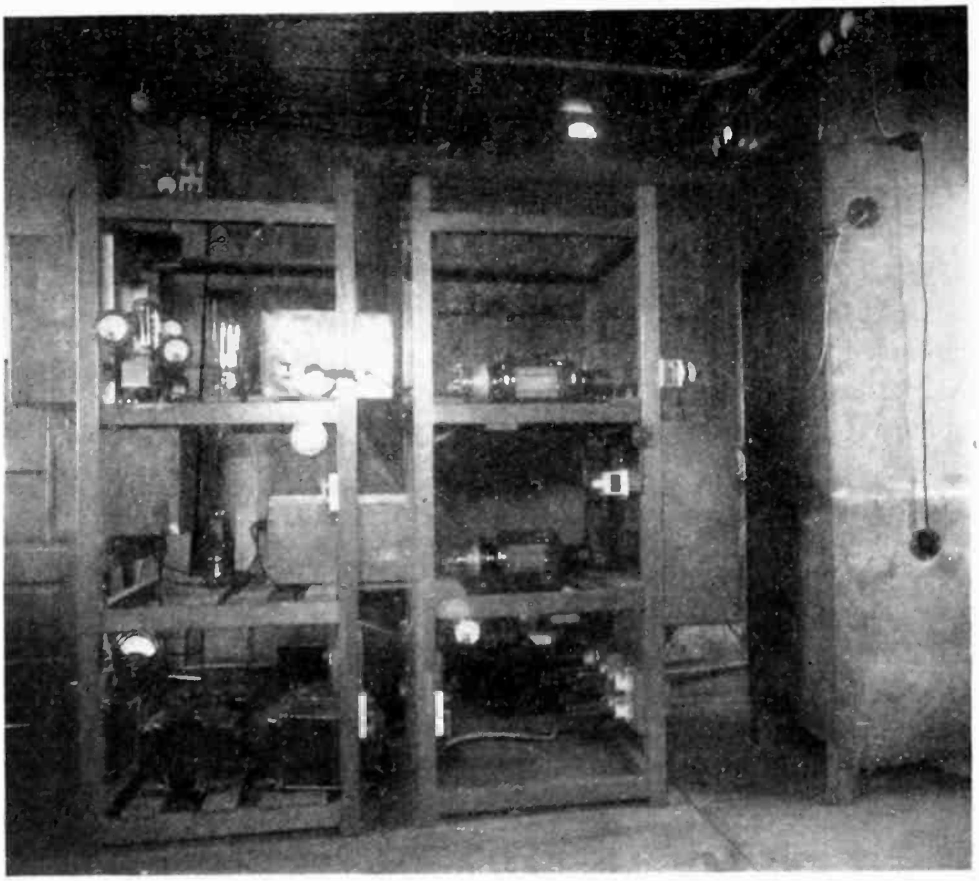

Fig. 2—Photograph of experimental 12-course range transmitter installed at College Park, Md.

respectively. Using the first indicator, he will find four courses at M, N, O, and P. With the second course indicator he will observe four courses at Q, R, S, and T. Similarly, with the third reed indicator, four courses at W, X, Y, and Z will be obtained. Two of each set of four courses (for example, M and N) have an equisignal zone of 1 to 1.5 deg. wide, while the width of this zone for the other two courses, (viz., O and P) is from 2 to 3 deg.

[2] F. W. Dunmore, "Design of tuned-reed indicators for aircraft radio beacon," *Bureau of Standards Journal of Research*, November, 1928. Research Paper No. 28.

To simplify the use of this beacon, a special three-reed indicator for receiving all twelve courses of the beacon without confusion has been developed.[3] This indicator is shown in Fig. 6.

II. Theory of 12-Course Radio Range

Since the goniometer stator windings are not at 90 deg. with each other, (see Fig. 1), undesirable intercoupling between S_1, S_2 and S_3 will exist, resulting in a combined beacon space pattern which cannot be used. Direct inductive coupling between stator windings is eliminated by the use of the three section rotor system, each stator winding being placed in a separate shielded compartment. Indirect coupling between stator windings by virtue of their mutual induction with the rotor system still exists. This coupling is somewhat more difficult to eliminate. By setting up the circuit equations for the goniometer system when no precautions are taken to prevent this coupling, and solving for the currents in S_1, S_2, and S_3, two facts may be determined:

(1) that the radio-frequency current in each stator winding is modulated to all three modulation frequencies of the beacon.

(2) that the amount of coupling between stator windings is independent of the rotor setting.

Several arrangements are possible for preventing this undesirable ntercoupling.

1. Link Circuit Arrangement

One arrangement consists of neutralizing this coupling by introducing inductive coupling between stator windings of opposite sense, as shown in Fig. 7. Thus S_1 is coupled to S_2 by coils $L_1L_1''L'_2L_2$ and to S_3 by coils $L_1L_1'L_3''L_3$. Similarly, S_2 is coupled to S_1 so noted and to S_3 by coils $L_2L_2''L_3'L_3$. The link circuits $L_1''C_1L_2'$, $L_2''C_2L'_3$, and $L''_3C_3L_1'$ are each tuned to the beacon carrier frequency. It is evident that by virtue of these link circuits a current flowing in any stator induces in each of the other stators an e.m.f. 180 deg. out of phase with that current. On account of the indirect coupling between stators by way of the rotor system, a current flowing in any one stator induces in each of the other stators a voltage exactly in phase with that current. By properly adjusting the amount of coupling due to the link circuits, an exact neutralization may be obtained.

The carrier-frequency currents in S_1, S_2, and S_3 (Fig. 1) have been assumed to be in time phase. Since the stator windings are displaced by 120 deg. space phase, the resultant carrier transmitted is zero. A circu-

[3] F. W. Dunmore, "A tuned reed course indicator for the 4- and 12-course aircraft radio range," this issue of the Proceedings, page 963. To be published in forthcoming issues of the *Bureau of Standards Journal of Research*.

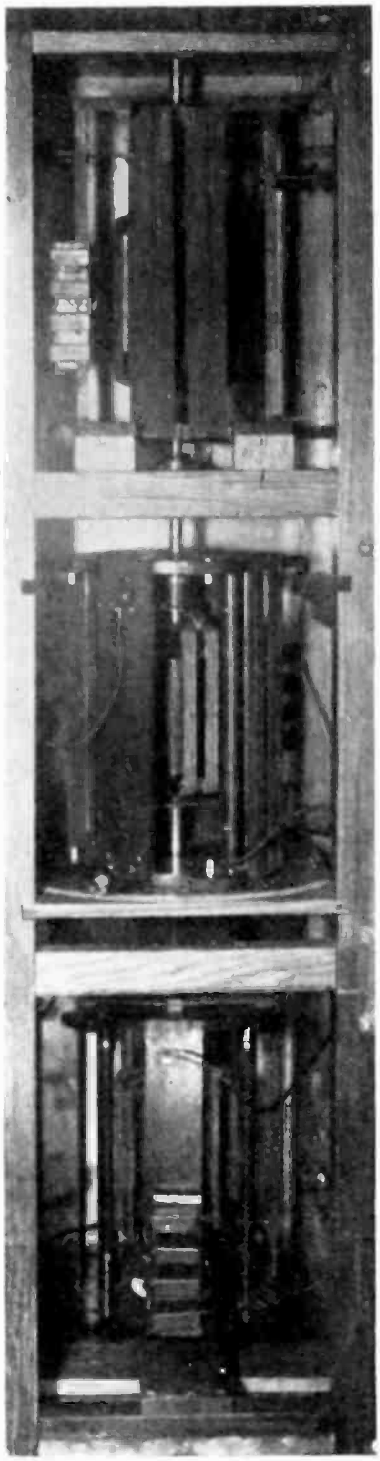

Fig. 3—Photograph of special 3-stator goniometer employed with 12-course radio range.

lar carrier can, however, be supplied by the use of a vertical antenna, extending along the beacon tower and coupled through a 50-watt amplifying tube to the master oscillator of the beacon transmitting set. For optimum results, this antenna must be in accurate tune to the beacon carrier frequency. The polar patterns shown in Figs. 4 and 5 are for this transmitting arrangement.

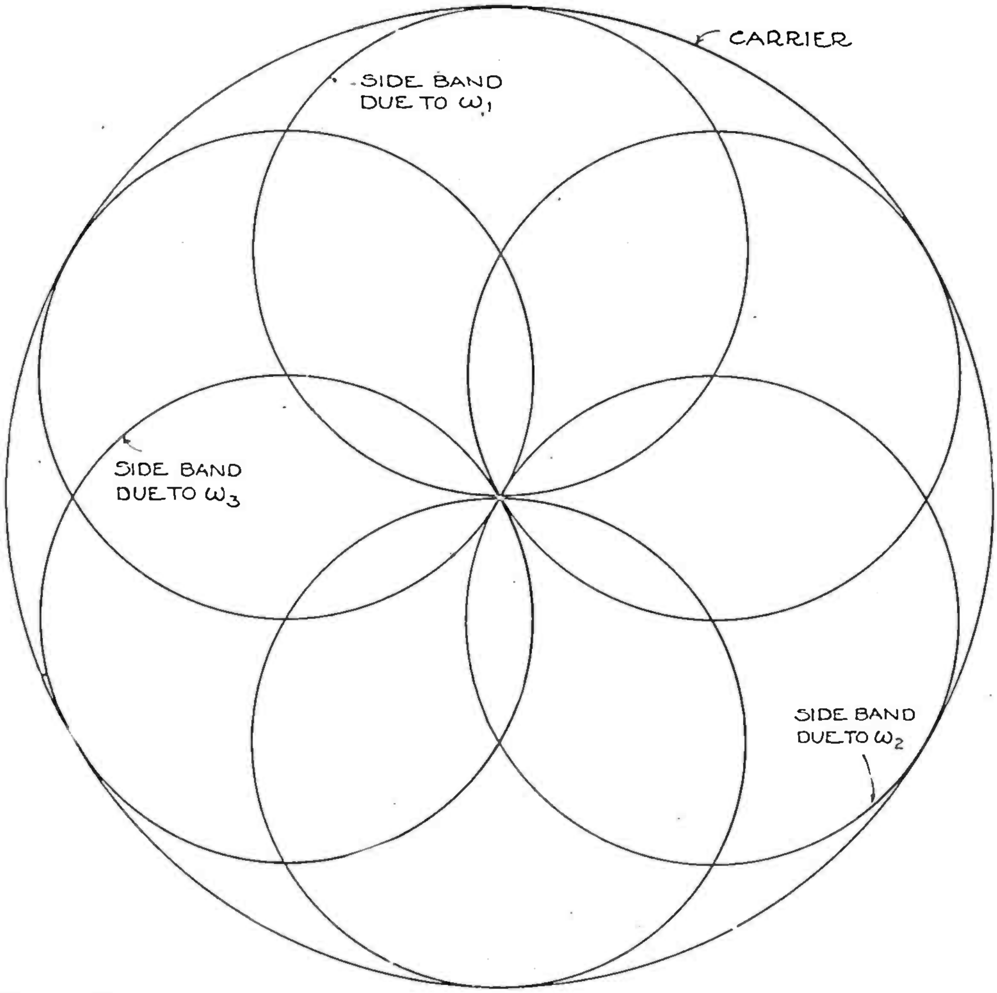

Fig. 4—Space pattern radiated by 12-course radio range (link circuit type).

2. Automatic Switching Using Three-Phase Grid Biasing

The use of link circuits to eliminate coupling between stator windings and the need for supplying an auxiliary carrier to replace the one suppressed by the goniometer system may both be precluded if means are provided for exciting but one stator winding at a time. The complexity of the scheme outlined above led to experiments toward devising such means.

An obvious arrangement for accomplishing this consisted of inserting a commutator carrying three segments occupying successive 120-deg. arcs in the supply leads from the master oscillator to the grids of the intermediate power amplifier tubes. In this way voltage was supplied to each amplifier train only during one third of each revolution of the commutator. The switching was performed at a sufficiently rapid rate so that in spite of the interruptions in the signals transmitted, the reeds maintained their vibration amplitudes due to mechanical inertia.

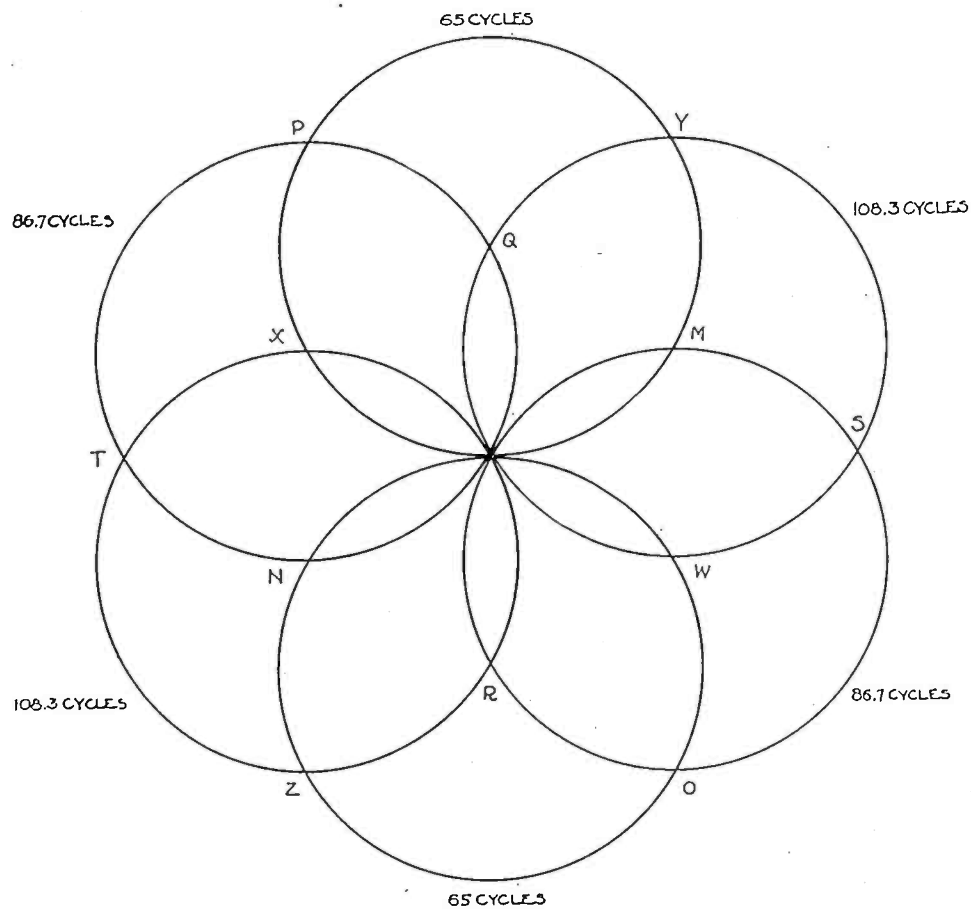

Fig. 5—Polar pattern as received on the reeds.

This method had two distinct disadvantages, (1) the power transmitted by each amplifier train was reduced to one-third normal, and (2) surges in the transmitted wave due to high-speed switching resulted in shock excitation of the reeds used for reception, the net effect in the beacon operation being a virtual widening of the equisignal zones.

The next circuit arrangement used proved considerably more successful, the switching of power being performed at an audio-frequency rate by means of the circuit shown in Fig. 8. A d-c biasing voltage, common to the grid circuits of the three intermediate amplifying tubes, is employed and is of such magnitude that (in the absence of the three-phase 300-cycle supply) no power is transmitted through these tubes. In series with this common d-c biasing voltage, each grid circuit has induced in it one of the phase voltages of the three-phase 300-cycle supply. The resultant biasing voltages impressed on the grid of each of the three intermediate amplifying tubes is indicated by curves A, B, and C, respectively, of Fig. 9. Since $-E_c$ is the cutoff voltage, each tube passes power only during the positive half-cycle of its a-c biasing voltage. As a result no two amplifier tubes transmit power simultaneously except during the small intervals of time a-b, c-d, etc., as shown

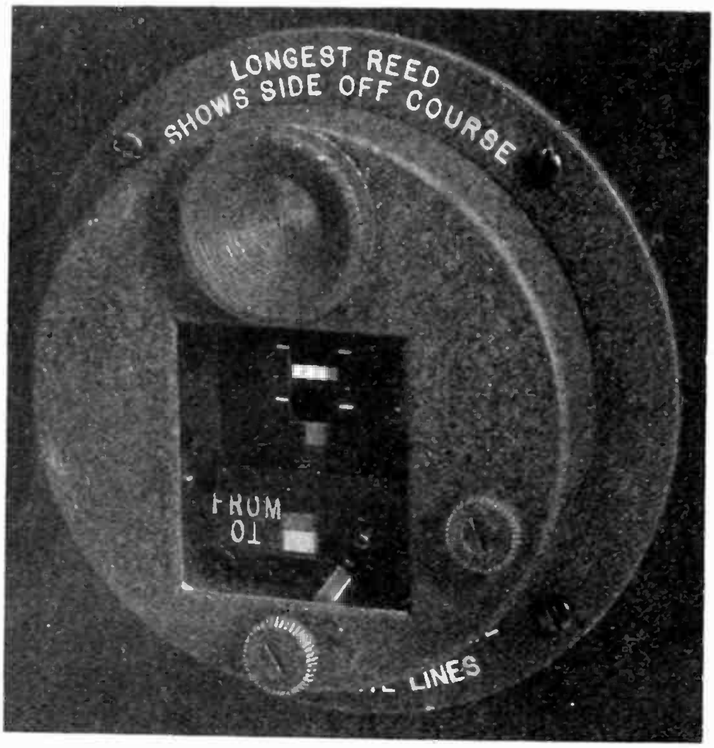

Fig. 6—Photograph of special course indicator for 12-course radio range.

in Fig. 9. The proportion of power transmitted during these intervals is less than the ratio of shaded area aob to total area of transmission per cycle for one tube $E_c aob$, since the tube is then operating on the knee of its characteristic curve. The amount of coupling still present may be further reduced by increasing the common d-c bias voltage, but this results also in a decrease of total power transmitted. In practice this further reduction is unnecessary, the beacon performance proving satisfactory without it.

Neglecting the small amount of coupling present, the beacon space pattern becomes as shown in Fig. 10. Note that, since but one stator

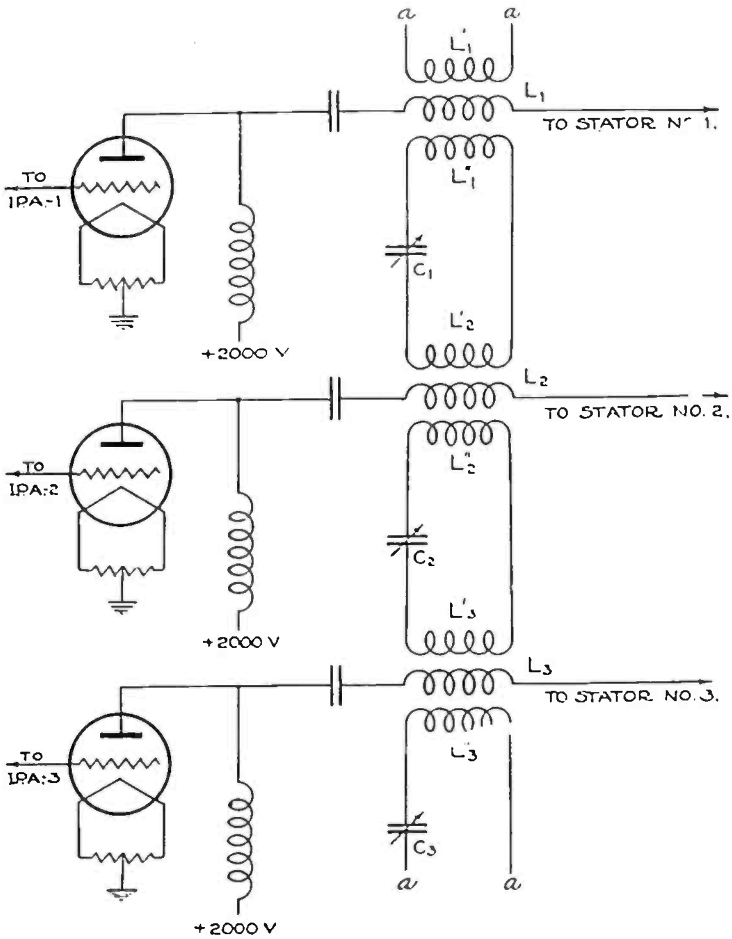

Fig. 7—Link circuit arrangement for preventing intercoupling between primary goniometer windings.

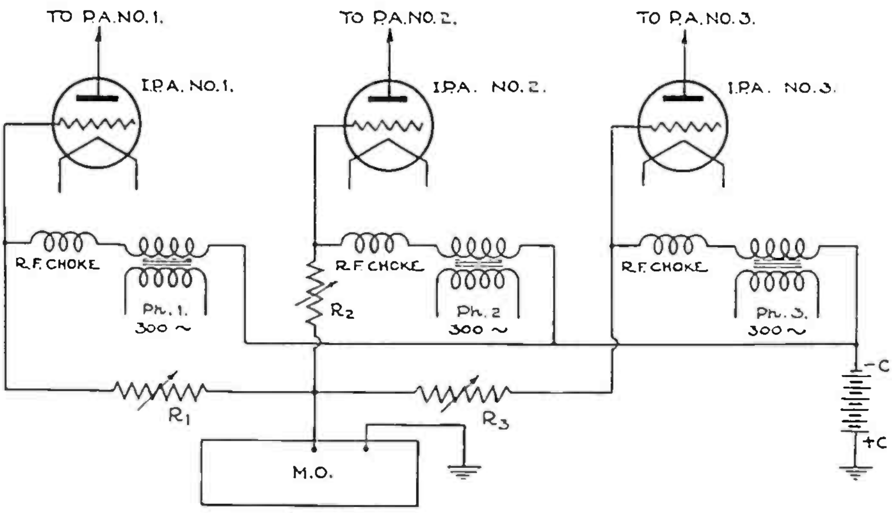

Fig. 8—Grid-biasing arrangement for switching radio-frequency supply successively to each amplifier train.

winding is excited at a time, there are three independent carrier waves in the beacon space pattern. Assuming square-law detection, the polar pattern as received on the reeds is shown in Fig. 11.

The choice of frequency of the three-phase alternator used for the grid bias switching is important. The fundamental and second harmonic of this frequency must not be closer than approximately 20 cycles to any of the harmonics of the modulation frequencies used in the beacon. To illustrate the reason for this requirement assume that a frequency of 262 cycles were chosen for the three-phase alternator. The fourth harmonic of 65 cycles and the third harmonic of 86 2/3 cycles are each 260 cycles. Under this condition a beating of the reed indicator at the difference frequency of 2 cycles is obtained.

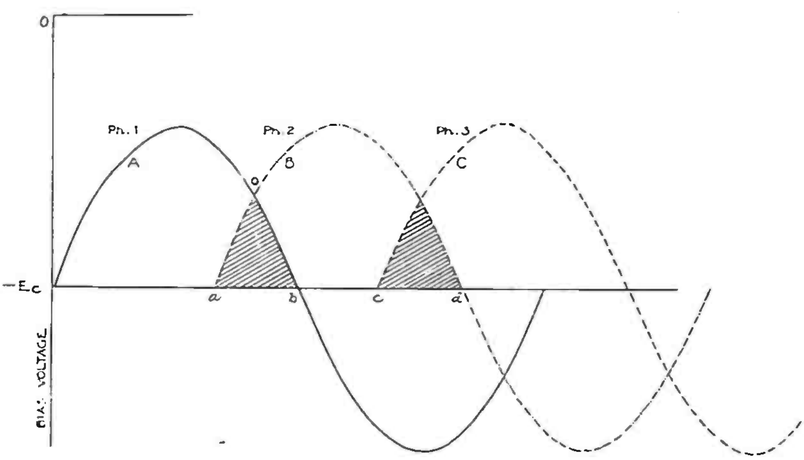

Fig. 9—Resultant biasing voltages on grid of each intermediate-amplifier stage.

The net power transmitted by each amplifier branch when using this method of switching is, of course, reduced considerably below normal; primarily, because power is transmitted only during approximatly one-half the time, and secondly, because during that time the operating point travels along the tube characteristic from the point of cut-off to the point of normal operation and back to the cut-off point. The net reduction in power is, however, not as great as might be expected from the above, since it is possible to force the tubes consideraly above normal during the half cycle of operation and still maintain safe average plate dissipation. Using this system, with 8 amperes in each loop antenna, an operating range of 100 miles was obtained.

3. Three-Phase Radio-Frequency Switching.

The use of three-phase audio frequency for switching the radio-frequency power as described above suggested that the same results could be obtained if a source of three-phase radio frequency were available. Each phase of this three-phase radio-frequency supply could then be used for supplying carrier voltage to the grid of one intermediate amplifier, the three-phase source thus serving as a master oscillator as well as a switching device. The advantages to be gained are fourfold: (a) a decided increase in the power transmitted by each amplifier branch; (b) the elimination of a three-phase audio-frequency unit; (c) the operation of the intermediate amplifier tubes under normal conditions and not with the high grid voltages necessary in the method of

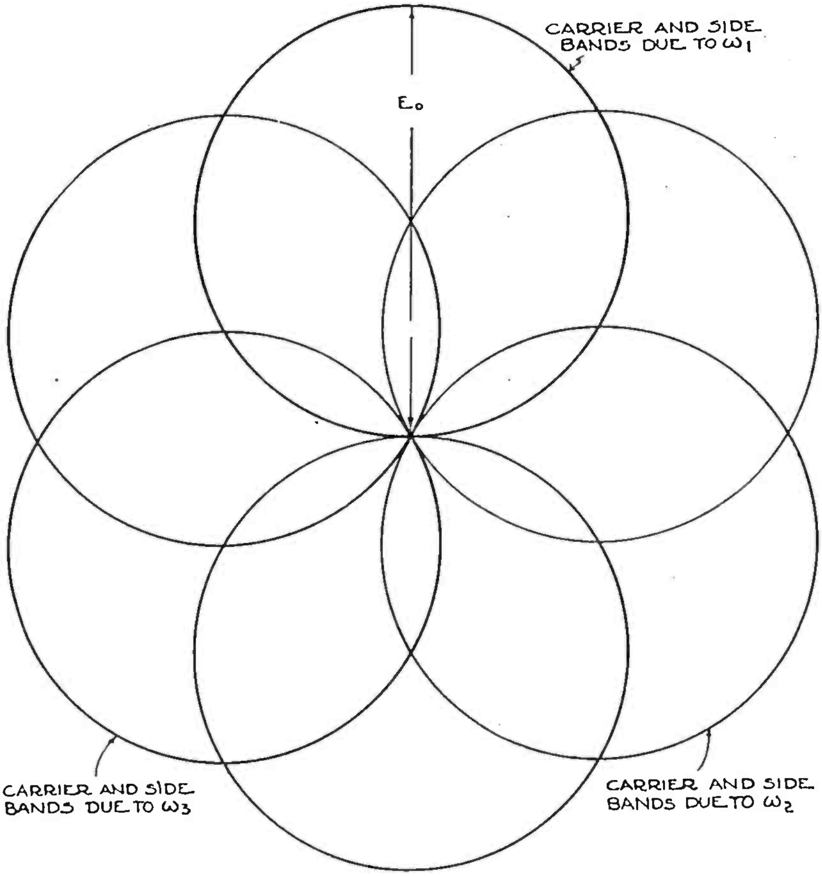

Fig. 10—Space pattern radiated by beacon when using grid-biasing arrangement for preventing intercoupling between primary goniometer windings.

grid-bias switching; and (d) the elimination of interference caused by the audio-frequency used for grid biasing.

A three-phase radio-frequency oscillator of the type described by R. Mesny[4] was constructed to serve as the three-phase source. However, advantages of simplicity in design and operation led to the development of the phase-splitting arrangement shown in Fig. 12.

[4] R. Mesny, "Generation of polyphase oscillations by means of electron tubes," Proc. I. R. E., 13, 471–476; August, 1925.

Referring to Fig. 12, a single-phase Colpitts oscillating circuit is employed. Advantage is taken of the fact that the voltages from points a and b to ground are approximately 180 deg. out of time-phase. It is desired that the voltages from points m, n, and p to ground have a 120 deg., time-phase displacement. The values of the condenser, C_1, and the inductance, L, are so chosen that the voltage from m to ground leads the voltage from q to ground by 60 deg. and the voltage from n to ground lags the same voltage by 60 deg. The condenser, C_o, is so chosen that the voltage from q to ground is 180 deg., out of phase with

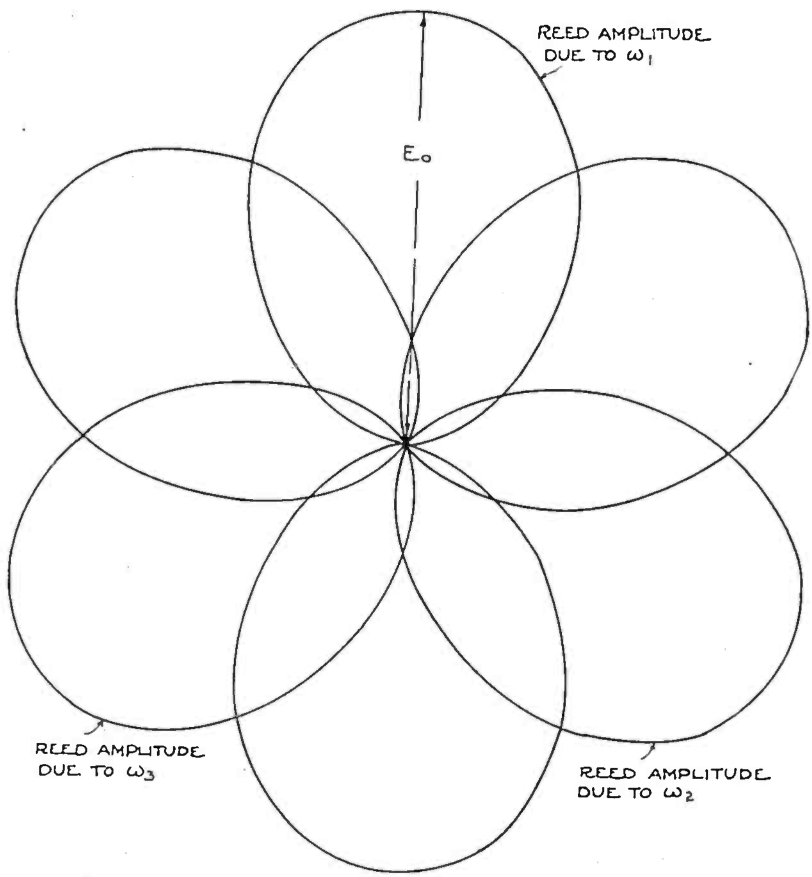

Fig. 11—Received pattern corresponding to space pattern of Fig. 10.

the voltage from p to ground. The variable resistances shown aid materially in making these adjustments. The impedances of the various branches must be such that the voltages from m, n, and p to ground are equal in magnitude as well as 120 deg. out of phase. The voltage vector diagram corresponding to this adjustment is shown in Fig. 13.

As will be explained below, the time-phase displacement between the voltages applied to the grids of the three intermediate amplifying tubes must remain constant to prevent a shifting of the beacon courses in space. A variation of the grid to filament tube impedances would tend to cause such a displacement. The resistances, R_g (having values of 7500 ohms) are connected in parallel with the grid to filament impedances to minimize such variation. With this arrangement, a 20 per cent change in the grid to filament impedance of any tube results in but a 2 deg. variation in the time-phase displacement. Similarly, a change in the oscillator frequency of 2-kc results in a 0.2 deg. time phase variation. These are well within the permissible limits. A description of the means employed for adjusting the voltages from points m, n, and p to ground to their proper phase relationships is of interest.

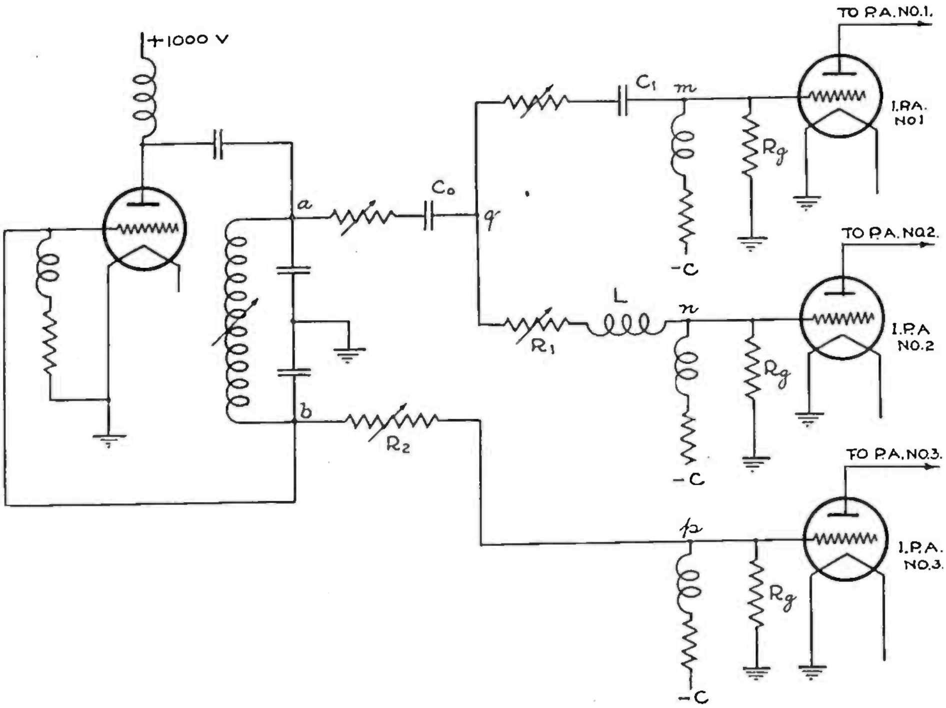

Fig. 12—Phase-splitting arrangement for obtaining three-phase radio-frequency switching.

The use of 7500-ohm resistors in parallel with the tube input impedance has already been mentioned. With these resistors in use, the load impedances of the phase divider remain sensibly constant and adjustments can be made on that basis.

First, condensers C_o and C_1 are short-circuited and the output of the master oscillator is adjusted by raising or lowering its plate voltage until the grid meter in the branch mo reaches a convenient reading, I_o. The current in branch no will then read, say I_1. Condenser C_1 is then inserted in the circuit and its capacity adjusted until the current in the circuit mo is reduced to one-half of its former value. At the same time the master oscillator output is kept constant by keeping the grid current in branch *no* at I_1. When the current is halved, assuming a purely resistive circuit in the first instance, the phase angle of the voltage *mo* has been adjusted to a 60 deg. leading angle with the voltage *qo*.

Next, with C_o still short-circuited, the process is repeated in the branch *no*, adjustment of L being made until the desired angle is secured. At the same time the resistance R_1 is varied to keep the total circuit impedance the same as that of the condenser branch. This is readily noted from the grid ammeter readings which should read the same in both branches.

The voltages *mo* and *no* are now in the correct phase relationship. There remains the selection of the proper value for C_0 in order that *qo* be in 180 deg. time phase with *po*. This adjustment is not as straightforward as the preceding ones.

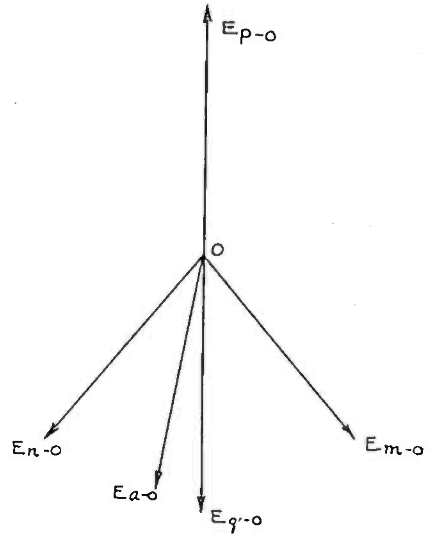

Fig. 13—Voltage vector diagram corresponding to circuit of Fig. 12.

As previously explained, when the phase relations are correct, the carrier may be considered as a rotating figure-of-eight of constant magnitude. This effect is utilized in making the final adjustment.

With d-c excitation on the plates of the intermediate power amplifier tubes and the condenser C_o in the circuit, the output of each amplifier branch is separately balanced in magnitude by adjustment of R_2 until all three are equal as shown by the antenna currents. All the branches are then excited through the phase divider and the goniometer rotated. When C_o is of the correct value the antenna current remains constant for all positions of the rotor. C_o is consequently varied until this constancy of antenna current occurs. A readjustment of R_2 is necessary after each change of C_o in order that the magnitudes of the three voltages remain equal.

The three voltages are now in correct time phase and will not change appreciably with changes in amplifier load or tube constants encountered in normal operation.

There is one important difference between the radio-frequency method of switching and the audio-frequency grid-bias method previously described. In the latter method coupling between stator windings was eliminated because the stators were never excited simultaneously. In the radio-frequency method the coupling is eliminated chiefly due to the time-phase displacement.

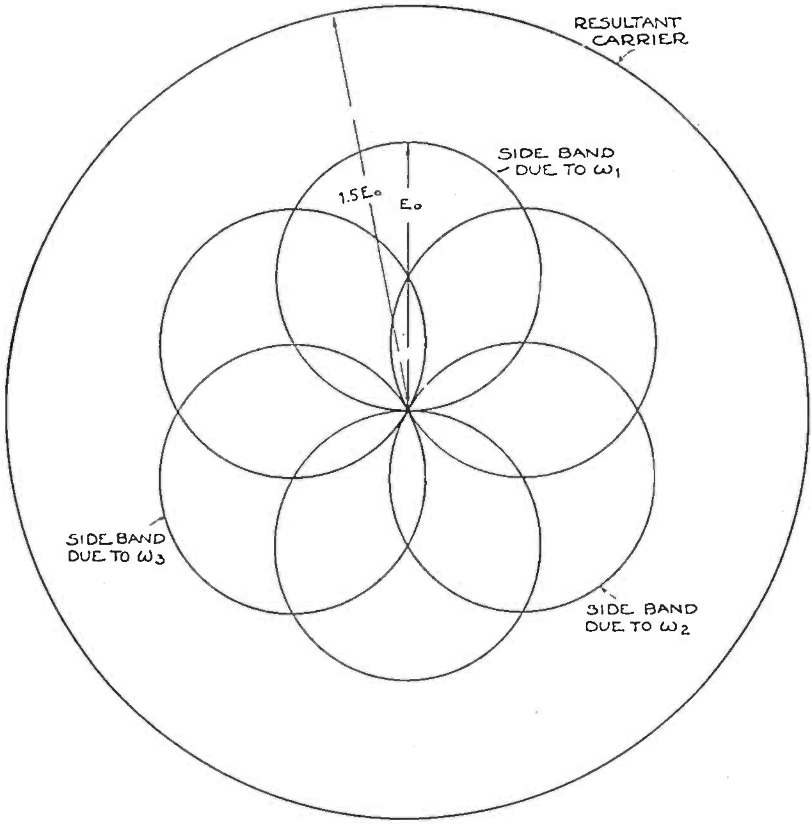

Fig. 14—Beacon space pattern when using radio-frequency switching.

The beacon space pattern obtained when using this system and the corresponding received polar diagram are shown in Figs. 14 and 15 respectively. Note that even for the same power transmitted by each amplifier branch this system yields 50 per cent greater received signal than the system employing grid-bias switching. The patterns of Figs. 14 and 15 are for the special case when the three stator windings are exactly 120 deg. apart in space phase and the three voltages applied to these stators exactly 120 deg. apart in time phase. The general expression for the space pattern for any time-phase relationship between these three voltages is given in (1). The first term to the right of the equality sign represents the carrier, while the second, third, and fourth terms represent the side bands due to modulation frequencies, ω_1, ω_2, and ω_3, respectively. As will be observed, the carrier comprises a figure-of-eight revolving in space at the carrier-frequency rate. The curve representing the carrier in Fig. 14 is in reality the locus in space of successive maxima of the rotating figure.

$$
\begin{aligned}
e_p = KE_o &\left\{ \begin{aligned} &\cos(\omega t+\Delta_1)\cos\theta+\cos\left(\omega t-\frac{2\pi}{3}+\Delta_2\right)\cos\left(\theta-\frac{2\pi}{3}\right)\\ &+\cos\left(\omega t-\frac{4\pi}{3}+\Delta_3\right)\cos\left(\theta-\frac{4\pi}{3}\right)\end{aligned}\right\}\\
&+\frac{KE_1}{2}\left[\left\{\cos(\omega t+\Delta_1-\omega_1 t)+\cos(\omega t+\Delta_1+\omega_1 t)\right\}\cos\theta\right]\\
&+\frac{KE_2}{2}\left[\left\{\cos\left(\omega t-\frac{2\pi}{3}+\Delta_2-\omega_2 t\right)\right.\right.\\
&\qquad\left.\left.+\cos\left(\omega t-\frac{2\pi}{3}+\Delta_2+\omega_2 t\right)\right\}\cos\left(\theta-\frac{2\pi}{3}\right)\right]\\
&+\frac{KE_3}{2}\left[\left\{\cos\left(\omega t-\frac{4\pi}{3}+\Delta_3-\omega_3 t\right)\right.\right.\\
&\qquad\left.\left.+\cos\left(\omega t-\frac{4\pi}{3}+\Delta_3+\omega_3 t\right)\right\}\cos\left(\theta-\frac{4\pi}{3}\right)\right]
\end{aligned}
\tag{1}
$$

where e_p is the field intensity at any point, P, in space as a polar function of the angle θ.

$E_1/E_0\times 100$ is the percentage modulation in amplifier branch 1 due to ω_1

$E_2/E_0\times 100$ and $E_3/E_0\times 100$ have similar meanings.

Δ_1, Δ_2 and Δ_3 are the time-phase displacement angles of the three carrier components from exact 120-deg. relationship.

The corresponding expression for the received polar pattern is given by equation (2), assuming square-law detection. The three terms to the right of the equality sign represent the reed amplitudes due to ω_1, ω_2, and ω_3, respectively.

$$
e_r = KK' \left[\begin{aligned}
& E_oE_1 \cos \omega_1 t \left\{ \cos^2 \theta + \cos \alpha_1 \cos \theta \cos \left(\theta - \frac{2\pi}{3} \right) \right. \\
& \left. + \cos \alpha_3 \cos \theta \cos \left(\theta - \frac{4\pi}{3} \right) \right\} \\
& + E_oE_2 \cos \omega_2 t \left\{ \cos^2 \left(\theta - \frac{2\pi}{3} \right) \right. \\
& + \cos \alpha_1 \cos \theta \cos \left(\theta - \frac{2\pi}{3} \right) \\
& \left. + \cos \alpha_2 \cos \left(\theta - \frac{2\pi}{3} \right) \cos \left(\theta - \frac{4\pi}{3} \right) \right\} \\
& + E_oE_3 \cos \omega_3 t \left\{ \cos^2 \left(\theta - \frac{4\pi}{3} \right) \right. \\
& + \cos \alpha_3 \cos \theta \cos \left(\theta - \frac{4\pi}{3} \right) \\
& \left. + \cos \alpha_2 \cos \left(\theta - \frac{2\pi}{3} \right) \cos \left(\theta - \frac{4\pi}{3} \right) \right\}
\end{aligned} \right] \qquad ()
$$

where e_r is the signal received on the reeds as a polar function of the angle θ.

$$\alpha_1 = \frac{2\pi}{3} + \Delta_1 - \Delta_2$$

$$\alpha_2 = \frac{2\pi}{3} + \Delta_2 - \Delta_3$$

$$\alpha_3 = \frac{2\pi}{3} + \Delta_3 - \Delta_1$$

Reference to Fig. 16 will show that α_1, α_2, and α_3 are, respectively, the time-phase displacement in degrees between the three carrier voltage vectors of the transmitting system. The amplitudes of these voltage vectors are in turn space functions of the angle θ. It should be noted that each set of side bands transmitted by one amplifier branch beats with its own carrier and also with the in-phase components of the other two carriers of the system.

As noted above, a departure of the three carrier voltages from exact 120-deg. time-phase displacement results in a shifting of the beacon courses in space. The extent of this shift can be determined by substituting special values for the time-phase displacements in equations (1) and (2) above.

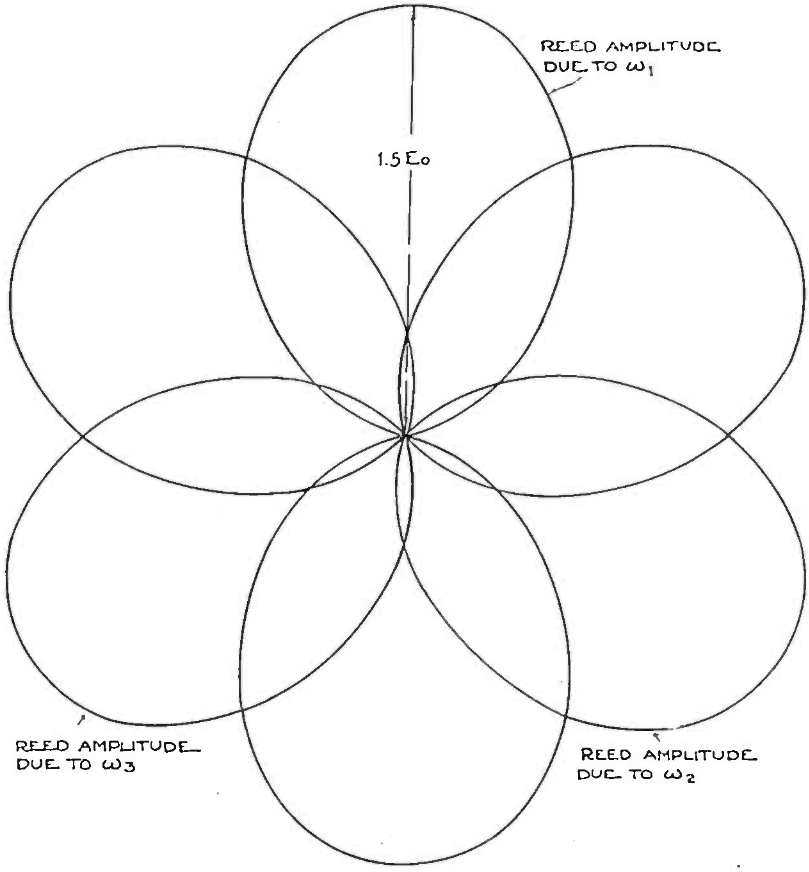

Fig. 15—Received pattern corresponding to space pattern of Fig. 14.

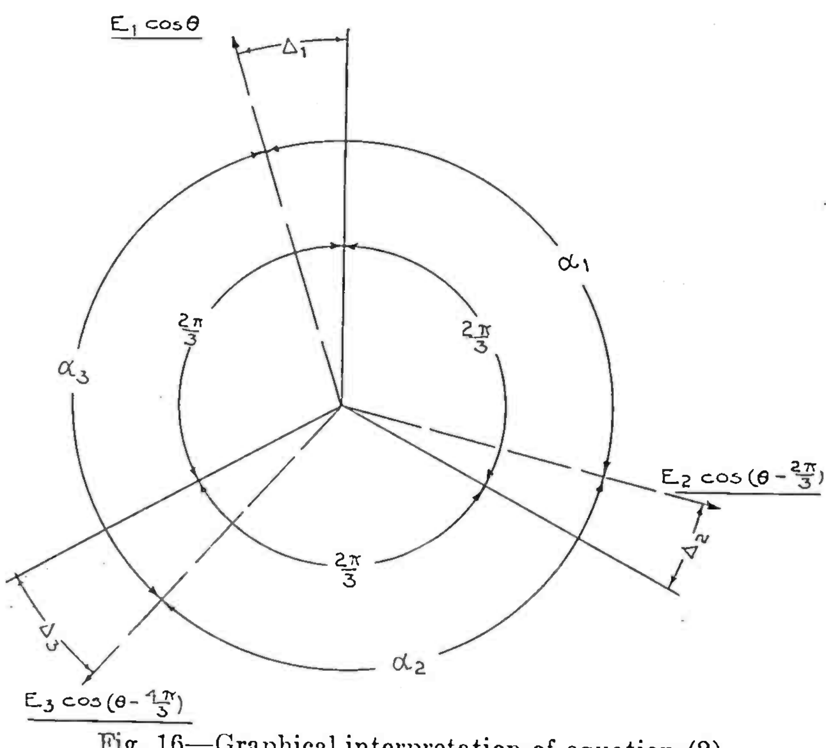

Fig. 16—Graphical interpretation of equation (2).

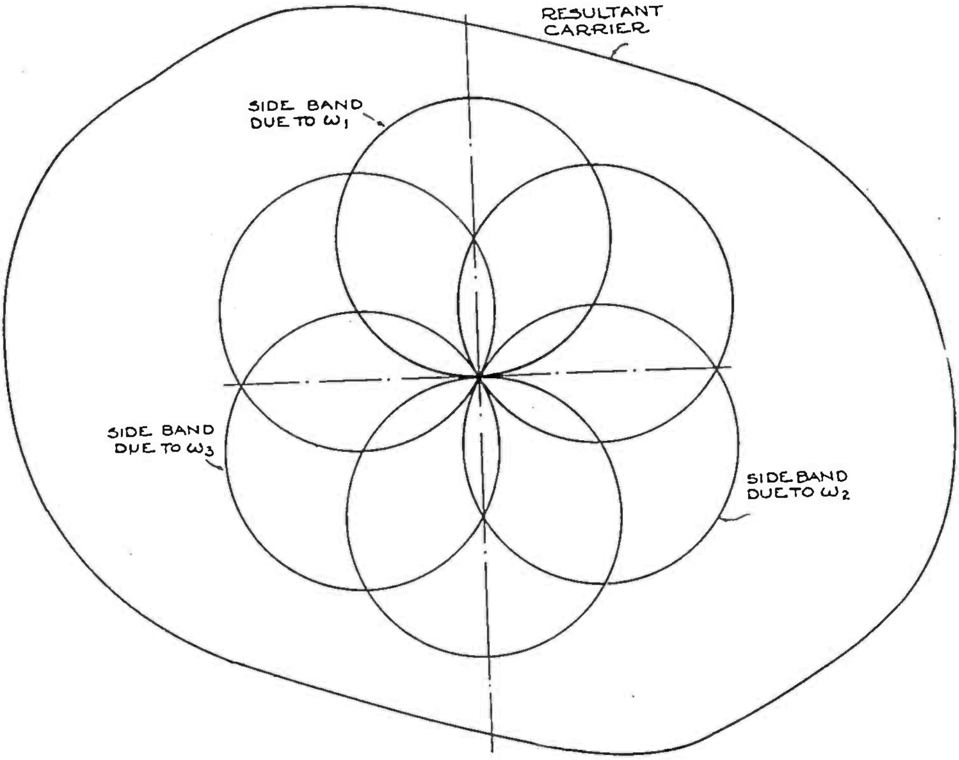

Fig. 17—Beacon space pattern when the time phase of one of the three carrier-frequency voltages of the system is displaced by 30 deg. from its normal value.

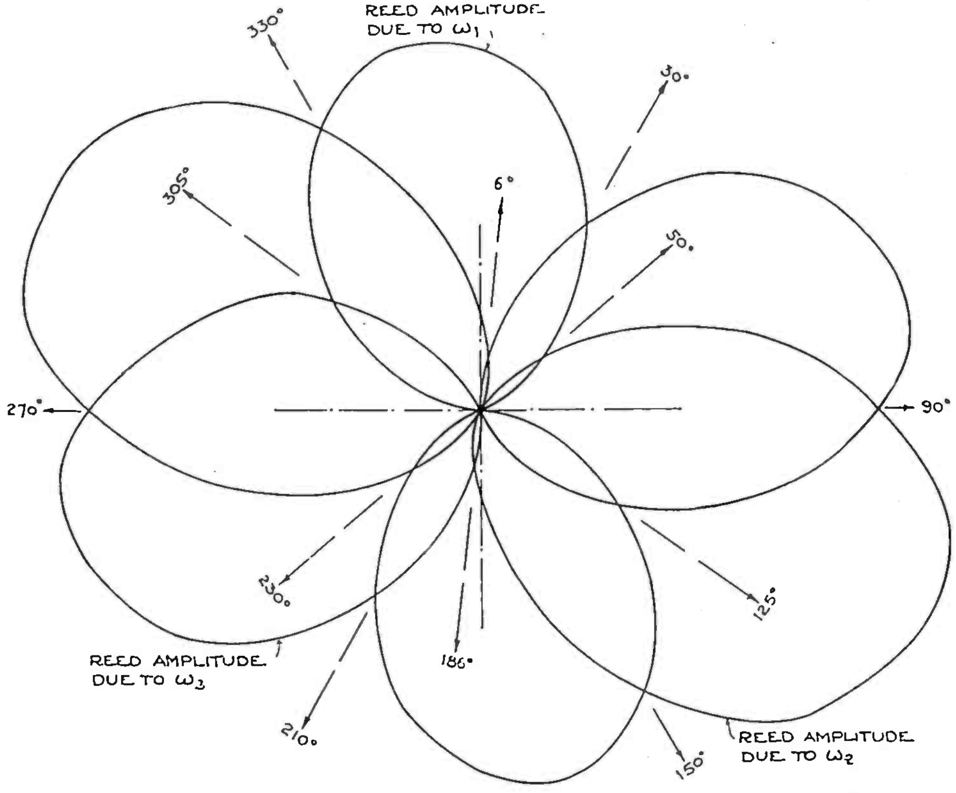

Fig. 18—Received pattern corresponding to space pattern of Fig. 17.

For the conditions

$$\alpha_1 = 120 \text{ deg.}$$

$$\alpha_2 = 150 \text{ deg.}$$

$$\alpha_3 = 90 \text{ deg.}$$

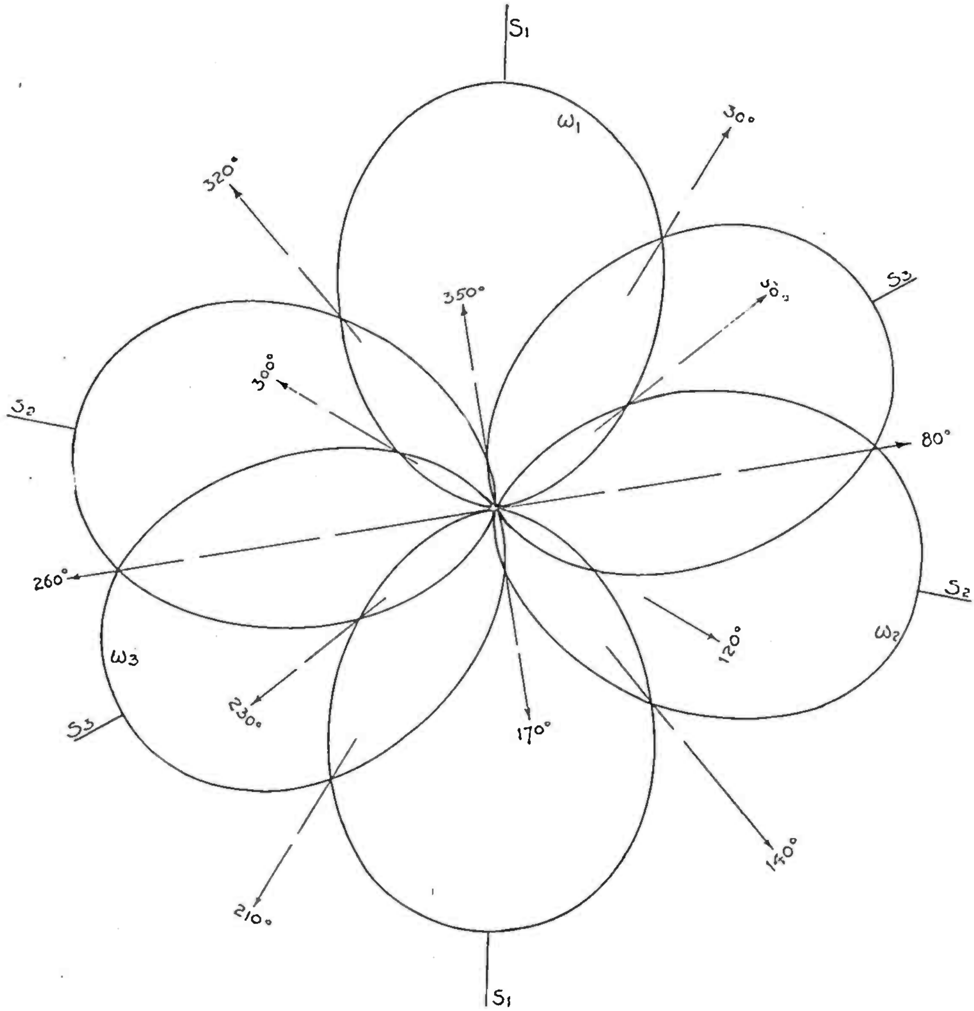

Fig. 19—Displacement of beacon courses obtained when shifting stator No. 2 by 20 deg. from its normal position.

which is equivalent to a 30-degree time-phase displacement of one of the voltage vectors, the beacon space pattern becomes as shown in Fig. 17 and the received pattern as shown in Fig. 18. Note that the maximum shift of any course from its normal position is 10 deg. For normal variations of time-phase displacements of the three voltage vectors, the shifting of the beacon courses in space is negligible.

III. Methods of Shifting the Beacon Courses

The same means are available for shifting the beacon courses from their 30-deg. space relationship (in order to align them with the airways emanating from a given airport) as have been previously described for application to the four-course radio range.[5] A simpler method, applicable only to the 12-course beacon, is, however, preferable.

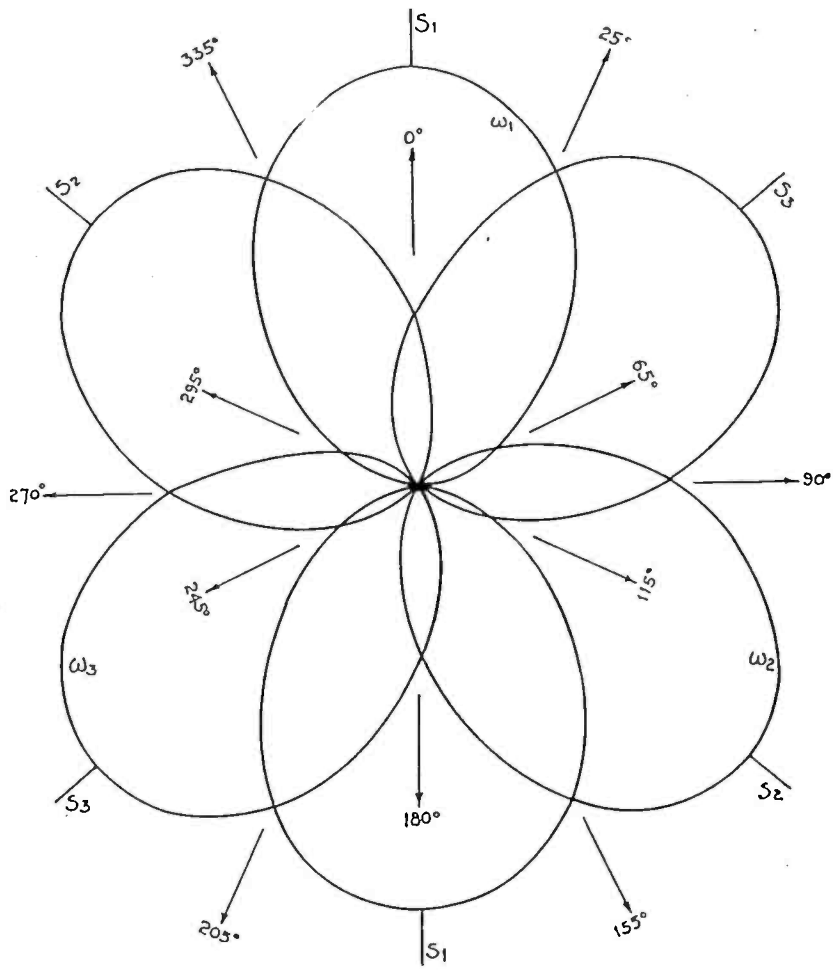

Fig. 20—Displacement of beacon courses when S_2 and S_3 are separated by 130 deg. from S_1 instead of the normal value of 120 deg.

This method consists of displacing the stator windings from their normal 120-deg. positions, and may be used when either audio-frequency grid-bias switching or radio-frequency switching is employed.

When audio-frequency switching is used, as the stators are excited one at a time, displacement of a given stator results in an equivalent

[5] H. Diamond, "Applying the visual double-modulation type radio range to the airways," *Bureau of Standards Journal of Research*, February, 1930. Research paper No. 148. Proc. I.R.E., 17, 2158–2189; December, 1929.

displacement of the field pattern due to that stator. Thus, Fig. 19 is the received polar pattern when stator No. 2 is displaced by 20 deg. from its normal position. The pattern due to S_2 is similarly displaced. (Compare with Fig. 11). The eight courses formed by the intersection of the pattern due to stator S_2 with the patterns due to S_1 and S_3 are all shifted by 10 deg. in the direction of displacement of

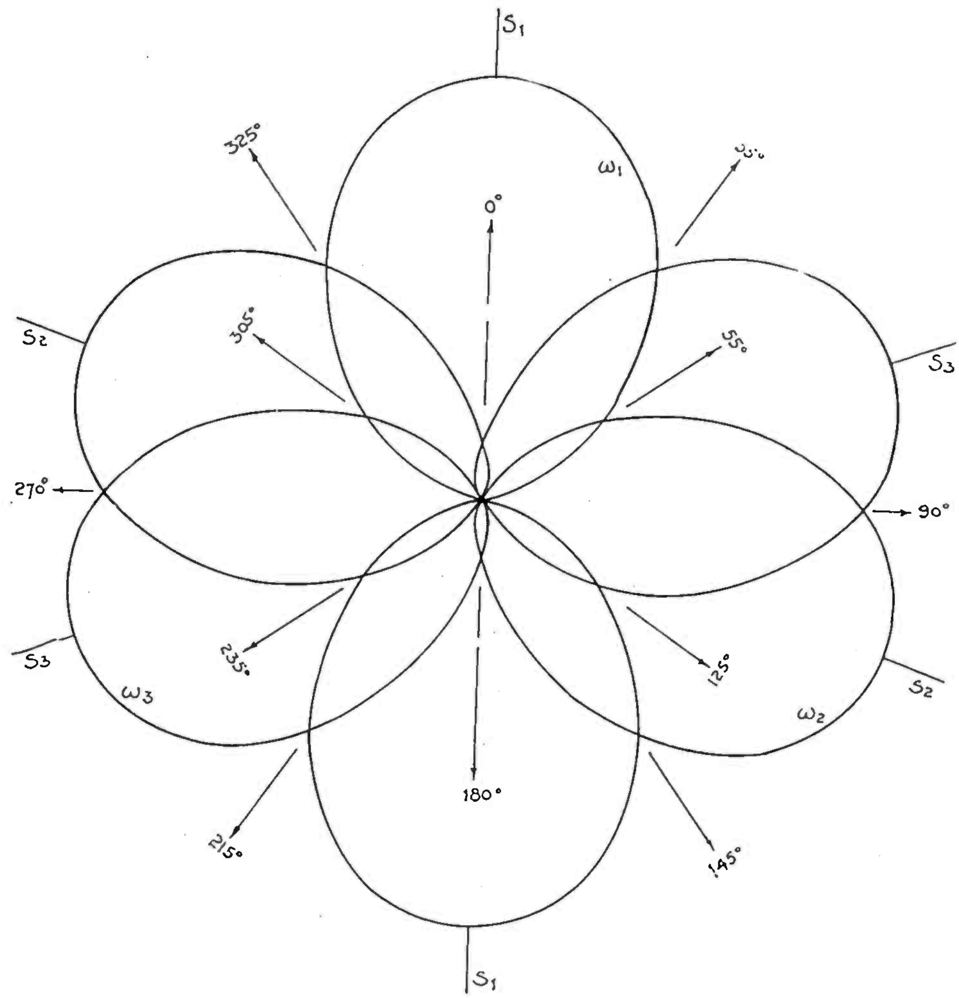

Fig. 21—Displacement of beacon courses when S_2 and S_3 are separated by 110 deg. from S_1.

S_2, while the four courses due to the patterns of S_1 and S_3 remain fixed in their normal positions. A greater variation of the angles between courses may be obtained by displacing two of the three stators in equal amounts, but in opposite directions. Fig. 20 corresponds to a case of this kind, S_2 and S_3 being separated from S_1 by 130 deg. instead of the normal value of 120 deg. The four courses formed by the intersection of the patterns due to S_1 and S_2 are here rotated 5 deg. clockwise. The four courses due to the patterns of S_1 and S_3 are rotated 5 deg. counter-clockwise, while the four courses due to S_2 and S_3 remain fixed. Fig. 21 is for the case when S_2 and S_3 are separated from S_1 by 110 deg. In this case the four courses due to S_1 and S_3 are rotated 5 deg. counter-clockwise, the four courses due to S_1 and S_2 are rotated 5 deg. clockwise, and the four courses due to S_2 and S_3 again remain fixed. Because of the change in direction of rotation the angles between

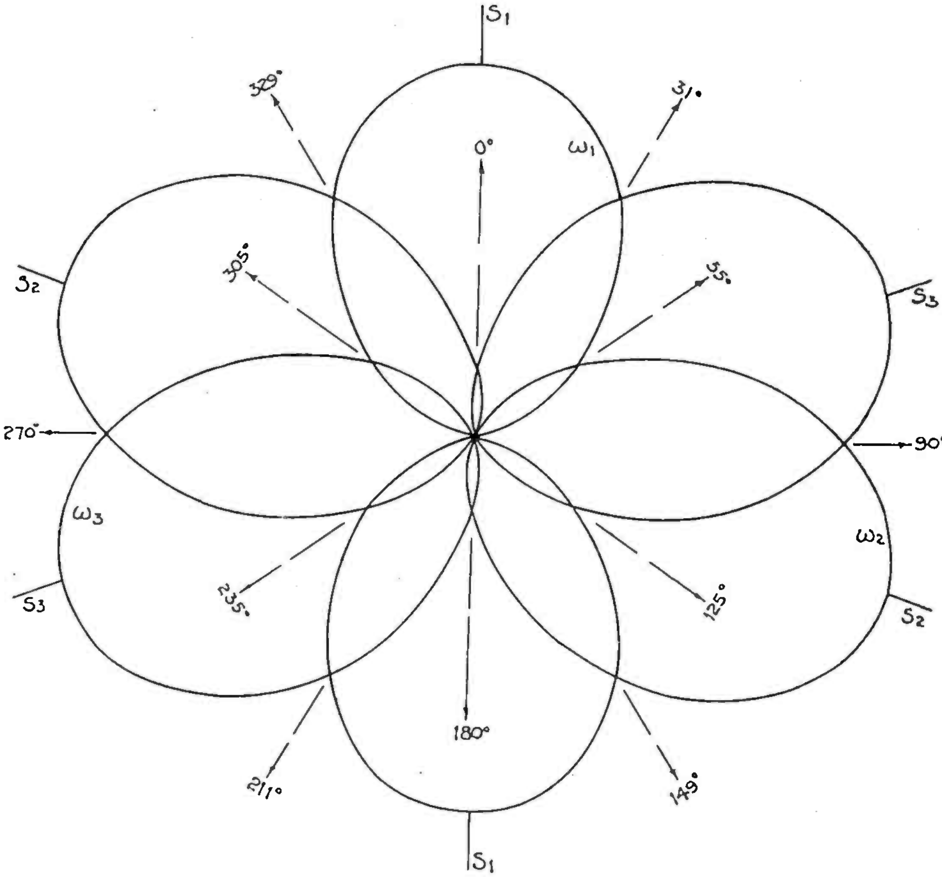

Fig. 22—Stator positions are the same as for Fig. 21 except that radio-frequency rather than audio-frequency switching is employed.

courses are quite different from that obtained in Fig. 20. Note also that the majority of the courses are somewhat more useful in so far as signal strength and sharpness of course is concerned.

Using this method of attack, it becomes possible to align the beacon courses with the airways at a great number of airports. In certain special cases, however, it may become necessary to resort to the other methods of course shifting described in the previous paper.[5]

In the system using radio-frequency switching of power, the fact that displacing the stators from their normal position serves to distort the carrier renders this method of course shifting a little less powerful. Fig. 22 is for the case when S_2 and S_3 are separated from S_1 by 110 deg. Note that the variation of the courses from their normal positions is not exactly the same as for Fig. 21. Enough variation can be obtained with this method, however, to meet the requirements at most airports.

IV. Tests

Numerous test flights have been made on the 12-course beacon described above. The distances obtainable on all 12 courses have been determined. Under average day-time conditions a distance of 100-150 miles on the weak courses, and a correspondingly greater range on the strong courses is obtained. Ground tests were made at Media, Pennsylvania, 105 miles from College Park, Maryland, all 12 courses being oriented through that point by rotating the goniometer at the beacon station, and the angles between the courses thus determined. These tests were repeated at a later date, and the same results were obtained. These results checked also the data on the angles between courses as obtained in the air.

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A TUNED-REED COURSE INDICATOR FOR THE FOUR- AND TWELVE-COURSE AIRCRAFT RADIO RANGE*

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Summary—*For the 12-course radio range system, in which three modulation frequencies are used, a type of reed indicator has been developed to indicate when the aircraft is on any one of the twelve courses, and if off, approximately how many degrees and whether to the right or left. In addition, it indicates to the pilot in case he becomes lost which is his nearest course, how to turn to get on it, and which way he is flying on it. This is accomplished by the use of three reeds in the visual indicator, each reed being tuned to one of the modulation frequencies sent out by the radio range, namely 65 cycles, 86.7 cycles, and 108.3 cycles. Unequal amplitudes of vibration of the reeds indicate the plane is off the course to the side of the reed having the greatest amplitude. A simple shutter with windows, in front of the vibrating reeds, exposes any two at a time. The correct two for a given course is determined by a color system which is exposed by the window to correspond to the color of the particular radio range route marked on the map. A second shutter and color system is provided so that the rule, "longest reed indicates side off course," may be made to hold regardless of the course being flown or the direction of flight.*

The 4-course indicator is the same as the two-course indicator with the exception of a shutter and color system on its face to adapt it to any one of the four courses.

A pilot using these indicators may hold a plane in a given radio-range course with an accuracy of approximately 2 deg.

I. Introduction

THE TUNED-REED type of visual indicator is used to give a pilot a visual indication as to whether or not he is flying on a specified double-modulation radio-range course and, if not, to which side and how much he has deviated. The indication is given continuously by two vibrating reeds, the relative amplitudes of which indicate the position of the airplane with respect to the radio-range course. In order to observe the reed vibration, each reed carries a white tab on its free end. These two tabs produce two adjacent white lines when the reeds vibrate. It is the relative length of these two lines which the pilot observes. Each reed is tuned to one of the frequencies of modulation used at the radio range. The course is a zone in space where the strengths of the radio-range modulation frequencies are equal, each zone being indicated to the pilot by equality of amplitude of vibration of the two reeds. A deviation from the course is indicated by an increase in that reed amplitude on the side to which the airplane has deviated and an equivalent decrease in the other reed amplitude.

* Dewey decimal classification: R526.1.

A tuned-reed indicator designated as type *F* for the double-modulation (2-course) radio range has been previously described.[1] There has recently been developed a 12-course radio-range system[2] in which three modulation frequencies are used and which gives twelve courses about 30 deg. apart. This requires a tuned-reed indicator useful on any one of the twelve courses. This paper describes such an indicator, which when used in conjunction with this radio range, serves to give a pilot the following information: (1) indicates when he is on any of the twelve courses; (2) indicates when off the course and approximately by how many degrees and whether to the right or left; and (3) indicates when he becomes lost, (which is hardly possible when using the radio range), which course he is nearest, how to turn to get to it, and which way he is flying on it; i.e., whether "To" or "From" the radio range.

In this paper there is also described a 4-course indicator which was designed to meet the requirements of the 4-course radio range, which uses two modulation frequencies but produces four courses which may be oriented at will.[3] These requirements, while not as difficult to meet as those of the 12-course radio range, did necessitate, however, the use of new features on the face of the 2-course reed box in order to adapt it to any one of the four courses.

II. The 12-Course Reed Indicator

A. Details of Design

1. Reeds and Driving Elements

This indicator, shown in Figs. 1, 2, 4, 5, and 6, contains three reeds tuned to frequencies of 86.7, 108.3 and 65 cycles, respectively, the three frequencies of modulation used at the radio range. These reeds are made of elinvar which makes their natural period of vibration independent of temperature. Steel reeds may be used, in which case a weighted bimetallic compensation strip should be fastened to their free end. This strip, when bending because of a temperature change, moves the weight on its end a sufficient amount to change the tuning of the reed by an amount which compensates for the change in its tuning due to the effect of temperature on its elastic constant.

[1] F. W. Dunmore, Design of tuned-reed course indicators for aircraft radio-beacon. *Bureau of Standards Journal of Research*, 1, 751–769; November, 1928. Reprinted as Research Paper No. 28.

[2] H. Diamond and F. G. Kear, A 12-course radio range for guiding aircraft with tuned-reed visual indication." *Bureau of Standards Journal of Research*, 4, 351, R. P. No. 154. Proc. I.R.E., this issue, see pp. 939.

[3] H. Diamond "Applying the visual double-modulation type radio range to the airways," *Bureau of Standards Journal of Research*, 4, 265, R. P. No. 245; February, 1930.

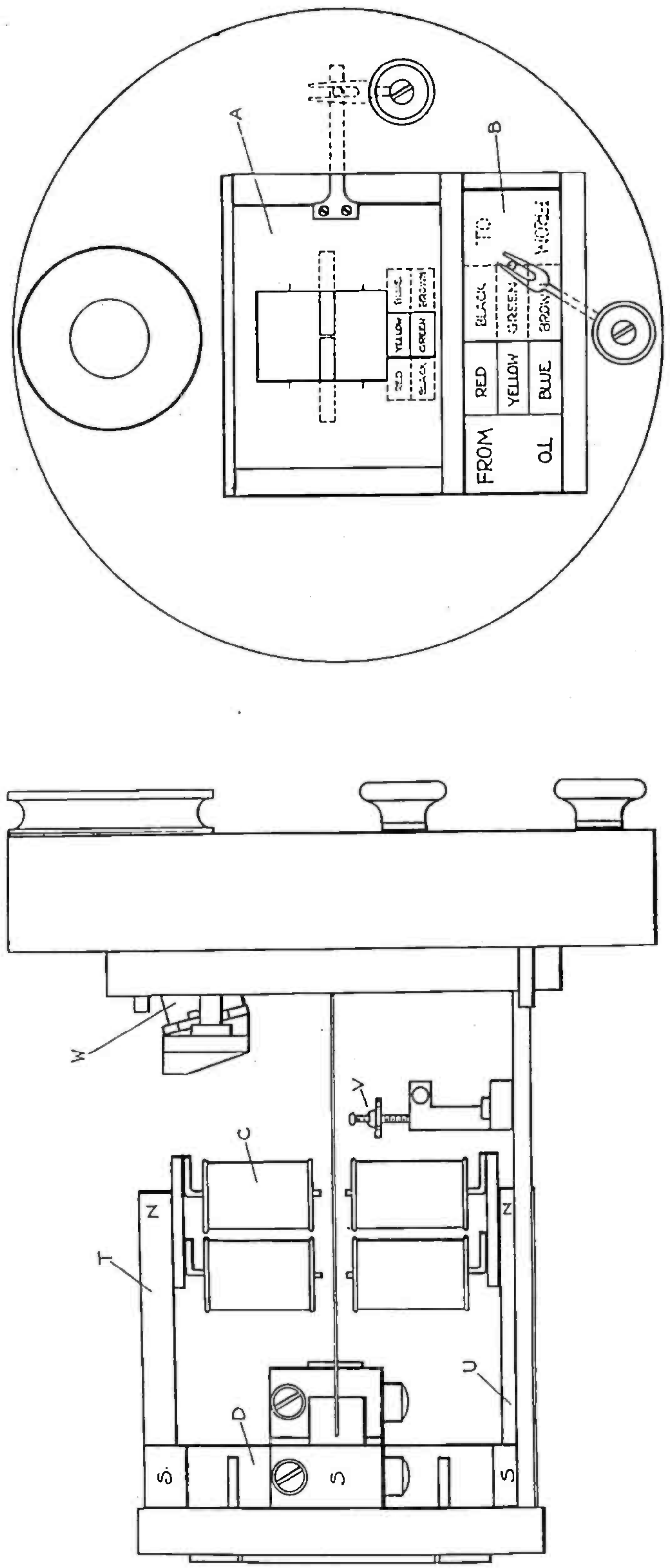

Fig. 1—The tuned-reed course indicator for the 12-course radio range.

Each reed is polarized by a set of permanent magnets, T and U, Fig. 1, common to all reeds. D is a soft iron yoke connecting two like poles of the magnets. Each reed has a separate set of driving electromagnetics, C, similar to those used in telephone receivers, the windings of which are all connected in series in the proper polarity to operate the polarized reeds. The terminals of these electromagnets are connected to the output of the radio-range receiving set. The housing, W, contains a lamp for illuminating the reeds; V is a bumper to hold the reed vibration within bounds; A and B are shutters over the reeds and color system. Fig. 2 is a plan view of a portion of Fig. 1.

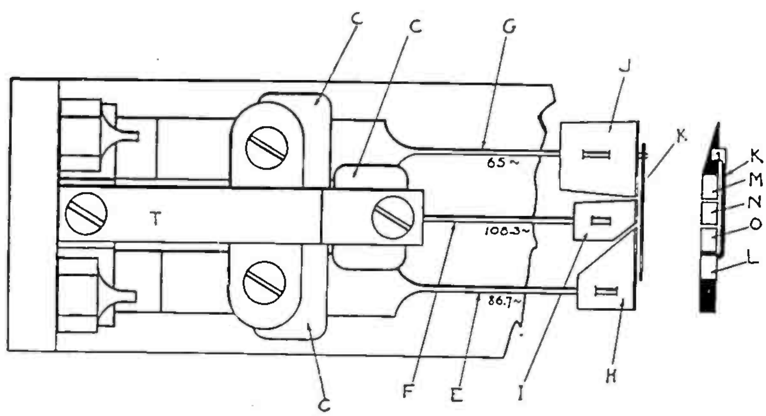

Fig. 2—Plan view of 12-course indicator showing reeds, dampers, extension arm on 65-cycle reed, and whitened tabs.

For each course the vibration of adjacent reeds must be observed. The 108.3-cycle reed is placed between the 86.7- and 65-cycle reeds. For one set of courses, therefore, in order to observe the 86.7-cycle reed adjacent to the 65-cycle reed, a light arm extension, K, Fig. 2, is fastened to the free end of the 65-cycle reed. This arm carries a white tab, L, on the far end, which tab vibrates adjacent to the tab, O, on the upturned front end of the damper, H, on the 86.7-cycle reed. The other two air dampers, I and J, also have upturned whitened ends or tabs, as shown at N and M. L, O, N, M is an end view of the reeds or tabs as the pilot would see them with the shutter removed. There are three sets of reed combinations which go to make up the twelve-course indications. For one set of courses, tabs M and N are observed; for another, tabs N and O; for a third, tabs O and L. As previously stated, tabs L and M both vibrate with the 65-cycle reed, G.

2. Shutter and Color System

Since it is necessary to observe any two adjacent whitened tabs on the reeds for a given course, without seeing the others, a shutter, A, Figs. 1 and 3, with a window, is provided. This window may be moved to expose any two adjacent tabs depending on what radio-range course is to be flown. This same window also exposes two different colors at each setting in order to facilitate the choice of the proper two reeds for a given course.

Another shutter, *B*, is provided with a color system to simplify the operation of the indicator in connection with its use when flying "To" or "From" the radio range. The use of both of these shutters will be explained in more detail under "Application of reed indicator to the 12-course radio range."

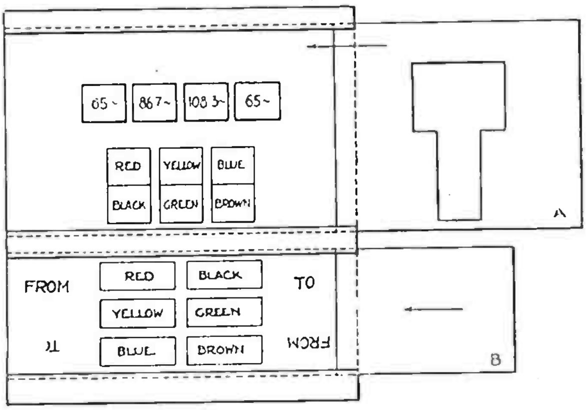

Fig. 3—Face of 12-course reed indicator showing shutters removed to expose the color system and whitened tabs attached to the reeds.

3. Cylindrical Type of Indicator and Shockproof Mounting

Since it is not necessary to plug in different indicators for different radio-range courses with the 12-course indicator, the one indicator serving all courses, this instrument and mounting may be made more in keeping with the rest of the aircraft instruments. A cylindrical shape for the indicator and mounting, as shown in Figs. 4, 5, and 6, may therefore be adopted. The reed unit shown in Fig. 4 is designed to rotate within the inner cylinder shown in Fig. 5, this cylinder being held within an outer cylinder by means of eight springs. This outer cylinder is fastened to the instrument board. The spring mounting is necessary to prevent the mechanical vibration from the airplane from operating the reeds at certain engine speeds. Slip rings shown in Fig. 4 on the rear of the indicator and brushes on the rear of the inner cylinder serve to carry the current for operating the reeds and the light for illuminating them.

To show the words "To" and "From" on the indicator the proper side up, the reed unit shown in Fig. 4 is turned through 108 deg. by revolving it in its mounting. A covering with a glass window is placed over the reeds, shutter, and color system, to protect them from dirt and rain. Two knobs extending through this cover provide means for operating the shutters. A front view of the instrument, as seen by the pilot, is shown in Fig. 6. The indicator and mounting shown in Figs. 4 and 5 weighs about $1\frac{3}{4}$ pounds. It is $3\frac{1}{2}$ inches in diameter and 5 inches long.

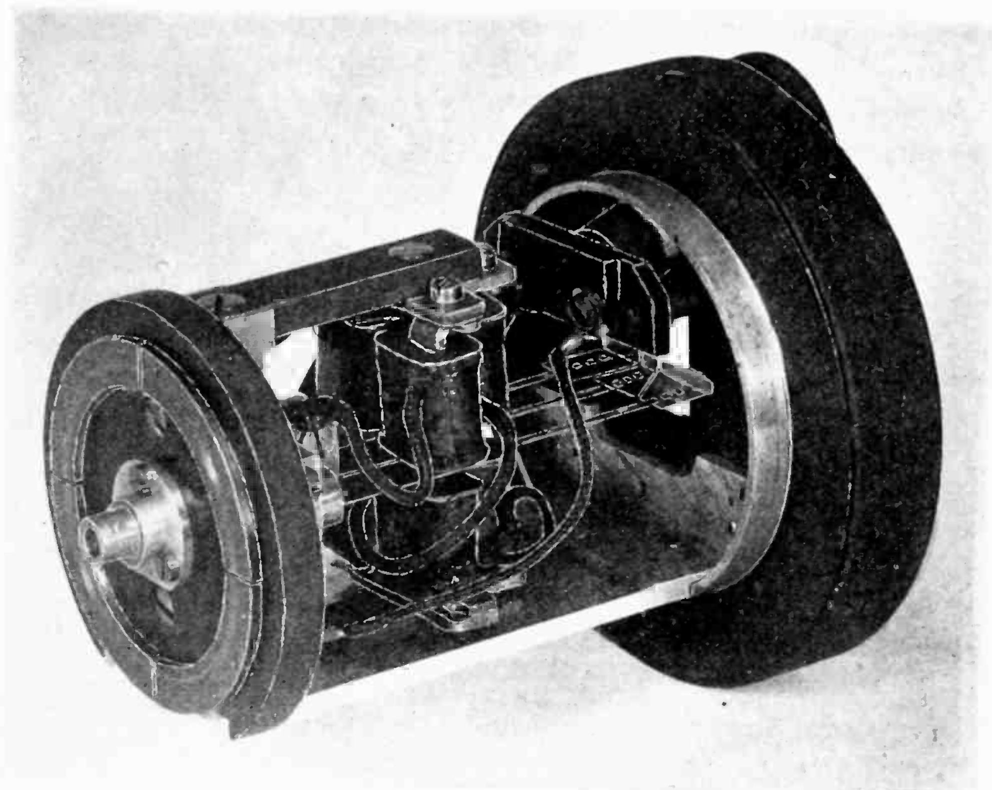

Fig. 4—Rear inside view of tuned-reed 12-course indicator shown in Fig. 6.

B. Operating Characteristics

1. Sensitivity

The three reeds are adjusted to be equisensitive by changing the air gap between the electromagnet pole pieces for each reed. Fig. 7 shows the reed deflections in millimeters, as seen by an observer, plotted against the voltage applied to the terminals of the reed indicator. At the amplitudes of vibration normally used; that is, 4 to 9 mm, it will be noted that essentially a straight line relation exists between the deflection and applied voltage, which is quite necessary in order to prevent any apparent shift in course with adjustment of volume control on the radio receiver operating the reeds. At the normal deflection of 8 mm the current in the driving coils is 1.4 ma. This sensitivity has been obtained by means of a switch operated by shutter *A*, which short circuits the two driving coils for whichever reed is not in use. Additional sensitivity may be obtained over that shown in Fig. 7 by the use of a large one-piece permanent magnet.

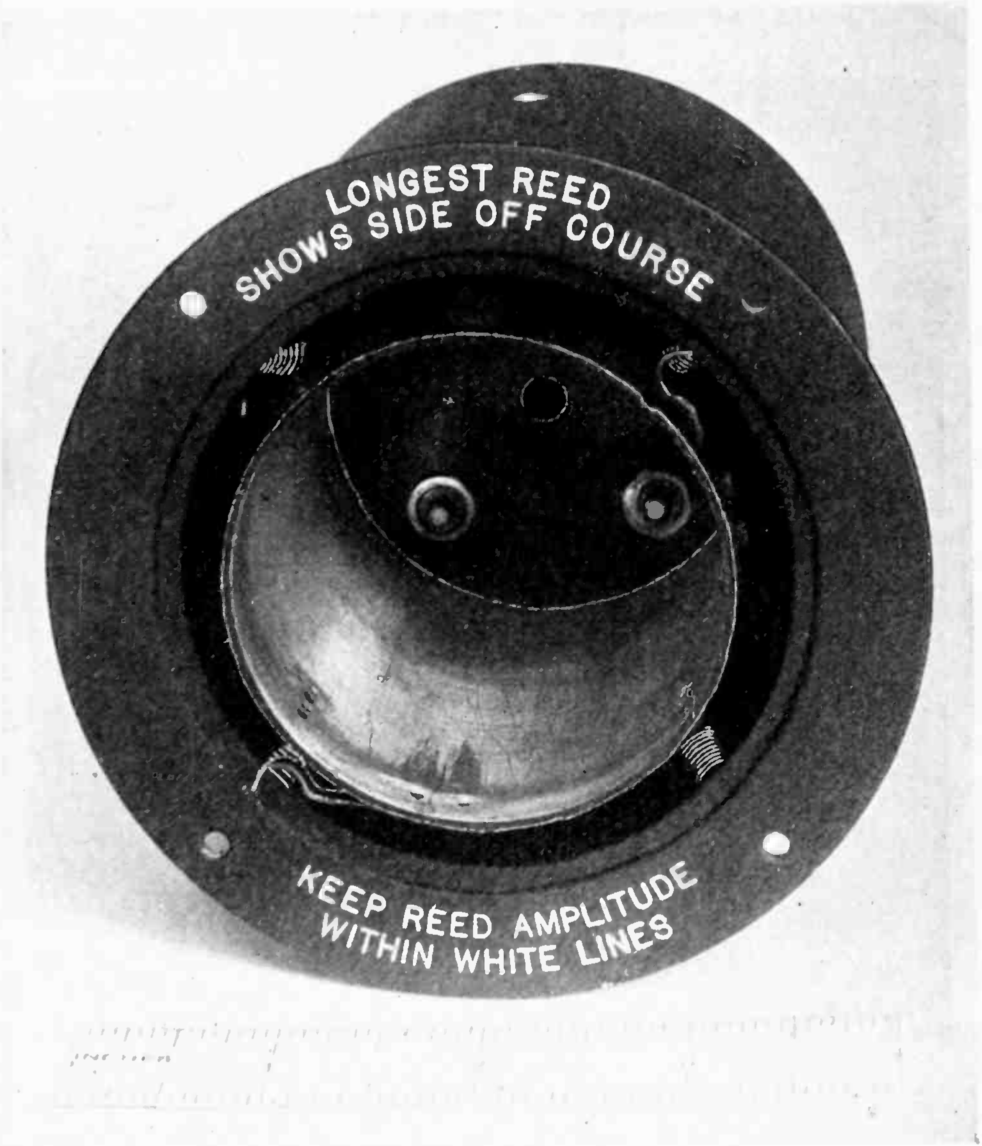

Fig. 5—Cylindrical type of shockproof mounting for 12-course tuned reed indicator shown in Fig. 6

2. Selectivity

As in the case of the 2-course indicator each reed in the 12-course indicator is insensitive to any frequency other than its natural frequency. This is a very valuable feature, since it practically eliminates the effects of interfering signals unless these signals are severe enough to block the tubes in the receiving set, or unless they are very near the same frequency to which the reeds are tuned. This interference may come from many sources, such as engine ignition and atmospheric disturbances, marine beacon signals and radio-range signals of the aural type. In many cases it was found that where radiotelephone sig-

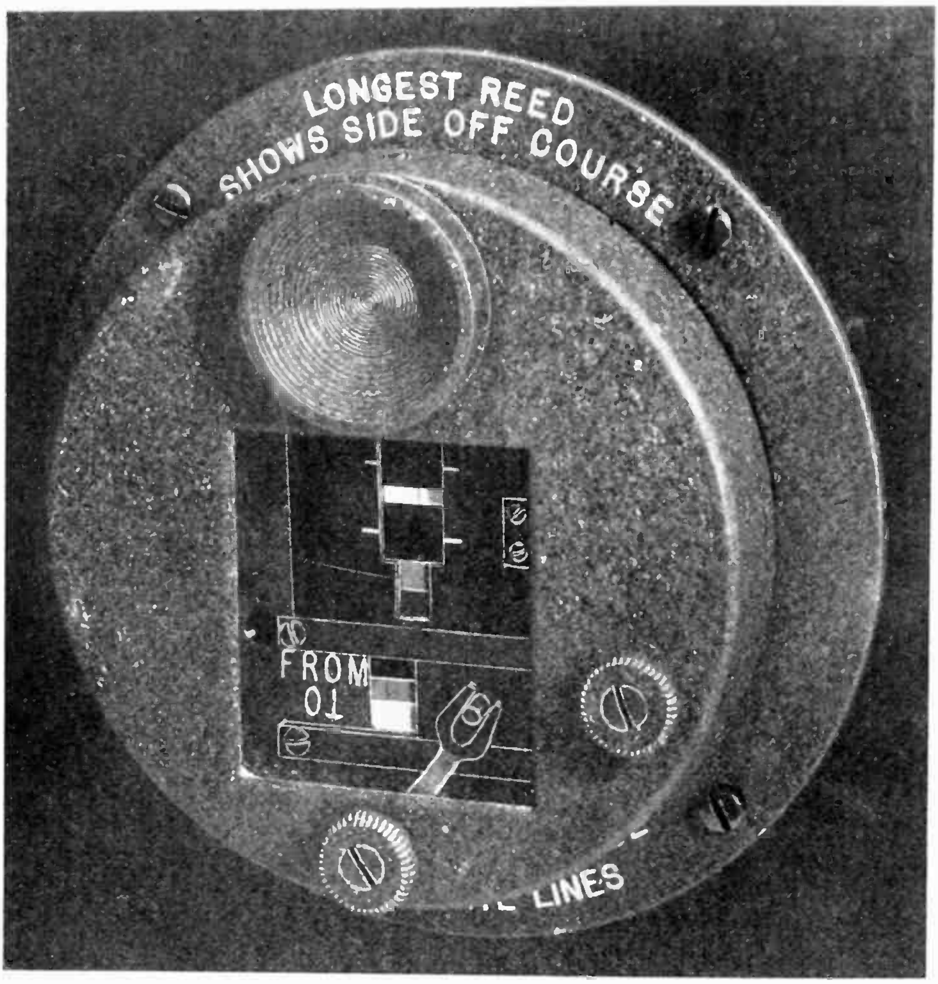

Fig. 6—Latest type of tuned-reed course indicator for the 12-course aircraft radio range.

nals were coming in stronger than the aural radio-range signals the latter were entirely unintelligible on account of interference while under the same conditions the reeds functioned satisfactorily.

3. Effect of Damping the Reeds

Light aluminum air dampers, *J*, *H*, *I*, Fig. 2, are placed on the end of each reed in order to broaden the tuning, to prevent any appreciable change in reed amplitude should the modulation frequency shift by as much as 0.3 per cent. The damping is so proportioned that the relative reed amplitude will not change appreciably even though the frequency varies by as much as 0.5 per cent. The resonance curves for

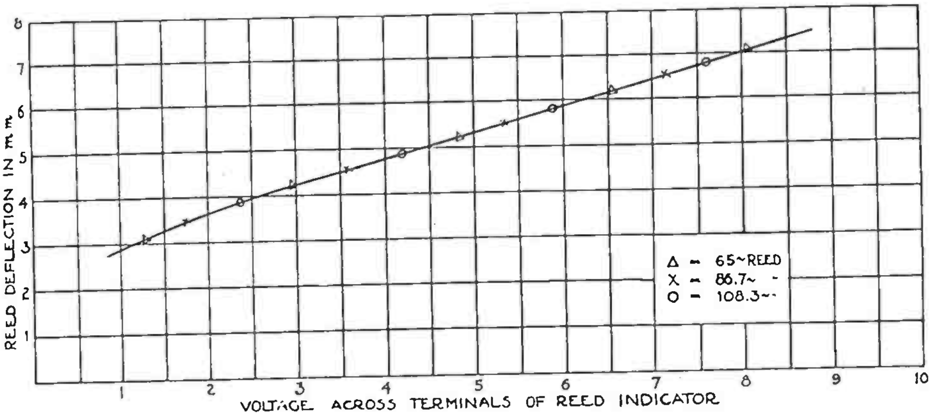

Fig. 7—Voltage required to operate the reeds on the 12-course indicator.

the three reeds are shown in Fig. 8. Since the three frequencies of modulation at the radio range are obtained from three generators with 6, 8, and 10 poles with shafts directly connected, the three frequencies must vary in this fixed ratio, and the resonance curves are propor-

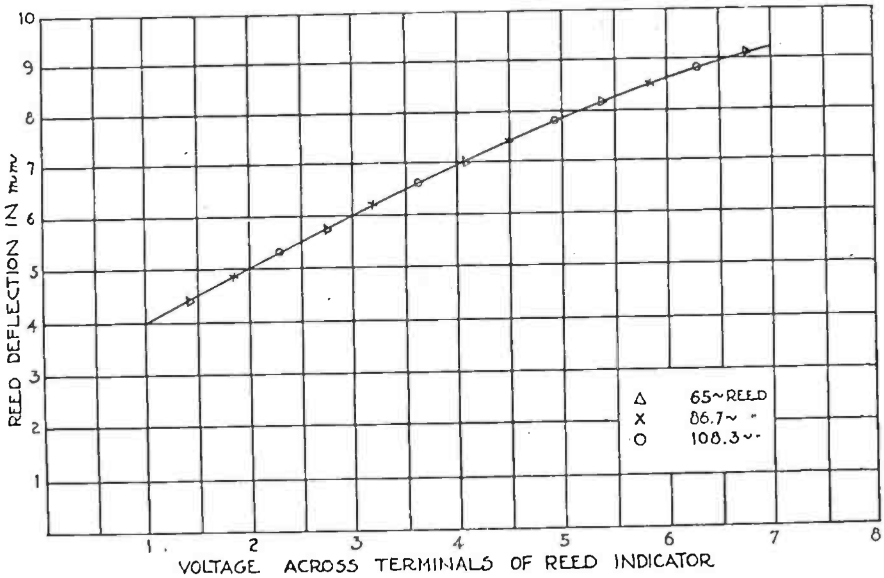

Fig. 8—Resonance curves for the three reeds in the 12-course indicator, showing effect of correctly proportioning the damping to keep reed amplitudes the same relatively as frequency changes.

tioned, i.e., so that if, for example, a 0.3-cycle variation occurs in the 65-cycle frequency, a 0.4-cycle variation will occur in the 86.7-cycle frequency, and a 0.5-cycle variation will occur in the 108.3-cycle frequency. From the curves in Fig. 8 it will be seen that for such a variation in each frequency the reeds will all drop in amplitude approximately the same amount, i.e., 1.8 mm. Since the relative amplitudes of the reeds do not change, an apparent shift in the course therefore is not obtained.

Since the data for the curves shown in Fig. 8 were obtained, means have become available for holding the modulation frequencies to the correct values with greater accuracy, so that it is possible to use less damping on the reeds. This not only increases their sensitivity but also their selectivity, making them even less subject to interfering signals of frequencies near that to which they are tuned.

C. Application of the Reed Indicator to the Twelve Courses

One of the features which simplifies the use of the 2-course reed indicator is the one simple rule which the pilot must remember, i.e.,

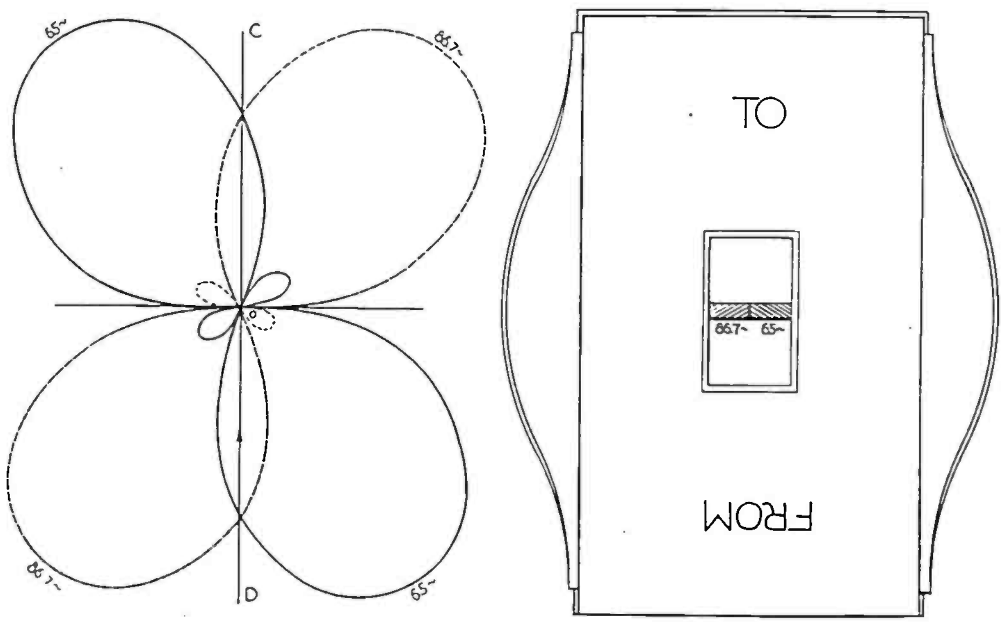

Fig. 9—Radio-range transmission characteristic and face of reed indicator showing necessity for turning the unit upside down when reversing the direction of flight.

"longest reed shows side off course." For example, if the right-hand reed vibrates with greater amplitude than the left-hand reed, the plane has drifted off the course to the right. In order that this rule will hold true regardless of the direction of flight, the reed box is used as follows:

Referring to Fig. 9, which shows a typical radiation characteristic of the double-modulation radio range and also the front of the 2-course reed box, when the pilot is on the course flying in a certain direction, say *toward* the radio range located at O, along the line DO, the zone of greatest 65-cycle modulation is on his right and the zone of greatest 86.7-cycle modulation on his left. When drifting off the course to the right, therefore, the 65-cycle reed would vibrate with greater amplitude. This reed should, therefore, be on the pilot's right since the one rule should hold, "longest reed shows side off course." The words "To" and "From" are so engraved on the face of the reed box as shown that when the word "To" is right side up, the 65-cycle reed is on the pilot's right. Should the pilot make a 180-deg. turn and fly *from* the radio range, the location of the zones of greatest 65- and 86.7-cycle modulation reverses with respect to his right and left. This is also true if he passes over the radio range and flies from it along the line OC. It is therefore necessary to turn the reed box upside down, i.e., so the "From" is right side up. This revises the reed location and places the 86.7-cycle reed on his right, in accordance with the reversal of the zones of modulation with respect to the pilot's right and left.

With the 12-course indicator the problem of maintaining this simple rule becomes more difficult, as will be seen by Fig. 10, which shows the distribution of the modulation frequencies used at the radio range for the different courses. The three figures-of-eight show the radiation characteristics of the 12-course radio range for each of the three frequencies of modulation. The colors indicate the courses or zones where two of the frequencies of modulation are present in equal amounts. This color combination was chosen to match the color system on the face of the reed box in order to simplify the operation of the reed indicator, as shown in Fig. 3, where two colors appear for each setting of the window A, and three colors for each setting of window B. The pilot's map has the radio-range courses in color so if he wishes to fly on a red radio-range course, as shown by the map, the shutter A on the face of the reed indicator is set to show red through part of the window. This exposes the 65- and 86.7-cycle reeds which, from Fig. 10, are the two frequencies of modulation used on the red radio-range course. A black course could also be flown with this same shutter setting. It will be noted when flying on a black course *from* the radio range (it being located at the intersection of all lines) that the 86.7-cycle signal is on the pilot's left while it is on his right when flying *from* the radio range on the red airway. This reversal is true of all the 90-deg. courses and upsets the fundamental rule for using the reed indicator, i.e., "longest reed indicates side off course." To overcome this, a second shutter, B, and color scheme, Fig. 3, are provided on the face of the indicator. This shutter reverses the "To" and "From," as shown, to compensate for the reversal of the location of the frequencies of modulation with respect to the pilot's right and left on the 90-deg. courses. The shutter system operates as follows: the pilot observes on his map the color of the radio-range airway course he desires to fly and which way he desires to fly on it, i.e., whether "To" or "From" the radio range. If he chooses a red course, he sets both shutter A and B to show red. This exposes the 65- and 86.7-cycle reeds,

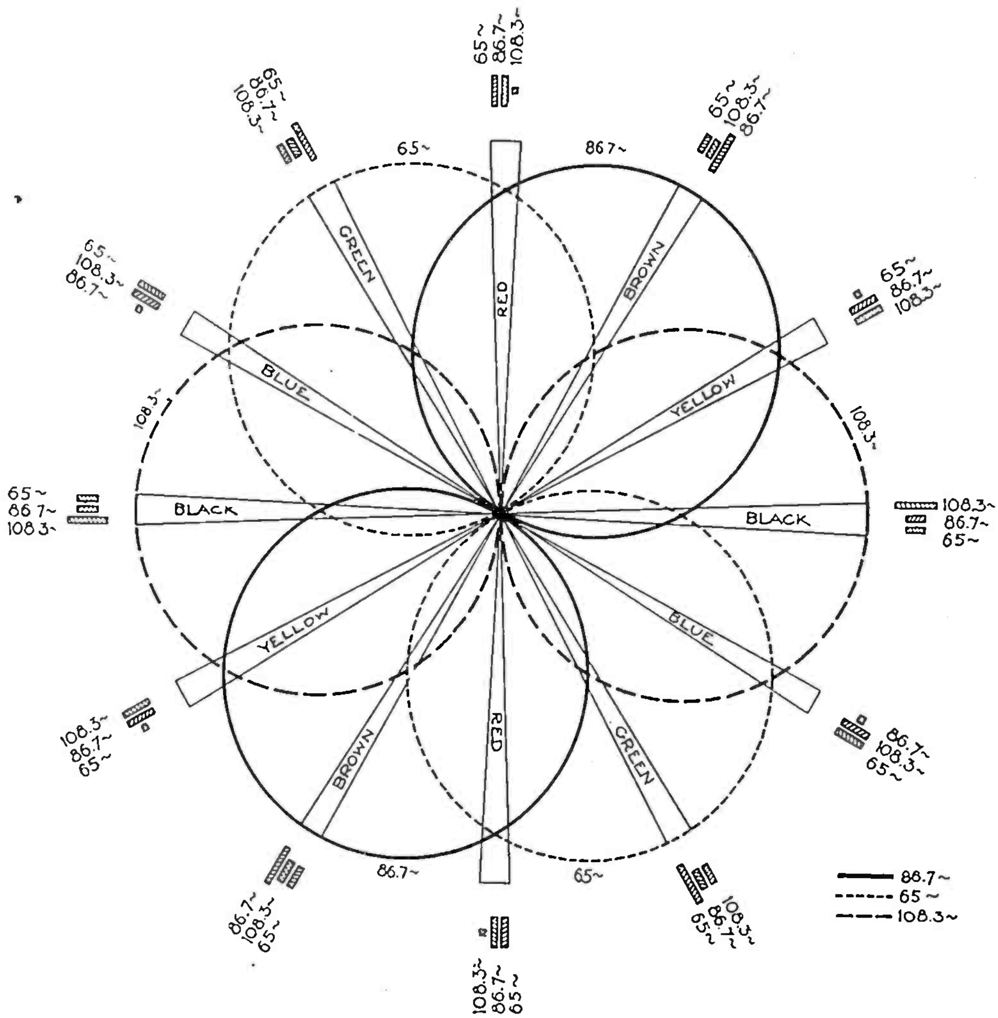

Fig. 10—The 12-course radio-range transmission characteristic. The color indicates the courses where two modulation frequencies are of equal strength and reeds vibrate with equal amplitudes.

which are the correct ones for the red course, as shown on Fig. 10. The lower shutter exposes the words "From" and "To," one of which is upside down. If he desires to fly "From" the radio-range, he rotates the indicator unit in its mounting so the word "From" is right side up (when the red is exposed by both shutters). This puts the 86.7-cycle reed on his right and the 65-cycle reed on his left. From a glance at Fig.

10, on the red course, it will be noted that the 86.7-cycle modulation is on the pilot's right when flying "From" the radio range, the 65-cycle on his left, so the rule will hold, since if he turns to the right the 86.7-cycle signal will become stronger and the 65-cycle signal weaker. Therefore the 86.7-cycle reed indication will appear longer, and the 65-cycle reed indication will appear shorter. A similar test may be made, using Figs. 3 and 10, on any one of the twelve courses, and it will always be found, if the reed box is rotated to have the correct side up, that the longest reed will always indicate the side off course.

The pilot's instructions for operating the indicator may be condensed to the following:

(1) Set both shutters to show the color, according to the map, of the airway to be flown.
(2) Turn reed box to show "From" or "To" right side up, depending upon whether the desired direction of flight is *from* or *to* the radio range.
(3) Longest reed indicates side off course.

A further application of this type of reed indicator is its use by a pilot when lost in fog, to guide him in the right direction to the nearest radio range. There are many instances when a pilot navigating by magnetic compass in fog without radio-range facilities has been completely lost. With the 12-course reed indicator used in conjunction with the 12-course type of radio range, a pilot should have no occasion to become lost, but if he should, it is a rather simple matter for him to "find himself," i.e., he is able to get on a radio-range course and determine definitely which way he is flying along that course.

This feature is made possible by the fact, as will be seen from Fig. 10, that the courses alternate in their relative signal strength, i.e., there are six courses of given signal strength and six more between these of 58 per cent of the signal strength. The amplitudes of vibration of the three reeds for each course are shown opposite each course in Fig. 10. A pilot, therefore, if lost, may make use of his third reed to determine what course he is on, in the following manner. First, he moves shutter *A*, Fig. 3, and finds the two adjacent reeds which are nearest equal, and navigates until they are equal. This places the airplane on one of four courses, say, either of the two red or black courses, since from Fig. 10 it will be seen that a given course, its 180 deg. course, and the two 90 deg. courses, have the same modulation frequencies, which would cause the same two reeds to vibrate. Two of the courses may be eliminated by observing the third reed, i.e., the reed adjacent to the two equal reeds. If this reed is vibrating with greater amplitude than the two equal reeds, then, as seen from Fig. 10, the airplane is on one of the black courses, since the 108.3-cycle signal is nearly twice the 86.7- and 65-cycle signal operating the two equal reeds. Should the airplane have been on a red course, the third reed would have had zero amplitude. Having determined that the airplane is on one of the black courses, the shutter *B*, Fig. 3, is set to show black. There still remains the ambiguity as to which of the black courses the airplane is on, and as to the direction of flight. Assume the black courses extend in a north and south direction as shown in Fig. 11. The airplane is flown by means of the magnet compass in one of these directions, say, north,

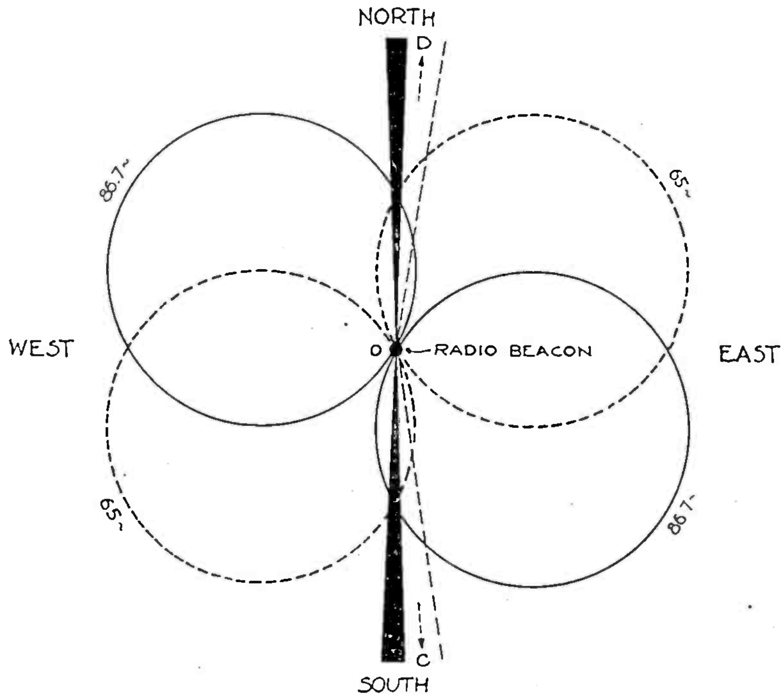

Fig. 11—Chart showing method of determining the direction of flight relative to the radio range when such direction is unknown.

and flown off course to the right. The reed box is turned in its mounting so that the right-hand reed is longest when off course to the right. If the word "From" exposed by shutter *B* is right side up, the airplane is flying north *from* the radio range on the northern black route *OD*. It cannot fly north *to* the radio range along this line. If the word "To" exposed by shutter *B* is right side up the airplane is flying north *to* the radio range on the southern black route *OC*. It cannot fly north *from* the radio range along this line. Thus a pilot may definitely establish his location with respect to the radio range. The above system of procedure may be condensed into a few simple rules for the pilot to follow without any technical knowledge on his part of the radio-range system. These rules, which a pilot should seldom find necessary to use, are as follows:

(1) Move shutter *A* to show the two reeds of nearest equal amplitude and navigate airplane until they are equal.
(2) Note amplitude of reed adjacent to the two equal reeds.
(3) If amplitude of this reed is greater than that of the two equal reeds, set shutter *B* to show black, green, and brown; if less, set it to show red, yellow, and blue.
(4) Then the common color exposed by both shutters is the course being flown.
(5) Note the directions of this course on the map and fly according to the magnetic compass in one of these directions, deviating to the right until the equal reeds become unequal.
(6) Turn reed box so that the longest reed is on the right; then, whichever of the words "To" or "From" is right side up indicates the general direction of flight relative to the radio range, and the magnetic compass indicates the absolute direction.

A single 12-course indicator may be used on any number of 12-course radio ranges since neighboring radio ranges operate on the same modulation frequencies but on a slightly different carrier frequency. A change in tuning of the radio receiver, therefore, is all that is necessary to cause the indicator to operate from signals from another radio range. This tuning should be done when the plane reaches a point approximately midway between the two radio ranges being used. The courses of two neighboring radio ranges are oriented where possible so that courses with the same modulation frequencies will be in a straight line. In this case the reed box need only be turned upside down to show "To" instead of "From" at the midpoint between the two radio ranges when the radio receiver is tuned to the radio range being approached.

The 12-course reed indicator gives a continuous indication to the pilot as to the position of his airplane with respect to the radio-range course. This feature is of great advantage when used to guide an object moving as fast as a modern airplane. This is especially true when approaching a radio range located on a landing field. As the airplane nears the radio range, any slight movement of the airplane from one side to the other is immediately noticed with only a glance at the reeds. In fact, when over the field the indication is sharp enough so that a pilot is able to keep the airplane within the width of the average runway if the course is oriented down the center of it.

III. The 4-Course Reed Indicator

A. Details of Design

1. Reeds and Driving Elements

These features of the 4-course indicator are identical to those of the type *F*, 2-course indicator referred to in the introduction.

2. Shutter and Color System

It is the addition of a shutter with a color system that changes the 2-course indicator to the 4-course indicator. This shutter is the same

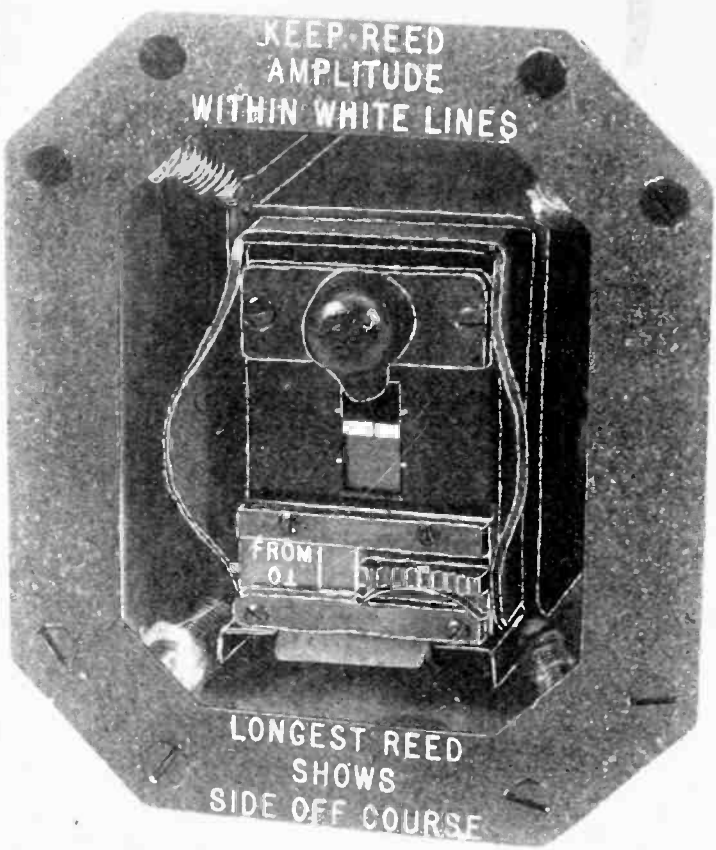

Fig. 12—The 4-course tuned-reed indicator and shockproof mounting.

as shutter *B*, Figs. 1 and 3, except only two colors are used, one for each position of the shutter. Shutter *A* and its color system are not used, as there are but two reeds in this indicator. Fig. 12 shows indicator with mounting.

3. Shockproof Mounting

Since it is necessary to plug in a different indicator of this type if radio-range courses using different modulation frequencies are to be used, the shockproof mounting is made the same as that used with the type *F*, 2-course indicator.

B. Application of the Reed Indicator to the 4-Course Radio Range

A typical radiation characteristic for the 4-course radio range and the face of the 4-course reed box with shutter *B*, are shown in Fig. 13. The purpose of this shutter is the same as the similar shutter used on the 12-course reed box, i.e., it reverses the "To" and "From" so the rule, "longest reed shows side off course," will hold, regardless of the course being flown and direction of flight. In other words, it compensates for the reversal of the location of the frequencies of modulation with respect to the pilot's right and left on the 90 deg. courses.

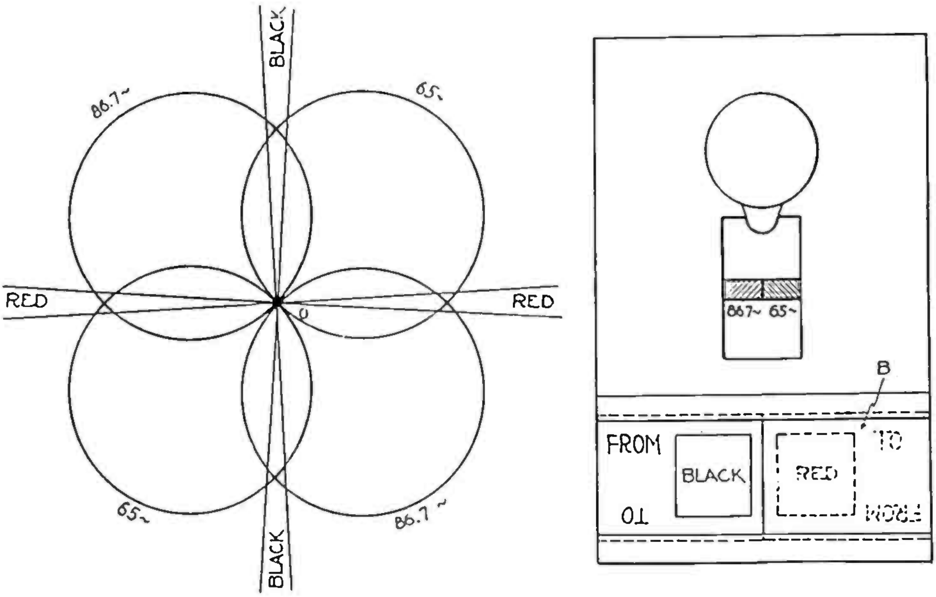

Fig. 13—The 4-course radio-range transmission characteristic and face of reed indicator for use on any one of the four courses.

To adjust the reed box for use on a given course the pilot merely sets the shutter to show the color of the radio-range course he is to fly as shown on his map, and plugs the indicator into its holder the proper side up to show the "To" or "From" right side up, depending upon whether he is flying "To" or "From" the radio range. In Fig. 13 the shutter is set for a black airway, and the reed box is in a position for flying *from* the radio range.

C. Application of 4-Course Reed Box to the 12-Course Radio Range

Aside from its use with the 4-course radio range, the 4-course reed indicator may be used with the 12-course radio range. For example, the 4-course indicator described above with 65- and 86.7-cycle reeds, may be used on the two black and two red courses of the 12-course radio range. With the two reeds in the 4-course indicator tuned to 65 and 108.3 cycles and a brown and blue color scheme used on its face, this indicator may be used on the two brown and two blue courses of the 12-course radio range. When the reeds are tuned to 108.3 and 86.7 cycles and the colors green and yellow used, the remaining courses of the 12-course radio range may be utilized.

Thus, a pilot chooses the reed box having the same color as the radio-range course to be flown and uses it for flights on those courses.

In some instances, therefore, when an airplane flying on a 12-course radio range is used on a fixed route, as is often the case with mail airplanes, a 4-course reed box may be used in place of the 12-course indicator.

The 4-course indicator plugs into its mounting so that another may be quickly substituted for use on another set of four courses. Thus three 4-course reed boxes may be used in place of one 12-course reed box in case of necessity.

With a 4-course indicator with reed-driving coils shunted by a potentiometer which changes the relative sensitivity of the two reeds, a single radio-range course may be made effective over an angle of 30 deg. or more, i.e., a course may be flown with equal reed deflections along any line making an angle of up to 15 deg. on either side of the true course.

In this way of using three 4-course indicators with shunting potentiometer, the 12-course radio range becomes effective over practically the full 360 deg., i.e., a course may be held with equal reed deflections at any angle of flight towards or away from the radio range.

A 10,000-ohm potentiometer is connected to the reed-driving coils, U, V, W, and X, of the 4-course indicator, as shown in Fig. 14. A front view of this course deviometer with uncalibrated scales is shown in Fig. 15.

As a direction of movement of the sliding contact on the potentiometer reverses for the 90-deg. courses, a color system with a double pointer is again used. In this way the pointer which is over the correct color scale is moved to the right or left, depending upon which side of the course the pilot desires to fly. The scales may be calibrated approximately in degrees deviation from the course. When calibrated for the 12-course radio range the lower scale is calibrated for the black, brown, and green courses, while the upper scale, is calibrated for the yellow, blue, and red courses. (See Fig. 10).

The deviometer may be used with the 12-course indicator, in which case shutter A, Fig. 3, may be made to operate a second switch, connecting the moving contact of the deviometer to the center connection between the two sets of driving coils in circuit for the particular setting of shutter A.

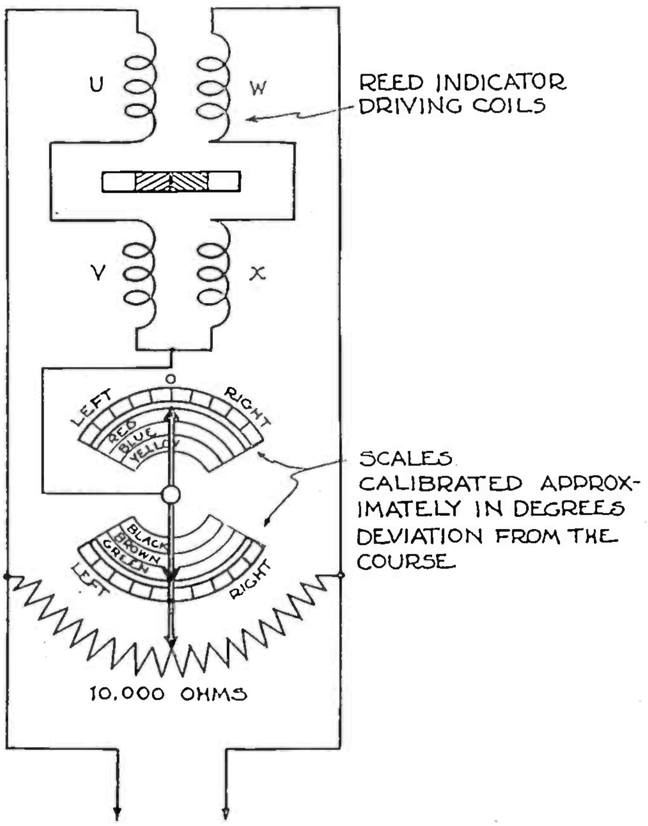

Fig. 14—Circuit diagram for the deviometer or course-shifting device for use with the two- and four-course tuned-reed indicators.

IV. Conclusion

The 12-course reed indicator described herein contains three reeds tuned to the three frequencies of modulation used in a 12-course radio range. It has been so designed as to permit the guiding of an airplane along any one of the twelve courses without confusion.

The relative amplitude of vibration of any two adjacent reeds indicates continuously the position of the airplane with respect to a given course. Equal amplitudes of vibration indicate that the airplane is on the course. Unequal amplitudes of vibration of the reeds indicate that the airplane is off the course to the side of the reed having the greater amplitude. A simple shutter with windows, in front of the vibrating reeds, exposes any two at a time. The correct two for a given course is determined by a color system which is exposed by the window to correspond to the color of the particular radio-range route marked on the map. A second shutter and color system is provided so that the rule, "longest reed indicates side off course," may be made to hold regardless of the course being flown or the direction of flight.

The 4-course indicator is the same as the 2-course indicator with the exception of a shutter and color system on its face to adapt it to any one of the four courses. The 4-course indicator may be used on the 12-course radio range, three such indicators being necessary to cover all twelve courses, each indicator having reeds tuned to match the frequency of modulation of the different courses.

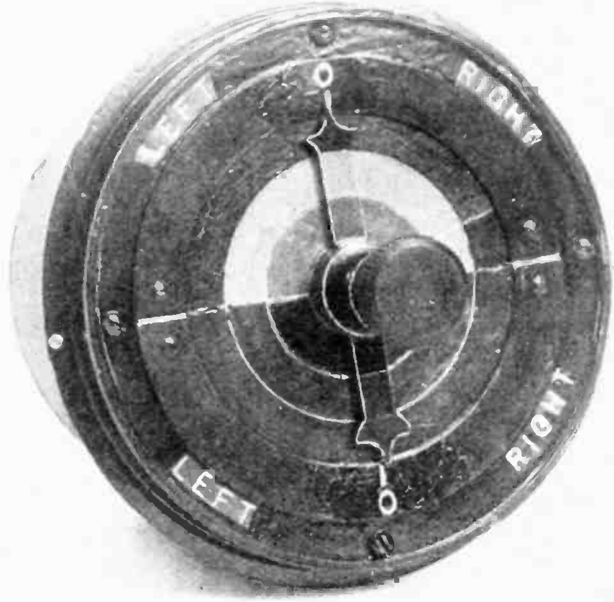

Fig. 15—The deviometer, or course-shifting device, used with the reed indicator to enable the pilot to fly off the course with equal reed deflections.

A pilot using these indicators may hold an airplane in a given radio-range course with an accuracy of approximately 2 deg. By changing the relative sensitivity of the reeds with shunt resistances a given radio-range course may be made effective over an angle of 30 deg., thus greatly increasing the service area of the radio range.

The 12-course indicator and mounting weighs about $1\frac{3}{4}$ pounds and is $3\frac{1}{2}$ inches in diameter and 5 inches long. The 4-course indicator and mounting weighs about $1\frac{1}{2}$ pounds and is $3\frac{1}{2} \times 2\frac{3}{4} \times 4\frac{1}{2}$ inches.

The author is indebted to H. Diamond for helpful suggestions in connection with the design of these indicators, and to R. R. Gessford for constructing the indicator models and for suggestions pertaining to their mechanical design.

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SINGLE- AND COUPLED-CIRCUIT SYSTEMS*

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Summary—*This article discusses networks based upon singly-tuned circuits, and upon coupled circuits with primary and secondary both resonant to the same frequency. Transmission equations are developed and it is shown that for a desired possible transmission-curve shape, the sum of all decrement coefficients must be a certain amount, readily computable for coupled circuits as well as for single circuits.*

Coupled circuit transmission-curve shapes may be developed from single-circuit curves by a multiplication process as in staggered-cascade amplification, or by a vector difference process, employing two staggered single circuits with opposite couplings from a power source. A special case of the vector difference method is the coupled circuit itself, with primary current the vector sum and secondary current the vector difference of two single-circuit currents. This property permits a suitable coupled system to be used for radiating energy of two closely adjacent channels from a single antenna without cross reactions on the power sources.

Complex networks are handled by transfer equations by which a branch consisting of a voltage source and resistor in series coupled to a network by a transformer device is replaced by an equivalent voltage and impedance within the network. Application is made to computation of interstage amplifying transformers, and of single- and two-circuit filters with resistance loading. A brief treatment is given of the impedance and power-factor loading of generator circuits which is of especial importance when tuned networks are output devices of power tubes operating at high plate efficiency.

I. Scope of the Article

THE PURPOSE of this article is to present some of the important relations in single-circuit systems, in coupled-circuit systems, and between single- and coupled-circuit systems, with attention to use of such systems in intertube couplings as well as in circuits for which the output is power consuming.

To simplify the treatment, attention at first will be directed to basic arrangements in which the driving voltage and the power-consuming devices are in series relationships with the tuned-circuit elements. Later, consideration will be given to more complex arrangements with the tuned systems fed from amplifying or power tubes, and loaded by resistance either directly or inductively coupled to one of the circuit reactance elements.

In the treatment here given details of mathematical processes

* Dewey decimal classification: R140.

which are matters of algebraic manipulation will be minimized. Certain statements will be made and justified on physical bases, without giving the details of mathematical and experimental checks which have been made. Further, because of the complexity into which studies of this kind could extend if carried out in full detail, restrictions will be made in order to handle cases of especial interest in a manner correct to the first degree of approximation. In some cases information will be given for determining the magnitude of second-order effects. More attention will be paid to the treatment for single circuits in order to show the processes involved with a fairly simple arrangement and similar processes for coupled circuits will be presented in abstract only.

II. Curve Shapes of Transmission

The chief interest in tuned circuits is in the production of selectivity, and the most important study is perhaps the general nature of transmission curves for single and coupled circuits of the simplest types. In actual tube amplifier and in filtering circuits with power load the transmission shapes will be substantially the same as in the elementary basic arrangements which have scarcely more than a mathematical existence.

A. Single-Circuit Arrangement

In contrast with coupled-circuit systems, for single-circuit study there is but one basic arrangement, as shown in Fig. 1, with the im-

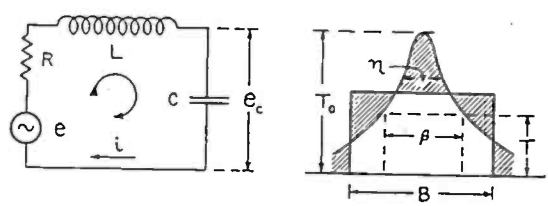

Fig. 1—Single circuit.

pressed voltage e, resistance R, inductance L, and capacitance C, all carrying the same current i. Interest centers upon the voltage e_c established across the condenser C for different frequencies of the voltage source. We assume a linear network with currents and voltages, which for given frequencies are proportional, and describe the selective properties of the circuit by a transmission ratio T given by the ratio of the condenser voltage to the driving voltage: $T = e_c/e$. The numerical capacitance and inductance is assumed to be independent of frequency. However, the dissipative resistance is assumed for mathematical convenience to be proportional to the frequency of currents through it. There is a physical justification of this assumption to a certain degree, and mathematical treatment with this assumption is more likely to be correct than if the resistance is assumed independent of frequency. Further, the voltage source is assumed to be maintained the same for all values of current drawn from it. An ultra-exact treatment would hardly give results of greater physical and engineering value than a treatment with the assumptions here made which give simpler equations.

1. Notation

For single-circuit notation, let

e = value of impressed voltage
Ω = the angular velocity at resonance
F_0 = the frequency at resonance
ω = vector angular velocity of voltage e
f = frequency of voltage e
$r = f/F_0$ ratio of impressed to resonance frequency
X = reactance of L, and of C, at resonance
R = lumped circuit resistance at resonance
η = decrement coefficient at resonance $= R/X$
e_c = voltage across C
$T = e_c/e$ = transmission factor of the circuit
f_1 and f_2 = two values of f for which T is the same numerically
$(f_2 - f_1) = \Delta f$ = the difference of the two equal transmission frequencies
$\beta = \Delta f/F_0$ the relative difference of frequencies for equal transmission.

2. General Formula for Transmission

Subject to the conditions imposed on e, R, L and C the solution is

$$T = \frac{-j}{r^2\left[\eta + j\left(1 - \frac{1}{r^2}\right)\right]} \text{ vectorially;}$$

$$\frac{1}{r^2\sqrt{\eta^2 + \left(1 - \frac{1}{r^2}\right)^2}} \text{ numerically.} \qquad (1)$$

The quantity η is here termed decrement coefficient because of its simple relation to the logarithmic decrement of the circuit for each cycle of the current with the system oscillating freely. In many cases it is approximately numerically equal to the "power factor," or, as many prefer, the "phase defect" of the inductance coil. The total η however includes that due to dielectric and shield losses, and due to space conduction in vacuum tubes associated with the circuit. Since the entire η determines the breadth of the resonance curve, and the decrement idea is somewhat archaic, a terminology based upon the resonance idea would be preferable.

3. Symmetrical Approximate Solution

For frequencies close to resonance, the value of r^2 by itself is of minor importance as it does not change rapidly in the region of resonance, but the value of $(1-1/r^2)$ is of considerable importance as it changes rapidly for values near resonance, that is, near $r=1$. To express the curve in terms of β the relative difference of the two frequencies for equal transmission, we have $\beta=(r_2-r_1)$ in which r_1 and r_2 are values of r less and greater than unity, respectively, which make $(1-1/r^2)^2$ the same quantity, say A^2, that is,

$$\left(1-\frac{1}{r_2{}^2}\right)=A, \quad \text{and} \quad \left(1-\frac{1}{r_1{}^2}\right)=-A. \tag{2}$$

From which we find $(r_2-r_1)=A+A^3/8+7A^5/128$. Neglecting the higher order terms, $\beta=A$ very closely. Wherefore in the general formula for numerical transmission, by considering r^2 as it appears by itself to be unity, and replacing $(1-1/r^2)$ by β which allocates both values of frequency for equal transmission we have the simplest possible solution

$$T=\frac{1}{\sqrt{\eta^2+\beta^2}} \text{ approximately.} \tag{3}$$

Three values of T are of especial interest.

For the value $\beta=0$, that is at resonance, the transmission is $T_0=1/\eta$, and the transmission curve as a whole with respect to transmission at resonance is

$$\frac{T}{T_0}=\frac{\eta}{\sqrt{\eta^2+\beta^2}} \text{ approximately.} \tag{4}$$

For values of frequencies which make $T/T_0=\frac{1}{2}\sqrt{2}$ we have simply $\beta=\eta$. This relation is the basis of determining the value of η without evaluating R and X by noting the relative difference of the two frequencies for which transmission is 0.7 times the maximum transmission.

If the single-circuit system is to represent an imitation of a symmetrical band-filter system over a range of frequencies of relative breadth B, (e.g. for 10-kc range at 500-kc mean value $B=0.02$) then the ratio of transmission at the extremes of the band to that at the center is

$$\frac{T_B}{T_0} = \frac{\eta}{\sqrt{\eta^2 + \beta^2}} \text{ approximately.} \tag{5}$$

For single-circuit systems then, the curve shape of transmission is specified on a relative basis by a single quantity η. For a transmission range B, equality over the range can be approached only by using a large value for η in comparison with B, but selectivity against transmission of undesired frequencies can be approached by using a relatively small value of η. There are insufficient shape constants to permit single-circuit transmission curves to meet both requirements of a band-pass filter in passing desired frequencies fairly uniformly, and in excluding undesired frequencies fairly completely. It is of interest to note how many independent shape constants there will be for coupled-circuit systems, by an adjustment of which ideal band-pass conditions may be to some extent approached.

B. Coupled-Circuit Arrangements

1. Types of Coupled Circuits

Types of possible coupled circuits are far greater in number than the single basic circuit of single-circuit systems. Several diagrams are shown in Fig. 2. In Fig. 2a the primary circuit containing the driving

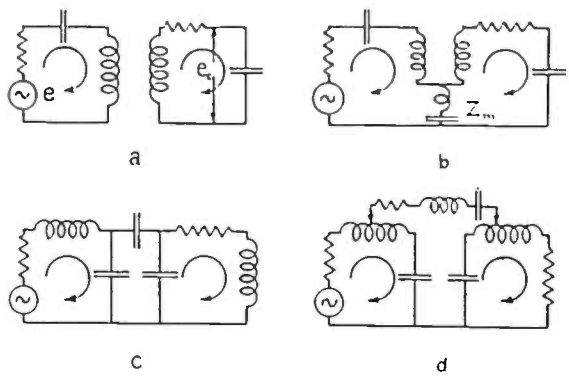

Fig. 2—Coupled-circuit arrangements.

voltage source is inductively coupled magnetically to the secondary circuit which includes the voltage operated or power consuming devices. Here but two currents must be considered in setting up the equations of flow, and this may perhaps be considered the most important as well as the simplest type. In Fig. 2b, voltage from each circuit into the other is inserted by impedance drop across the generalized impedance Z_m, which in special cases may be inductive, capacitive, or resistive. In both of these circuits, the coupling impedance is numerically small in comparison with reactance of the circuit elements. For Figs. 2c and 2d the coupling impedances are between points on primary and secondary circuits, and they are of large numerical value in comparison with circuit reactances. Circuits of types shown in Figs. 2a and 2b have the common property that with equal primary- and secondary-resonance frequencies, and with a coupling impedance which does not change rapidly in the range of transmission, then the transmission curve is fairly symmetrical with respect to the frequency to which each circuit is tuned. If the coupling impedance of Fig. 2b changes rapidly in the region of resonance, then the network does not exhibit the properties of a coupled-circuit system, and if the coupling impedance is also tuned to the same frequency as the circuits which are coupled an interesting three-circuit case arises. Transmission curves for networks of the types Figs. 2c and 2d are not symmetrical with respect to equal frequency tuning of primary and secondary, and this fact, combined with the difficulty often met in providing sufficiently high coupling impedances, makes these types of less importance. Often in actual equipment combinations of networks such as Figs. 2a and 2c occur due to undesired stray couplings.

Mathematical attention will be directed to the type of Fig. 2a, of the simplest possible arrangement, but results will be applicable in some respects to all types in which the coupling does not change rapidly in the transmission range.

By inspection of the simplest system, Fig. 2a, it may be judged that the transmission-curve shape is determined by four quantities, the primary decrement coefficient, the secondary decrement coefficient, the coupling between the circuits, and the relative difference of the frequencies to which the circuits are individually tuned. For simplification of the problem, treatment is given only for the case of the two circuits tuned each by itself to the same frequency F_0, the most important case in practice. However, the two decrement coefficients in general will be made unequal.

2. Additional Notation

The notation for coupled system, Fig. 2a, follows along the same general lines as for the single ciucuit, with the following additions:

X_1 = reactance of primary inductor or condenser at F_0

X_2 = reactance of secondary inductor or condenser at F_0

η_1 = primary decrement coefficient at resonance $= R_1/X_1$

η_2 = secondary decrement coefficient at resonance $= R_2/X_2$

$$\tau = \frac{M}{\sqrt{L_1 L_2}} = \frac{X_m}{\sqrt{X_1 X_2}} = \text{coupling coefficient.}$$

$T = e_{c_2}/e =$ transmission ratio of secondary condenser voltage to primary driving voltage.

3. General Transmission Formula

The general formula for absolute transmission is exactly

$$T = \frac{1}{r^2}\sqrt{\frac{X_2}{X_1}}\,\frac{\tau}{\sqrt{\left[\eta_1\eta_2 + \tau^2 - \left(1 - \frac{1}{r^2}\right)^2\right]^2 + \left[(\eta_1 + \eta_2)\left(1 - \frac{1}{r^2}\right)\right]^2}} \tag{6}$$

4. Symmetrical Approximate Solution

As in the single-circuit case consider $r^2 = 1$ over the range of interest, and identify $(1 - 1/r^2)^2$ with β^2, in which β is the relative difference of the symmetrically located frequencies for equal transmission and expanding:

$$T = \sqrt{\frac{X_2}{X_1}}\,\frac{\tau}{\sqrt{\beta^4 + (\eta_1^2 + \eta_2^2 - 2\tau^2)\beta^2 + (\eta_1\eta_2 + \tau^2)^2}} \tag{7}$$

The shape of the transmission curve as the frequency or its representative β varies is dependent solely upon the values of the coefficients $(\eta_1^2 + \eta_2^2 - 2\tau^2)$ and $(\eta_1\eta_2 + \tau^2)^2$. For a specified shape, the values of η_1, η_2 and τ must be so related as to make these coefficients definite numerical quantities. Because three variable parameters exist and only two numerical values are to be produced, it is evident many combinations of η_1, η_2 and τ will result in the same curve shape. Maximum possible transmission with a given shape requires a choice of τ, η_1 and η_2 which produces the desired shape with the largest possible value of τ.

However, it is not convenient to specify shapes of curves by assigning values to the coefficient of β^2 and to the constant term, since these have no direct physical or pictorial meaning. It is more logical to specify shapes of curves in terms of properties of the curves themselves, wherefore the shapes of the curves possible must be examined in more detail.

5. Classification of Curve Shapes

Examination for maxima and minima of T of (7) with changing β shows

(1) T is maximum for a set of values τ, η_1 and η_2 for the particular

value $\beta=\sqrt{\tau^2-(\eta_1^2+\eta_2^2)/2}$ provided $\tau>\sqrt{(\eta_1^2+\eta_2^2)/2}$. At the same time there is a minima at $\beta=0$ corresponding to a dip in the curve at frequency F_0.

(2) If $\tau\leqq\sqrt{(\eta_1^2+\eta_2^2)/2}$, then T is maximum for $\beta=0$ at the center of the transmission band.

The relation $\tau=\sqrt{(\eta_1^2+\eta_2^2)/2}$ may be called critical-shape coupling relation. For values of τ less than $\sqrt{(\eta_1^2+\eta_2^2)/2}$, then τ, η_1, and η_2 are so related as to give one-hump transmission curves, and the coefficient of β^2 in (7) is positive. For relations of τ, η_1, and η_2 such that τ is greater than $\sqrt{(\eta_1^2+\eta_2^2)/2}$, the curves are of the two-hump type and the coefficient of β^2 is negative.

a. Critical Shape Coupling.

For the boundary condition $\tau^2=(\eta_1^2+\eta_2^2)/2$, the coefficient of β^2 in (7) is zero, and the transmission shape is

$$T=\sqrt{\frac{X_2}{X_1}}\,\frac{\sqrt{\dfrac{\eta_1^2+\eta_2^2}{2}}}{\sqrt{\beta^4+\dfrac{(\eta_1+\eta_2)^4}{4}}}. \qquad (8)$$

The transmission at F_0 is $T_0=\sqrt{X_2/X_1}\ \sqrt{2(\eta_1^2+\eta_2^2)/(\eta_1+\eta_2)^4}$, and with this as a reference value the curve shape reduces to

$$\frac{T}{T_0}=\frac{\dfrac{(\eta_1+\eta_2)^2}{2}}{\sqrt{\beta^4+\dfrac{(\eta_1+\eta_2)^4}{4}}}. \qquad (9)$$

b. Sharper than Critical Shape Coupling.

There is no need for detailed discussion of curve shapes for less than critical shape coupling. For the symmetrical condition $\eta_1=\eta_2=\eta$ say, and with $\tau=k\eta$ in which k is a quantity less than 1,

$$T_0=\sqrt{\frac{X_2}{X_1}}\,\frac{k}{(1+k^2)\eta} \qquad (10)$$

$$\frac{T}{T_0}=\frac{\eta^2(1+k^2)}{\sqrt{[\beta^2+(1-k^2)\eta^2]^2+4k^2\eta^4}}. \qquad (11)$$

As k is made small, the amount of transmission becomes small and the selectivity greater. While coupled circuits were originally devised for the purpose of improving selectivity, the most important purposes at present are combined improvement of selectivity and improvement of uniformity of transmission, and chief interest in coupled circuits is therefore on the two-humped side of critical coupling rather than on the one-peak side.

c. Double-Hump Curve Shapes.

For describing a two-humped curve shape as in Fig. 3 resulting for $\tau > \sqrt{(\eta_1^2+\eta_2^2)/2}$, we use two important ratios. First is the breadth re-

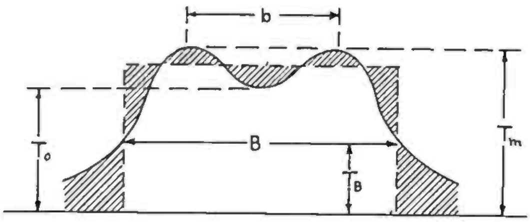

Fig. 3—Double-hump curve specification.

lation b conveniently measured by the relative difference $(fm_2-fm_1)/F_0$ of frequencies for maximum transmission. This is the value β assumes for these peaks and is given numerically by

$$b = \sqrt{\tau^2 - \frac{\eta_1^2 + \eta_2^2}{2}}. \tag{12}$$

The other important ratio is a height relation $h = T_m/T_0$ which gives the maximum transmission relative to the transmission at the center of the band for $r=1$ or $\beta=0$, and this height ratio is

$$h = \frac{\tau^2 + \eta_1\eta_2}{(\eta_1 + \eta_2)\sqrt{\tau^2 - \frac{1}{4}(\eta_1 - \eta_2)^2}}. \tag{13}$$

These two relations may be considered to take the place of the numerical values of the coefficient of β^2 and of the constant term of (7) in describing the curve shapes.

Elimination of τ between the equations for b and h gives the following relation of interest

$$(\eta_1 + \eta_2) = b\sqrt{2}\sqrt{\sqrt{\frac{h^2}{h^2 - 1}} - 1} = bf(h). \tag{14}$$

That is, if both b and h are specified, the sum of the two decrement coefficients must have a definite numerical value proportional to b

and to a function of h. For several values of h there are listed below the values of $f(h)$ by which b is to be multiplied to give the necessary sum of decrement coefficients.

TABLE I
REQUIRED SUM OF DECREMENT COEFFICIENTS

h	$f(h)$	h	$f(h)$	h	$f(h)$
1.01	3.50	1.07	1.905	1.3	1.06
1.02	2.75	1.10	1.675	1.4	0.935
1.03	2.52	1.15	1.43	1.5	0.825
1.04	2.30	1.20	1.275	1.7	0.69
1.05	2.14	1.25	1.16	2.0	0.55

In terms of b and h the complete curve shape is

$$T = \sqrt{\frac{X_2}{X_1}} \frac{\tau}{\sqrt{(\beta^2 - b^2)^2 + \frac{b^4}{h^2 - 1}}} \tag{15}$$

subject to $\tau = \sqrt{b^2 + (\eta_1{}^2 + \eta_2{}^2)/2}$ and $(\eta_1 + \eta_2) = bf(h)$ given above. For the peak transmission at $\beta = b$, the value T_{max} is

$$T_{max} = \sqrt{\frac{X_2}{X_1}} \frac{\tau\sqrt{h^2 - 1}}{b^2} = \sqrt{\frac{X_2}{X_1}} \frac{1}{(\eta_1 + \eta_2)} \sqrt{\frac{b^2 + \left(\frac{\eta_1{}^2 + \eta_2{}^2}{2}\right)}{b^2 + \frac{(\eta_1 + \eta_2)^2}{4}}} \tag{16}$$

For a given b and h and therefore a given sum for $(\eta_1 + \eta_2)$, maximum possible value of T_{max} occurs when the required decrement coefficients are equal, giving say $\eta_1 = \eta_2 = \eta$, for which

$$(T)_{max} = \frac{1}{2\eta} \sqrt{\frac{X_2}{X_1}}\,. \tag{17}$$

This value is independent of the values of b and h, so that the maximum transmission on two humps is quite independent of the exact nature of the transmission curve.

The quantity $\sqrt{X_2/X_1}$ appearing as a factor in several of the formulas is numerically equal to the ratio of secondary turns to the primary turns provided the windings in the two cases are the same dimensions and shape. In general, it is not helpful to compare a loosely-coupled transformer as in the present article with a tightly-coupled power or audio transformer. For example, it would be a mistake to conclude that a two-to-one winding ratio in both cases indicates a two-to-one voltage ratio.

On a relative basis with the transmission T_0 at the center of the transmission band as a basis, the entire curve shape is

$$\frac{T}{T_0} = \frac{\dfrac{b^2 h}{\sqrt{h^2 - 1}}}{\sqrt{(\beta^2 - b^2)^2 + \dfrac{b^4}{(h^2 - 1)}}} \tag{18}$$

which reduces properly to unity for $\beta = 0$ and h for $\beta = b$.

Wherefore for coupled-circuit systems with two-peak transmission, the whole curve is specified when the shape values of b and h are specified, and a definite sum of primary- and secondary-decrement coefficients is required so that when τ is suitably chosen the specified shape will be attained. The symmetrical equation (18) will suffice for most computations of shapes, but if more exact computations are required, or transmission is to be examined far from resonance, attention must be directed to the more complex equation (6).

III. Relations of Single- and Coupled-Circuit Transmission Curves

In comparing selectivity of coupled circuits with that of single circuits, it is evident from (4) and (9) that for off-resonant selectivity with β large in comparison with any of the η's, or τ, the transmission for single circuits is roughly inversely as the departure from resonance, while for coupled circuits it is roughly as the square of departure from resonance. This suggests that possibly simple relationships may exist between single- and coupled-circuit shapes at least in the simpler approximate solutions. Three relations known to exist will be described in brief.

A. Cascading of Two Staggered Single Circuits*

By the cascading of circuits we mean causing the output voltage of one circuit to determine a proportional input to second circuit without any reaction of the second circuit upon the first. Thus in cascading two curves one obtains the final overall transmission curve shape by multiplication of the individual curve shapes. By staggering is meant that one curve is centered upon one frequency and the other is centered on a slightly different frequency. The amount of staggering is described by the relative difference of the two frequencies F_1 and F_2 about which the original curves are symmetrical, and is given numerically by $a = (F_2 - F_1)/F_0$, in which F_0 is the mean value.

* See also Hazeltine, Proc. I.R.E., 14, 395; June, 1926.

It is physically evident that with the original curves identical, and with zero staggering, cascading must give the same type of transmission curve as coupled circuits with very weak coupling.

If, however, we set up a single-circuit curve of decrement coefficient centered at $F_0 = (1 + a/2)$ and a similar curve centered at $F_0(1 - a/2)$ each with unity maximum amplitude, then the product curve upon cascading is centered at F_0, and in terms of β is

$$T = \frac{\eta^2}{\sqrt{\beta^4 + 2(\eta^2 - a^2)\beta^2 + (\eta^2 + a^2)^2}}. \tag{19}$$

By reference to (7), the quantities under the radical sign are equal in (19) and (7) provided $\eta_1 = \eta_2 = \eta$ and $\tau = a$. Wherefore, as regards shape of transmission curves, two staggered single-circuit curves of the same shape cascade into a coupled-circuit transmission curve with a numerical correspondence between τ, the coefficient of coupling between equal circuits in the coupled-circuit case, and a the relative separation of the staggered circuits in the cascaded case. However by making a small to produce greater selectivity, the amount of transmission is improved in the cascaded case, but by making τ small the amount of transmission is diminished for the coupled-circuit case.

B. Vector Difference of Two Staggered Single Circuits

Instead of taking the product of two single-circuit transmission curves of like shape but staggered, we may take the vector difference and obtain a coupled-circuit type curve. This may be done experimentally by an arrangement as in Fig. 4. Here circuits I and II of equal

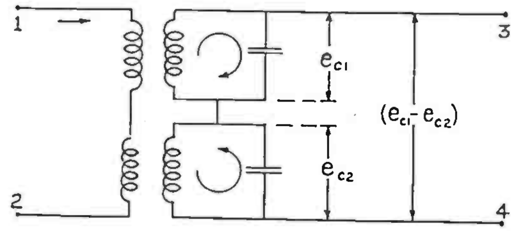

Fig. 4—Selectivity by vector difference.

constants, except slightly staggered as to resonance frequency and without coupling, are driven by equal couplings from power sources across terminals 1 and 2 supplying a current i which is assumed inversely as the frequency to maintain the voltage induced into the tuned systems constant independent of the reactions they may produce upon the power source. The two couplings from the power source are in an opposite sense whereby the voltage across the terminals 3 and 4 on open circuit is the vector difference of the voltages across the two condensers that would exist if the two were similarly driven. If circuits I and II are of the same frequency, it is obvious there will be no transmission, similar to the case of coupled circuits of extremely weak coupling; if however the circuits are of sufficiently different frequencies two-humped transmission results with high efficiency on each peak.

By taking the vector difference of two single-circuit curves of the same decrement coefficients but staggered in frequency by an amount a with unity maximum value for each curve, the overall transmission for a system of Fig. 3 numerically in terms of β becomes

$$T = \frac{2a\eta}{\sqrt{\beta^4 + 2(\eta^2 - a^2)\beta^2 + (\eta^2 + a^2)^2}}. \tag{20}$$

By comparison with the coupled-circuit shape (7) it is seen that (20) transmission curves correspond very closely to curves of coupled-circuit systems with a replacing τ in the numerator of the equation as well as where it occurs in the denominator.

For illustrating the two methods of obtaining coupled-circuit transmission curve shapes from single-circuit curve shapes, there are plotted in Fig. 5, *a*, *b*, and *c* three sets of curves based upon a single-circuit

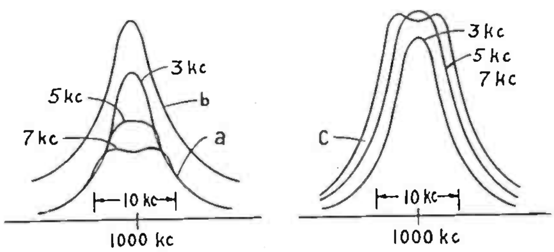

Fig. 5—Production of coupled-circuit transmission.
a. Staggered cascade b. Original circuit
c. Vector difference (Coupled circuit)

curve of unity peak transmission of decrement coefficient 0.005 at 1000 kc. The single-circuit curve shape is shown by itself in Fig. 5b. A nest of three coupled-circuit type curves produced by the cascading method is shown in Fig. 5a, corresponding to separations of 3000, 5000 and 7000 cycles, respectively, of the single circuits from which they are derived, with $a = 0.003$, 0.005, and 0.007. These curves are computable from (19). The nest of curves of Fig. 5c is for the same conditions with the combining by the vector-difference method, and computable from (20). Critical shape is illustrated by the curves for 5000-cycle separation, sharper than critical by 3000, and broader than critical by 7000-cycle separation curves. The set of curves *a* for cascading arrangement shows clearly that increase of coupling or of staggering up to limits makes for a transmission curve approaching that of an ideal band-pass filter. The set of curves c for vector-difference method indicates what happens in actual coupled-circuit conditions when the coupling coefficient is varied only. Considering the center of the transmission band, the transmission is a maximum for $\tau = a = \eta$. Since in these curves $\eta_1 = \eta_2$, the condition for critical shape coupling $\tau = \sqrt{\eta_1^2 + \eta_2^2/2}$ becomes identical with the well-known critical coupling relation $\tau = \sqrt{\eta_1 \eta_2}$ referring to the condition of maximum transmission

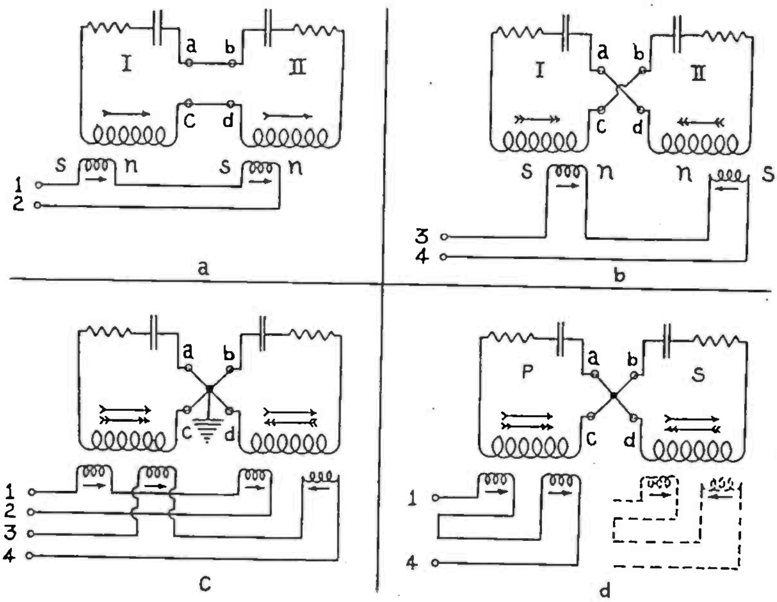

Fig. 6—Coupled circuits based on single circuits.

for F_0 at the center of the band as β is varied. For values of $\tau > \eta$, with two-peak transmission the amount of peak transmission is precisely the same as for the central frequency F_0 on critical coupling, and according to the simplified formulas, this is true, independent of how great τ may be. That is, the value T_{max} for each of the two humps is the same as for T_{max} for critical coupling, and for the basic coupled-circuit arrangement, transmission on each of two humps may be as efficient as transmission on a single hump.

C. Coupled Circuits with Primary and Secondary Excitation

A most interesting case of coupled-circuit selectivity produced by a vector difference of two single circuits is the case of a coupled circuit itself. This paradoxical statement is explained with reference to Figs. 6a to 6d.

In Fig. 6a are two equal L-C-R systems P and S with mutual inductance between them, all elements being connected in series by connectors a to b and c to d such that the inductive effects are additive, with lumped inductance $2(L+M)$, and capacity $\frac{1}{2}C$. The individual inductances are excited from a power source connected to terminals 1 and 2 by means of coils m and n additive in their effects. It is evident that this series arrangement is tuned as a whole to a frequency $F_0/\sqrt{1+\tau}$ in which F_0 is the frequency corresponding to L, C, and R, and τ is the coupling coefficient. Points a, b, c, and d are by symmetry equipotential for all currents of all frequencies in the tuned system. Each half of the circuit is excited by the same applied voltage from m and n respectively, and each half further has the same impedance for every frequency. Therefore no change in circuit currents will occur when the four linkage points are connected together. This produces from a single-circuit system a coupled circuit with primary and secondary excitation in such a manner as to excite them most effectively with single-circuit selectivity on frequency $F_0/\sqrt{1+\tau}$.

Similarly in Fig. 6b are the same two circuits with connections a to d and b to c with currents instantaneously opposing in direction in the circuit inductances, so that they may be correctly excited by couplings p and q from a circuit 3–4, coupling q being in a reverse sense to p. The frequency to which the system is responsive with single-circuit selectivity in this case is $F_0/1\sqrt{-\tau}$. It is further evident in this figure also that no change in currents for any frequency will occur when the linkage points are connected together. This makes a coupled circuit with primary and secondary excitation capable of being excited with single circuit selectivity with peak efficiency for frequency $F_0/\sqrt{1-\tau}$.

Further it is evident that with both power sources 1–2 and 3–4 coupled to Fig. 6a circuits, or 6b circuits, currents in the tuned system due to power source 1–2 will not react on the power source 3–4 and vice versa, whether connected as single- or coupled-circuit systems. Such a combination with linkage points joined and grounded is shown in Fig. 6c. This is termed a "flux filter" since directions of flux in primary and secondary circuits are utilized to permit power sources connected to 1–2 and 3–4 respectively to deliver power to each of the tuned circuits with no reaction of one source upon the other. This arrangement is described in patent to Dr. E. L. Chaffee, U. S. No. 1,601,109 in its applications to duplex transmission on a single antenna. In this application of the principle, an antenna and loading coil replaces one of the L-C-R circuits. This arrangement has been subjected to laboratory tests with even overlapping radiations without reactions. It is evident that circuits connected to 1–2 and 3–4 may contain tuned circuits so that the principle may be utilized with coupled circuit or higher degrees of selectivity on each channel with appropriate design on each channel to be transmitted. It may find a very useful application in the future for example in simultaneous sound and picture broadcast on adjacent channels without transmitter reactions.

It has been shown that if terminals 1–2 are driven at any frequency f, the currents in P and S are precisely the same when all points are tied together, as when a is connected to b and c to d. Further if terminals 3–4 are driven at the same frequency f the currents in P and S are precisely the same when the points are tied together as when a is connected to d, and b to c. If now we join 2 to 3 and impress but one power source across 1 to 4, the windings m and n will produce primary and secondary currents corresponding to the low-frequency resonance curve with like direction of flow of primary and secondary currents. Further windings p and q will produce currents corresponding to the high-frequency resonance curve with opposite direction of flow in the two closed circuits. The total primary current will be found by vector addition of the two individual primary currents, and the total secondary current by a vector subtraction. But actually the current from the driving source traverses all four windings, with n and q producing exactly opposite effects. Wherefore the net coupling in reality is to the primary circuit only as shown in Fig. 6d, corresponding to the circuit of Fig. 2a. Wherefore the coupled circuit system itself is a case of staggered single-circuit selectivity, with primary-current conditions determined by a vector summation of two single-circuit currents, and secondary currents determined by a vector difference.

A laboratory arrangement of circuits as in Fig. 6, with suitable links for making various connections and with primary and secondary meters has been used in experimental checks of the theory of the "flux filter" here presented, and it is believed such experiments would be of considerable value in showing the essential relationships of single and coupled circuits.

IV. Transfer Equations for Simplification of Problems

In single- and coupled-circuit tube systems, in practice, the driving voltages are not in a series arrangement with the circuit elements. This is because the driving circuits are incorporated in an impedance which is high compared with the reactance or series resistance it is feasible to use in the tuned circuits. As a result transformer action is required to adapt the circuits to the tube. The purpose of transfer equations is to determine what mathematical substitute may be used in a series arrangement to produce the same result as the actual tube connection.

For amplification study in the region of linear operation a vacuum tube is represented mathematically as a resistor of value R_0 which includes an alternating-voltage source μe_g. Here R_0 is the internal plate impedance of the tube, μ is the amplification constant and e_g is the voltage impressed upon the grid circuit. This is a mathematical abbreviation only, giving the correct result for circuits external to the tube, but of no physical use in explaining the tube operation, for example, as to heating of the tube anode during operation delivering power. When an external circuit of impedance Z is imposed upon the tube, this determines the ratio of alternating voltage across plate to filament of the tube to the alternating current drawn from the plate and returned to the filament of the tube. That is $Z = e_z/i_z$. The mathematical abbreviation gives the correct value for alternating voltage from plate to filament and therefore across the external load as different currents are drawn by the load, in accordance with the equation $e_z = \mu e_g - R_0 i_z$. This may be considered the voltage-regulation curve of the device as a generator, in the same manner as voltage-regulation curves are required for other generators for determining the actual performance with loading. For a given e_g, maximum value of $e_z i_z$, the power output of the tube to the external circuit occurs when the current drawn reduces the plate to filament voltage to half the no-load value, and this occurs with the external impedance a resistance of value equal to the mathematical internal impedance of the tube. In power transmitters for which the linear region of operation is greatly exceeded the regulation curve of the tube as a generator is no longer linear and the determination of effect of an external load is much more complex.

The most important cases of circuit connections for vacuum tube amplifiers are shown in Fig. 7, with R_0 representing the internal impedance and e representing the fictitious voltage μe_g inserted into the plate circuit. For other networks these diagrams are also useful as in any generating device in which the voltage source is inextricably associated with an internal impedance that is a resistance. In the first diagram Fig. 7a1 the coupling is direct capacitive by condenser X_c, with Z_1 an impedance of any nature which need not be specified. It is desired to know what fictitious voltage e' and what fictitious impedance Z_1, both in a series arrangement with X_c and Z_1, will produce the same identical current i through Z_1 of Fig. 7a2 as the real voltage source e would produce through Z_1 of Fig. 7a1. For 7b1 and 2 the feed is by magnetic coupling with primary reactance X_p, and mutual reactance X_m between X_p and the entire closed circuit. For Fig. 7c the transfer is by direct magnetic coupling with the coupling inductance X_1 itself coupled in general by mutual reactance X_m with the remainder of the circuit. In these cases also it is desired to know values e' and Z' which would replace as series voltage and impedance the

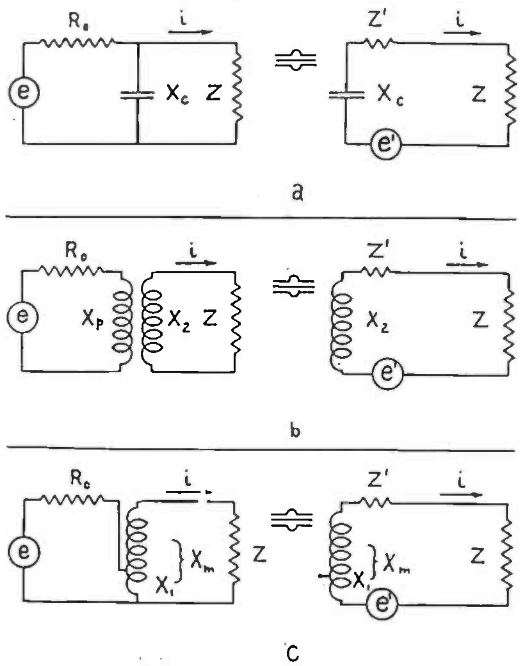

Fig. 7—Transfer equations.

actual voltage e and impedance R_0 in their effects upon Z_1. Neglecting resistance of the coupling devices, the results of computations are as in Table II.

TABLE II
TRANSFER EQUATIONS GENERAL

Fig.	Coupling	e'	Z'
7a	Direct capacitive	$\frac{-jX_c}{R_0-jX_c}e$	$\frac{X_c^2}{R_0-jX_c}$
7b	Inductive magnetic	$\frac{jX_m}{R_0+jX_p}e$	$\frac{X_m^2}{R_0+jX_p}$
7c	Direct magnetic	$\frac{j(X_1+X_m)}{R_0+jX_1}e$	$\frac{(X_1+X_m)^2}{R_0+jX_1}$

In case other voltages exist in the network due to power sources other than e, these equations may be used to compute the portion of current at any point due to e only, and a separate computation for the other voltage source may be made, assuming of course such other currents do not interfere with the proper functioning due to overloading of the devices represented by e and R_0. In particular the transfer equations are applicable with e = zero to determine effects when a power-consuming resistance loads a closed circuit with other than a series arrangement.

A. Approximations for Selective Systems

For approximate work with sharply selective circuits R_0 in practice is large compared with X_c, X_p, and X_1 in the three cases, with the result that Z' is nearly a resistance, but with slightly inductive phase defect for capacitive coupling, and slightly capacitive phase defect for the two magnetic couplings. Let the lumped reactance of a tuned system driven by any of the methods shown be X. We may then conveniently record approximate numerical values for: voltage transfer ratio e'/e, equivalent series resistance R', decrement coefficient η' due to the resistance R_0, and change of resonant frequency $\Delta f/F_0$ due to the reactive component of Z', as follows:

TABLE III
Approximations for Selective Systems

Fig.	Coupling	e'/e	R'	η'	$\Delta f/F_0$
7a	Direct capacitive	$\frac{X_c}{R_0}$	$\frac{X_c^2}{R_0}$	$\frac{X_c^2}{R_0 X}$	$-\frac{1}{2}\frac{X_c^3}{R_0^2 X}$
7b	Inductive magnetic	$\frac{X_m}{R_0}$	$\frac{X_m^2}{R_0}$	$\frac{X_m^2}{R_0 X}$	$+\frac{1}{2}\frac{X_m^2 X_p}{R_0^2 X}$
7c	Direct magnetic	$\frac{X_1+X_m}{R_0}$	$\frac{(X_1+X_m)^2}{R_0}$	$\frac{(X_1+X_m)^2}{R_0 X}$	$+\frac{1}{2}\frac{(X_m+X_1)^2 X_1}{R_0^2 X}$

By means of these transfer equations, tuned-circuit problems are readily simplified with a sufficient degree of accuracy, and coupled-circuit problems which rigorously would involve as many as six branch networks may be reduced to problems with but two currents to be considered.

V. Single-Circuit Tube-Circuit Design

Three important cases of tube-circuit design arise. In the first case, tuned circuits are used to receive power from the plate circuit of an amplifying tube and deliver voltage selectively to the grid circuit of the next amplifying tube, with practically all the power received by the tuned system expended in circuit losses. In the second, the circuits transfer power selectively from a power source associated with an internal impedance to another power-consuming device. In the third case as in power tubes for transmission with tuned output, there is not a linear relationship between output voltage of the generating source and the current drawn from it. These three cases will be discussed in turn, first for single circuits for which the computations and statement of the problem are fairly simple.

A. Voltage Amplification (Single Circuit)

The case of voltage amplification with the load not power consuming is illustrated with reference to a magnetically-coupled transfer from tube to tuned circuit shown in Fig. 8a and 8b, commonly used in amplifier circuits. Interest centers in the ratio e_c/e_g of the voltage across the condenser divided by the impressed grid voltage, and the solution is readily reached by use of the reduced diagram Fig. 8b,

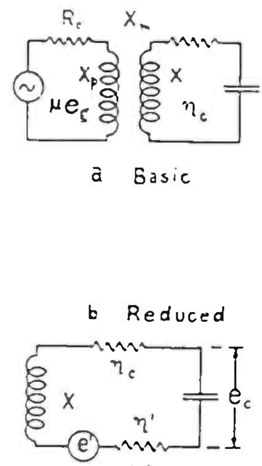

Fig. 8—Voltage amplification.

with e' and η' representative of the voltage μe_g and internal impedance of the tube, as tabulated in Table III. As shown in II.A.3., the condenser voltage of Fig. 8b is the impressed voltage e' divided by the sum of the two decrement coefficients η_c due to coil-condenser losses, and η' due to the presence of the tube impedance. By using the transfer equations, we find for the amplification of the stage

$$A = \frac{e_c}{e_g} = \frac{\mu \dfrac{X_m}{R_0}}{\eta_c + \dfrac{X_m^2}{R_0 X}}. \tag{21}$$

If not particular as to the shape of the transmission curve, maximum amplification by increasing X_m results when the decrement due to the tube is made to equal the coil-condenser decrement, giving

$$A_{\max} = \frac{\mu}{2\sqrt{R_0}}\sqrt{\frac{X}{\eta_c}} \tag{22}$$

requiring

$$X_m = \sqrt{\eta_c R_0 X}. \tag{23}$$

For voltage amplification we are limited in the amount of gain per stage by the ratio X/η_c which it is feasible to use for the coil-condenser system, and in some cases by the feasible amount of coupling. Also, if the coefficient of coupling between the coil X_p and X is too weak, the value of X_p required for maximum amplification may be so great as to be not negligible in comparison with R_0, requiring slight change of the condenser to give maximum amplification at the desired frequency.

The condition for maximum amplification with a given coil-condenser system, that the coupling from tube to circuit doubles the effective breadth of resonance, is entirely equivalent to making the load impedance upon the tube equal to its internal impedance. Under these conditions the alternating voltage from plate to filament is half the value for no-load with very high external impedance. This is for the central frequency only. At off-resonance frequencies the impedance is less than the tube impedance, the plate current becomes correspondingly greater. The broadening of resonance therefore is explained by the greater voltage impressed from primary to secondary for off-resonant frequencies.

If it is not desired to accept a breadth of resonance equal to double that of the circuit itself, then maximum amplification cannot be attained. If on the other hand a greater breadth of resonance is desired, then preferably from the standpoint of maximum amplification under the specified shape conditions the additional decrement should come from increased coupling to the tube rather than from resistance purposely inserted in the coil-condenser system.

B. Selective Power Transfer

In many applications the output is not of the voltage type with power consumed in the coil-condenser system only, but rather the useful output is consumed in a load coupled to the tuned system. Such a case is illustrated in Fig. 9a and 9b, with R_l the useful load resistance shunting the condenser of the tuned system. In this case the tuned system is interposed as a selective circuit which also may act as a transformer to adapt the load to the tube. If this circuit acts as an ideal transformer without losses the voltage output across R_1 cannot possibly exceed the value $\mu e_g/\sqrt{R_l/R_0}/2$ for which condition the maximum possible power capable of being delivered external to the tube goes to the resistance R_l.

For determining the proper choice of circuit constants for a given R_0, R_l and μ, the circuit is reduced to the basic arrangement of 9b by use of the transfer equations of Table III. The condenser voltage for this figure is $e'/\eta_c+\eta_t+\eta_l$ in which η_c is the coil-condenser decre-

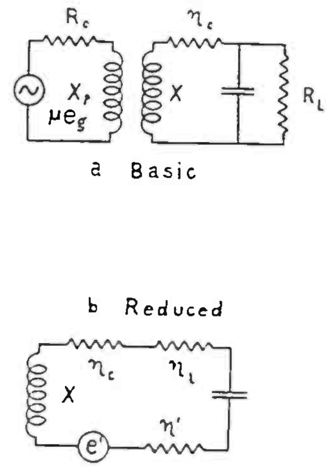

Fig. 9—Power transfer.

ment coefficient, η_t is that due to the tube and η_l is that due to the load. By using the transfer equations noting that the voltage in the resistor R_l is zero, the voltage amplification is

$$A = \frac{e_l}{e_g} = \frac{\mu \dfrac{X_m}{R_0}}{\eta_c + \dfrac{X_m^2}{R_0 X} + \dfrac{X}{R_l}}. \tag{24}$$

If then we specify that the total breadth of resonance corresponds to a certain amount of decrement coefficient η, and further make the reasonable assumption that the coil-condenser decrement may be maintained η_c independent of what X shall ultimately be, then it turns out that maximum amplification results when the tube and load decrements are equal, and therefore $\eta_t=\eta_l=(\eta-\eta_c)/2$. Whence the values of tube coupling and of circuit reactance are both uniquely determined, giving

$$X = \left(\frac{\eta - \eta_c}{2}\right) R_l; \quad X_m = \left(\frac{\eta - \eta_c}{2}\right) \sqrt{R_0 R_l} \tag{25}$$

The amplification under these conditions is

$$A_{\max} = \frac{\mu e_g}{2} \sqrt{\frac{R_l}{R_0}} \left(\frac{\eta - \eta_c}{\eta}\right) \tag{26}$$

whereby the gain ratio at resonance is cut down by a factor $(\eta-\eta_c)/\eta$ from that which would result from using a circuit of no losses.

It will be noted that the load impedance upon the tube for maximum voltage to the resistance R_l is not equal to the internal impedance of the tube. Of course if the circuit X, η_c and R_l is specified, and no regard need be given to overall decrement, maximum amplification would occur when the tube coupling is adjusted to make tube impedance and load impedance equal. But with the curve shape restriction, best design calls for dividing the difference between coil-condenser decrement and shape decrement equally between the tube and the ultimate load irrespective of what the impedance relation may be.

C. Tuned Power Transmission

When a vacuum tube operates into a selective system with the region of linear operation greatly exceeded, the representation of the tube as a generator source can no longer be in terms of an amplification constant and an internal impedance. With a given bias, and a given alternating voltage impressed upon the grid circuit, and an output impedance which is small for harmonic frequencies, compared with the fundamental, as in usual power-transmitter circuits, the distortion is chiefly in the current wave from tube to load rather than in the voltage-wave form from plate to filament across the load. The load impedance for the fundamental frequency determines the ratio of plate-filament alternating voltage to fundamental load current. Lowering the load impedance to draw more plate current causes a lowering of terminal voltage, but the regulation of the tube as a generator for different load impedances and power factors cannot be expressed in terms of a voltage drop due to a non-varying internal resistance, and for accurate work resort must be made to curves for solutions of problems. Such complications are beyond the scope of the present article.

In transmitter-tube circuits much greater attention must be paid to obtaining the proper load impedance and power factor than in receiver circuits. In linear operation of tubes requirements of shape of transmission curves, especially with a power-consuming loading, may cause considerable mismatch between tube internal impedance and load impedance, but in transmitter circuits, especially operating at high efficiency and maximum output consistent with a power loss rating of the tube structure, considerable care must be exercised to make the load impedance correct. For example, a hypothetical tube may operate at recommended values 2000 v, 5 amperes, plate supply, with 7.5-kw plate output, and therefore 2.5-kw plate dissipation. Tube characteristics may be such as to make this possible with 1250 v a-c from plate to filament, and 6 amperes plate fundamental current to the load, requiring a load impedance of 208 ohms. If the impedance due to the load differs from this value by fifty per cent, or if the power factor of the load is less than 0.50 for example, grid-circuit adjustments would be required to prevent excessive heating, with an extremely great falling off in output power compared with the loss under similar conditions in receiver tubes in the linear region.

Wherefore for study of tuned systems in connection with power transmitters, first center interest upon the power factor and impedance of the load rather than upon transmission curve shape. Moreover, if transmission curve shapes are obtained by varying the frequency of impressed grid voltage, without change of bias or numerical amount of voltage or plate-circuit conditions, with due regard to avoiding overheating, the resulting curves are inaccurate for determining the performance of the tube with composite wave form such as a carrier wave and two side bands. The effect of the presence of the tube is to broaden the transmission curve, because of the falling nature of the regulation characteristics, but it is beyond the scope of the present article to discuss these effects in detail.

The tuned systems in use at present are usually of the coupled-circuit type, and will be discussed later. But for completeness it is desirable to record the impedance load values due to single circuits with couplings of the three types of Fig. 7, in which the impedance Z_1 in connection with the coupling device makes a tuned circuit of decrement coefficient η with circuit lumped reactance at resonance X. For the three types of coupling the impedance values at resonance are measured by the ratio of voltage across X_c, X_p, and X_1 respectively to the current from source e through R_0. The values are

direct capacitive	inductive magnetic	direct magnetic
$\dfrac{X_c^2}{X\eta}$	$\dfrac{X_m^2}{X\eta}$	$\dfrac{(X_m + X_1)^2}{X\eta}$

In the important case (Fig. 7a) in which Z_1 is an inductance, of decrement coefficient η, then $X_c = X$, and the impedance is simply X/η.

VI. Coupled-Circuit Tube-Circuit Design

In coupled-circuit tube systems for receiver amplifiers, stress is usually upon transmission-curve shapes rather than amount of amplification. In the earlier uses of coupled circuits before cascaded amplification with tuned systems, and even before vacuum tubes themselves were available, the purpose was attaining selectivity over that possible with single-circuit tuning, and conditions were such that attention was concentrated upon transmission forms corresponding to critical or less than critical coupling. In the present-day applications, however, attention is chiefly directed rather to the two-humped side of critical coupling. The requirements are two-fold, that the system pass fairly well all frequencies within a desired range, and that it will not pass well frequencies outside the desired range. To satisfy both conditions, a double-humped curve is unquestionably superior to a single-humped curve of single- or of coupled-circuit selectivity.

The shape of double-humped curves has been specified already by a breadth relation b with reference to the relative difference of frequencies for peak transmission and a height relation h with reference to the peak transmission relative to that at the center of the curve. For band-pass design, we have further to consider B the relative difference of the highest and lowest frequencies desired to be passed.

Let a further ratio be k the transmission at the center of the band relative to the transmission at the extremes of the band for $\beta = B$. Then numerically we have the following relation

$$b = \frac{B}{\sqrt{1 + \sqrt{\dfrac{k^2h^2 - 1}{h^2 - 1}}}} \tag{27}$$

For specification of a band-pass filter, there is little chance for disagreement as to numerical value of B. For example if a transmission band 10 kc wide at 1000-kc mean value, then $B = 0.01$; however many opinions might arise as to what values of k and h should be chosen. If $k = h$, the transmission varies by a factor not greater than h from the value at the center of the band. The greater the value of $k = h$, the less uniform will the transmission be over the desired range, but the greater will be the exclusion of the undesired frequencies. If uniformity in the center of the band is more important than uniformity at the edges, h will be not much greater than unity, but k may be considerably greater than unity. Further consideration of course must be paid to characteristics of any other selective devices of the amplifier system, as a choice of k and h for one stage alone would be different than if single or coupled circuits are cascaded with the coupled circuit under consideration.

On a mathematical basis, k and h might be determined numerically by a method of least squares, such that the corresponding transmission curve has the least possible average of the squares of the departures from the perfect shape for band filtering. However, such a numerical determination would presumably not serve as a standardization of coupled-circuit band-pass filter shape because of the many non-mathematical factors involved.

As a concrete example, consider the circuit arrangement of Fig. 10, which is desired to give a transmission shape $h=1.1$, $k=1.3$ to represent a band filter of 10-kc breadth at 200-kc mean frequency, that is, for $B=0.05$.

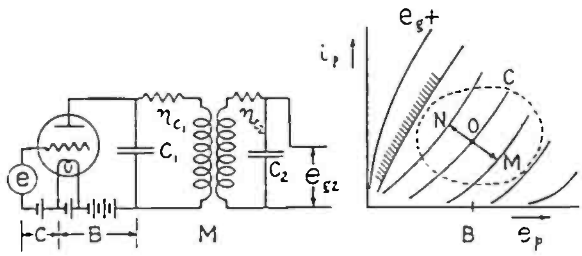

Fig. 10—Amplifier circuit, linear operation.

Let $R_0=10{,}000$, and $\mu=8$ for the tube, and $C_2=0.0005$, $\eta_2=0.01$ for the secondary circuit. Let it be assumed that whatever value is required for primary reactance, we will make $\eta_1=0.01$ for the coil-condenser system.

Having specified completely the curve shape by choice of B, h and k, it is desired to determine

what choice shall be made for primary-circuit constants?
what will be the amount of amplification?
what must be the value of τ?
how much will the primary-circuit tuning frequency be off from the value determined by L and C values?

what would the amplification be for single-circuit selectivity with maximum-power transfer to a circuit of the characteristics of the secondary chosen?

For computation purposes the circuit reduces to Fig. 2a, in which

$$e = \mu e_{g1} \frac{X_{c1}}{R_0}$$

$$\eta_1 = \frac{X_{c1}}{R_0}$$

$$\frac{\Delta f}{F_0} = -\frac{1}{2}\left(\frac{X_{c1}}{R_0}\right)^2$$

The computation proceeds along the following lines.

TABLE IV

Computation of Band-Pass Selector-Coupled Circuit

$b = \dfrac{B}{\sqrt{1+\sqrt{\dfrac{k^2h^2-1}{h^2-1}}}}$	(1)	0.0278
$(\eta_1+\eta_2) = b\sqrt{2}\sqrt{\sqrt{\dfrac{h^2}{h^2-1}}-1}$	(2)	0.0466
$\eta_t = (\eta_1+\eta_2)-(\eta_1+\eta_{c1})$	(3)	0.0266
$X_1 = R_o\eta_t$	(4)	266 ohms
$\tau = \sqrt{b^2+\dfrac{\eta_1^2+\eta_2^2}{2}}$	(5)	0.0385
$T_o = \dfrac{e_{c2}}{e_t} = \sqrt{\dfrac{X_2}{X_1}}\;\dfrac{\tau\sqrt{h^2-1}}{b^2}$	(6)	56
$A_o = \dfrac{e_{g2}}{e_{g1}} = \dfrac{\mu X_{c1}}{R_o}T_o$	(7)	12
$\dfrac{\Delta f}{F_o} = \dfrac{1}{2}\left(\dfrac{X_{c1}}{R_o}\right)^2$	(8)	0.000350
$A_{\max} = \dfrac{\mu}{2\sqrt{R_o}}\sqrt{\dfrac{X_2}{\eta}}$	(9)	16

From (4), (5), (7), (8), and (9) of above table we find the primary condenser should be 0.003 μf, and inductance 208 μh; the coefficient of coupling 0.0385; the amplification at center of the band 12; the error of primary circuit tuning approximately 0.035 per cent; and the maximum amplification with single-circuit selectivity is 16. Whence the amplification at the center of the band is 75 per cent of what it could have been with single-circuit design disregarding selectivity and breadth of resonance.

For producing a band-pass filter of given shape and numerical width adjustable by a control for a wide range of central frequencies, as in broadcast receivers, many factors are involved. The value of B and therefore b for a specified shape must be smaller for the higher-central frequencies than for the lower-central frequencies. This means that the sum of the circuit and tube decrements must be inversely as the frequency, and even if circuit decrements at high frequencies could be maintained the same as at low frequencies, the tube decrement must necessarily be made to fall off at the higher frequencies, resulting in very much less amplification as the price to be paid for constant band width. Not only must the coupling from tube to primary be changed in accordance with the frequency in a correct manner, but also the coupling from primary to secondary must be changed in order to maintain overall transmission shape constant.

For the staggered-cascade amplification method, the staggering must be relatively less for the higher-frequency settings than the low, but it should prove more economical to provide for such staggering than to provide both for equalizing primary and secondary frequencies and for varying the coupling between them in a correct manner. In addition to staggering, the equivalent decrement coefficient of tube and circuit must be made less for the higher frequencies than for the lower, whereby an arrangement for changing the coupling from tube to circuit must be made in the staggered case as well as in the coupled circuit case.

It would appear then that coupled-circuit methods of maintaining uniform band width require, (1) equality of primary and secondary tunings over the range, (2) weakening of coupling from primary to secondary as the frequency is increased, (3) weakening of coupling from tube to circuits as the frequency is increased. On the other hand, maintaining uniform band width by the staggered-cascade method requires, (1) correct staggering over the range, for example, by condenser-plate shapes or other adjustments, (2) weakening of the coupling from tube to circuit at the higher frequencies.

In commercial designs for the broadcast band, the desire for uniform band width which must be in accordance with these requirements, must be tempered by consideration of other desires, including maximum sensitivity irrespective of band width or transmission-curve shape, simplicity of construction, and permanence of adjustment.

Presumably a given accuracy of approach to band-pass selectivity over a wide range of frequencies is more easily worked out by use of staggered circuits than by coupled circuits, and coupled-circuit systems are justifiable over the cascade-staggering method only on the basis of added selectivity in the early stages of amplification, and on the basis that more tuned circuits are desired than corresponding amplifying tubes.

B. Selective Power Transfer

In some instances the coupled-circuit system is desired to transmit power selectivity from a voltage source in series with a resistance to another resistance. The resistance including the voltage source is associated with the primary circuit and the load resistance with the secondary. If the voltage-source resistance is R_0 and the load resistance i R_l then the maximum voltage across R_l cannot possibly exceed the lue $e\sqrt{R_l/R_0}/2$ in which e is the voltage source value. The design roblem is to secure the proper transmission-curve shape as may be ecified by a choice of B, h, and k as in the previous section, and at e same time arrange for maximum transmission. Once a choice shape is made, the sum of the total decrement coefficients of primary nd secondary circuits is uniquely determined. For voltage amplificaon the difference between the decrement available in coil-condenser sses and that required for the given shape specifications is made up y choice of primary circuit to give the proper tube decrement.

For power transfer purposes the proper distribution of decrements in general quite complex, but for the important special case of qual decrement coefficients of primary and secondary coil-condenser ystems, the balance of the required decrements should be equally istributed between decrement to the primary due to the source resisance and that to the secondary due to the load resistance. In general, he ratio of output to input voltage is given for both peaks by

$$\frac{e_l}{e} = \frac{1}{2}\sqrt{\frac{R_l}{R_0}}\sqrt{\left(\frac{\eta_1 - \eta_{c1}}{\eta_1}\right)\left(\frac{\eta_2 - \eta_{c2}}{\eta_2}\right)} = \frac{1}{2}\sqrt{\frac{R_l}{R_0}}\sqrt{\frac{\eta_t}{\eta_1} \times \frac{\eta_l}{\eta_2}} \tag{28}$$

vith η_t and η_l representing the contributions of source and load derements to the total decrements η_1 and η_2 of primary and secondary ircuits respectively. If therefore we let $\eta_1 = \eta_2 = \eta$ representative of he curve shape, and $\eta_{c1} = \eta_{c2} = \eta_c$ representative of the coil-condenser lecrements, we have for the important symmetrical case which gives naximum possible transmission

$$\frac{e_l}{e} = \frac{1}{2}\sqrt{\frac{R_l}{R_0}}\left(\frac{\eta - \eta_c}{\eta}\right) \tag{29}$$

Wherefore as in the single-circuit case for power transfer, the maximum possible voltage output value $\frac{1}{2}\sqrt{R_l/R_0}$ is to be multiplied by the factor $(\eta - \eta_c)/\eta$ to take account of the coil-condenser losses. The decrements due to source and load must be high compared with coil-condenser decrements if high efficiency of power transfer is desired. The manner of coupling to give the desired source and load decrements may be chosen to suit individual cases. If the resistance values R_0 and R_l are small naturally they will be inserted in circuits with a series relationship, but if high, couplings involving transformer action will be required for the computation of which the transfer equations of Table III will be found useful.

C. Tuned Power Transmission

As described under the discussion of tuned single-circuit power transmitters with linear region of operation greatly exceeded, it is of primary importance in design that the load impedance on the tube be of proper value to a considerable accuracy if good performance is to be attained. Most of the modern tube transmitters use coupled-circuit output with the primary a tank circuit coupled to the transmitter tube and the secondary including the antenna system. The secondary circuit usually is of considerably greater decrement coefficient than the primary and with good design a large portion of the power from the tube is transferred to the secondary. The well-known advantages of tank circuit transmitters are elimination of harmonic frequencies through superior selectivity, certainty of operation with different antenna constants, and flexibility of impedance matching.

The computation of impedance load upon a voltage source due to coupled circuits is not especially difficult for the case of primary and secondary circuits tuned separately to the same frequency. Referring back to the basic system of Fig. 2a the impedance load upon source e is in general readily proved to be

$$Z = \frac{e}{i_1} = rX_1 \left\{ \frac{\left[\eta_1\eta_2 + \tau^2 - \left(1 - \frac{1}{r^2}\right)^2\right] + j\left[(\eta_1 + \eta_2)\left(1 - \frac{1}{r^2}\right)\right]}{\eta_2 + j\left(1 - \frac{1}{r^2}\right)} \right. \tag{30}$$

The impedance Z is a resistance for the value $r=1$, which causes the value of Z to become for the central frequency

$$Z_0 = X_1\left(\eta_1 + \frac{\tau^2}{\eta_2}\right). \tag{31}$$

Also the impedance Z is a resistance for any value of r which makes the phase angle of the numerator equal to that of the denominator of the general expression for Z, that is:

$$\frac{\eta_1\eta_2 + \tau^2 - \left(1 - \frac{1}{r^2}\right)^2}{\eta_2} = (\eta_1 + \eta_2) \tag{32}$$

giving

$$\left(1 - \frac{1}{r^2}\right) = \pm\sqrt{\tau^2 - \eta_2{}^2}. \tag{33}$$

Therefore not only is the impedance a resistance for $r=1$, at the requency at which both primary and secondary circuits are tuned, ut also for frequencies above and below with relative difference given y

$$\frac{\Delta fr}{F_0} = \sqrt{\tau^2 - \eta_2{}^2}. \tag{34}$$

The three frequencies for resistance impedance are designated F_-, F_0, nd F_+.

It is interesting to note three critical or borderline relations among he decrement coefficients and the coupling coefficient. First is the vell-known critical-coupling relation $\tau = \sqrt{\eta_1\eta_2}$ referring to the conlitions of maximum power transfer from primary to secondary for requency F_0, the second is the critical shape relation $T = \sqrt{(\eta_1^2 + \eta_2^2/2}$ eferring to the borderline condition for appearance or disappearance of two-hump transmission curve, the third is the critical triple-resisance relation $\tau = \eta_2$ referring to the borderline condition determining vhether the impedance is a resistance for one frequency or for three. For the third, it is evident that the coupling relation cannot depend ipon the value of the primary decrement coefficients because once onditions are such that a resistance impedance exists for any freuency, then changing by addition of resistance can merely add to he impedance value without introducing phase defect. Wherefore onditions may be such that the impedance is a resistance for three requencies if the voltage is impressed in one tuned circuit, while it is a esistance for but one frequency if impressed in the other circuit.

The frequencies for resistance impedance other than F_0 coincide with the frequencies for maximum transmission only if $\eta_1 = \eta_2$, but they coincide approximately if the coupling coefficient is large compared with the decrement coefficients, with sharply-peaked transmission curves. The numerical values of impedance for the three frequencies for resistance impedance are approximately

frequency	resistance
F_-	$X_1(\eta_1 + \eta_2)$
F_0	$X_1(\eta_1 + \tau^2/\eta_2)$
F_+	$X_1(\eta_1 + \eta_2)$.

In connection with tube circuits the impedance load upon the tube is a resistance for practically the same frequencies as if the voltage were impressed into the primary in a series arrangement. Referring to Fig. 7, replacing Z_1 of the figure by an inductance to which a second-ary circuit is coupled, then the impedance load resistance values given by the ratio of voltage across the coupling device to the current through power source e is given by

TABLE V

RESISTANCE LOAD VALUES

Frequency	Capacity coupling	Inductive magnetic	Inductive magnetic
F_-	$\dfrac{X_c}{\eta_1+\eta_2}$	$\dfrac{X_m}{\eta_1+\eta_2}$	$\dfrac{X_1+X_m}{\eta_1+\eta_2}$
F_0	$\dfrac{X_c}{\eta_1+\tau^2/\eta_2}$	$\dfrac{X_m}{\eta_1+\tau^2/\eta_2}$	$\dfrac{X_1+X_m}{\eta_1+\tau^2/\eta_2}$
F_+	$\dfrac{X_c}{\eta_1+\eta_2}$	$\dfrac{X_m}{\eta_1+\eta_2}$	$\dfrac{X_1+X_m}{\eta_1+\eta_2}$

These values are of the same general nature for the three types of coupling, and illustration by capacity coupling will serve to show the general properties.

Overall efficiency of operation requires that the impedance load upon the tube be a suitable value so that it may deliver power well to its load, and further that the decrement referred to the primary due to the presence of the secondary shall be large compared with the primary decrement by itself. With transmission at F_0 there is consider-able leeway in adjusting the circuit to optimum value for tube opera-tion through variation in primary- to secondary-coupling coefficient. For efficient operation it is necessary to make X_c/η_1 several times the desired ultimate impedance, and then to bring about the desired load-ing by coupling the secondary with sufficient closeness to bring about the proper impedance condition. But if transmission at $F-$ or $F+$ is desired no flexibility exists, and if it is desired to use the two-humped properties to permit transmission in an efficient manner on either of two frequencies without change of tube output circuit values, then considerable care must be exercised in the choice of circuit constants, and in designing the primary circuit to be of low decrement coefficient.

The tank-circuit efficiency is measured by the ratio of power de-livered to the secondary divided by the power received from the tube, and is numerically measured by the decrement on the primary due to the secondary divided by the same increased by the true primary-decrement coefficient. For the two cases we have

frequency	circuit efficiency
F_0	$\dfrac{\tau^2}{\tau^2+\eta_1\eta_2}$
$F\pm$	$\dfrac{\eta_2}{\eta_1+\eta_2}$

For a given η_1 and η_2, the circuit efficiency for F_0 always exceeds iat for $F\pm$, since when the latter frequencies exist $\tau>\eta_2$, which in ie above equations makes $\tau^2/(\tau^2+\eta_1\eta_2)>\eta_2/(\eta_1+\eta_2)$. For the threshold alue $\tau=\eta_2$ with equal efficiency on all three frequencies because they re extremely close together, the numerical efficiency may be conderably greater than fifty per cent, since $\eta_2>\eta_1$ usually. For highest fficiency however the circuits will be adjusted to have resistance

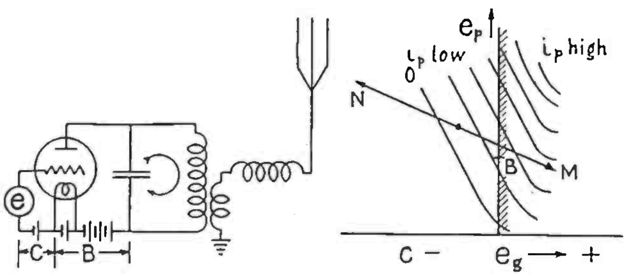

Fig. 11—Tank transmitter, non-linear operation.

npedance for three frequencies, but the impedance value will be iade correct and the circuit operated on the central frequency. For he side values $F+$ and $F-$, the impedance value is higher. It is to e remembered that the impedance being a resistance for three freuency values does not mean that the transmission will be doubleeaked, because of the unknown broadening of resonance due to the ube-regulation characteristics.

As a numerical example, consider a tank-circuit transmitter as hown in Fig. 11, for a central frequency 1,000-kc, secondary decrement oefficient 0.04, primary tank-circuit decrement coefficient 0.01, with ank condenser and coupling from primary to secondary to be chosen o give impedance value 800 ohms for F_0 and 1,000 ohms for $F\pm$. Ve find readily

$$\tau = 0.0458; \quad X_1 = 50 \text{ ohms}; \quad \frac{\Delta f}{F_0} = 0.0224; \quad C_1 = 0.00318\mu\text{f}$$

$$L_1 = 12.3\ \mu\text{h}; \quad F_+ - F_- = 22{,}400 \text{ cycles}.$$

Circuit efficiency at $F_0=84$ per cent.
Circuit efficiency at $F_\pm=80$ per cent.

The impedance and power factor computed in detail for the above example are shown in Fig. 12, and this shows how the circuits would react upon a tube over a range of applied grid voltages. The remainder of the computation of how the entire system would behave over a

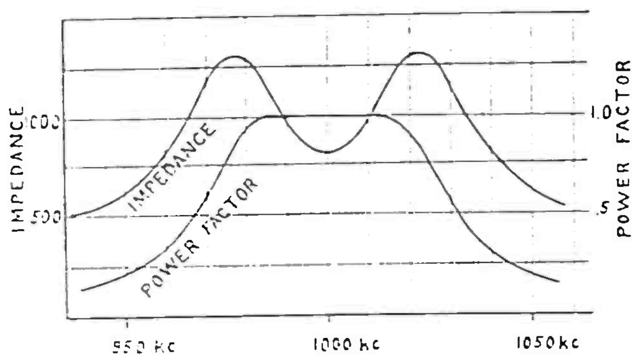

Fig. 12—Impedance and power-factor loading by coupled circuit.

frequency range would involve knowing the tube regulation characteristics as to plate—filament voltages for all impedances and power-factor combinations presented by the tank circuit, and a computation of secondary currents from the frequency and voltage values across the primary condenser. Such a computation, however, has already been stated to be beyond the scope of the present article, which deals fundamentally with the properties of coupled circuits only.

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THE ESTABLISHMENT OF THE JAPANESE RADIO-FREQUENCY STANDARD*

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Summary—*This paper describes the standard frequency equipment established at this laboratory. A valve-maintained tuning fork of Elinvar steel, made by H. W. Sullivan Co., Ltd., London, has been taken as the working standard. Some of the important characteristics of its frequency variation have been measured and adjustments have been made in order to minimize the variations which have been observed. The absolute value of its frequency has been measured through a long series of tests and the following value obtained as the mean frequency:*

999.770 cycles per sec. at 36.0 deg. Cent.

The precision of measurement is estimated as being well within one part in one million, but, because of the effect of atmospheric pressure, the constancy of maintenance is a few parts in one million. Two stages of multivibrators multiply the standard frequency by one thousand and the calibration can be made with an accuracy of one part in several hundred thousand. In order to determine the constancy of maintenance of the standard, a piezo-electric oscillator has been calibrated from time to time over a period of five months and the results have been satisfactory.

Introduction

THE PRECISE determination of high frequencies has become a most important technical problem. Especially in radio communication, the growing demand for radio channels will necessitate a more precise allocation of frequencies for different services and a more accurate maintenance of the assigned frequencies. To secure a high-grade standard of radio frequencies is, accordingly, an essential requirement.

The work of establishing accurate frequency standards has been in progress in Japan since 1918. First, a standard consisting of inductance and capacity was established by comparing it with either rotating machines or Lecher wires. Subsequently, on account of the need for frequency comparisons between the Japanese Army, Navy, and Ministry of Communications, and in line with the recommendations of the International Union of Scientific Radiotelegraphy (U. R. S. I.) the fundamental frequency standards were reestablished.

The ideal to be sought in working on this problem is a vibrator whose frequency can be indefinitely maintained constant to a very high

* Dewey decimal classification: R210. This is based on a paper on the same subject presented before the World Engineering Congress in Japan in November, 1929.

degree. Many kinds of vibrators, either mechanical, electrical, or electromechanical, can be suggested, but perfect constancy in the generated frequency cannot be expected in any of them.

Among many kinds of vibrators, the following general classes may be mentioned;

(1) mechanical; as a pendulum of a clock,

(2) electrical; as a special kind of a valve oscillator,

(3) electro-mechanical; as a valve-maintained tuning fork, a magnetostriction oscillator, or a piezo-electric oscillator.

Each of these classes has its own distinguishing features, and their frequency characteristics have been described more or less satisfactorily by many workers. The proper choice of the method having the greatest constancy and ease of maintenance, however, calls for further investigation.

Assuming that a suitable frequency standard has been set up, the next problem is to obtain an adequate number of separate frequencies, spread over a sufficiently wide frequency band. This object can be attained by the use of a frequency multiplying system. There are several possible methods, among which the harmonic amplifier and the controlled distorted-wave generator are to be preferred.

The system which has been adopted in this laboratory consists of a valve-maintained tuning fork and a controlled multivibrator. The object of this paper is to describe the equipment used for maintaining the standard frequency.

Standard Frequency Equipment

As a source of the standard frequency, a valve-maintained tuning fork, thermostatically controlled, has been chosen on the basis of both constancy and ease of the frequency measurement. The properties of this vibrator have been thoroughly studied, experimentally and theoretically by many investigators.[1] The most important effect upon the frequency is that of the variation of temperature. The ordinary steel fork is not sufficient, because the temperature variation in a thermostatic chamber will be, at the least, of the order of one-tenth of a degree Centigrade, and the constancy of the fork frequency can only be attained to the order of one part in one hundred thousand, as the temperature coefficient of the ordinary fork is about one part in ten thousand per degree Centigrade. A tuning fork made of Elinvar steel, invented by Guillaum,[2] has, therefore, been adopted as the standard.

[1] D. W. Dye, *Roy. Soc.*, Series A, 103, 240; 1923. W. H. Eccles, *Phys. Soc.*, 31, 260; 1919. S. Butterworth, *Phys. Soc.*, 32, 345; 1920. T. G. Hodgkinson, *Phys. Soc.*, 38, 24; 1925.

[2] Ch. Ed. Guillaum, *Compt. Rend.*, 170, 1433; 1920.

A fork made by H. W. Sullivan Co., Ltd., London, was used, and its frequency was adjusted, as closely as possible, to the value of 1 kc.

The method of supporting the fork is a factor which may give rise to a variation in frequency, on account of the mechanical coupling between the fork and the surrounding bodies, and, therefore, the fork was mounted firmly and permanently on a suitable cushion to reduce the effect.

The capacity of a shunting condenser connected across the anode coil of the exciting circuit of the fork was adjusted so as to give a minimum variation of the fork vibration. Fig. 1 shows the amplitude of vibration for various values of this capacity. Though there remains the problem of compensating for the frequency variation due to the

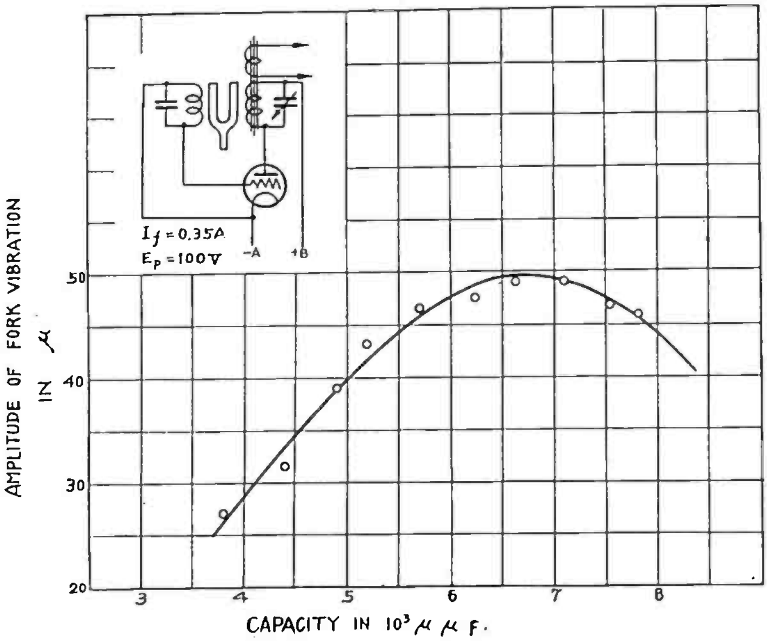

Fig. 1

unsteadiness of the electric power sources, as was pointed out by Dye, the order of this variation is very small and it is an easy matter to keep it well within the accuracy of measurement. The variation actually resulting from varying the plate voltage and the filament current was -2.1×10^{-7} for plate voltage and $+8.0 \times 10^{-7}$ for filament current, with one per cent change of these two sources of power.

The value of the output from the fork affects the value of its frequency, if the output coil is fairly tightly coupled to the exciting circuit, since the back electromotive force due to the output current causes a change in the phase angle of the exciting current. Particularly, the d-c polarization of the anode coil due to an external load was found to give a change to the electro-mechanical constant of the exciting circuit. Accordingly, the output coil has been shunted by a fixed high resistance, and suitable power is drawn from the output side of an amplifier to remove the mutual action. This method was recently adopted by Clapp.[3] The circuit is shown in Fig. 2.

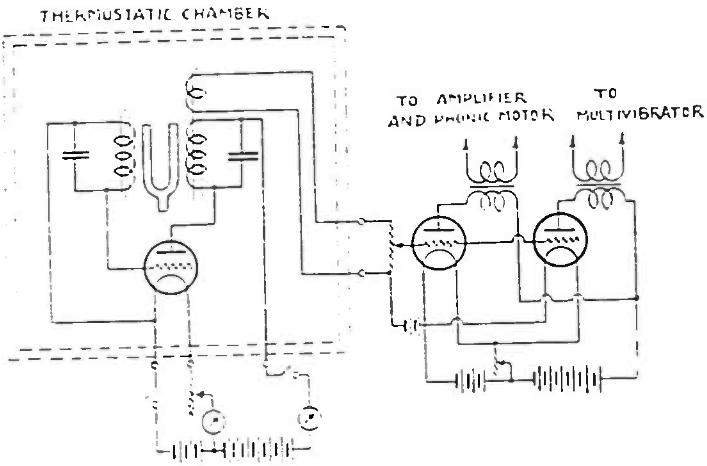

Fig. 2

As pointed out above, an Elinvar fork has a very low temperature coefficient, of the order of 1×10^{-5}, but still the temperature effect is the factor of greatest importance. A thermostatic chamber has, therefore, been provided. A schematic diagram of this chamber is

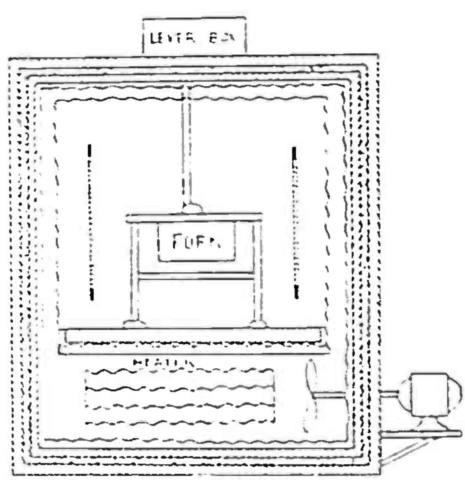

Fig. 3

shown in Fig. 3. Three layers of cork and silicate cotton insulate the chamber from the outside, and double distributing layers of heavy corrugated copper give space for the circulation of air. The regulator includes a bimettalic element, which consists of a thin aluminium tube

[3] J. K. Clapp and L. M. Hull, Proc. I.R.E., 17, 252; February, 1929.

and an Invar rod, with a high-ratio lever and a tungsten-platinum contact. A valve relay is used to reduce the sparking at the contract. The regulation of temperature is quite satisfactory and well within 0.1 deg. Cent. The schematic-circuit diagram is shown in Fig. 4.

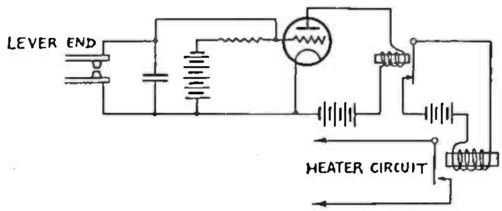

Fig. 4

The Determination of Fork Frequency

The determination of the absolute value of frequency is another difficult problem to be solved after the constant frequency source has been established. There are several methods which have been tried by many authorities, as discussed in the paper of Krein and Rouse.[4] The most accurate methods may be conveniently considered as falling in one of the following two classes:

(1) integrating method,

(2) method of comparison with a rotating machine.

The first method is preferable, because it requires no intermediate vibrator, and the measurement can be most directly conducted. The second method, on the other hand, utilizes the frequency of an alternator whose rotation is controlled by another vibrator, such as in the method employed by How and Wenner,[5] and, in addition, the process of time measurement is to be repeated twice. Consequently, the precision of the measurement will be reduced.

A phonic motor of Rayleigh and La Cour was adopted, and as the frequency of the motor exactly synchronizes with an introduced frequency, it functions as a frequency-counting mechanism. The motor which is introduced is made by the General Radio Co., Cambridge, Mass., and has been provided with a commutator which makes a mark every 12 seconds on the tape when a frequency of exactly 1000 cycles per second is used.

The accuracy of frequency measurement depends chiefly upon the accuracy of the time-measuring devices and also the length of time intervals taken in the measurement. It should be pointed out that the values, thus obtained, always give the mean value of frequency during the time interval.

[4] E. Krein and G. F. Rouse, *Opt. Soc.*, 14, 263; 1927.
[5] R. H. How and F. Wenner, *Phys. Rev.*, 24, 535; 1907.

The clock maintained at this laboratory is that of Clemens Riefler, Class B, No. 468, and its daily rates are measured every day by comparing its time with the standard time signals sent out at 11 A.M. from the Astronomical Observatory at Mitaka, Tokio district.

The absolute measurements were conducted, at first, over comparatively short time intervals, i.e., 1 hour (3600 seconds) and afterwards over intervals of 24 hours (86,400 seconds). In the latter case, the accuracy of the clock could be neglected from the measurement and the time was compared directly with the standard time signals, which

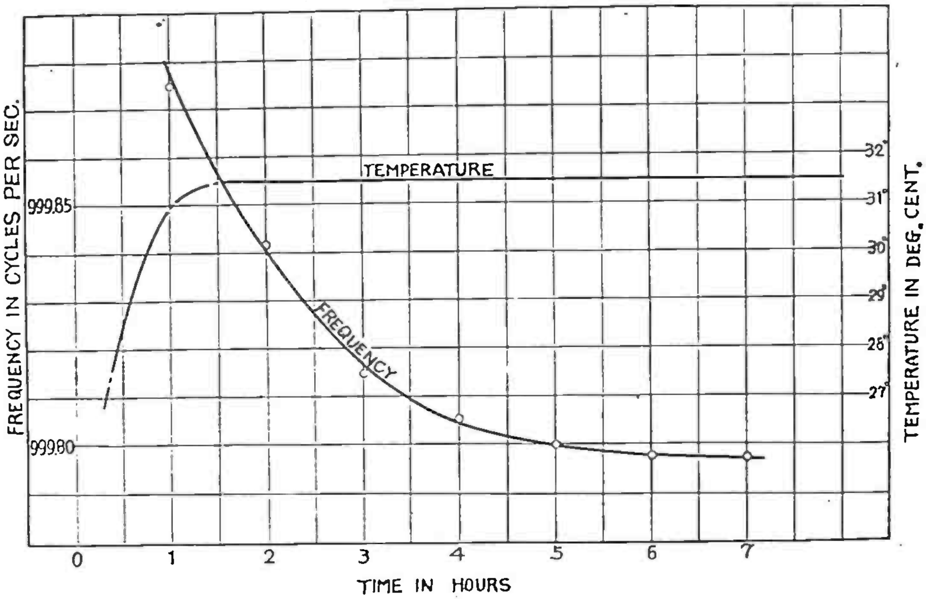

Fig. 5

are given according to the mean solar day. The probable error of time measurement in the course of the frequency measurement is estimated as ± 0.002 second. As a final result, the following has been obtained as the value of the mean frequency:

999.770 cycles per second at 36.0 deg. Cent.

Since the greatest effect on the frequency variation of the fork is the temperature, it is of prime importance to determine its temperature coefficient. A long series of measurements was conducted in the range of temperature between 25 deg. Cent. and 38 deg. Cent. It has been shown that the second-order term is practically negligible over the above temperature range and the relation is perfectly linear within the accuracy concerned, i.e.,

$$\text{Coeff.} = 0.000018.$$

It should be pointed out here that, as the fork is of heavy construction, the thermal capacity is so great that a remarkable lag in the temrature rise resulted. In Fig. 5 is shown an example of the variation the frequency of the fork in a case where the fork was started at the om temperature of 24 deg. Cent. and raised to a temperature tween 31 deg. and 32 deg. Cent. It may be seen in the figure that e temperature of bath reached a constant value in one hour and a alf, while the frequency of the fork reached a constant value only ter five or six hours. This may be one of the practical disadvantages the use of the fork as the standard.

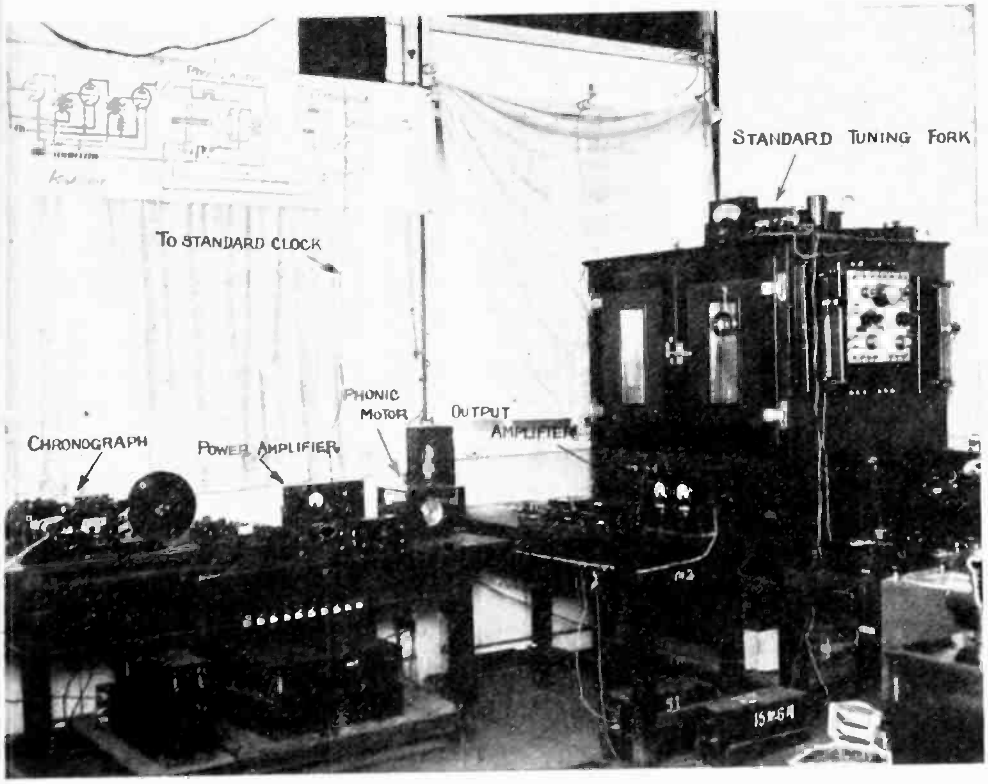

Fig. 6

The effect of atmospheric pressure ought to be considered because fork produces an appreciable amount of acoustic energy when it is ibrating, and the degree of air-loading will vary the damping of the ork. As a result of an experiment with a pressure chamber, the fracional change of frequency has been observed to be approximately -3×10^{-7} per millimeter of mercury.

Fig. 6 is from a photograph of the general arrangement of the tandard fork and its frequency-measuring devices.

Frequency-Multiplying Equipment

Since the frequency standard has been established as a single requency, it is necessary to multiply this frequency so as to cover the required frequency band. To accomplish this, several methods have been tried by many workers. From the standpoint of accuracy and ease of manipulation, the methods can be divided into two classes as explained above.[6,7] The multivibrator system, as used by Dye, has been adopted in this laboratory, because the system can produce harmonics of high order and of uniform intensity, and in addition, the

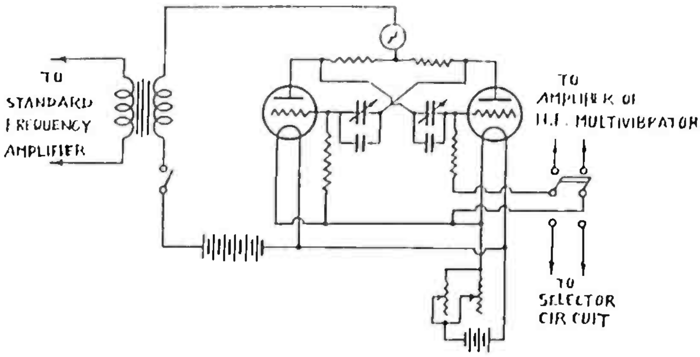

Fig. 7

sub-multiplication of frequency is possible, thus widening the frequency range of the standard equipment.

The first stage of the multivibrator has a fundamental which can be varied throughout a given range in the vicinity of 1000 cycles per second and is controlled by its fundamental. Harmonics up to the 100th can be easily selected. To increase the separation between the

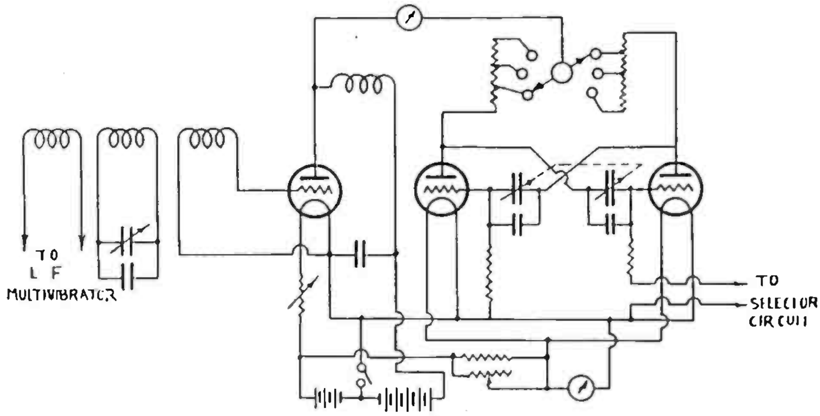

Fig. 8

neighboring harmonics of high order, a second stage of the multivibrator has been provided whose fundamental can be chosen to correspond with the 19th, 20th, or 21st harmonic of the original stage. A change in the harmonic frequencies can be secured by varying the plate resistance step by step and a further minor adjustment can be

[6] J. W. Horton, N. H. Ricker, and W. A. Marrison, *Jour. A. I. E. E.*, 42, 730; 1923 C. B. Jolliffe and G. Hazen, *B. O. S. Scic.*, 530; 1926.

[7] D. W. Dye, *Phil. Trans. Roy. Soc.*, 224, 259; 1924. H. Abraham and E. Bloch, *Ann. de Phys.*, 12, 237; 1919.

performed by the variable condenser inserted in the grid circuit. This latter adjustment is a good one by which to keep the intensity of the successive frequencies uniform. It was found that the effect of humidity on the operation of the multivibrator is so great that it sometimes entirely destroys the condition of synchronism. In Figs. 7 and 8 are shown the circuit diagrams of the low-frequency and high-frequency multivibrators, respectively.

The order of the harmonics can be determined by the pre-calibrated scale of the selector circuit when they are in beat with the auxiliary oscillator. The characteristic tone heard in a telephone receiver

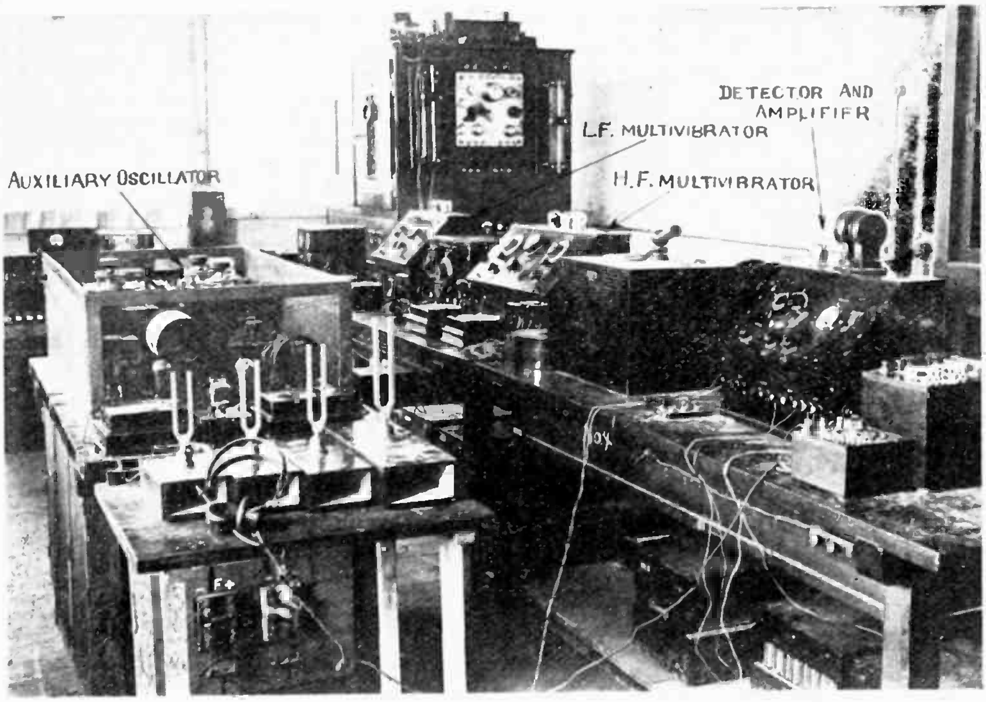

Fig. 9

clearly indicates the synchronizing point of the selected harmonic. The standard frequency can be multiplied by the whole system up to 1,000 times. Extension to higher frequencies up to 10,000 kc can easily be secured by the harmonic relation between the frequencies produced and those of the auxiliary oscillator without lowering the accuracy of calibration. Fig. 9 is from a photograph of the general arrangement of the frequency-multiplying system and the equipment for frequency calibration.

Method of Calibration

The aural method of calibration has been preferred, a heterodyning valve oscillator of high stability being used, because it will take the least amount of power from the standard set, and by the principle of double beating, any subsidiary harmonic can be obtained within a kilocycle band, which considerably increases the precision of calibration.

The utmost care should be taken in the construction of this auxiliary oscillator in order to minimize its frequency variation, but the effect of variation of the oscillator frequency can be excluded from the result by making the time of measurement as short as possible, for the oscillator is used only for substitution.

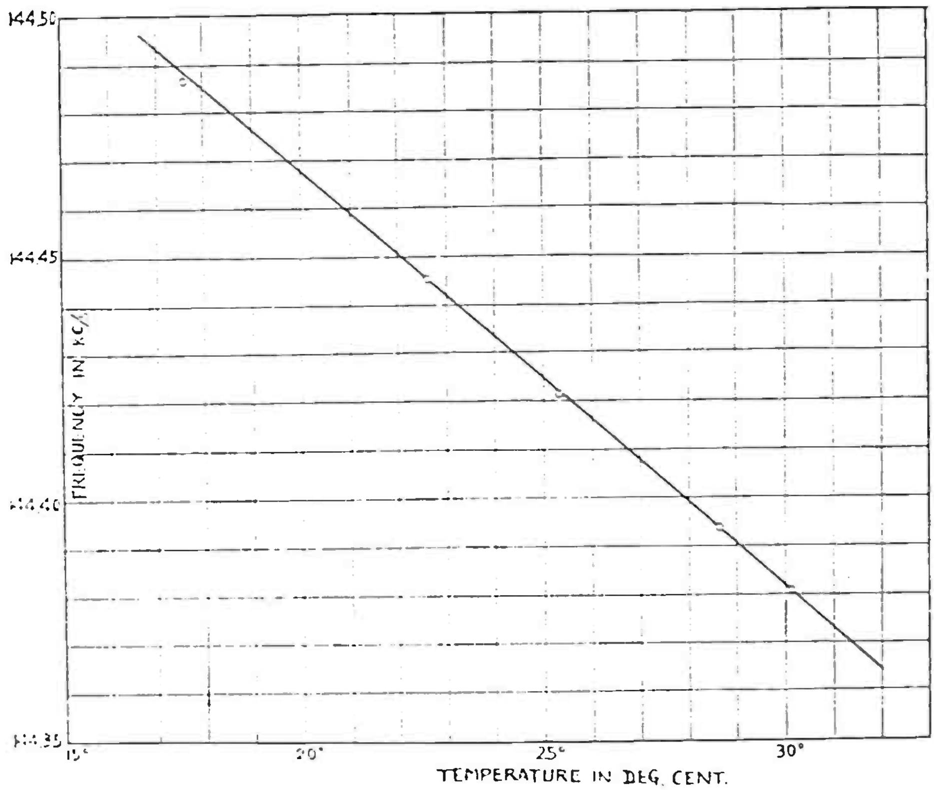

Fig. 10

The precision of calibration depends largely upon the percentage variation of the oscillator frequency. A bank of condensers in the oscillation circuit is combined in series or parallel, and by the aid of a vernier condenser, precision of reading of a few parts in one million can be easily attained. The condensers used should be of a precision type of high-grade material and workmanship.

For the purpose of checking the overall constancy and the accuracy of calibration of the standard set, a series of frequency measurements has been carried out on a quartz oscillator. It has been circulated for other purposes, among other similar laboratories, since August, 1928.

The quartz oscillator is not of a special kind and the temperature of the quartz plate is not thermostatically controlled. The quartz plate is of the circular Curie cut and the frequencies due to diameter and thickness are approximately adjusted to the values 144 kc and 1466 kc, respectively.

The following table shows the results of the experiments carried out over a period of five months in our laboratory.

TABLE I

MEASURED FREQUENCIES OF A QUARTZ PIEZO-ELECTRIC OSCILLATOR

	Quartz, 144 kc		Quartz, 1466 kc	
Date	Temperature in deg. Cent.	Frequency in cycles per sec.	Temperature in deg. Cent.	Freqcuency in cycles per sec.
August, 1928	25.3	144.422	28.7	1466.66
	28.6	144.394		
	30.1	144.381		
October, 1928	22.6	144.446	24.6	1466.78
			23.0	1466.81
November, 1928	17.6	144.486	18.2	1466.97

Curves showing the values of frequency of this quartz plate over a range of temperature are shown in Figs. 9 and 10. These results are fairly good illustrations of the constancy of maintenance of the standard

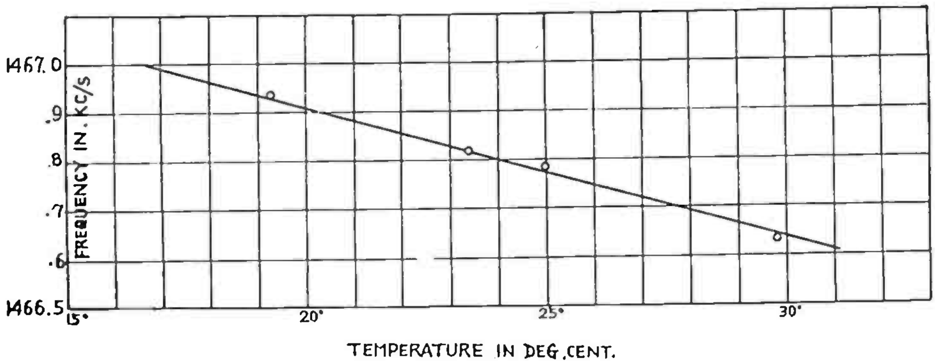

Fig. 11

and the precision of calibration, and they also show that the constancy of the generated frequencies of the quartz oscillator is equally excellent. Furthermore, it can be seen that the frequency variation is quite in a linear relation with temperature within a few parts in one hundred thousand. The following temperature coefficients were found:

Coefficient of oscillation due to Xx excitation -2.6×10^{-5}

Coefficient of oscillation due to Yy excitation -8.0×10^{-5}.

In conclusion it may be stated that the standard equipment described above is only the beginning, and the measurements of the standard frequency and various improvements of the equipment are still to be undertaken. Acknowledgment is due to R. Yoneda who gave the author valuable advice, and to S. Namba, S. Ishikawa, and S. Kanzaki of this laboratory for cooperation in the long and laborious work.

Proceedings of the Institute of Radio Engineers
Volume 18, Number 6 *June, 1930*

THE AMPLIFICATION AND DETECTION OF ULTRA-SHORT ELECTRIC WAVES*

By
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Summary—*In this paper, experimental results and theoretical considerations are discussed which concern the amplification and detection of electric waves shorter than one meter in length, in a system wherein diodes and triodes are connected so as to produce oscillations of the Barkhausen and Kurz type. A simple theory of the electronic effect in the detector action is given. In some of the circuits use is made of a magnetic field applied in the direction of the axis of the electrodes within the tube.*

I. INTRODUCTION

IN ANOTHER communication[1] the author has pointed out that the ultra-short electric waves, (below one meter), could be amplified and detected simultaneously with diodes and triodes which are connected so as to produce the oscillations of Barkhausen and Kurz (or Gill and Morrell type). For convenience's sake, the author will call the receiving circuit of this kind "Electronic Ampli-Detector."

Assistant Professor Uda of Tohoku Imperial University, Japan, succeeded recently in telegraphic communications through the distance of 30 km and telephonic communications through the distance of 10 km with a modulated wave of 45-cm wavelength, due to the development of the "electronic ampli-detector."

The vacuum-tube detector used by Barkhausen and Kurz seems to correspond to one of these "electronic ampli-detectors" although no amplifying action was mentioned.[2] The vacuum-tube detectors used by Hollmann seem to be quite different in principle from the present ones.[3]

It may be noted that most of the present receiving circuits are comparatively easy to adjust. This may be supposed from the theoretical point of view.

II. "ELECTRONIC AMPLI-DETECTORS" OF VARIOUS KINDS

Most circuits which will produce oscillations in this range of wavelengths can now be used as "electronic ampli-detectors." We can

* Dewey decimal classification: R343. The present research was carried out at the Research Laboratory of the Tokyo Electric Company, Kawasaki, Japan.
[1] *Jour. I. E. E.* (Japan), January, 1929.
[2] *Phys. Zeits.*, 21; 1920.
[3] *E. N. T.* 7, Bd. 5; 1928.

receive the waves successfully with the various circuits, such as shown in Figs. 1, 2, 3, 4, 5, and 6, with which the oscillations of the Barkhausen and Kurz type can be obtained. Satisfactory reception of waves can be obtained while the oscillations are present.

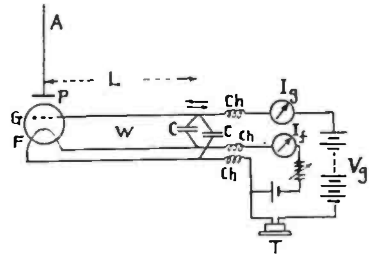

Fig. 1—A-circuit

In the figures, P, G, and F respectively represent the anode, the grid, and the cathode of vacuum tube. In the case of Fig. 1, P is placed at the outside of the vacuum tube, so that it becomes a diode. A, W, C, Ch, T, and M respectively represent receiving antenna,

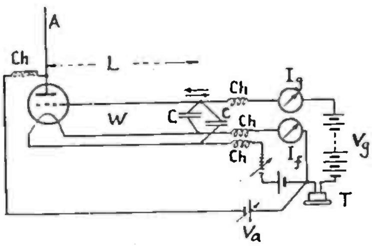

Fig. 2—B-circuit

parallel wires, short-circuit condenser, choke coil, telephone receiver, and magnetic field coil. V_g, I_g, V_a, I_f, and R_t denote the grid source voltage, the grid current, the anode voltage, the anode current, the filament current, and the d-c resistance of T respectively, while L

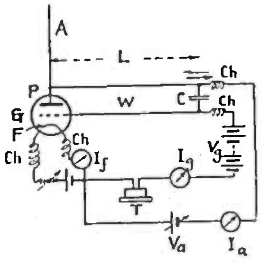

Fig. 3—C-circuit

denotes the distance between the center of the electrodes of vacuum-tube and the short-circuit condenser.

In the cases of Figs. 1, 2, 5, and 6, three parallel wires have been used. In all cases, the telephone receiver T (4000 Ohms) has been connected to the grid circuit, although similar results have been obtained when it has been connected to the anode circuit. Of course an audio-frequency amplifier must be used in practical communications.

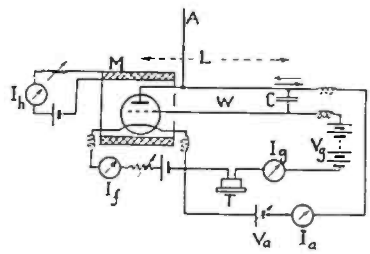

Fig. 4—D-circuit

B-circuit (Fig. 2) represents an improvement of the A-circuit (Fig. 1). The other circuits are quite the same except for the point of the application of the magnetic field.

III. Earlier Experiments

Experiments regarding the sensitiveness of the "electronic amplidetectors" and those tending to prove the existence of an amplifying

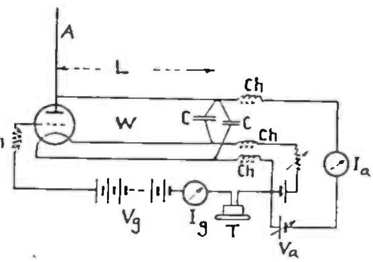

Fig. 5—E-circuit

action had often been successfully carried out in the earlier course of the present research.

The author could receive the modulated ultra-short waves very satisfactorily at a distance many times that of a Hertz resonator-

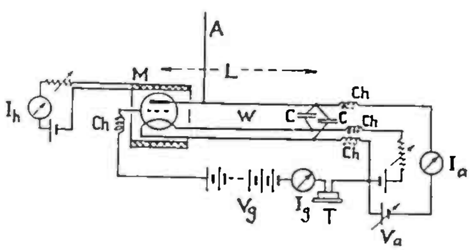

Fig. 6—F-circuit

detector (a crystal detector and two wire arms). The existence of an amplifying action was proved experimentally with the circuit by connecting a Hertz resonator instead of the receiving antenna. The signal intensity in the telephone receiver, which was connected to the terminals of the crystal detector of the above resonator, was increased abruptly by gradually increasing the amount of filament current and it was decreased suddenly by further increasing the same current. During this experiment the grid voltage and the position of

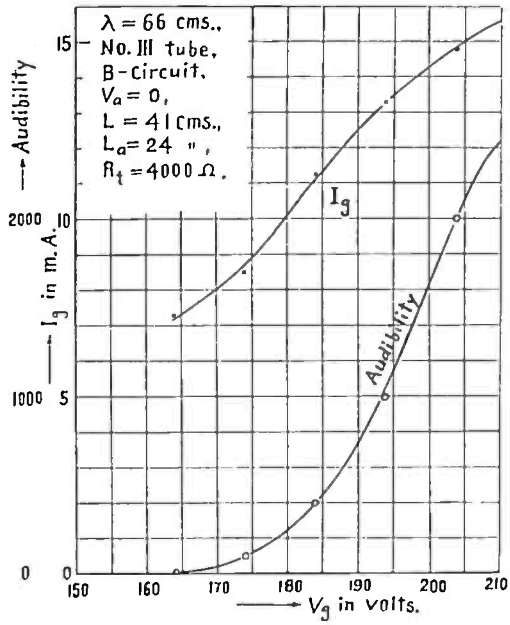

Fig. 7

the short-circuit condensers, which had been decided carefully, were kept constant. The author finds no reasonable way to explain this result other than attributing it to the existence of an amplifying action of this particular character. Similar experiments with other circuits could be easily accomplished. In general these receiving circuits were easily adjusted.

IV. Further Experiments

Further experiments have been carried out with vacuum tubes shown in Table I.

TABLE I

No.	Diameter of Anode	Diameter of Grid	Name
I	0.75 cm	0.25 cm	UF-101 (commercial name,
II	0.9 "	0.4 "	UE-102 "
III	0.75 "	0.25 "	US-101 (laboratory name)

The No. III tubes have been specially manufactured so as to fit the B-circuit, and the anode-lead was located at the side opposite the other leads.

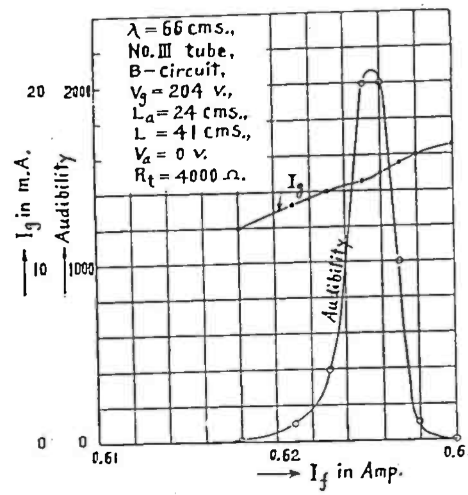

Fig. 8

All of the following experiments, were made by receiving a modulated wave of 66 cm. This wavelength is denoted by λ in the accompanying diagrams.

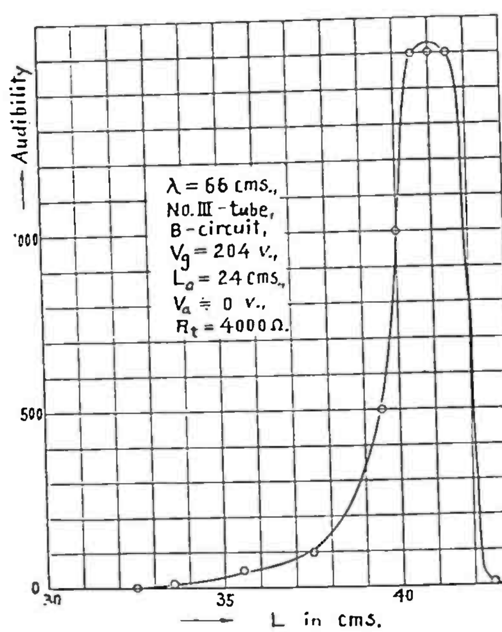

Fig. 9

Next, the experimental results obtained with various circuits will be given.

A-circuit, as shown in Fig. 1, may be of interest, as we can amplify these ultra-short wavelengths with diodes, but it has been found that the B-circuit is superior in the points of sensitiveness and ease of adjustment.

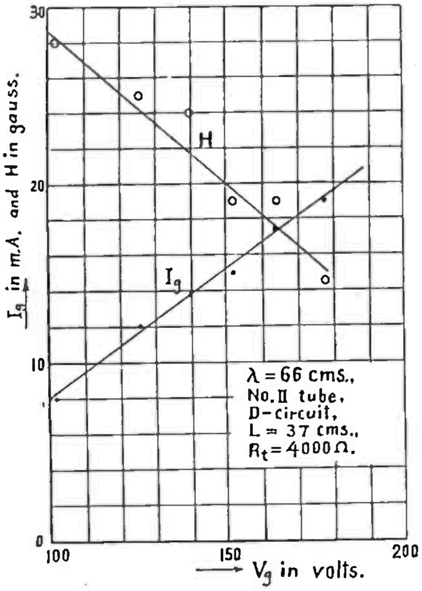

Fig. 10

The author could obtain very satisfactory results with B-circuit Fig. 2, with respect to adjustment, sensitiveness, selectivity and stability.

Fig. 7 shows the relation between V_g and I_g with which the best reception could be obtained. The corresponding audibility curve is

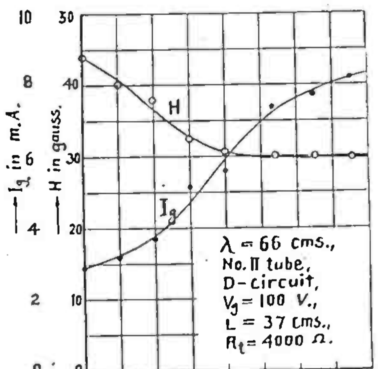

Fig. 11

also shown in the same figure. In this case, V_a, L, and L_a (the length of the receiving antenna) are kept constant at the values shown in the figure. The lengths of L and L_a are to be selected properly. Fig. 8

shows the change of audibility when only I_f varies, together with the change of I_g, which is naturally caused by the change of I_f.

Fig. 9 shows the change of audibility when only L varies. Similar results can be obtained when L_a varies instead of L.

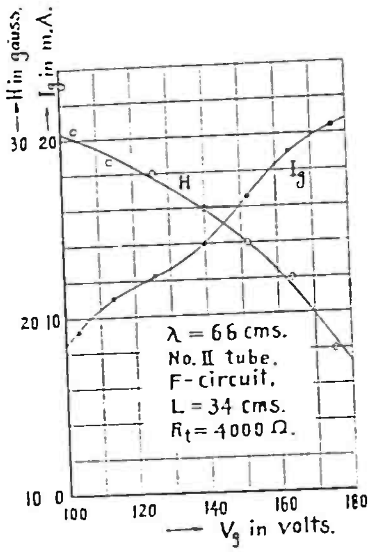

Fig. 12

C-circuit in Fig. 3 is also excellent. The author has recognized in most cases that better reception of waves can be obtained by applying a proper magnetic field in parallel to the electrode axis. As seen in Figs. 3 and 4, D-circuit utilizes this fact.

We could obtain very satisfactory results with D-circuit. Fig. 10 shows the change of optimum values of I_g and H, resulting from a change of V_g. In this case L is kept constant at 37 cm while V_a was so adjusted at each measurement as to afford best results.

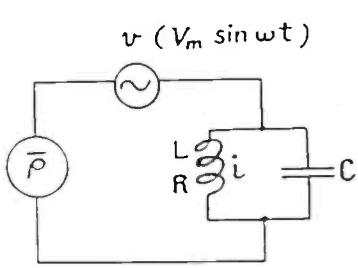

Fig. 13

Fig. 11 shows the change of optimum values of I_g and H, required by a change of V_a. In this case V_g and L are kept constant through the measurement.

E-circuit, as shown in Fig. 5, gave results which were not so satisfactory. The modified circuit which utilizes a magnetic field is shown in Fig. 6. F-circuit in Fig. 12 shows the change of optimum values of I_g and H when V_a varies in this circuit. In this case L is kept constant at 34 cm while V_a is adjusted as to give the best results.

Many things remain yet to be studied regarding the E- and F-circuits. We could obtain satisfactory results with No. 1 tube in various circuits.

V. A Simple Theory Regarding Amplifying Action

In all cases of "electronic ampli-detectors," the existence of a negative resistance is very probable.[4] Then for simplicity's sake let us admit the following circuit, (Fig. 13), instead of considering the actual circuit. In the figure, $\bar{\rho}$ denotes the negative resistance, L, R, and C denote the inductance, the resistance, and the capacity respectively, V and i denote the impressed e.m.f., (which corresponds to that induced by incoming waves), and the current respectively.

Then the following equation can be obtained.

$$\frac{d^2i}{dt^2} + \left(\frac{R}{L} + \frac{1}{C\bar{\rho}}\right)\frac{di}{dt} + \frac{1}{LC}\left(1 + \frac{R}{\bar{\rho}}\right)i + \frac{Q}{LC} = \frac{-V_m \sin \omega t}{LC\bar{\rho}} \tag{1}$$

where Q is a constant and can be determined by assuming the negative resistance characteristic.

When $Q=0$, (1) is solved as follows.

$$\begin{aligned} i = {} & e^{-\frac{A}{2}t}\left\{C_1 \cos \omega' t + C_2 \sin \omega' t\right\} \\ & + \frac{D}{\sqrt{A^2\omega^2 + (B - \omega^2)^2}}\left\{\frac{(B - \omega^2)\sin \omega t}{\sqrt{A^2\omega^2 + (B - \omega^2)^2}} - \frac{A\omega \cos \omega t}{\sqrt{A^2\omega^2 + (B - \omega^2)^2}}\right\} \end{aligned} \tag{2}$$

where

$$A = \frac{R}{L} + \frac{1}{C\bar{\rho}}\,; \quad B = \frac{1}{LC}\left(1 + \frac{R}{\bar{\rho}}\right) \quad D = \frac{-V_m}{LC\bar{\rho}}\,;$$

$$\omega' = \frac{\sqrt{(4B - A^2)}}{2}; \quad C_1 \text{ and } C_2 = \text{constants}.$$

By putting the condition, which makes i infinity, into (2), we obtain

$$i = C_3 \sin(\omega' t + C_4) + KV_m \sin \omega t \tag{3}$$

[4] Tonks, *Phys. Rev.*, October, 1929.

where

$$\left.\begin{aligned} \omega' &= \omega \\ K &= \infty \end{aligned}\right\} \tag{4}$$

C_3 and C_4 are constants.

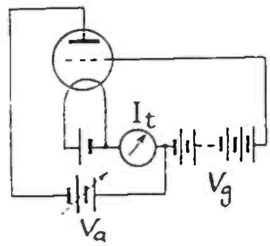

Fig. 14

Equations (3) and (4) give the following results:

(a) the best reception may be obtained, when ω and ω' are same, i.e., the frequency of the incoming wave and that of the self-sustained oscillation are equal:

(b) in practical cases a slight difference between ω and ω' is not serious, as K cannot be infinity at all;

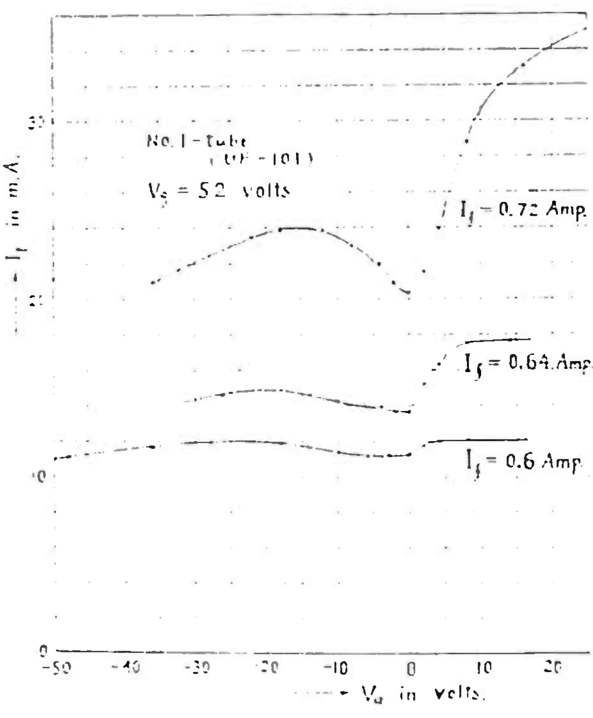

Fig. 15

(c) the self oscillation has no direct effect on the amplifying action but it may arise generally whenever any satisfactory reception of waves is obtained.

These results are quite consistent with the experiments, except in a ew cases.

VI. An Electronic Effect on the Detecting Action

Fig. 14 shows the electrical connection with which the result shown n Fig. 15 has been obtained. As seen from the figure, the total electron urrent I_t decreases abnormally at the neighborhood of V_a (anode voltage) $=0$. During the experiment the grid voltage V_g has been kept constant.

These characteristics can be expected by considering the increase of space charge caused by the returning electrons in the neighborhood of the cathode.

In special cases this phenomenon may give some effect to the detecting action of the present circuits. But, in most of the cases, the detecting action may be illustrated in the usual way, ignoring this phenomenon in the considerations.

VII. Conclusion

In the foregoing, the author has given a general idea regarding the "electronic ampli-detectors." There have been left, of course, many things to be studied further both from theoretical and experimental points of view, especially with respect to a theoretical consideration of amplifying action, since the author has only considered a very simple case differing from the actual circuits in some degree.

The author wishes to express his thanks to Professor Yagi and to Professor Nukiyama of Tohoku Imperial University for their valuable instructions and suggestions which were given him at the beginning of the present research.

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THE RADIO ENGINEER AND THE LAW*

By
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Summary—*This paper presents the difficulties inherent in the development of a body of law for radio. It points out that that body of law must be based upon sound scientific principles as presented by radio engineers. The radio engineer as an expert witness may be able to avoid the difficulties encountered by experts of other classes and take advantage of important opportunities for the education of courts and commissions. As legislative advisers the members of the engineering profession have tasks of great responsibility. Radio engineers can bring about a healthy growth of law through cooperative efforts and group consciousness.*

IN ORDER that the radio industries may properly grow and make their full contribution to civilization, they require a body of law which is consistent with that growth. The radio engineering profession must supply the impetus, the principles, and the leadership for its development. This is necessarily true because radio is essentially a scientific, not a legal enterprise. Rules of law enter into it only for the purpose of insuring stability and either preventing or adjudicating conflicting claims of right.

The dependence of legal development upon engineering principles is well shown by the inclusion in the latest report[1] of the standing committee on radio law of the American Bar Association. The radio engineering profession can lose sight neither of this relationship nor of its consequences.

II

Most important of the engineer's contacts with law is that arising from actually litigated disputes. Whether the decisions in these disputes are to be correct and far-sighted will depend in large measure upon the practitioners of radio engineering.

The rôle of the radio engineer both as a consultant and as an expert witness in litigated disputes is an important one. Innumerable cases involving patents on radio apparatus, conflicting rights in radio communications, claims to station licenses, the soundness of station allocations, the organization of radio communication systems, are already before the courts and the Federal Radio Commission.

* Dewey decimal classification: R002.
[1] Report of the Standing Committee on Radio Law, presented at the meet-*casting Station, Incorporated*, a corporation, *Coyne Electrical School, Inc.*, a*I* corporation, and *J. Louis Guyon*, defendants, Circuit Court of Cook County,a S-34.

The famous decision of Judge Francis S. Wilson in the Tribune case[2] was based upon a consideration of principles of radio engineering presented to the Court by expert testimony. Judge Wilson, in his opinion, specifically passed on the question of the frequency separation necessary between the contesting broadcast stations to prevent undue interference.

In the leading case of *United States vs. American Bond and Mortgage Company*,[3] wherein Judge Wilkerson upheld the validity of the Radio Act of 1927, engineering principles are discussed. Heterodyne and cross-talk interference, "blanketing," service area, and "nuisance area" are all dealt with.

In *General Electric Company vs. Federal Radio Commission*,[4] the Court of Appeals of the District of Columbia also passed upon engineering considerations, and the opinion balances, as against other considerations of public interest, the interference existing between KGO, 10 kw., Oakland, California, and WGY, 50 kw, Schenectady, New York, both at a frequency of 790 kc.

These are but examples. As radio jurisprudence grows, such instances must increase indefinitely.

The determinations and decisions of the Federal Radio Commission are based upon engineering considerations to a much greater extent. Engineering testimony is being offered in almost all hearings before the Commission and the Commission's decisions are being increasingly based upon testimony of that character.[5]

III

Judicial and popular reaction to the testimony of expert witnesses has not been entirely favorable.

In 1876, in the case of *Whitaker vs. Parker*,[6] involving expert testimony as to the signature on a promissory note, the Supreme Court of Iowa said:

> "The testimony of experts is of the lowest order and of the most unsatisfactory character . . . . . We believe that in this opinion experienced laymen unite with members of the legal profession."

[2] *The Tribune Company*, a corporation, complainant, *vs. Oak Leaves Broadcasting Station, Incorporated*, a corporation, *Coyne Electrical School, Inc.*, a corporation, and *J. Louis Guyon*, defendants, Circuit Court of Cook County, Illinois, number B-136864, November 17, 1926.

[3] In the District Court of the United States for the Northern District of Illinois, Eastern Division, March 1, 1929. 31 F. (2d) 448.

[4] February 25, 1929. 31F. (2d) 630.

[5] See particularly the Statement of Facts and Grounds for Decision of the Commission filed in the Court of Appeals of the District of Columbia, August 15, 1929, in the Intercity, Wireless, R. C. A., and Mackay cases.

[6] 42 Iowa 585, per Beck, J.

In the New York case of *People vs. Kemmler*,[7] involving the mental responsibility of a man charged with killing his paramour, the Court said:

> "The frequent spectacle of scientific experts differing in their opinions upon a case according to the side upon which retained, tends much to discredit such testimony, or to impair its force and usefulness . . . . ."

Handwriting and medical experts, who frequently appear as witnesses, familiarize themselves with the rules of law governing their qualifications and testimony. There is an extensive literature available for their study. If radio engineers are to be useful witnesses and if they are to exercise a beneficial influence upon the development of the laws under which they are to practice, they too will need to familiarize themselves with the rules.

It is, of course, true that in any particular litigation, an engineer without courtroom experience, will in preparation of the case consult with counsel, and be prepared to testify properly as to his qualifications and to state his conclusions effectively. But a familiarity with the rules governing expert testimony is desirable in advance of this.

Primarily an expert with this knowledge is more apt to be consulted. In cases where a conflict of expert opinion is to be presented, other factors being equal, success will more probably come to the litigant whose experts are best able to state their qualifications, to present their opinions, and to extend their reasons therefore. Moreover, in the preparation of a case, the work of the expert in the laboratory or in the field is without legal guidance and he should direct his experiments and work along lines which will be proper for statement at the time he gives his testimony. Also, once the expert takes the witness stand he is, to a large measure, cut off from legal advice. Counsel inquiring of him on direct examination are not permitted to ask leading questions. The witness must determine how fully he shall answer a question and it is well for him to know how fully he is permitted to answer it. And when the stage of cross-examination is reached, a witness's knowledge of the rules under which that cross-examination must be conducted is invaluable.

It is not the purpose of this paper to present a discussion of the applicable rules of evidence, but rather to point out the necessity for their study.

The preservation of judicial respect for radio engineers also requires that testimony be presented fairly and without partisanship, and above all that expert witnesses maintain high standards of scientific integrity in order that the bickering and recriminations of

[7] 1890. 119 N. Y. 580, per Gray, J.

expert witnesses in other fields may be completely avoided by radio experts from the outset.

It is perhaps unnecessary to point out that one of the greatest pitfalls for an expert witness lies in the temptation to cover the weak spots of a presentation by a false appearance of learned mystery. No principle or rule of the radio sciences is beyond explanation insofar as that explanation is necessary in judicial proceedings. It cannot, therefore, assist the presentation of a case to state considerations in metaphysical or obscure language. The efforts of one witness who recently testified on the general subject of "dead spots" in reception to gloss over his lack of familiarity with the problem by an involved distinction between the propagation of electromagnetic and electrostatic impulses is an example of this weakness. Plainly, neither his case nor his standing as an expert were improved by this legerdemain.

IV

An equally important function of the radio engineer is his assistance in the formulation of legislation. The report of the Bar Committee above referred to contains this language:[8]

> "Radio law is constantly open to the danger, because of incomplete or inaccurate information, that legislators will base their enactments on unsound scientific principles. The field is a fruitful one for plausible utterances, to check the truth of which the average legislator has neither the time nor trustworthy means. In the light of the best available information, obtained by the committee from radio engineers of recognized standing, the committee is led to the conclusion that misconceptions of radio facts and principles have already found their way into legislation now in force and also are inherent in certain bills introduced into the 70th Congress."

Much good work has already been done by the Institute of Radio Engineers and by its individual members in making sound knowledge available to legislative and quasi-legislative tribunals. Mention need only be made of the part played by the Institute in the development of General Order No. 40 of the Federal Radio Commission providing a skeleton for broadcast allocation, of the testimony of Hogan, Dellinger, and Jansky before the Commission in July of 1928, of the testimony of many engineers before the various committees of Congress, and of the excellent studies of Dellinger.[9]

But much remains to be done, as can be seen from the following examples.

Under the provisions of the fourth section of the Radio Act of 1927 as amended, the Federal Radio Commission has powers which may be

[8] Page 8.

[9] J. H. Dellinger, "Radio broadcasting regulation and legislation," Proc. I.R.E., 17, 2006–2010; November, 1929. "Analysis of broadcasting station allocation," Proc. I.R.E., 16, 1477–1485; November, 1928.

described as quasi-legislative, which is to say that within limits, Congress has given the Commission the power to make certain rules and regulations for the government of radio communication which regulations, when properly made, have the force and effect of law. Generally speaking, these regulations are to deal with the classification of radio stations, the service of the stations, the allocation of frequency bands, the kind of apparatus with respect to its effect, the prevention of interference, the establishment of service areas, chain broadcasting, station records and railroad rolling stock stations. Plainly most of the questions involved are scientific. It is equally plain however that these regulations must be framed and enforced as laws. A case for the greatest cooperation between the radio engineering profession and the Commission is at once made out. Sound regulations can be based only upon sound engineering.

In view of the recent action of Congress making the Commission a permanent administrative body, it at once becomes apparent that important regulations must be promulgated fixing standards for the measurement of station power, for the measurement and suppression of harmonic radiations, fixing standards of geographical separation, fixing standards for the approval or disapproval of automatic frequency-control apparatus, outlining a technical program for frequency-maintenance standards, establishing a system for stable and ascertainable reassignments as frequency separations narrow in the telegraphic service, and many similar matters which at once suggest themselves. It may be that many of these problems cannot be solved at once but it is equally true that they deserve the impartial attention of radio engineers to the end that the work of the American Standards Association, the Institute's Committee on Standardization, and the associations of radio manufacturers may be extended to the assistance of the regulatory bodies.

In order that this work may insure a sound body of law for radio, it must, at least at the outset, be disassociated from the representation of particular clients, the advancement of particular schools of thought and the preference of particular types of patented apparatus. The regulations must be impartial both in intention and effect, narrow enough to provide firm discipline and yet broad enough to permit wide development.

While Congress and its committees are of course without attached technical assistance, the Federal Radio Commission and the Department of Commerce have their own engineering staffs. The willingness of these staffs to encourage the assistance of the profession generally, appears from the following announcement made by the Commission on December 21, 1929:

"The Commission will hold an engineering conference on January 17, 1930, for the purpose of obtaining recommendations from engineers regarding broadcasting operation.

"The two main subjects on which information and discussion are desired are the following:

(a) The possibility of developing an antenna system which will increase the service range near the transmitter and decrease the interference at a distance, that is, an antenna system that will reduce the 'sky wave' and strengthen the 'ground wave.'

(b) Discussion of synchronism of broadcasting stations. Information on this subject desired as to reliability and advantages and disadvantages of the various methods of synchronism.

(c) Any other engineering principles to be discussed that would aid broadcasting stations to render better service.

"The conference will be entirely devoted to engineering considerations and it is desired that those interested send engineers, so far as they may be available, who will be qualified to discuss the above questions."

This statement indicates, of course, a wise realization that truly proper regulation of the radio communication industries must come from wide consultation and general consensus.

Many other industries under whole or partial government regulation have developed far without that consultation and consensus, the realization for their necessity coming at a late stage when fixation had occurred and a complete change of habits and points of view was difficult to obtain. The radio industries can avoid this error.

V

The principles of radiophysics and their practical application constitute such a body of learning with such far-reaching effects upon the national welfare that the practice of radio engineering has become characterized by a fine distinction. The professional service in the development of radio jurisprudence and the consequent professional obligation that that jurisprudence shall be sound must ever be borne in mind. As consultants, as moulders of the policies of great industrial organizations, as expert witnesses and legislative advisers, the radio engineers, with their group consciousness, can be of service as great and important as with their individual ingenuity in the laboratories.

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NOTE ON VARIATIONS IN THE AMPLIFICATION FACTOR OF TRIODES*

By

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Summary—*It is pointed out that variations in the amplification factor of triodes are due to different portions of the tube having different values of μ. The resulting μ variations over the operating range of the tube characteristic are very considerable in commercial tubes, and give numerous undesirable operating features.*

WHILE IT is generally understood that the amplification factor of three-element tubes is not absolutely constant, there is very little appreciation of the extent of the μ variations in ordinary commercial tubes. The situation shown in Fig. 2, in which μ varies between 10.4 and 8.5, or about 20 per cent, over the operating range, is representative of what may ordinarily be expected.

The variations that are observed in the amplification factor of ordinary tubes are caused by edge effects and other dissymmetries that result in different parts of the tube having different μ's. These different parts are all in parallel, so that the behavior of the equivalent μ of an ordinary tube can be considered as the result of a parallel combination of unlike tubes, the analysis of which is available.[1]

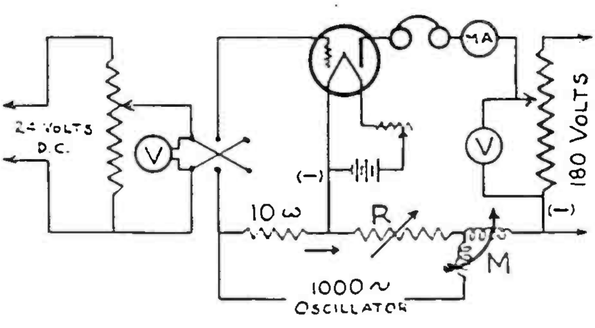

Fig. 1—Modified Miller bridge for measuring μ, in which a mutual inductance M is used to balance out grid-plate capacity current.

With tubes having an equipotential cathode, μ always decreases as the grid becomes more negative and as the plate voltage is reduced. This results from the fact that when such changes are made, the plate current cuts off first in those parts of the tube having the highest μ, leaving only the low μ portions of the tube in operation. This behavior is clearly shown in Fig. 2.

* Dewey decimal classification: R 131.

[1] R. V. L. Hartley, "Vacuum tube amplifiers in parallel," PROC. I.R.E., 9, 250; June, 1921.

With filament type cathodes the situation is complicated by the fact that different portions of the cathode not only have different μ's but also have different effective grid biases as a result of the voltage drop in the filament. The result is that although there is a general tendency for μ to decrease as the grid is made negative, this effect is sometimes masked by other tendencies. Thus some filament type tubes have a μ which increases as the grid becomes more negative, and in at least one case a tube was found in which as the grid was made negative μ first increased, then decreased, next increased, and finally decreased as cutoff of all plate current was approached.

One of the most important effects of μ variation is the distortion that it introduces in amplifiers. As a consequence of μ variation, an ordinary amplifier tube will not give distortionless amplification even

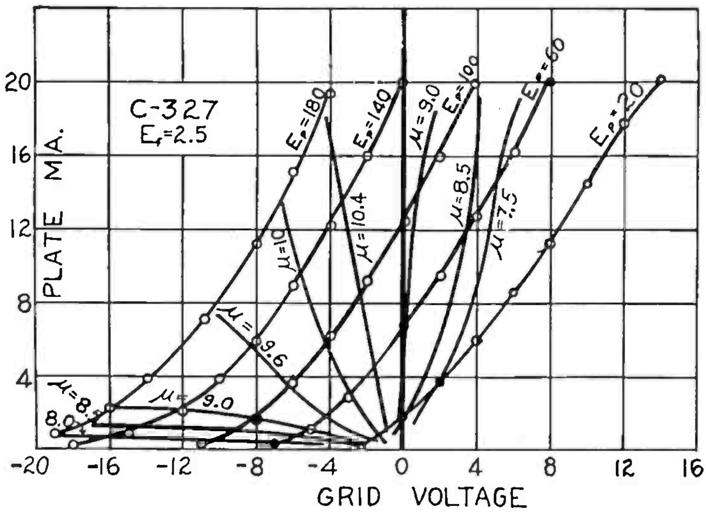

Fig. 2—Typical μ characteristic of commercial three-element tube.

though operated on a straight line portion of a plate-voltage, plate-current characteristic, and furthermore, under these conditions the distortion will be greatest when the load impedance is very large. Inconstancy of μ also causes crosstalk when a tube is simultaneously used to amplify several communication channels. These effects have been discussed in detail elsewhere and will not be taken up here.[2]

The reduction of μ as the grid is made negative reduces the completeness of rectification in plate detectors, and can very materially reduce the detector output. A low μ at a very high grid bias also reduces the efficiency of oscillator tubes by allowing a small residual of plate current to flow when the plate voltage is high.

[2] Peterson and Evans, "Modulation in vacuum tubes used as amplifiers," *Bell Sys. Tech. Jour.*, 6, 442; July, 1927.

The standard method of measuring μ is by means of the Miller bridge. This arrangement does not give a sharp balance point, however, because of the capacity current flowing between the plate and grid of the tube. The modification of the Miller bridge which is shown in Fig. 1 overcomes this difficulty by balancing the charging current out of the telephone receivers by means of the variable mutual inductance. The telephone receivers should also be brought to ground potential

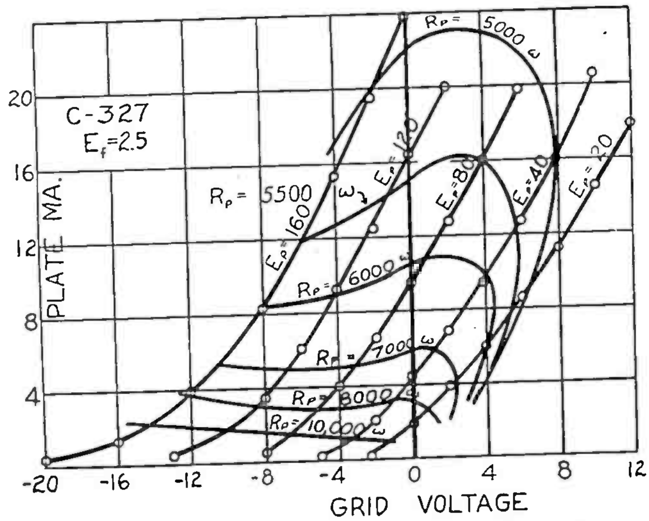

Fig. 3—Typical set of plate resistance contour lines for commercial three-element tube.

either by the use of a Wagner earth connection or by grounding a suitable part of the circuit.

The method of representing tube constants by means of contour lines such as used in Fig. 2 is a very useful method of giving the entire characteristic at a glance. Fig. 3 shows the results obtained by representing plate resistance in this manner, and indicates very clearly how the plate resistance varies from point to point over the characteristic curves.

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RADIOTELEGRAPHY AND RADIOTELEPHONY ON HALF-METER WAVES*

By

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Summary *—Communication tests in radiotelegraphy and radiotelephony on half-meter waves are described. The first section of the paper is devoted to the description of a new type of receiver for 40–80-cm waves. Experimental results are given to show the action of this receiver. It is pointed out that a sort of regenerative amplifying accompanies the detection of such extremely short waves, and a brief explanation of this action is given.*

The next section of the paper contains the results of experiments on a half-meter transmitter, with special regard to the modulation system.

In the latter portion of the paper, actual tests of communication to the distance from 10 to 30 km are described. The possible application of our wave collecting system to direction finding with such extremely short waves is pointed out.

INTRODUCTION

COMMUNICATION with ultra-short waves below one meter has not hitherto been accomplished, so far as the author is aware, even to the short distance of a few kilometers. Difficulties lie chiefly in the reception, and a receiver for such extremely short waves has not yet been satisfactorily developed.

The author's short-wave receiver[1] is now found to serve the purpose, and actual tests on telegraphic and telephonic communication with 45–50-cm waves have yielded very promising results.

RECEIVER FOR 40–80-CM WAVES

Apparatus

According to the author's experience, a heterodyne reception is difficult with these extremely short wavelengths, and is perhaps not practicable for technical purposes. Consequently, in telegraphy the wave must generally be modulated at a single audible frequency.

The simplest means for reception is to use a crystal detector and a multi-stage audio amplifier as shown in Fig. 1. This circuit is found not well suited to the reception of very weak signals from a great distance, because the detecting action of crystals is not sensitive enough in detecting weak signals.

* Dewey decimal classification: R402.

[1] S. Uda, "A new type of receiving set for extremely short waves," *Jour. I. E. E.* of Japan, June, 1929.

Fig. 2 shows the connection diagram of the author's receiver, in which *A* is the receiving antenna, and *L*, *L'* are parallel wires, on which the tuning condenser $C(0.0003\mu f)$ is bridged. Cymotron UF-101 was used as the detector tube. Chokes *Ch* were used to minimize the disturbing effects of the leads. A telephone receiver or an audio-frequency transformer is to be placed in the plate circuit of the detector

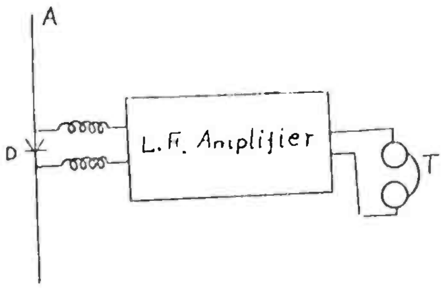

Fig. 1—Receiver with crystal detector and low-frequency amplifier.

tube. This is practically very advantageous, since the direct current flowing to the plate in this case is very small in comparison with the current in the grid circuit.

The negative voltage applied to the plate can be varied by the potentiometer *R*, and by adjusting it the wavelength may be slightly

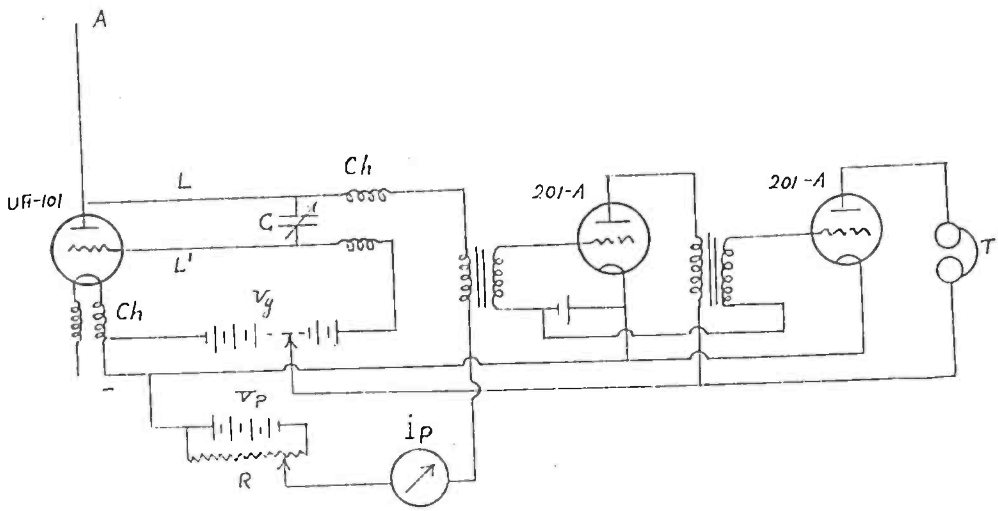

Fig. 2—Connection diagram for 40–80-cm wave receiver.

changed. A closer adjustment can be effected by the variable air condenser *C* having the usual vernier dial. It is interesting to note that even in cases of such extremely short waves as below 80 cm, a very effective control of tuning may be made by ordinary means.

In our experiments, 45–50-cm waves were generally used. The grid voltage was about 145, the plate voltage was from about -10 to -30, nd the grid current was about 20 ma. The apparatus is shown in Figs. 3 and 4. In Fig. 4, the receiving antenna system is shown, in

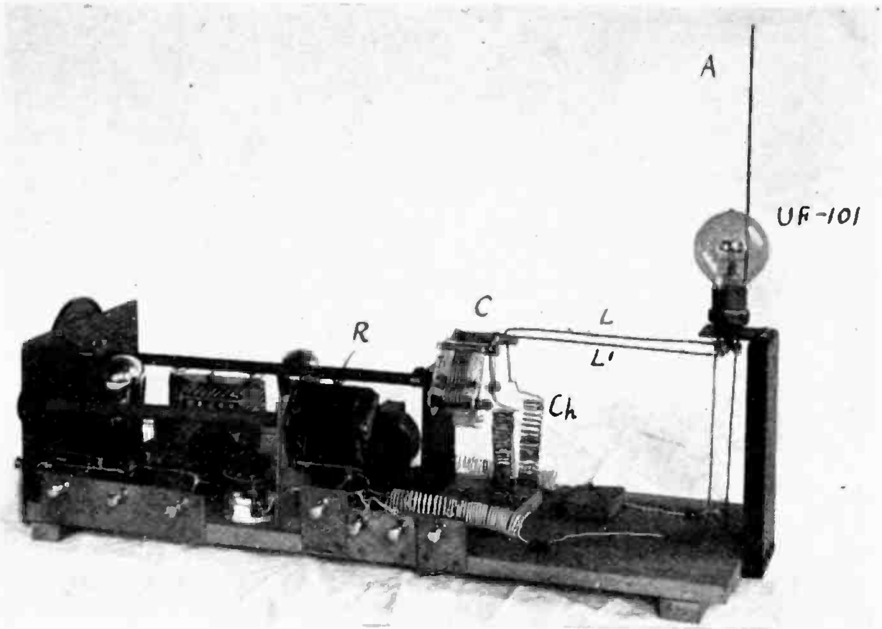

Fig. 3—Receiver for 40–80-cm waves. Cymotron UF-101 is used.

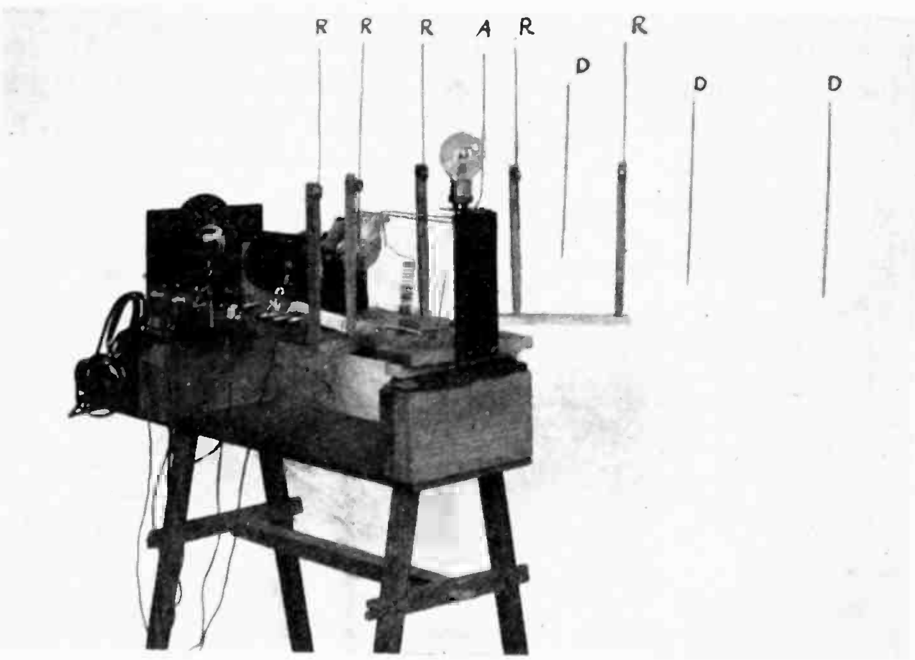

Fig. 4—Receiving antenna system.

which the parabolic reflector of 5 rods and wave-director rods can be seen.

If it is required to increase the range of the wavelength, the grid voltage may be varied, and one can thus easily obtain the range from 40 cm to 80 cm with this single radio receiver.

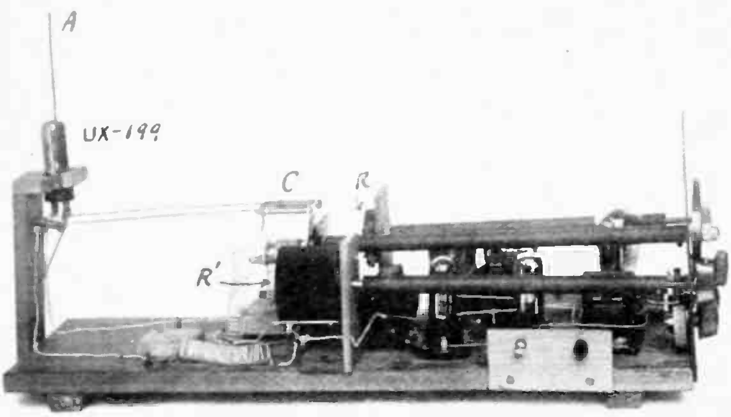

Fig. 5—Receiver for 50-cm waves. UX-199 tube is used.

Afterwards, the detector tube Cymotron UF-101 was replaced by Cymotron UX-199 tube, and the grid voltage was then varied with a potentiometer. For 45–50-cm waves, the grid voltage was then kept at about 90–100, while a few negative volts were applied to the plate, and the grid current was about 10 ma. This set could be operated with a much smaller number of dry cells in both the grid and the plate

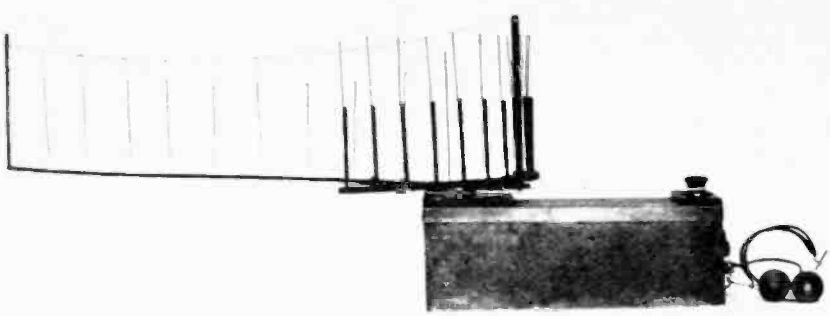

Fig. 6—Fifty-cm wave receiver with shielded box.

circuits than in the former receiver. Figs. 5 and 6 show the actual portable sets of this receiver. In Fig. 6, the antenna arrangement is shown, and Fig. 7 shows its interior view.

Fig. 7—Interior view of Fig. 6 receiver.

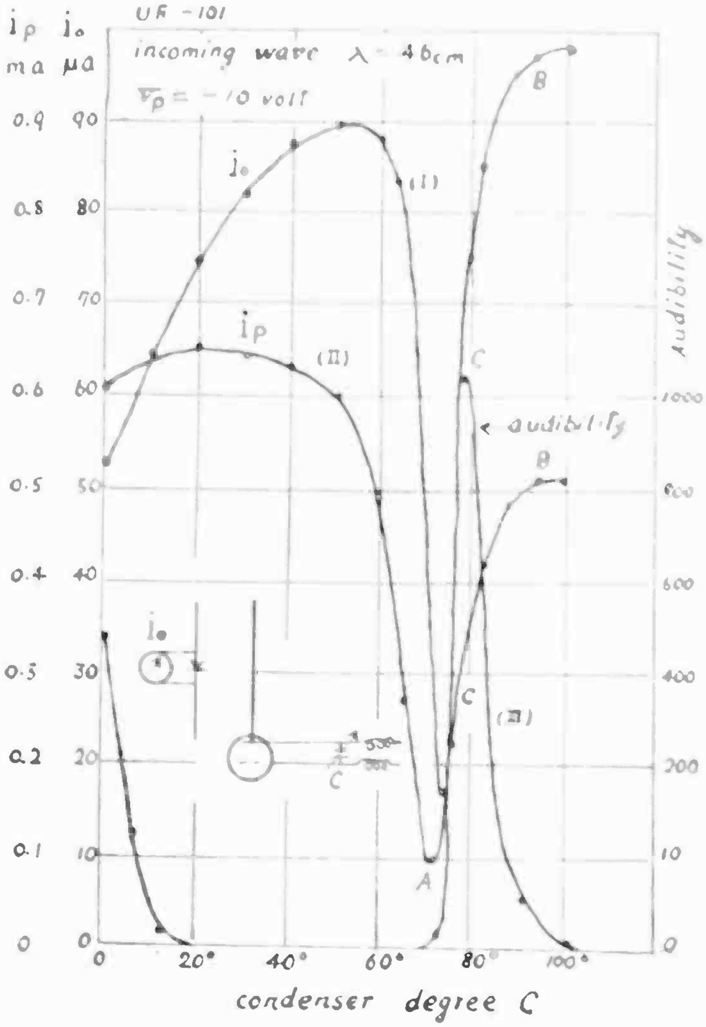

Fig. 8—Relation between condenser reading and audibility of detected signals, oscillation intensity, and anode current.

Detection

The author stated in his previous paper that a kind of regenerative amplifying action can exist with the detecting action of an electron oscillator. This probably is the cause of the great sensitiveness of our receiver.

As the condenser capacity $C(0.0003\mu f)$ is gradually increased the oscillation current usually suffers a sudden decrease at a certain point, and after passing through a minimum, rapidly increases again, gradually returning to its original value. Typical examples are shown by curves (I) in Figs. 8 and 9, in which i_o represents the readings of

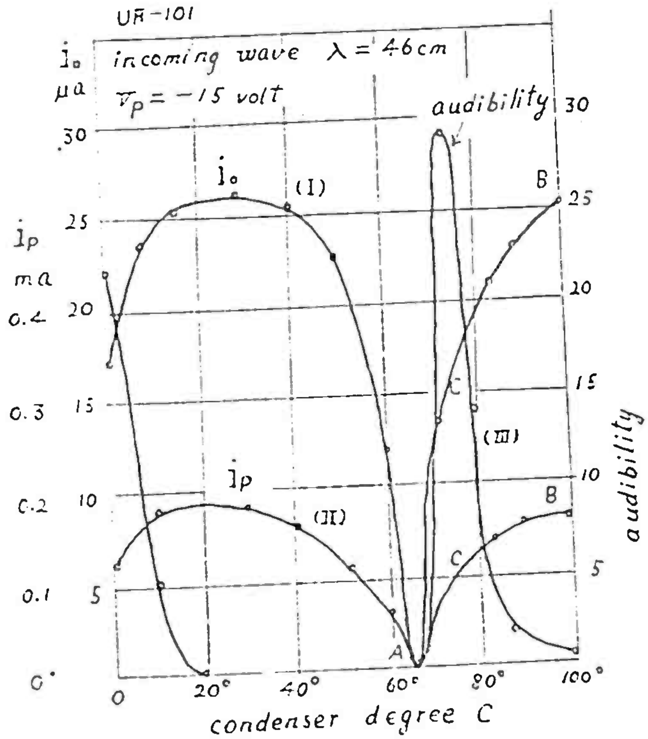

Fig. 9—Relation between condenser degree and audibility of detected signals, oscillation intensity, and anode current.

the microammeter attached to the center of a straight rod, placed near the receiver as shown in Fig. 8. Curves (II) in the same figures show the corresponding plate current i_p. In these experiments, measurements were made of the sound intensities of the received signals. The audibilities are thus given by curves (III) in these figures.

In order to explain the phenomena clearly, let us now examine the following experiment, in which the variable air condenser is replaced by the short-circuit condenser C_s, which can be slid on the Lecher wires as shown in Fig. 10. Now Fig. 11 shows the high-frequency current i_o, the plate current i_p, and the wavelength λ, as they vary with the distance d along the Lecher wires. Sound intensity of the detected signal was also recorded at the same time, and is plotted

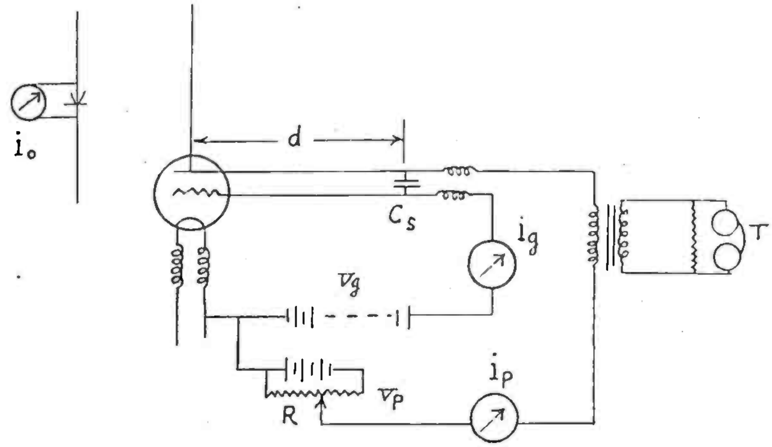

Fig. 10—Circuit connection used in tests for detection.

as curve (III) in the same figure. In this case, the source was modulated at 1000 cycles, and the exact wavelength was equal to 46 cm.

It is obvious that the change of the condenser capacity in the former experiments (Figs. 8 and 9) corresponds to the sliding of the

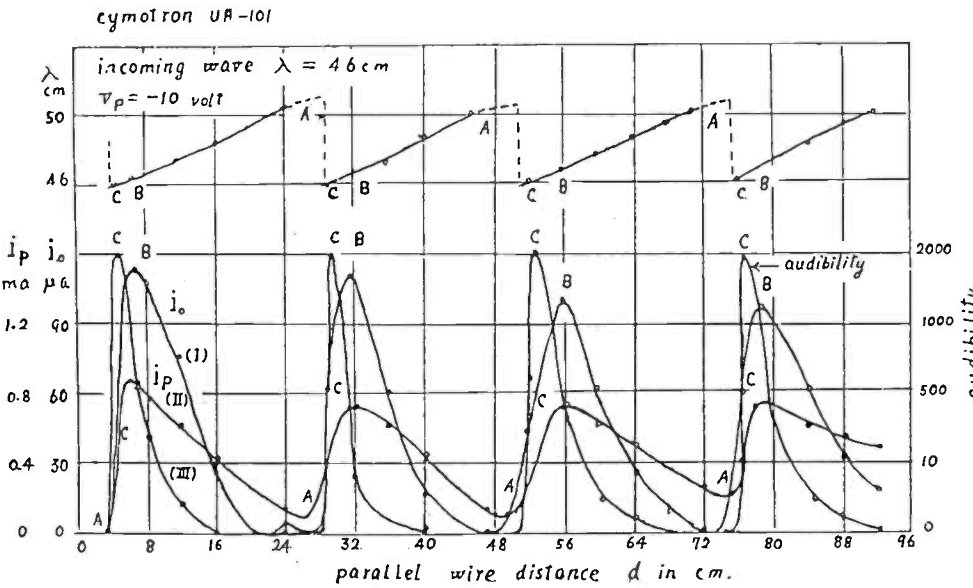

Fig. 11—Variation of audibility of detected signals, oscillation intensity, wavelength, and anode current with parallel wire distance.

short-circuit condenser C_s in the latter. (Fig. 11.) The points A, C, B as indicated in Fig. 11 correspond respectively to those indicated in Figs. 8 and 9. In the region near the point A, as Prof. Hollmann has stated,[2] weak Barkhausen-Kurz oscillations may usually exist, or else, in some cases, there will be no oscillations in this region.

Suppose that the negative plate voltage of the detector tube is so adjusted by the potentiometer R (in Fig. 2) that the frequency of the incoming waves nearly coincides with that at the C state in the detector oscillation regions. (This may be effected conveniently, in some cases, by changing the impressed grid voltage with a potentiometer.) At this state C, the frequency of the oscillation is determined by the external tuning system, as pointed out by Hollmann.

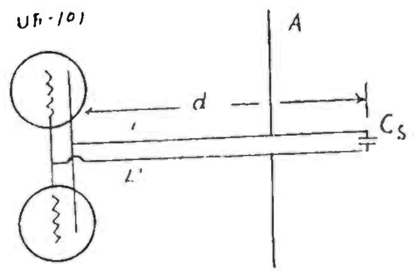

Fig. 12—Two tubes connected in parallel.

Therefore, if this external oscillation system is tuned to the incoming waves by varying the tuning condenser (Fig. 2) or by sliding the short-circuit condenser (Fig. 10), then electron oscillations in the detector tube will be affected and controlled by these waves, and will very likely be pulled in to the same modulation as the incoming waves. In other words, the amplitude of the oscillations will be changed according to the signal strength, and their frequency will also depend upon the incoming waves. This fact is plainly recognizable, especially when the signal strength is comparatively strong. It seems that quite the same action is accomplished as long as the field intensity is not below a certain critical value. A regenerative amplifying action of this receiver is probably due to this controlling effect of the incoming waves. Hence if the strength of the incoming signals is too weak to affect the internal electron oscillations, the detection will not be

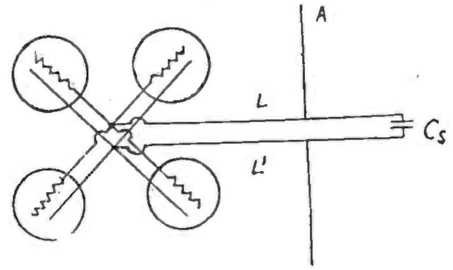

Fig. 13—Four tubes in parallel.

[2] H. E. Hollmann, Proc. I.R.E., 17, 229; February, 1929.

efficiently accomplished, and then signals may not be heard at all. But, on the contrary, if the signal strength is stronger than a certain critical value, pulling action will take place, and, consequently, a regenerative amplifying detection will be very efficiently accomplished. This characteristic has often been noticed in our electron detector.

The above explanation seems to be confirmed by the experiments above described (Figs. 8, 9, and 11), in which the points of detection with the highest sensitivity occurred always near the point C, just after the transition from the A state to the C state. It is again noticeable that when the first transition takes place at a parallel wire length d of 5 cm, the process repeats itself at $d=29$ cm, $d=52$ cm, etc. (Fig. 11.) Evidently, in the later C regions we have overtones of the parallel-

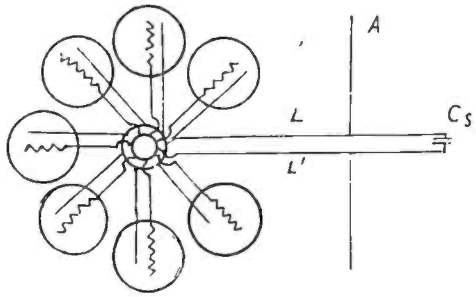

Fig. 14—Seven tubes in parallel.

wire system. In Figs. 8 and 9, too, two points of sensitive detection were found.

Since no special care was taken to keep the frequency of the sender exactly constant, a certain degree of "frequency modulation" may have resulted. Although no audible heterodyne beats could usually be heard by adjusting this receiver, it was difficult to distinguish them from other weak noises originating in the receiver. Perhaps for this reason the detected signals were not disturbed at all by audible beats, although the detector tube was at its most sensitive detection point, always in the oscillation region. This is one of the merits of this electron regenerative detector, and is different from an ordinary regenerative autodyne receiver of a few meters.

TRANSMITTER

Parallel Operation

To increase the intensity of oscillation, a number of triodes were connected in parallel to a common tuned oscillation system. In Fig. 12, two tubes are connected in parallel. C_s is a bridging short-circuit condenser, and A is the sending antenna connected to the parallel wires a quarter-wave distant from C_s. The antenna current of this transmitter was about 15–20 ma. Cymotron UF-101 tubes were used.

Four tubes are connected in parallel in the case of Fig. 13, and seven tubes in Fig. 14, the antenna current of the former being about 30 ma, and of the latter about 50 ma. The actual apparatus with seven tubes is shown in the photograph of Fig. 15. The wavelength of

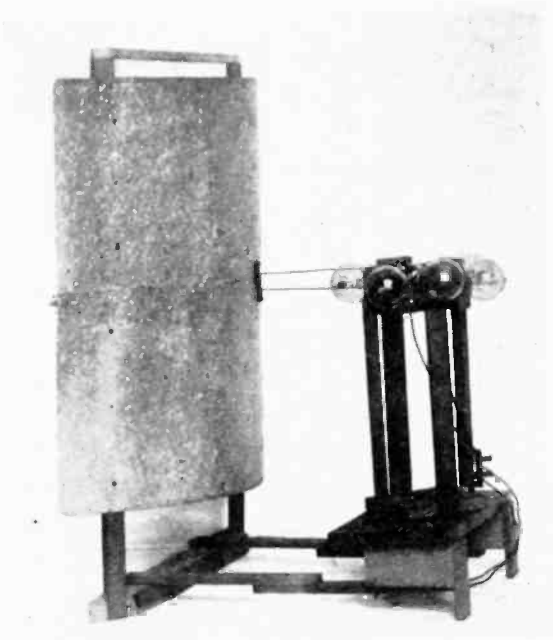

Fig. 15—Fifty-cm wave transmitter with seven tubes.

this transmitter was about 50 cm, and was found not to be much elongated due to the increase of the number of tubes. It is certainly possible that some inconvenience will be experienced in practice due

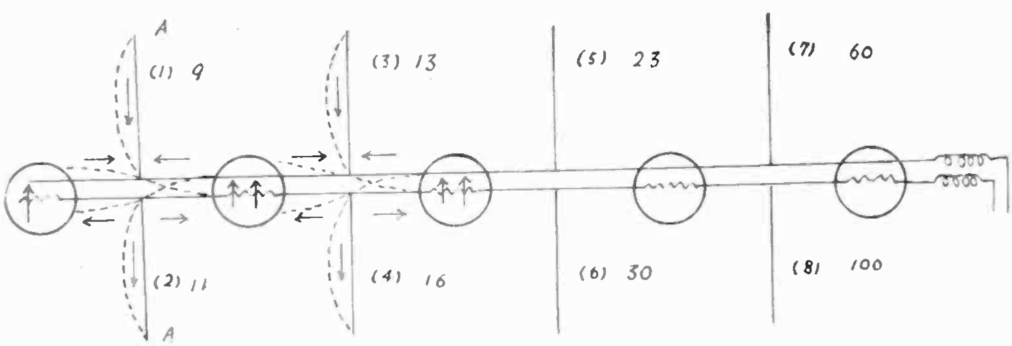

Fig. 16—Directive antenna system with parallel excited tubes.

to the failure of one of the many tubes employed, thereby upsetting the condition of successful parallel operation. Yet a 2- or 4-tube oscillator was found to be very convenient, according to the author's experience.

Fig. 16 shows another form of parallel connection, in which A_8 are the antennas erected vertically at the middle points between adjacent tubes. If the distance between adjacent oscillators be properly

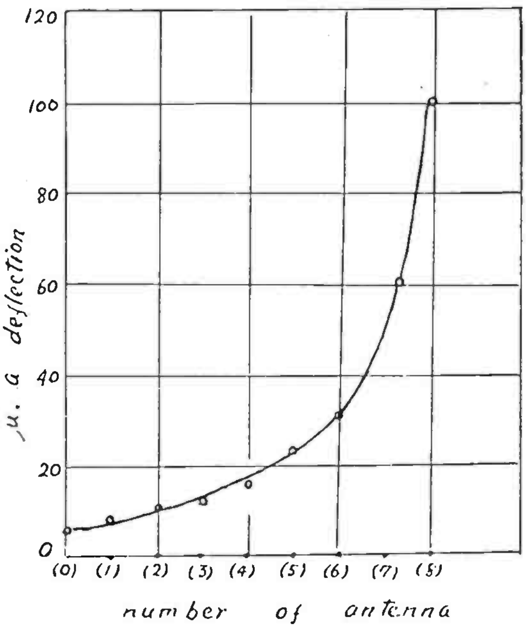

Fig. 17—Variation of received energy with number of antenna.

chosen, the effects of all the antennas will become additive and the radiated energy will be much increased with the number of the an-

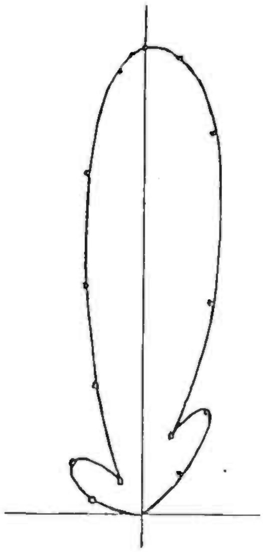

Fig. 18—Polar curve in horizontal plane.

tennas. The numbers in parentheses in Fig. 16 indicate the order of introducing the antennas, and the annexed numerical figures show the meter readings in the receiver. The increase of the received current with the number of the antenna in this case is also shown in Fig. 17. Fig. 18 shows the horizontal polar distribution due to this system. But, in this system, if the distance between adjacent oscillators be not suitably chosen, the antènna will not oscillate in the normal manner as shown in Fig. 16, and this will cause an undesirable result.

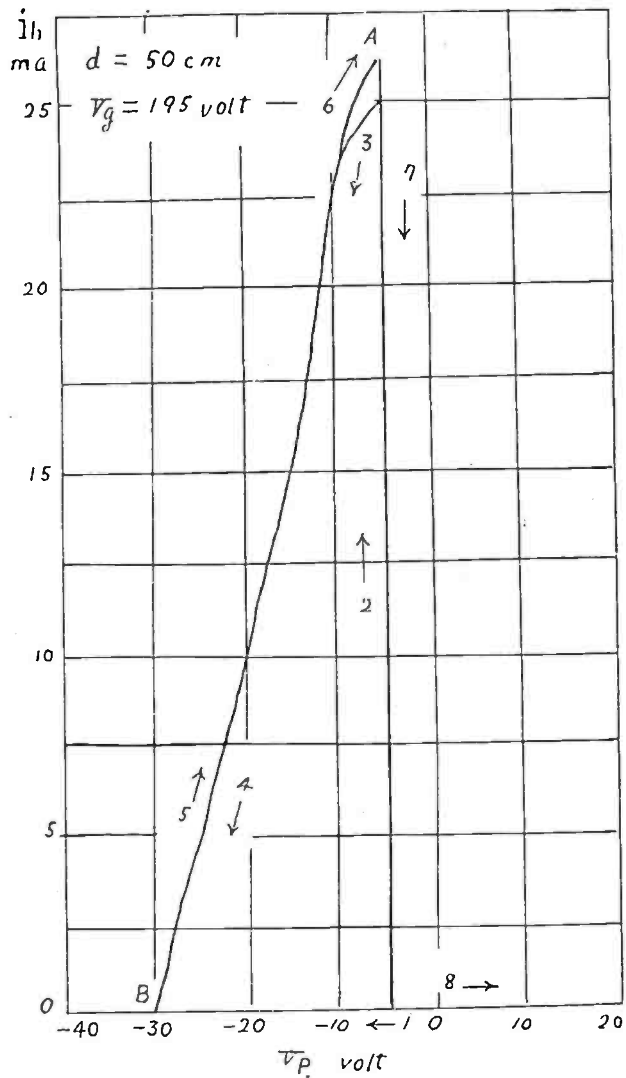

Fig. 19—Variation of oscillation intensity with applied plate voltage.

Modulation

In ordinary radiotelephony, "amplitude modulation" is generally used. In the case of an electron oscillator, pure amplitude modulation in a strict sense can hardly be maintained, and usually a certain small amount of "frequency modulation" is present, even in the region of the so-called Gill and Morrell oscillations. According to our experience, however, this causes no difficulty in actual telephonic communication.

Various systems of modulation were investigated, such as the grid-voltage or the plate-voltage control system, or the resistance control in the plate circuit of the transmitter. Among them, the plate-voltage control system was found to give the best results.

Fig. 19 shows how the intensity of oscillation changes with the plate voltage, when it is varied according to the order indicated by the

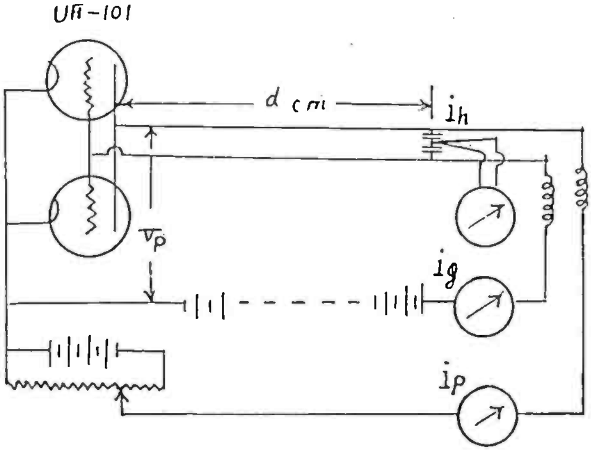

Fig. 20—Circuit connection for parallel-excited 2-tube oscillator.

arrows. The circuit connection in this experiment is shown in Fig. 20. The magnitude of i_h represents the oscillation current as measured by a thermocouple. In the part *AB* of the curve (Fig. 19), the intensity of oscillation is nearly proportional to the negative plate voltage, and therefore this region may conveniently be used for the distortionless modulation of the waves. The actual arrangement of this system is

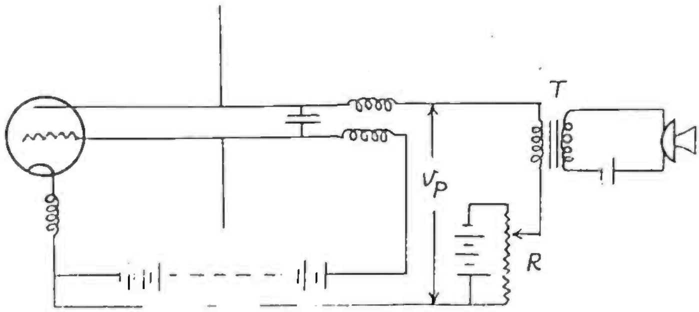

Fig. 21—Modulation system.

shown diagrammatically in Fig. 21. In this method, the low-frequency transformer is to be placed in the plate circuit, which has the merit that the direct current flowing into the transformer coil can be kept very small. In order to get satisfactory results, the applied plate bias voltage must be suitably chosen by adjusting the potentiometer *R*.

Fig. 22 shows the connection diagram of our transmitter, in which the plate modulation system with three-stage speech amplifier is used.

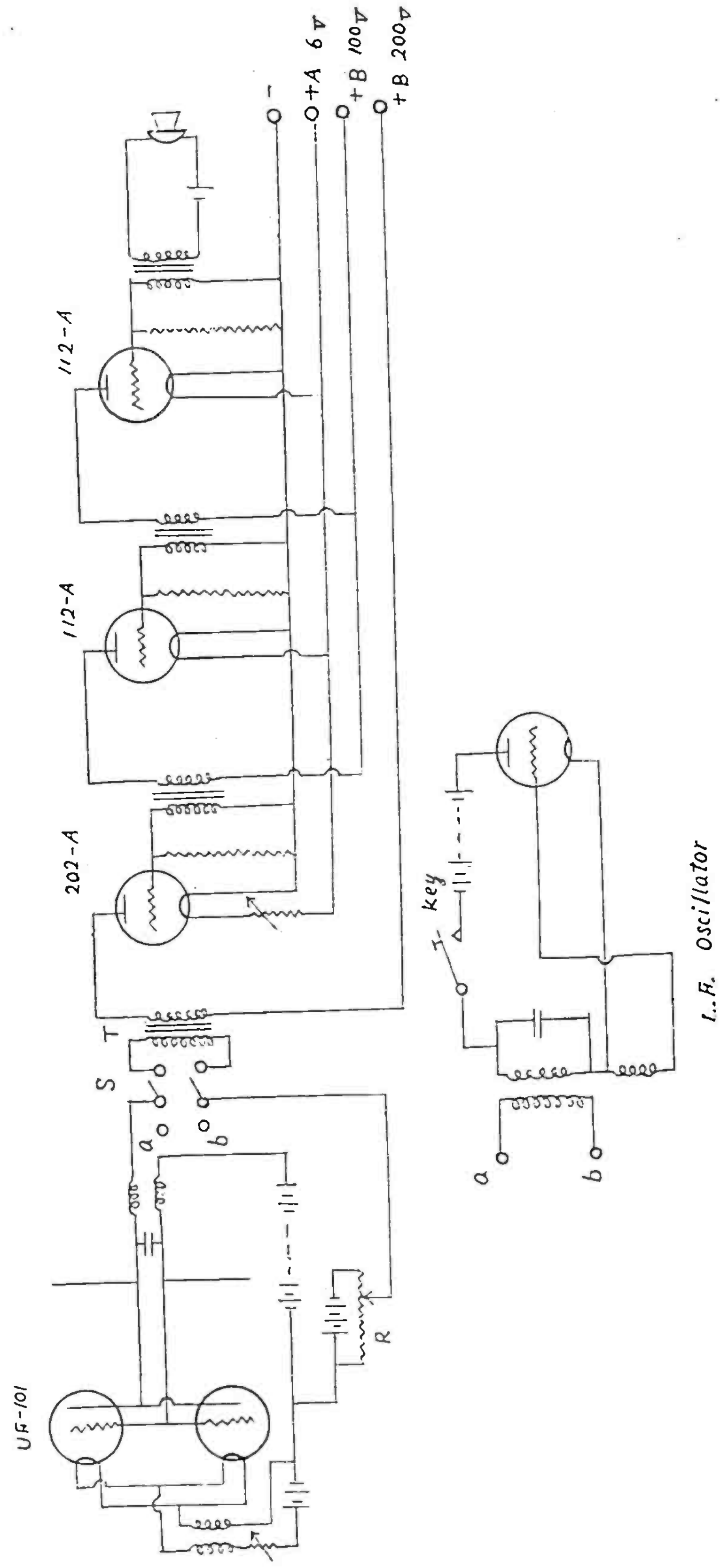

Fig. 22—Circuit diagram of transmitter; 45–50 cm.

By changing over the switch *S*, either telephonic or telegraphic communication can be easily carried out (Fig. 22). The actual set is shown in Fig. 23.

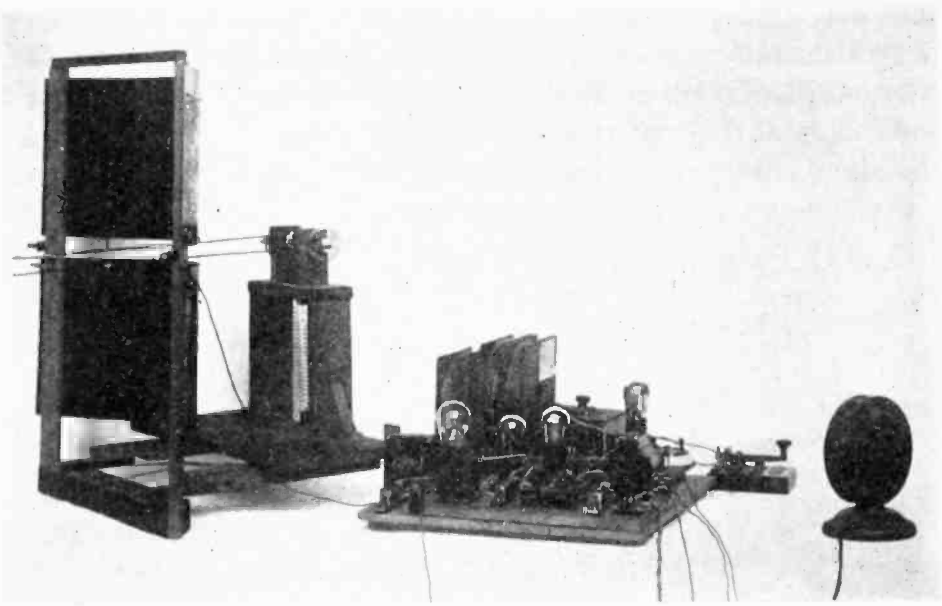

Fig. 23—Apparatus for 45 50-cm wave transmitter with modulation system.

Communication Tests

Tests for reception were carried out at a distance of 10 km. Both the transmitter and the receiver were located on the hills with a visual line between them. Fifty-cm waves were used. A director chain of about 7 meters was used both on the sending and the receiving sides as illustrated diagrammatically in Fig. 24. In telegraphy, the trans-

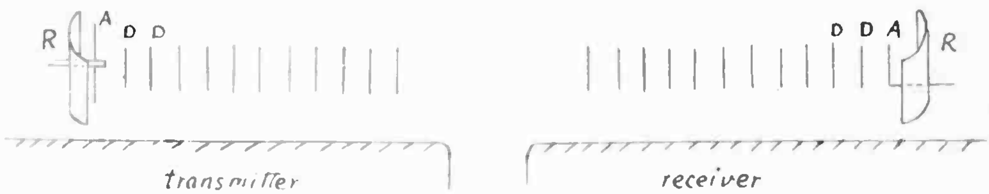

Fig. 24—Sketch showing the locations of the transmitter and receiver.

mitter was modulated with a single frequency of 1000 cycles, and, in this case, the signal strength was 2000 in audibility (using the General Radio audibility meter, type 164B). Reception of radiotelephony, in which the waves were modulated by speech, has also yielded a very satisfactory result. No distortion was experienced, and clear speech could be loudly heard.

It must be remembered that the short-wave transmitter used in the above experiment was not so strong as a magnetron oscillator. The antenna current was about 10–15 ma when the waves were modulated. Notwithstanding this, the signal was very loud. This is certainly due to the excellent characteristics of the receiver, and partly due to the application of the wave-director chains, which proved to be astonishingly advantageous in concentrating and collecting the wave energy at these short wavelengths. A 7-meter director chain contains about 40 wave directors, and concentrates the radiated energy

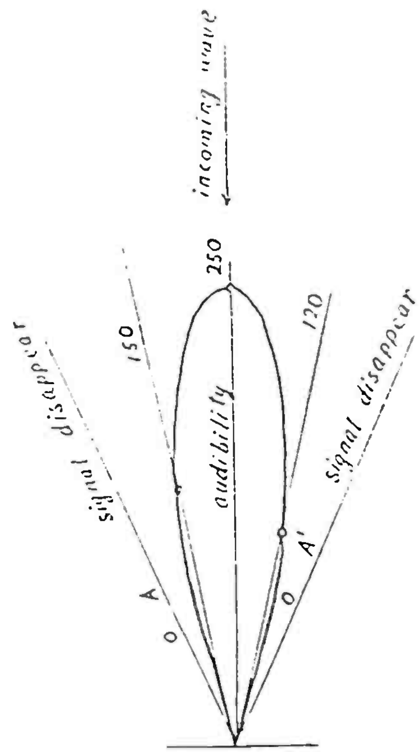

Fig. 25—Audibility curve of detected signals at 10 km.

within an extremely sharp beam with only a few degrees of divergence, not only in the horizontal plane but also in the vertical plane. Afterwards, the communication distance in telegraphy could be increased to a distance of 30 km. In this case, the signal strength was about 120 expressed in audibility.

Direction Finding

In the 10-km test, we tried to revolve the director chain on the receiving side around the main antenna as a center. (Fig. 6) When the direction of the director chain nearly coincided with the direction of the incoming waves the signal was of course very strong, and as it was removed from this direction of wave propagation the signal strength decreased rather rapidly and then abruptly faded. The result of this experiment is shown in Fig. 25, where A and A' indicate the vanishing points of the signal. The direction of the incoming waves can be found by taking the middle direction between these two points, A and A'. The method seems to have a very small error, because the points of vanishing signals can be quite distinctly determined, and moreover the angle between them may be narrow, owing to the sharp directivity of this direction-finding system. The fact that the signals disappear so distinctly and abruptly, seems to prove that, in our electron receiver, the regenerative detection becomes suddenly ineffective when the signal strength becomes too weak to control the electron oscillations within the detector tube.

Conclusion

From the above experiments, it may be concluded that radiotelegraphy and radiotelephony on half-meter waves can be successfully executed to a distance of more than 10 km. This achievement is chiefly due to the development of the special receiver for such extremely short waves, and thus there remains no question of the possibility of the practical application of these extremely short waves. It is hoped that this paper may serve in some measure to promote future progress in special communication over a comparatively short distance, as in secret communication, etc., using these extremely short waves.

The author wishes to express his thanks to Prof. H. Yagi, under whose direction the work was carried out, and to Prof. S. Chiba for his valuable advice. We are much indebted to the Saito Gratitude Foundation in Sendai for the financial support given our work.

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THE EFFECT OF RAIN AND FOG ON THE PROPAGATION OF VERY SHORT RADIO WAVES*

By

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(Department of Electrical Engineering, Massachusetts Institute of Technology, Cambridge, Massachusetts)

Summary—*The purpose of this paper is to investigate theoretically the effect of rain, fog, or clouds on the propagation of short radio waves. The theory of the propagation of electromagnetic waves in a medium in which are suspended spherical particles of an arbitrary material is first reviewed. The available physical data on fog and rain is then referred to. The conclusion is arrived at that for waves greater than 5 cm in length, the effect of ordinary rain or fog on the absorption is negligible.*

Introduction

The steady trend towards higher frequencies in radio communication brings with it a host of new problems on the means of their generation and the manner of their propagation. At present there appears to be very little data available on the feasibility of point-to-point communication at frequencies exceeding about 60 megacycles per second, and in view of conditions at the upper surface of reflection, the possibility of regular transmission between points on the surface of the earth seems questionable. Frequencies lying in this region of the spectrum might, however, be utilized to great advantage for beacon work and beam transmission from the ground to nearby aircraft. In order to concentrate electromagnetic energy into a true beam, the dimensions of the emitting system must be very much greater than the wavelength of the radiation. Parabolic reflectors of the proper dimensions are entirely impractical for the frequencies employed in radiotelegraphy, and in the directional systems now in use the energy is concentrated into planes, not beams. At frequencies greater than 100 or 200 megacycles per second reflector systems enter the domain of practicability, and therein lies the principal advantage of these frequencies.

The greatest impediment to the experimental investigation of frequencies exceeding 100 megacycles per second is the absence of a technique for generating them with any considerable amount of power. There seems to be, however, no *a priori* reason why this technique cannot be developed. Assuming that the gap in the spectrum between the infra-red and short radio waves can be suitably bridged, what will

* Dewey decimal classification: R113.5. Contribution from Round Hill Laboratories.

be the influence of the state of the atmosphere on the propagation? Visible light is scattered in rain and fog; infra-red radiation is scattered and suffers selective absorption as well. The present paper gives a theoretical investigation of the question:

At what point in the radio spectrum does absorption in rain and fog due to scattering and selective absorption become appreciable?[1]

Simple Solution

If only the absorption due to the scattering of energy by rain drops is desired, the problem is simply solved by the method employed by Lord Rayleigh in his investigation of the scattering of visible light in the atmosphere. This method is based on the assumption that the scattering particle may be replaced by an equivalent dipole whose dimensions are small compared to the length of the exciting wave, and that the average distance between drops is such that different drops scatter incoherently.

Suppose that the incident wave is plane, its electric vector polarized along the X axis and its direction of propagation parallel to the Z axis. The axis of the equivalent dipole which replaces the rain drop is parallel to X. The Poynting vector in the radiation field of an oscillating dipole of moment p excited by a wave of angular frequency ω is given by the real part of

$$S_1 = \frac{p^2\omega^4 e^{2i\omega t}}{4\pi c^3 r^2} \sin^2 \theta. \tag{1}$$

In this expression c stands for the velocity of light, r the distance from the dipole to the point of observation, and θ the angle that r makes with the axis of the dipole. Taking the time average of (1) we have

$$\bar{S}_1 = \frac{p^2\omega^4 \sin^2 \theta}{8\pi c^3 r^2}. \tag{2}$$

Since the energy scattered by different drops is by hypothesis incoherent, energies and not amplitudes are to be added when summing up the contributions of the N dipoles in unit volume.

$$\bar{S} = \frac{\omega^4 \sin^2 \theta N p^2}{8\pi c^3 r^2} \tag{3}$$

where N is the number of scattering particles per unit volume.

The polarization per unit volume of dielectric is given by the Lorenz-Lorentz formula

[1] The author gladly acknowledges his indebtedness to E. O. Hulburt, of the U. S. Naval Research Laboratory in Washington, for the suggestion that scattering by clouds or rain might influence communication with waves of the order of 5 meters or less.

$$P = \frac{3}{4\pi}\frac{m^2 - 1}{m^2 + 2}E. \tag{4}$$

Here E is the electric force acting on the dipole and m the index of refraction of the polarized dielectric, in this case water. m is in general complex, but since it is assumed in this solution that the rain drops have zero conductivity, m has no imaginary part and may be placed equal to the square root of the dielectric constant.

From (4) it follows directly that the polarization of a single drop is

$$p = \frac{m^2 - 1}{m^2 + 2}\rho^3 E, \tag{5}$$

where ρ is the radius of the drop. The total energy scattered per unit volume of air containing N drops is found by integrating the Poynting vector (3) over the unit sphere.

$$\begin{aligned} S &= \frac{Np^2\omega^4}{8\pi c^3}\int_0^{2\pi}\int_0^{\pi}\int_0^{1}\sin^3\theta d\phi d\theta dr \\ &= \frac{16}{3}\frac{\pi^4 cN\rho^6}{\lambda^4}\left(\frac{m^2 - 1}{m^2 + 2}\right)^2 E^2. \end{aligned} \tag{6}$$

The energy density of the incident wave is

$$P = \frac{c}{4\pi}E^2. \tag{7}$$

The scattering coefficient σ is defined as that fraction of the primary intensity P which is lost in penetrating through unit length of the scattering medium.

$$\sigma = -\frac{1}{P}\frac{dP}{dz}$$

or, upon integration,

$$P = P_0 e^{-\sigma z}. \tag{8}$$

This is equivalent to

$$\sigma = \frac{S}{P} = \frac{64\pi^5 N\rho^6}{3\lambda^4}\left(\frac{m^2 - 1}{m^2 + 2}\right)^2. \tag{9}$$

The restrictions on the validity of this formula have already been indicated. It does not apply when the length of the incident wave is of the same order of magnitude as the diameter of the rain drop. It does not take into account the effect of a possible conductivity or the presence of absorption bands in water. These absorption bands play a predominant role in the calculation of the effective dielectric constants at wavelengths less than about 40 mm. Finally, if the raindrop density becomes too great the scattering may not be incoherent, as has been assumed, and the velocity of propagation of the wave will be materially altered by the rain storm or cloud. To settle this point, we must have recourse to specific physical data.

Numerical computations based on formula (9) show that the absorption of electromagnetic waves due to scattering in rain or fog is negligible for waves greater than 30 or 40 cm in length. The present investigation was made, however, with a view to determining the utility of even shorter wavelengths for communication purposes, assuming that the technique for generating such waves can be developed. In order to derive a formula for the absorption coefficient which is valid for any frequency and any density of rain drops, as well as to determine the region of validity of formula (9), a more detailed analysis of the problem must be made.

Rigorous Solution

The problem of absorption of very short radio waves by suspended particles of water has an exact optical analogy in the absorption of light in colloidal solutions One of the first to treat this problem was Maxwell Garnett[2] in an investigation of the colors of metal glasses. His solution, however, gives only the absorption due to the finite conductivity of the suspended particles and does not take into account the effect of scattering. A detailed investigation of the optics of colloidal suspensions in water was made by Mie,[3] and this work was then extended in a paper by Gans and Happel.[4] The method developed in these last two papers may be applied with but slight modification to the present problem. Where applicable their results will be used directly and the reader referred to the original papers for proof.

Very briefly stated, the mechanism of the phenomenon is as follows. In air, characterized by unit dielectric constant and zero conductivity, is suspended a drop of water of radius ρ and index of refraction m. The origin of the coordinate system coincides with the center of the drop. A plane primary wave falls on the drop which in turn is excited to emit a spherical secondary wave. The resultant intensity at any point outside the drop is given by the sum of the intensities of the primary and secondary waves. The primary wave suffers a decrease in intensity

[2] J. C. Maxwell Garnett, *Phil. Trans.*, 203, 385; 1904.
[3] G. Mie, *Ann. d. Phys.* 25, 377; 1908.
[4] R. Gans and H. Happel, *Ann. d. Phys.* 29, 277; 1909.

due first to absorption in the water of the drop and secondly to the scattering involved in the emission of the secondary wave.

Mie has calculated the intensity of a secondary wave from a spherical particle.[5] Let the primary wave be plane, of unit amplitude, of frequency ν, and propagated along the negative z axis. In electromagnetic units

$$E_x{}' = e^{2\pi i\nu(t+z/v)}, \qquad H_y{}' = -\sqrt{E}\, e^{2\pi i\nu(t+z/v)}, \tag{10}$$

where $v = 1/\sqrt{\epsilon_0}$ is the phase velocity and ϵ_0 the dielectric constant of air in e.m.u. It can now be shown that if the sphere is sufficiently small with respect to the wavelength λ, namely so small that the fourth and higher powers of $\alpha = 2\pi\rho/\lambda$ may be neglected with respect to unity, it may, for calculating the radiation, be replaced by

1. An electric dipole of moment p_{1e} parallel to the X axis;
2. A magnetic dipole of moment p_{1m} parallel to the y axis;
3. Two electric quadrupoles with moments $p_{2e}{}^1$ and $p_{2e}{}^2$ directed as shown in Figs. 1 and 2.

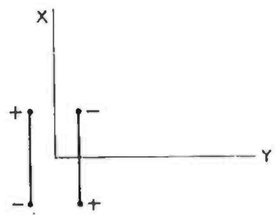

Fig. 1

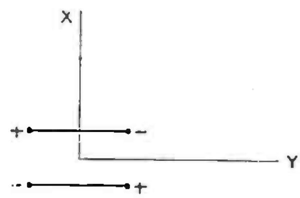

Fig. 2

$$\left|\overset{(1)}{p}_{2e}\right| = \left|\overset{(2)}{p}_{2e}\right| = e\,dx\,dz,$$

where $e\,dx$ and $e\,dz$ are the moments of the elementary dipoles which form the quadrupoles.

The assumption that $\alpha^4 \cdot << 1$ is a restriction on the ratio of the radius of the diffracting sphere to the wavelength. A second assumption concerns the concentration. It will be assumed that the average distance l between drops is so large that powers of ρ/l greater than the first may be neglected. In cases where this assumption is not fulfilled the suspension of spheres or drops will have a material influence on the velocity of propagation. Evidently the theory may be extended for shorter waves and greater concentrations by considering multipoles of higher order than the fourth. At this point we have recourse to the physical data of the problem to determine whether our assumptions are justified.

[5] For an excellent form of this calculation see Debye, *Diss. München*, 1908, and *Ann. d. Phys.* 30, 57; 1909, and the article by Sommerfeld in *Riemann-Weber*, "Differentialgleichungen der Physik," II, 495; 1927.

The following data on clouds and rain are taken from a standard treatise on meteorology.[6]

TABLE I

Type of Precipitation	Precipitation in mm per hour	Diameter of drops in mm	mgm of water per cu. meter of air
Fog	Trace	0.01	6.0
Mist	0.05	0.10	55.5
Drizzle	0.25	0.20	92.6
Light rain	1.00	0.45	138.9
Moderate rain	4.00	1.00	277.8
Heavy rain	15.00	1.50	833.3
Excessive rain	40.00	2.10	1851.9
Cloudburst	100.00	3.0 to 5.0	5401.4

A simple computation shows that if $\alpha^4 < 0.1$, λ must exceed 15 mm in the case of the heaviest rain and 5 mm in that of a moderate rain fall.

Further calculations of the concentration and average distance between drops result in the data of Table II.

TABLE II

Type	ρ in cm	No. drops per cu. meter	l in cm	ρ/l
Fog	0.0005	1.14×10^7	0.445	0.001
Moderate rain	0.05	530	12.3	0.004
Cloudburst	0.15	380	13.8	0.011

It will be seen from Table II that the condition $\rho << l$ is always fulfilled.

In Mie's paper the explicit expressions for the equivalent dipoles and quadrupoles are found to be

$$P_{1e} = \frac{\rho^3 u_1}{c^2} \frac{m^2 - v_1}{m^2 + 2w_1} E_x', \tag{11}$$

$$P_{2m} = \rho^3 u_1 \frac{1 - v_1}{1 + 2w_1} H_y', \tag{12}$$

$$P_{2e} = \frac{i\pi\rho}{c^2\lambda} u_2 \frac{m^2 - v_2}{m^2 + \frac{3}{2}w_2} E_x', \tag{13}$$

where u, v and w are relatively simple functions of α to be discussed below. An examination of (13) now shows that since ρ/λ does not exceed 0.1 for moderate rainfall and wavelengths greater than 5 mm, the contribution of the quadrupole p_{2e} may be neglected. In this approximation, therefore, the raindrop may be simulated by an electric and a magnetic dipole.

The functions u_1, v_1, w_1 are calculated from formulas to be found in the paper previously referred to.[7]

[6] W. J. Humphreys, "The physics of the air," 1920.
[7] *Ann. d. Physik*, 25, 400; 1908.

$$u_1 = \frac{3}{2\alpha} e^{i\alpha} \frac{\left(\sin\alpha + \dfrac{\cos\alpha}{\alpha} - \dfrac{\sin\alpha}{\alpha^2}\right)}{1 - \alpha^2 + i\alpha} \tag{14}$$

$$v_1 = m \frac{\dfrac{\sin\alpha}{\alpha} - \cos\alpha}{\dfrac{\sin m\alpha}{m\alpha} - \cos m\alpha} \cdot \frac{\sin m\alpha + \dfrac{\cos m\alpha}{m\alpha} - \dfrac{\sin m\alpha}{m^2\alpha^2}}{\sin\alpha + \dfrac{\cos\alpha}{\alpha} - \dfrac{\sin\alpha}{\alpha^2}} \tag{15}$$

$$w_1 = \frac{m\alpha}{2} \frac{1 + i\alpha}{1 - \alpha^2 + i\alpha} \cdot \frac{\sin m\alpha + \dfrac{\cos m\alpha}{m\alpha} - \dfrac{\sin m\alpha}{m^2\alpha^2}}{\dfrac{\sin m\alpha}{m\alpha} - \cos m\alpha} \tag{16}$$

The physical data of Table I show that except for waves of the order of a few millimeters α is small. We therefore expand (14), (15) and (16) according to increasing powers of α and in conformance with our original assumption, neglect terms of order higher than the third. The expansions give

$$u_1 = 1 + \frac{3}{10}\alpha^2 - \frac{2}{3} i\alpha^3 - \cdots \tag{17}$$

$$v_1 = 1 + \frac{\alpha^2}{10}(1 - m^2) + \cdots \tag{18}$$

$$w_1 = 1 + \alpha^2\left(1 - \frac{m^2}{10}\right) - i\alpha^3 - \cdots \tag{19}$$

$$a = u_1 \frac{m^2 - v_1}{m^2 + 2w_1} = \frac{m^2 - 1}{m^2 + 2}\left(1 + \frac{3}{5}\alpha^2 \frac{m^2 - 2}{m^2 + 2} - \frac{2i\alpha^3}{3} \frac{m^2 - 1}{m^2 + 2}\right), \tag{20}$$

$$b = u_1 \frac{1 - v_1}{1 + 2w_1} = \frac{\alpha^2}{30}(m^2 - 1). \tag{21}$$

From the expressions for the electric and magnetic dipole moments as explicit functions of the time may be calculated the secondary radiation from a drop of water. It will be seen upon examination of (20) and (21) that in the limiting case of very small drops or very long waves the magnetic dipole p_{lm} vanishes and the electric dipole p_{le} of (11) goes over into the form of (3). In that case the absorption was attributed entirely to scattering and was found by multiplying the secondary radiation from one drop by the number of drops per unit volume. Now the absorption coefficient will be determined from the complex dielectric constant and permeability of the suspension of water in air. The absorption of the radiation in water is taken into account through the complex form of the index of refraction and the scattering by the imaginary term in (20). It may be remarked that the fourth-order terms in the expansions (20) and (21) are both real.

It is necessary to distinguish carefully between the field intensities and the exciting forces in a medium. The electric and magnetic field intensities at any point determine the Poynting vector or energy flow at that point. They will be indicated by E and H, respectively. The exciting forces E' and H' are forces which excite the electric and magnetic dipoles into oscillation and emission of secondary waves. The two will be identical only in the case of vanishingly small concentrations.

The relation between the exciting forces and the field intensities, for any concentration, is

$$E' = E + \frac{4\pi c^2}{3} P. \tag{22}$$

$$H' = H + \frac{4\pi}{3} M. \tag{23}$$

Here P is the electric polarization per unit volume, or the product of the average electric dipole moment and the number of dipoles per unit volume. The magnetic polarization M is defined analogously. In the present notation

$$E_x' = E_x + \frac{4\pi}{3} N\rho^3 a E_x' \tag{24}$$

$$H_y' = H_y + \frac{4\pi}{3} N\rho^3 b H_y' \tag{25}$$

$$E_x' = \frac{E_x}{1 - \dfrac{4\pi}{3} N\rho^3 a} \tag{26}$$

$$H_y' = \frac{H_y}{1 - \dfrac{4\pi}{3} N\rho^3 b} \tag{27}$$

The relations between electric and magnetic displacements and their corresponding field intensities are

$$\begin{aligned} D &= \epsilon E = \epsilon_0 E + 4\pi P \\ B &= \mu H = H + 4\pi M \end{aligned} \tag{28}$$

ϵ_0 is the dielectric constant of air in e.m.u. and ϵ and μ the complex dielectric constant and permeability respectively of the suspension.

$$\epsilon = \epsilon_0\left[1 + \frac{4\pi N\rho^3 a}{1 - \frac{4\pi}{3}N\rho^3 a}\right] \tag{29}$$

$$\mu = 1 + \frac{4\pi N\rho^3 b}{1 - \frac{4\pi}{3}N\rho^3 b} \tag{30}$$

The equation for the plane, homogeneous waves considered here is satisfied by

$$E_x = E_0 e^{2\pi\nu i(t - \sqrt{\epsilon\mu}z)} \tag{31}$$

$\sqrt{\epsilon\mu}$ is the phase velocity only in the case of real values of ϵ and μ.

Put

$$\sqrt{\epsilon\mu} = \frac{n - i\kappa}{c} \tag{32}$$

$$E_x = E_0 e^{(-2\pi\kappa z/\lambda) + \omega i(t - nz/c)}. \tag{33}$$

From (32) it is easy to show that

$$\kappa^2 = \frac{c^2}{2}[\sqrt{[Re(\epsilon\mu)]^2 + [Im(\epsilon\mu)]^2} - Re(\epsilon\mu) \tag{34}$$

where R_a and Im stand for the real and imaginary parts respectively. In gaussian units, in which $\epsilon_0 = 1$, we have

$$\kappa^2 = \tfrac{1}{2}[\sqrt{[Re(\epsilon\mu)]^2 + [Im(\epsilon\mu)]^2} - Re(\epsilon\mu)]. \tag{35}$$

Putting $d = 4\pi N\rho^3/3$ and expanding ϵ and μ according to powers of d we obtain

$$\epsilon = 1 + 3da + 3d^2a^2 + \cdots. \tag{36}$$

$$\mu = 1 + 3db + 3d^2b^2 + \cdots. \tag{37}$$

Recourse to the physical data again shows that powers of d higher than the first may be neglected, and the terms retained give for $\epsilon\mu$

$$\epsilon\mu = 1 + 4\pi N\rho^3\frac{m^2 - 1}{m^2 + 2}\left(1 + \frac{3}{5}\alpha^2\frac{m^2 - 2}{m^2 + 2} + \frac{\alpha^2}{30}(m^2 + 2) - \frac{2i\alpha^3}{3}\frac{m^2 - 1}{m^2 + 2}\right). \tag{38}$$

In taking the real and imaginary parts of (38) it must be recalled that if the water absorbs the wavelength considered, m is complex. Experimental data to be given later on show that selective absorption may be neglected for wavelengths of the order of 50 mm or more and m is real. On the basis of this assumption we find

$$Re(\epsilon\mu) = 1 + 3d\frac{m^2 - 1}{m^2 + 2}\left[1 + \frac{3}{5}\alpha^2\frac{m^2 - 2}{m^2 + 2} + \frac{\alpha^3}{30}(m^2 + 2)\right]. \quad (39)$$

$$Im(\epsilon\mu) = -2d\alpha^3\left(\frac{m^2 - 1}{m^2 + 2}\right)^2. \quad (40)$$

Evidently $Re\ (\epsilon\mu) > Im(\epsilon\mu)$ and κ^2 may be expanded by the binomial theorem. Expanding and neglecting powers of d higher than the first we obtain

$$\kappa = \frac{Im(\epsilon\mu)}{2} = \frac{32}{3}\frac{\pi^4\rho^6 N}{\lambda^3}\left(\frac{m^2 - 1}{m^2 + 2}\right)^2. \quad (41)$$

The absorption coefficient due to scattering is then

$$\sigma = \frac{2\pi\kappa}{\lambda} = \frac{64\pi^5\rho^6 N}{3\lambda^4}\left(\frac{m^2 - 1}{m^2 + 2}\right)^2. \quad (42)$$

This is identical with formula (9), from which it may be concluded that the simple assumptions of the Rayleigh method are justified for wavelengths greater than 40 to 50 mm. For smaller values of λ the complex form of m must be taken into account in (38), and if the investigation is continued for values of λ less than about 5 mm, higher powers of α must be included in the derivation of (38).

Discussion of Results

In Table III are summed up the results of numerical computations of scattering coefficients. Three typical cases are chosen, representing the entire range of meteorological conditions. In the columns headed Z are given the distances in kilometers that the wave must travel in order that the intensity be reduced to one-tenth of the original value.

TABLE III

	$\lambda = 100$ cms		$\lambda = 50$ cms		$\lambda = 10$ cms		$\lambda = 5$ cms	
	σ	Z	σ	Z	σ	Z	σ	Z
Cloudburst	8.86×10^{-13}	2.6×10^{9}	1.4×10^{-11}	1.6×10^{6}	8.86×10^{-9}	2.6×10^{5}	1.4×10^{-7}	1.6×10^{2}
Moderate rain	16.9×10^{-16}		2.7×10^{-14}		16.9×10^{-12}	1.3×10^{6}	2.7×10^{-10}	8.5×10^{1}
Fog							5.7×10^{-18}	4×10^{12}

It is evident from these calculations that the scattering of radio waves of length greater than about 5 cm is absolutely negligible.

Beyond this point it increases very rapidly. For example, neglecting for the moment the selective absorption which also sets in in this region, we find corresponding to $\lambda = 1$ cm in moderate rain, a value of Z equal to 1.3×10^2 km.

In Table IV[8] is to be found data on dispersion of electromagnetic waves in water.

TABLE IV

λ in cm	n	κ
0.42	5.33	1.28
0.84	5.68	1.49
1.10	6.27	1.44
1.50	6.62	1.83
1.80	6.65	2.32
2.70	8.45	2.26

Here n is the index of refraction and κ the coefficient of absorption. They are related to the dielectric constant by

$$\epsilon = n^2 - \kappa^2.$$

It will be observed that water absorbs strongly in the region about $\lambda = 2$ cm. Absorption bands have been found for considerably lower frequencies but the magnitude of the corresponding absorption coefficients is small.

The general conclusions to be drawn are that rain and fog have no effect on the propagation of radio waves of the frequencies now employed. Rain first begins to markedly influence the propagation of waves less than 5 cm in length. Thereafter the absorption increases very rapidly with decreasing wavelength due both to scattering and the selective absorption of the water molecule. Infra-red radiation should not, therefore, prove satisfactory for communication purposes through heavy fog. The absorption coefficient for this case has not been calculated, but may be readily obtained from the formulas. Although the wavelengths in the infra-red region are very much shorter than those considered here, the condition that $\alpha^4 << 1$ is not violated since the radius of a rain drop is about 100 times that of a fog particle. For fog the formulas given in this paper are valid for wavelengths exceeding about 0.05 mm, provided the complex form of the index of refraction m is used.

From (29) may be calculated the coefficient of reflection of the bounding surface of a cloud or bank of fog. The amount of energy reflected in such a case proves to be negligible, as has already been shown by Pederson.[9]

[8] J. D. Tear, *Phys. Rev.* 21, 611; 1923. Cf. also the article by W. Romanoff in the *Handbuch der Physik*, XV, 1927.

[9] P. O. Pederson, "The Propagation of Radio Waves," Copenhagen, 1927.

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METEOROLOGICAL INFLUENCES ON LONG-DISTANCE, LONG-WAVE RECEPTION*

By

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Summary—*(1) The received field intensity was less affected by the variations of the meteorological elements in the environs of transmitting stations than by those of the receiving station. (2) The field intensities for both daylight and night reception varied inversely with the changes of atmospheric temperature and absolute humidity on the receiving side. This was in agreement with the daylight results obtained by Austin and Minohara. (3) The intensity-pressure relation was not found to be so apparent as in the case of (2), though in the monthly average variations the field intensity seemed to have direct relation with atmospheric pressure in summer and inverse relation in winter. (4) The influence of weather on the field intensity was found to be still less clear than in the cases of the above-mentioned three meteorological elements.*

GENERAL CONSIDERATIONS

THE QUESTION as to whether the meteorological elements, such as atmospheric temperature, humidity, and pressure, affect the propagation of electromagnetic waves has already been discussed by a number of physicists and radio engineers. Some correlations of the reception of radio signals with these elements have also been analytically studied by the authors from the results of a series of the field-intensity measurements[1] which were conducted respectively at a certain fixed time of day during the period of more than one year extending from the autumn of 1926 to January, 1928, for several stations whose transmitting wavelengths were from 10,000 to 20,800 m, the distances between the transmitting and receiving stations being as far as from 3,000 to 11,000 km.

To make a competent investigation of such a nature, however, it will be necesary to know the meteorological conditions over the entire route of wave propagation. It will be extremely difficult to gather such data, especially if the existence of waves propagating through an upper atmosphere is taken into consideration. In the present study, therefore, meteorological elements in the environs of transmitting and receiving stations are merely referred to.

* Dewey decimal classification: R113.5. Presented before the World Engineering Congress, Tokio, Japan, October, 1929.

[1] Yokoyama, E. and Nakai, T., "The measurements of the field intensities of some high-power long-distance radio stations. Part I: Bolinas and Bordeaux." Electrotech. Lab. Researches No. 299, June, 1928; Ditto, Part II: Malabar, Palao, and Rugby, No. 233, July, 1928; Ditto, Part III: Kahuku, Pearl Harbor, and Saigon, No. 238, Sept., 1928; Ditto, Part IV: Warsaw, Tananarive, and Monte Grande (in press.)

As the results of their analytical investigations, it has been found that the received field intensity was less affected by the meteorological elements in the localities of transmitting stations than by those of the receiving station. As one of the examples, the results of measurements for Kahuku are shown in Fig. 2 indicating that the field intensities vary markedly in accordance with the variations of the three meteorological elements on the receiving side in spite of no appreciable variations of those on the transmitting side. In this paper, the relation between the field intensities and meteorological elements on the receiving side is mainly considered.

Among many observed stations, only six stations—Bolinas, Bordeaux, Kahuku, Malabar, Rugby, and Warsaw—are selected by the authors for comparison; the remaining stations have been considered unsuitable because of some other factors masking the phenomena or because of lack of data sufficient for this particular analytical study.

For those stations selected above, the correlations of both the monthly average and the daily variations of field intensities with meteorological elements are considered. Of these two cases, the latter relations are not found so distinctly as the former. It seems to be a natural consequence to induce the latter relations to be of vaguer results than the former, as there are several other factors affecting field intensity, such as solar constant, sunspot, terrestrial magnetism, etc. If a study for the correlations of field intensities with meteorological elements could, however, be made throughout an entire path of wave propagation, they might be shown more clearly.

If a certain intercorrelation be found among the meteorological elements themselves, all the relations under consideration will be clearly obtained by finding only that of the field intensity with one of the elements. As temperature is closely related to humidity, the relation of field intensity with the latter element will be omitted from the present discussion, but as temperature is not closely related with pressure, the relation of field intensity with pressure will also be discussed in equal detail.

Correlations Between the Monthly Average Variations

The monthly average variations of the received field intensities from two of the six observed stations and corresponding three meteorological elements on the receiving side are plotted as shown in Figs. 1 and 2.

For Bordeaux, Rugby, and Warsaw, the measurements were made respectively at a certain fixed time of day when the influence of the seasonal shiftings of the sunrise and sunset times on the values of received field intensities had been found by some experiments to be practically negligible. In the cases shown in Fig. 1, for example, field intensity varied inversely with temperature, while it had a tendency

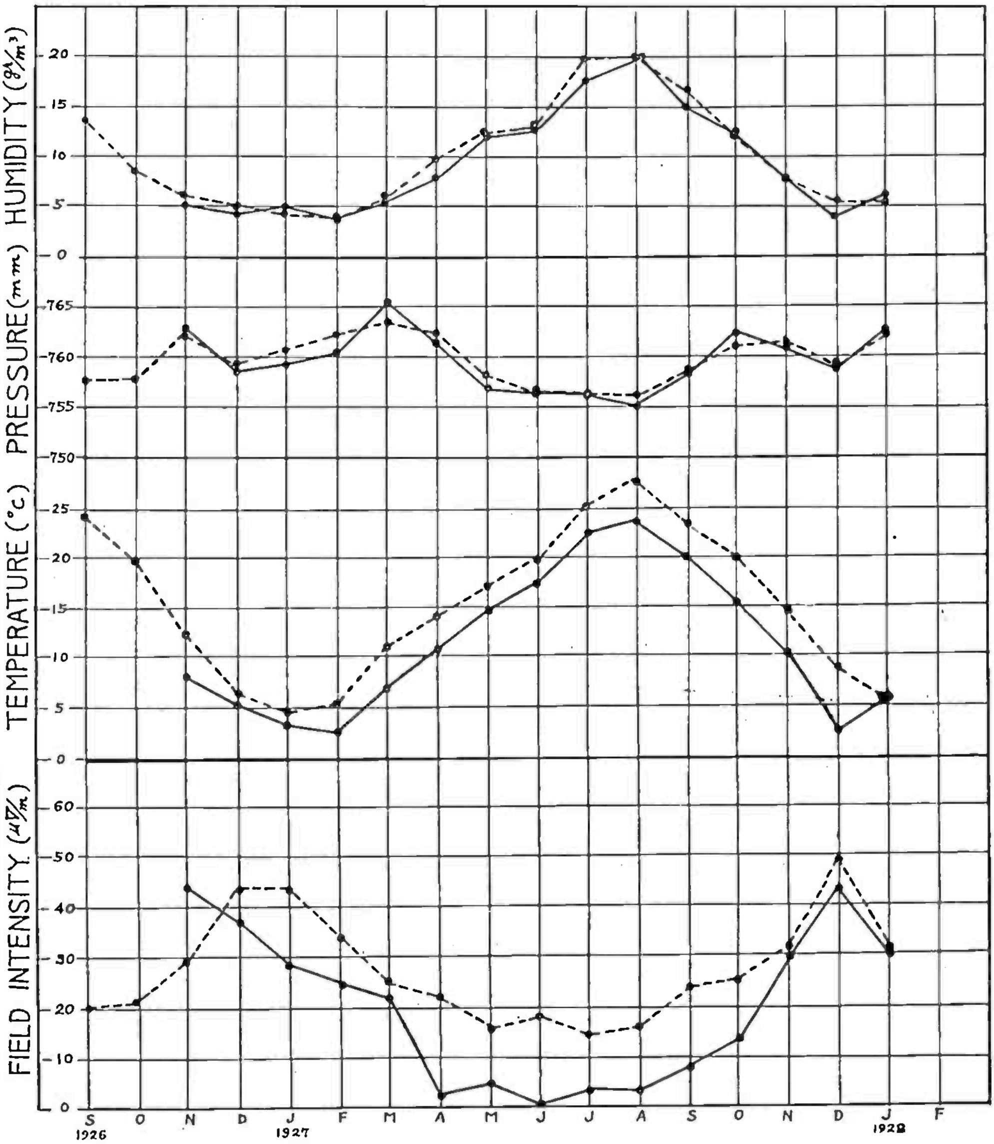

Fig. 1—Monthly average variations of field intensities and meteorological elements.

Bordeaux ●- - - - - - - -● 10:40 A.M., J. C. S. T.
●————————● 8:20 P.M., J. C. S. T.

to change with pressure directly in summer and inversely in winter. The intensity-pressure relation was not found so clearly as in the case of the intensity-temperature relation. The authors wish to point out here that each experiment mentioned above can be treated as all-day or all-night transmission, though the daylight transmission includes a slight portion of the night transmission at one end, and vice versa.

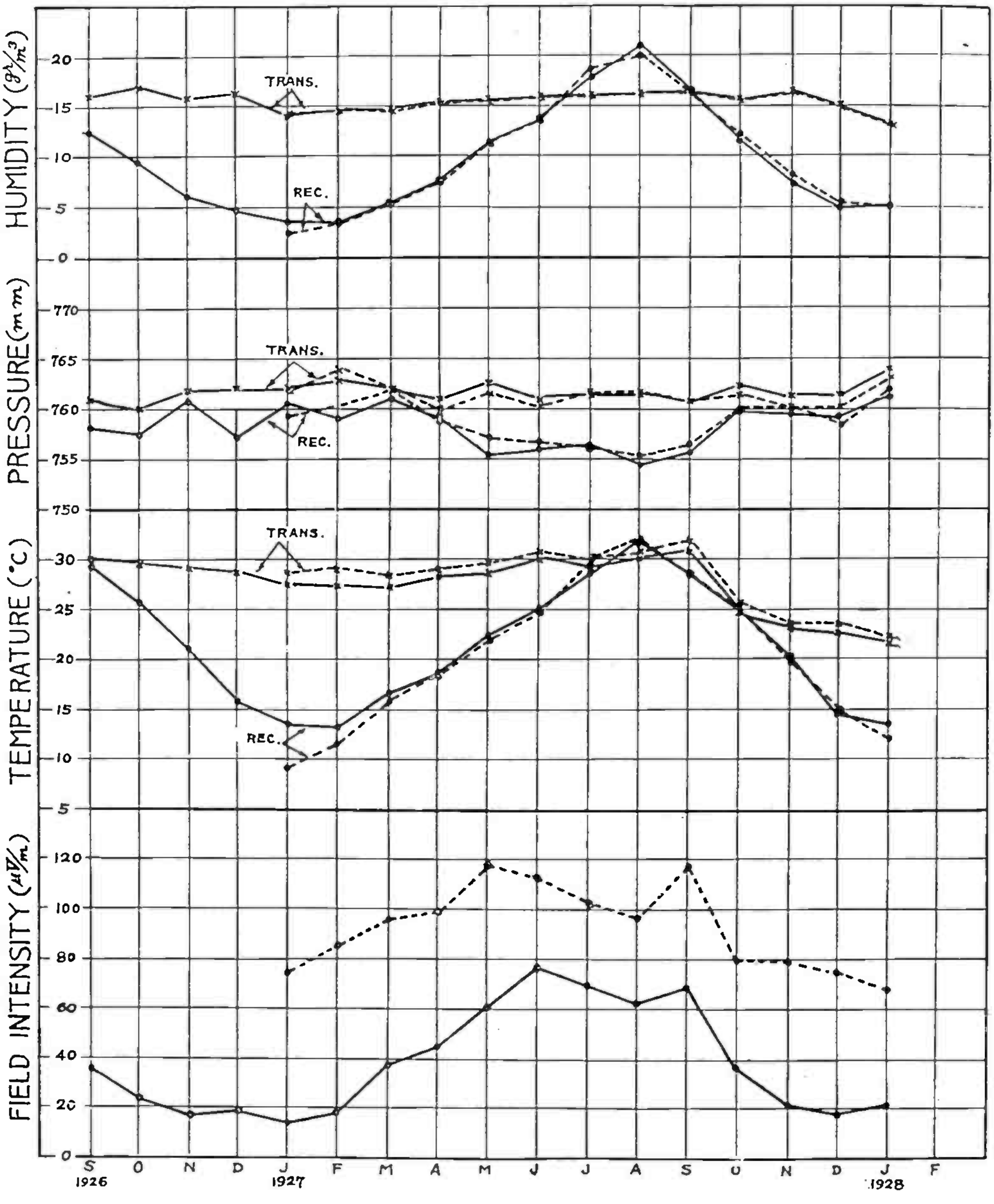

Fig. 2—Monthly average variations of field intensities and meteorological elements.

Kahuku ●- - - - ● , x- - - - x 0:00 P.M., J. C. S. T.; ●——● , x——x 2:00 P.M., J. C. S. T.

For another group of the stations, Bolinas, Kahuku, and Malabar, for example, as seen from Fig. 2, all the relations were just the reverse of those of Bordeaux and others. It seems to result from the fact that the measurements happened to be carried out at a time when the sunrise and sunset gave an appreciable influence on the values of field intensity. On the other hand there are, however, some reasons to attribute it to the fact that the reception from the former group of stations is of overland, whereas that from the latter group is of oversea. A further discussion on this particular point will be left for some other occasion. All the experiments for Bolinas and others were conducted under a condition of either all-day or all-night.

Though, as stated above, the measurements were carried out in both daytime and nighttime, no apparent difference was noticed in the relations between the two.

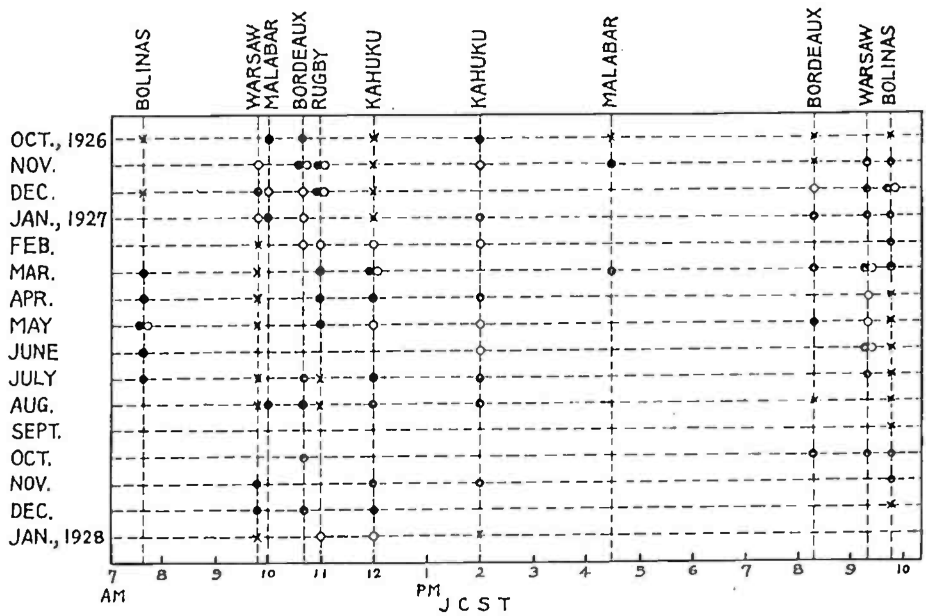

Fig. 3—Correlation of field intensitiy with temperature on the receiver side.
○ Cases in direct relation
● Cases in inverse relation
x No observation

Correlations Between the Daily Variations

For the comparative study of the daily variations, the deviations from the monthly average values are considered only in order to eliminate the effects of sunrise and sunset.

Figs. 3 and 4 are plotted to show respectively how the deviations in the field intensity and the corresponding temperature or pressure were, for month, and for day and night, distributed all over the observed period. When an inverse relationship was found for a certan number of days in a month, a black circle has been marked for that month regardless of the length of duration, while a white circle has been marked when direct relationship was found. The accompanying table shows the percentage ratios of the number of days on which the received field intensity varied directly or inversely with temperature or pressure, to the total number of days observed.

In view of the fact that some possible errors might come in from the field-intensity measurements, only such particular parts have been counted in Figs. 3 and 4, as well as in the table, that the differences in the field intensities observed on any two consecutive days were greater than 20 per cent during the period from June to September when the atmospherics are prevalent, and greater than 10 per cent during the remaining period of the year. This lasted over four consecutive days observed. The maximum range of variation in field intensity which is inversely related with temperature was found to be nearly 100 per cent of monthly averages. An example illustrating the field-intensity-temperature relation is shown in Fig. 5 where the inverse relation is recognized.

Name of Stations	Time of Measurements J. C. S. T.	I/T in Percent		D/T in Percent	
		Temperature	Pressure	Temperature	Pressure
Bolinas	7:40 A.M.	21	30	3	13
	9:50 P.M.	29	7	6	20
Bordeaux	10:40 A.M.	16	4	14	7
	8:20 P.M.	21	19	4	8
Kahuku	0:00 P.M.	39	6	10	13
	2:00 P.M.	30	27	12	11
Malabar	10:00 A.M.	6	11	2	5
	4:30 P.M.	9	13	0	28
Rugby	11:00 A.M.	16	15	9	10
Warsaw	9:50 A.M.	26	19	8	0
	9:20 P.M.	17	8	9	7

J. C. S. T. = Japan Central Standard Time
I = Number of days on which field intensity varied inversely with temperature or pressure
D = Ditto directly
T = Total number of days observed

As seen from Fig. 3, the intensity-temperature relations were not found so clearly as in the case of the monthly average variations, part changing directly, and part, inversely. It is, however, seen from the table that a greater percentage ratio has always been found in the inverse relation than in the direct relation, irrespective of the overland and oversea stations, which is not in agreement with the results obtained for monthly averages, though, for a majority of the total number of days observed no relation could be found between field intensity and temperature.

Also, no difference was noticed for the above relation in the observations in both day and night as in the case of monthly averages.

As shown in Fig. 4 and the table, the intensity-pressure relations were still vaguer, nearly one half changing directly and the other half changing inversely, and indicated no such similarity as in the case of the monthly average variations, where the field intensity varied directly with pressure in summer and inversely in winter. However, to venture a closer analysis, what is in inverse relation seem to be slightly greater in the daylight observations, while the case is reversed at night.

Although it might be one of the essential points of this investigation to determine whether the daily variations of the field intensities and of the meteorological elements be in leading or lagging phase relations, this kind of analysis could not be done, since the above

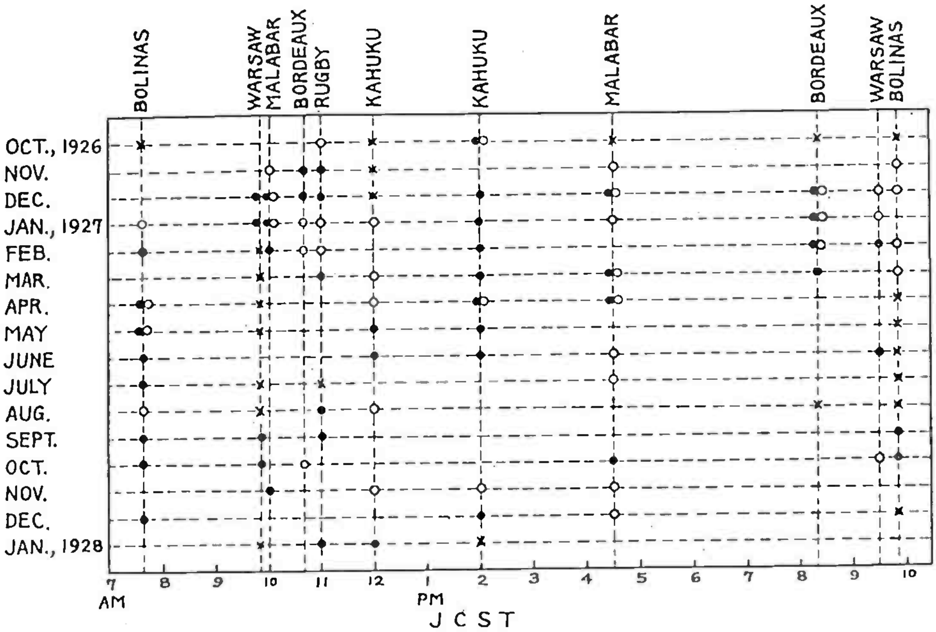

Fig. 4—Correlation of field intensity with pressure on the receiver side.
○ Cases in direct relation
● Cases in inverse relation
x No observation

measurements could not be conducted day by day but done desultorily about ten times a month.

It has already been known that in Japan, temperature and absolute humidity have similar tendencies in variations. Even in that relation the analytical study of the present cases reveals the fact that the number of days in the normal relation was only 37 per cent in daytime and 62 per cent at night of the total number of days observed, though the number of days in inverse relation was as small as 3 per cent and 0 per cent respectively. As such was the case even in the two elements which are on direct causual dependence, it may also be said that the relations between the field intensities and the three meteorological elements are unexpectedly clear.

Influence of Weather

The relations of field intensity with the change of weather were far less distinct than those with the three meteorological elements. Nevertheless, the increase of signal strength due to the change of weather from "cloudy" or "rainy" to "clear," and the decrease of the same on account of the change of weather from "clear" to "rainy" was noticed comparatively well, but even in those cases, they only reached a percentage as small as 17 per cent against the total number of times for the changes of weather. It must be mentioned that the definition distinguishing "clear," "cloudy" or "rainy" was not of a scientific

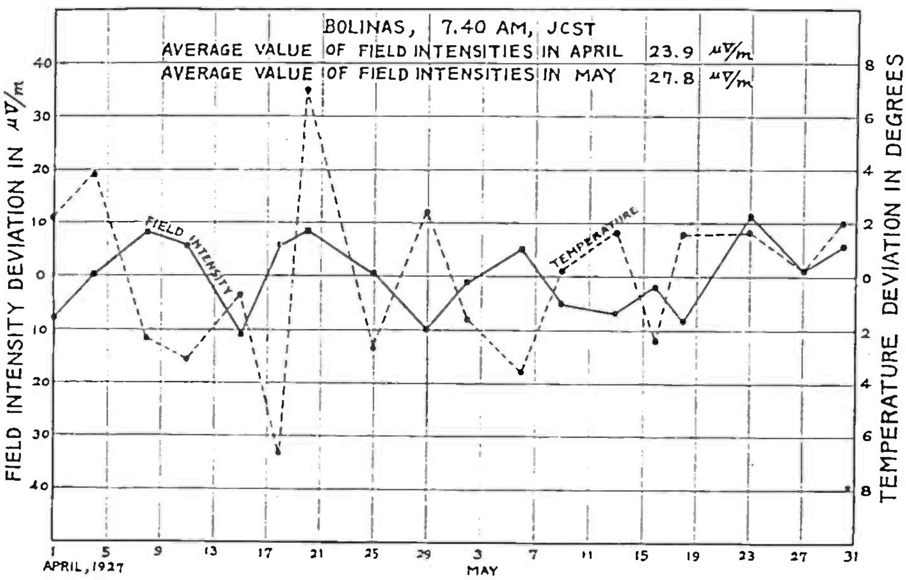

Fig. 5—An example of field-intensity temperature daily relation.

nature, but it was conventionally defined by mere amateur observations. Such obscure relations will naturally result from the fact that, though there is a close relation between weather and atmospheric pressure, a relation is scarcely found between field intensity and pressure.

Results Obtained by Other Investigators

One of the authors' conclusions that the field intensity varies inversely with atmospheric temperature on the receiving side, not only in monthly averages but also in daily variations, is in agreement with the results of Austin[2] who studied the phenomenon with daylight waves of 13,600 and 15,900 m at short distances of 281 and 251 km,

[2] Austin, L. W., and Wymore, I. J., "Radio signal strength and temperature," Proc. I.R.E., 14, 781; December, 1926.

and with those of Minohara[3] who did the same with daylight waves of 11,490 m at a distance of 6,400 km. Pickard,[4] however, obtained a different result with night waves of the broadcast frequency band at distances from 640 to 1,120 km, that the daily variations of field intensity are directly in accordance with those of temperature, and inversely with those of pressure on the receiving side. In the authors' experiments, the results were both on the contrary (Figs. 3 and 4), though the intensity-pressure relation was not found so clearly as stated above. The question must be left for a future study as to whether they are caused by the difference in wavelengths or not. On the other hand, Bureau[5] stated that field intensity changed when a receiving station or a transmitting station or both were attacked by anti-cyclons, or when the party stations were separated by a surface of discontinuity. This kind of study is also under the consideration of the authors.

The following basic questions are still unsolved: Is one of the three meteorological elements the direct cause to change the field intensity or do the other two affect it indirectly? Or is there anything else which will be the chief cause of the effect?

Acknowledgment is due to Prof. Fujiwara, meteorologist, who kindly gave the authors valuable advice in the course of the present study.

[3] Minohara, T., "On the radio field intensity of time signal sent out by the Pearl Harbor Station, observed at Tokio," *Jour. I. E. E.* (Japan), 47, No. 464, 225; March, 1927.

[4] Pickard, G. W., "Some correlations of radio reception with atmospheric temperature and pressure," PROC. I.R.E., 16, 765; June, 1928.

[5] Bureau, R., "Les influences meteorologiques sur la propagation des ondes," U. R. S. I., Document No. 51, December, 1924.

BOOK REVIEW

The Radio Manual, 2nd edition, by George E. Sterling, edited by Robert S. Kruse. Published by D. Van Nostrand Company, Inc., 250 Fourth Ave., New York. 797 pp., price $6.00.

"The Radio Manual" is a strictly non-mathematical book with the strongest appeal to those who expect to enter the radio profession as operators or inspectors, although it is addressed to engineers and service men as well. It answers a multitude of questions on the installation, upkeep, operation, and control of nearly every form of commercial radio apparatus and equipment. It affords detailed information on broadcast transmitting, commercial transmitting and receiving, and radio technique in general.

The first six chapters are devoted to a discussion of elementary electricity and magnetism, motors and generators, storage batteries, vacuum tubes, fundamental vacuum-tube transmitting circuits, and modulating systems. This is followed by a short chapter on frequency measurements and field-strength measuring apparatus. Chapters 8 to 11 deal with transmitters of commercial types, marine vacuum-tube transmitters, W. E. broadcast transmitters, arc and spark transmitters. Commercial radio receivers are described in the next chapter. The auto-alarm described in Chapter 13 is a device to waken the operator when no one is on watch in order that he may receive an SOS call. Radio direction finders and their operation as used by U. S. Lighthouse Service are next discussed. This is followed by a chapter on aircraft radio equipment, including the radio-beacon development of the Bureau of Standards, and different types of receivers and transmitters for aircraft. The remaining chapters are devoted to amateur short-wave apparatus, television and radio movies, radio interference, radio laws, and the handling and abstracting of traffic.

The general method of treatment is to give detailed descriptions, directions for operation, maintenance and trouble shooting for typical commercial apparatus such as R.C.A. marine transmitters, W.E. 5-kw broadcast transmitter, and Federal arc transmitters.

The principal errors in the book are the figures of Chapter 2 which show the field poles of the motors and generators incorrectly wound.

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BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

Two new bulletins of Jenkins and Adair, Inc., 3333 Belmont Ave., Chicago, Ill., have recently become available. Bulletin No. 11 describes a condenser transmitter control panel, with which the broadcast operator may control three condenser microphone positions. Bulletin No. 12 lists a line equilizer for 500-ohm broadcast or telephone transmission lines.

Norden-Hauck, Inc., Philadelphia, has available for distribution a data sheet describing their Super DX-5, an a-c operated high-frequency receiver using a pentode in the radio-frequency amplifier, and having a push-pull audio amplifier.

A copy of the data sheet describing a test panel for service men may be obtained from Norden-Hauck, Inc., Delaware Ave. and South St., Philadelphia.

A leaflet describing their new receiving equipment may be obtained by addressing a request to Pierce-Airo, Inc., 117 Fourth Ave., New York City.

The Condenser Corp. of America, Jersey City, N. J., issues a leaflet describing by-pass condensers for radio receiving equipment.

Bulletin 90 of the United Scientific Laboratories, 113 Fourth Ave., New York City, describes midget and ganged condensers for manufacturing assemblies.

A 20-page catalog illustrating the use of pyrex radio insulators, and a folder illustrating the more important pyrex insulators devoted to aviation installations may be obtained from the Corning Glass Works, Corning, N. Y.

"About Vitreosil", a 16-page booklet describing the properties and applications of fused silica, and a price list, "Vitreosil for the Laboratory" will be sent to anyone sending a request for these publications to the Thermal Syndicate Ltd., 58 Schenectady Ave., Brooklyn, New York.

"Polymet Engineering Manual" a loose-leaf folder giving dimensions, specifications, and engineering data on fixed condensers, fixed and variable resistors, and form wound coils, will be sent to executives requesting this material on their letterheads. Requests should be addressed to Polymet Manufacturing Co., 829-J E. 134th St., New York City.

Herbert H. Frost, Inc., of Elkhart, Ind., has for distribution an 8 page catalog describing standard Frost radio products as carried in stock and ordinarily distributed through jobbers. A 4-page catalog describing merchandise built to order for manufacturers is also available.

The numerous uses of variable resistors are outlined in the catalog of the Central Radio Laboratories, 16 Keefe Ave., Milwaukee, Wis. Copies of this booklet are available.

A folder containing sets of blueprints of the radio tube parts maufactured by Sigmund Cohn, 44 Gold St., New York, will be sent to any interested reader upon request.

A 32-page copy of "The Gateway to Better Radio" may be obtained by addressing the American Mechanical Laboratories, Inc., 285 North Sixth St., Brooklyn, N. Y.

Price List No. 21 of A. M. Flechtheim & Co., 136 Liberty St., New York City, lists fixed condensers for replacement and manufacturing purposes.

"Radio Simplified", a 32-page booklet which gives a popular outline of the manufacture and use of vacuum tubes, and bulletins listing the complete line of De Forest transmitting and receiving tubes, are available from the De Forest Radio Co., Passaic N. J.

Several leaflets have recently been received from the Dubilier Condenser Corp., 342 Madison Ave., New York, N. Y. One of these is a descriptive folder and price list of fixed condensers for receiving and transmitting equipment; a separate sheet lists replacement condenser blocks for power-packs and A and B eliminators. A filter network for the reduction of interference in radio receivers is described in Form No. 164. A mimeographed leaflet contains technical information concerning high-voltage condensers for radio transmitting equipment or carrier current transmission while another mimeographed leaflet describes an untuned radio-frequency transformer for use with broadcast amplifiers using screen-grid tubes.

A complete line of dynamic loud spearkers for the home, theatre, or auditorium is described in a folder issued by the Oxford Radio Corp., 2035 Pershing Place, Chicago, Ill.

"Audio Transformers by Sangamo" is the title of a booklet issued by Sangamo Electric Co., Springfield, Ill., and dealing with the characteristics of audio-amplifying transformers and chokes. Mica-insulated receiving and transmitting condensers of small capacitance are also listed in the booklet.

A custom-built receiver is described in the Hammarlund HiQ-30 Manual, obtainable for twenty-five cents from Hammarlund-Roberts, Inc., 424 W. 33rd St., New York, N. Y.

The International Resistance Co., 2006 Chestnut St., Philadelphia, Pa., publishes at intervals a number of engineering bulletins which will be sent to readers upon request as the bulletins are issued. The following bulletins have been issued to date:

"Concerning Precision Audio Resistance Amplifiers for Television and Laboratory Experimenters," Engineering Bulletin No. 4.

"Loud Speakers," Engineering Bulletin No. 5.

"Dynamic Loud Speakers," Engineering Bulletin No. 6.

"A New Precision Resistance-Coupled Amplifier and Associated Power Supply," Engineering Bulletin No. 7.

"Fixed Resistor for Replacement Problems," Eng. Bul. No. 8.

"Precision Wire-Wound Resistors," Engineering Bulletin No. 9.

A series of pamphlets is in preparation by the Allen D. Cardwell Manufacturing Corp., 81 Prospect St., Brooklyn, New York. The following pamphlets, which will be mailed upon request, are now ready for distribution: "Cardwell Receiving Condensers," "Cardwell Transmitting Condensers," "Cardwell High-Voltage Condensers."

Rubber-covered wire and cables are described in the "Insulated Electrical Wires and Cables" catalog of the Rome Wire Co. of Rome, N. Y. Their catalogue, "Super Service Cord and Cable" concerns rubber-covered cables intended to have long life under unusually severe conditions. Both catalogues may be obtained upon request.

The following extracts of articles which have recently appeared in various journals and magazines have been made available in convenient loose-leaf form and may be obtained from Lefax, Ninth and Sansom Streets, Philadelphia: "The Pentode Tube," "Constructing a Modulated Oscillator," "Audio-Frequency Amplification With Screen-Grid Tubes," "A Dry Cell Screen-Grid Tube," "Some Facts About Acoustics in Cabinets," "Revolutioning High-Frequency Tuner Design," "Public-Address and Centralized Radio Systems," "Additions and Revisions to Broadcasting Stations."

A 24-page "Vest-Pocket Manual of Phenolite," obtainable upon request from the National Vulcanized Fibre Co., of Wilmington, Del., describes some of the numerous applications of laminated bakelite.

Bulletin No. 830, of the Roller-Smith Co., 233 Broadway, New York, lists a number of recording instruments. Graphic ammeters, voltmeters, and millivoltmeters for direct current, and ammeters, voltmeters, wattmeters, and power-factor meters for alternating current are listed. A tester for dry cells is described in Supplement No. 2 of Bulletin No. 210 of the same firm.

The Eisler Electric Corp., 744 South 13th St., Newark, N. J., has recently issued a catalog listing the various machines employed in the manufacture of vacuum tubes, photoelectric cells, neon sign lamps, incandescent lamps, and similar glass products.

Rheostats and potentiometers, with current carrying capacities up to 150 watts, and with resistor units up to 2000 ohms are described in a recent bulletin of the DeJur-Amsco Corporation. Copies of this bulletin will gladly be sent to those addressing a request to 418 Broome St., New York N. Y.

A mimeographed sheet giving specifications for high-voltage transformers for vacuum-tube plate-supply systems and a blue-print showing the regulation of these transformers has been received from the Hilet Engineering Co., of Orange, N. J. Another mimeographed sheet gives specifications of choke coils for filter systems of plate-supply systems. These data will be mailed upon request to those interested in this equipment.

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE

THIS IS a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to the professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The various articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. Radio Communication

R007.1 Caldwell, L. G. Practice and procedure before the Federal Radio Commission. The Journal of Air Law, 1, pp. 144–185; April, 1930.

(The powers and duties of the Federal Radio Commission under the Radio Act of 1927 as amended are classified and reviewed. The procedure followed by the Commission in exercising its quasi-legislative and its quasi-judicial functions is explained at length. The treatment is critical as well as explanatory.)

R007.9 Première réunion du Comité Consultatif International de Radioelectricité. (First meeting of the International Consulting Committee on Radio.) *L'Onde Electrique*, 9, pp. 75–92; February, 1930.

(A report is given of the work of the International Consulting Committee on Radio at the Hague Conference, 1929. (Concluded.))

R100. Radio Principles

R111 Kaplan, C. and Murnaghan, F. D. On the fundamental constitutive equations in electromagnetic theory. *Phys. Rev.*, 35, pp. 763–777; April 1, 1930.

(The absolute significance of the constitutive relations in electromagnetic theory is investigated. Various assumptions as to the isotropic character of the space-time continuum are made to examine what simplifications are thereby introduced.)

R113 Potter, R. K. Transmission characteristics of a short-wave telephone circuit. Proc. I.R.E., 18, pp. 581–648; April, 1930.

(A method of observing and recording the audio-frequency transmission characteristics of a short-wave radiophone channel is described. The characteristics are found to undergo rapid changes, apparently the result of wave interference and progressive changes in the angle of rotation of the polarization plane with frequency over the signal band. It is indicated that changes in the transmission path are progressive rather than erratic. Various types of distortion and fading are discussed at length and seasonal effects are noted. Many specimen records of double- and single-side-band transmissions are given.)

R113 Kenrick, G. W. and Pickard, G. W. Summary of progress in the study of radio-wave propagation phenomena. Proc. I.R.E., 18, pp. 649–668; April, 1930.

(Recent progress in the study of radio-wave propagation phenomena is reviewed in the light of the history of the art. The development of the art through 1927 is outlined. The discussion of recent advances concerns publications on the Störmer-van der Pol echoes and their interpretation, progress in Kennelly-Heaviside layer height determinations, and experimental studies in transmission and magnetic and solar correlations. The need of further consistent observations and other means of investigation is pointed out. A bibliography is included.)

R113.3 Tanimura, I. Some experiments on night errors for long waves. Proc. I.R.E., 18, pp. 718–722; April, 1930.

(The results of experiments on night errors observed for a 19.7-kc station located at a distance of 148 km are described. They are compared with the results of a theoretical analysis following methods used by Eckersley, and good agreement is found. Cyclic variations of bearings are noticed at sunset and sunrise, the maximum shifts being about 30 deg. At the moments of maximum shift the bearings are distinct while they are broad at other moments.)

R113.4 Appleton, E. V. On some measurements of the equivalent height of the atmospheric ionized layer. *Proc. Royal Soc.* (London), 126(a), pp. 542–69; March 3, 1930.

(The results of a series of early morning measurements of the equivalent height of the ionized layer for 400-m waves are recorded and discussed. Measurements were made by the "frequency-change" method, the theoretical basis of which is analyzed. Deviations from the normal type of nocturnal variation of equivalent height are noted, and a theory of regional distribution of ionization is presented as a possible explanation of the deviations.)

R113.6 Strutt, M. J. O. Reflexionsmessungen mit sehr kurzen elektrischen und mit akustischen Wellen. (Reflection measurements with very short electrical and acoustical waves.) *Elek. Nach. Tech.*, 7, pp. 65–71; February, 1930.

(Field-intensity measurements were made in a vertical plane of the radiation from a high-frequency electric oscillator and from an acoustic oscillator. The effects of the ground conditions and intervening objects between the transmitter and receiver were observed. A short explanation of the relationship between the direct and reflected waves according to the theory of Sommerfeld is included.)

R125.1 Smith-Rose, R. L. Radio direction finding by transmission and reception. *Nature* (London), 125, pp. 530–532; April, 5, 1930.

(The principles of radio direction finding and their application to navigation are outlined. Errors encountered in practice due to spurious electromotive forces, antenna effects, and to special conditions in the transmission path are discussed. Results of tests are given to indicate the reliability of radio direction finding as applied to marine navigation. Special attention is given to night error and its causes.)

R133 Sears, F. W. Integraph solutions of electron orbits in the Barkhausen-Kurz effect. *Jour. Franklin Institute*, 209, pp. 459–472; April, 1930.

(The continuous integraph is used to trace the orbits of an electron oscillating within the filament-plate space of a 3-electrode vacuum tube, such oscillations being presumably the origin of the Barkhausen-Kurz effect. The effects of space charge, initial velocity of emission, and amplitude of oscillation are considered. The results check and extend those of Kapzov found by other methods.)

R133 Moore, W. H. Ultra-short radio waves. *Jour. Franklin Institute*, 209, pp. 473–483; April, 1930.

(A review is given of the work of various experimenters up to the present time in the production of ultra-short radio waves. Generators employed successfully are briefly described and the power obtainable with each is stated. Methods used in determining the exact frequency of the oscillations produced are outlined.)

R133 Ito, Y. Theorie der Zweielektrodenröhren und Erzeugung elektrischer Schwingungen von extra niedriger Frequenz. (Theory of two-electrode tubes and the production of oscillations of ultra-low frequency.) *Zeits. für Hochfreq.* 35, pp. 67–75; February, 1930.

(The theory and operation of a two-electrode tube is discussed as detector, ampifier, and oscillation generator. By coupling the plate circuit of the tube back into the positive filament lead, either by transformer, inductive or capacity coupling, variations in the space current are produced which are in phase with the plate current and hence oscillations are built up. A considerable mathematical treatment is given of the frequency and amplitude of the oscillation produced. (Concluded.))

R133 Hollmann, H. E. Zusammenfassender Bericht—Die Erzeugung kürzester elektrischer Wellen mit Elektronenröhren. (Summary report—Production of ultra-short waves with electron tubes.) *Zeits. für Hochfreq.*, 35, pp. 76–80; February, 1930.

(Methods are described for obtaining electric waves of lengths as low as 5 cm. Directive two-way telephone communication with a 17-cm wave over distances up to 25 miles is reported. (Concluded.))

R134.75 Watts, E. G. Jr. Considerations in superheterodyne design. Proc. I.R.E., 18, pp. 690–694; April, 1930.

(Factors involved in the suppression of the characteristic double response of the superheterodyne receiver are considered. Design details are given for the oscillator circuit, and oscillator-modulator coupling, as affecting the inherently uniform response characteristics. A method of aligning the circuits for single-control purposes is given.)

R135 Alder, L. S. B. Threshold howl in reaction receivers. *Experimental Wireless & W. Engr.* (London), 7, pp. 197–200; April, 1930.

(An explanation is given of the phenomenon known as threshold howl, a low-frequency modulation produced in a regenerative detector at or near the point of self oscillation. The frequency of modulation is shown to be dependent on the time constant of the plate circuit. Points in receiver design emerging from the explanation are summarized.)

R140 Reed, M. Electrical wave filters—Part II. *Experimental Wireless & W. Engr.* (London), 7, pp. 190–196; April, 1930.

(It is shown that the formulas for the propagation constant and the characteristic impedance of many complex types of symmetrical wave filters may be obtained by considering such structures to be derived from simpler types. The application of the general formulas so obtained to a specific structure is illustrated. All the necessary formulas for the design of a low-pass filter are found. (To be continued.))

R144 Fromy, E. La formulation de l'effet Kelvin. (The formulation of the Kelvin effect.) *L'Onde Electrique*, 9, pp. 69–74; February, 1930.

(A simple and general empirical formula for the ratio of the a-c ohmic resistance of a conductor to its d-c resistance has been proposed by Levasseur. The method of formulation is analyzed and the formula for the general case, as well as for a cylindrical conductor, is given.)

R145 Barclay, W. A. Applications of the method of alignment to reactance computations and simple-filter theory—Part II. *Experimental Wireless & W. Engr.* (London), 7, pp. 180–189; April, 1930.

(The principles of reactive filters are systematically presented. An expression for the filter reactance is derived in terms of the load and of characteristics depending solely on the filter. It is shown that the alignment method may be used to estimate rapidly the performance of any given reactive filter under given conditions. The method is illustrated by a particular case.)

R160 von Ardenne, M. and Schlesinger, K. Phasenberhaltnisse und Schwingungseinsatz bei einem Zweiröhrensystem nach Art der Leithäuser-Heegner-Schaltung. (Phase relations and the start of oscillation in the two-tube system of the type of the Leithäuser-Heegner circuit.) *Zeits. für Hochfreq.*, 35, pp. 60–67; February, 1930.

(A mathematical analysis is given of the Leithauser-Heegner circuit. The results of the analysis are subjected to experimental test.)

R161 Glover, R. P. Note on day-to-day variations in sensitivity of a broadcast receiver. Proc. I.R.E., 18, pp. 683–689; April, 1930.

(A series of sensitivity measurements on a highly sensitive broadcast receiver over a period of one month indicates that large variations in sensitivity may occur. Between relative humidity and sensitivity there was found a high degree of correlation (inverse) probably attributable to increased losses in the radio-frequency transformers during periods of high humidity. It is shown that changes in sensitivity may be delayed after the corresponding changes in humidity. The importance of these variations in sensitivity is pointed out with particular reference to the intercomparison of receiver measuring equipment and production testing of radio receivers.)

R200. Radio Measurements and Standardization

R201.7 von Ardenne, M. Bestimmung von Modulationsgraden und Gleichrichterkennlinien mit der Braunschen Röhre. (Determination of the degree of modulation and rectifier characteristics with the Braun tube.) *Elek.-Nach. Technik*, 7, pp. 80–84; February, 1930.

(The use of a Braun tube to estimate the degree of modulation of a carrier wave is explained for the case where the carrier is modulated with a constant frequency and for the case where the modulation is continuously varying. A method of obtaining the dynamic characteristic curve of any type of rectifier with a Braun tube is also given.)

R214 Lucas, H. J. Some developments of the piezo-electric crystal as a frequency standard. *Experimental Wireless & W. Engr.* (London), 7, pp. 201–203; April, 1930.

(Abstract of paper read before Wireless Section, Institution Electrical Engineers on March 5, 1930.)

R214 Watanabe, Y. The piezo-electric resonator in high-frequency oscillation circuits. Proc. I.R.E., 18, pp. 695–717; April, 1930.

(Cady's theoretical considerations regarding motional admittance of a piezo-electric resonator are experimentally verified. The effect of an air gap between the resonator and the electrodes upon the motional admittance is studied theoretically and experimentally. The determination of the equivalent electrical constants of the piezo-electric resonator with an air-gap is developed by a simple mathematical treatment. The relation between motional admittance and size of electrodes is discussed, as well as the experimental arrangement when a high voltage is to be impressed upon the resonator.)

R223 Drake, F. H., Pierce, G. W., Dow, M. T. Measurement of the dielectric constant and index of refraction of water and aqueous solutions of KCl at high frequencies. *Phys. Rev.*, 35, pp. 613–622; March 15, 1930.

(The dielectric constant of various liquids is determined using a vacuum-tube source of voltage throughout a range of periods extending from $T = 1.31 \times 10^{-8}$ to $T = 8.49 \times 10^{-8}$ sec. The method employed consists in the measurement of standing electric waves between a pipe and a wire concentrically located within the pipe, with the liquid forming the dielectric between the wire and the pipe. Accurate measurements on distilled water, and aqueous solutions of KCl are recorded. The temperature coefficient of dielectric constant for distilled water is given.)

R223 Wyman, J., Jr. Measurements of the dielectric constants of conducting media. *Phys. Rev.*, 35, pp. 623–634; March 15, 1930.

(It is shown that if the product of the natural period of a small rigid circuit and the conductivity of the medium in which it is immersed is sufficiently small, the period is proportional to the square root of the dielectric constant of the medium. On the basis of this fact, measurements have been made of the dielectric constant of water from 0 deg. to 100 deg. C to an accuracy of 0.2 per cent or better and covering a range of frequencies from $T = 1.4 \times 10^{-8}$ to $T = 81 \times 10^{-8}$ sec.

R261 Welikin, J. Röhrenvoltmeter zur verlustfreien Messungen hoherer Spannungen bei Gleichstrom und Wechselstrom. (A vacuum-tube voltmeter for the accurate measurement of high tension direct and alternating currents.) *Elek.-Nach. Technik*, 7, pp. 78–79; February, 1930.

(Vacuum-tube voltmeter circuits using an inverted vacuum tube are described. It is shown that if a single triode is used a condenser must be placed in the input on the plate side to regulate the input impedance. Other circuits using two tubes in parallel and another using a two-grid tube are described.)

R270 Mortimore, R. H. Measurement of intensity of high-frequency magnetic fields. *Phys. Rev.*, 35, pp. 753–762; April 1, 1930.

(Two methods are developed for the measurement of the intensity of magnetic fields produced by currents of frequencies from 10 to 25 megacycles. In the first method the current in a coil is determined from vacuum-tube voltmeter readings of potential drop across a standard inductance and the intensity of field produced is computed. In the second method the e.m.f. induced in a single-turn coil placed in the field is indicated by a vacuum-tube voltmeter and the field necessary to induce the e.m.f. is computed. Good agreement is found between the two methods. Information regarding current distribution in a coil is obtained.)

R300. Radio Apparatus and Equipment

R300 Clapp, J. K. Antenna-measuring equipment. Proc. I.R.E., 18, pp. 571–580; April, 1930.

(A self-contained equipment of portable design for measurement of apparent capacity, apparent resistance, and natural frequency of antenna-ground systems within specified limits is described. A substitution method is employed in the resistance and capacity measurements. The natural frequency is found by extrapolation of a curve plotted from determinations of the antenna frequency for various amounts of loading inductance. The sensitivity and accuracy of the apparatus are discussed.)

R300 Smith-Rose, R. L. Wireless apparatus. *Journal of Scientific Instruments* (London), 7, pp. 64–68; February, 1930.

(A description is given of radio apparatus exhibited at the 20th Annual Exhibition of the Physical and Optical Societies.)

R342.6 Feldtkeller, R. Theorie neutralisierter Verstärkerketten. (The theory of neutralization in multistage amplifiers.) *Zeits. für Hochfreq.*, 35, pp. 45–55; February, 1930.

(A theoretical discussion is given of the maximum amplification obtainable from a multistage amplifier over a given frequency range. It is shown that the performance of the amplifier may be improved by connecting a "neutralizing" circuit, consisting of an inductance, capacity, and resistance in series, between the grids of successive amplifier tubes.)

R342.7 Kafka, H. Ein Beitrag zur Darstellung der Frequenzabhängigkeit von Transformatorschaltungen für Niederfrequenzverstärkung. (A contribution to the preparation of the frequency characteristic curve of a transformer-coupled audio-frequency amplifier.) *Zeits. für Hochfreq.*, 35, pp. 56–60; February, 1930.

(The frequency characteristic of an amplifying stage consisting of a vacuum tube and an audio-frequency transformer is theoretically discussed. The frequency curve for the new Philips transformer is derived and carefully discussed.)

R342.7 Loftin, E. H. and White, S. Y. Cascaded direct-coupled tube systems operated from an alternating current. Proc. I.R.E., 18, pp. 669–682; April, 1930.

(An outline is given of the characteritics which are desirable in an audio-amplifier or detector-amplifier. A description is given of some direct-coupled cascaded tube systems operating from a-c supply. Among the features which are discussed are: the reduction of current drain on the filter, the elimination of "motor boating," stabilizing against drift of plate current, the elimination of hum and the provision of automatic change of grid bias with change of carrier input. The paper gives circuit constants and amplification-frequency characteristics for certain circuit arrangements.)

R500. Applications of Radio

R526.1 Dunmore, F. W. A tuned-reed course indicator for the 4- and 12-course aircraft radio range. *Bureau of Standards Journal of Research*, 4, pp. 461–474; April, 1930. Research Paper No. 160 obtainable by purchase from Government Printing Office, Washington.

(Tuned-reed indicators of the 4- and 12-course aircraft radio range are described. Details of design and operating characteristics are given. A shutter and color system are explained which enable the pilot to know which of the various courses of the range he is flying on and in which direction.)

R800. Non-Radio Subjects

621.313.73 Demontvignier, M. Les redresseurs à vapeur de mercure à haute tension continue; application à l'alimentation des postes émetteurs de T.S.F. (High-tension mercury-vapor rectifiers; application as power supply of radio transmitters.) *L'Onde Electrique*, 9, pp. 45–68; February, 1930.

(The necessary characteristics of a high-voltage plate supply for a radio transmitter are noted. The fundamental properties of mercury-vapor rectifiers are reviewed and circuits are presented whereby rectifiers may be adapted to give a steady high voltage of the desired characteristics. A number of mercury-vapor rectifier installations in existing radio-transmitting stations are described.)

CONTRIBUTORS TO THIS ISSUE

Cook, A. L.: Born March 3, 1905 near Clay Center, Kansas. Received E. E. degree, Stanford University, 1929. Transmission engineer, Pacific Telephone and Telegraph Co., 1929 to date. Non-member, Institute of Radio Engineers.

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Schelleng, J. C.: Born November 11, 1892 at Freeport, Illinois. Received A.B. degree, Cornell University, 1915. Instructor in physics, Cornell University, 1915–1918; since 1918 with Western Electric Company and Bell Telephone Laboratories working on radio transmitters and transmission. For past few years in charge of studies on high-power transmitters, transmitting antennas, and radio transmission. Associate member, Institute of Radio Engineers, 1923; Member, 1925; Fellow, 1928.

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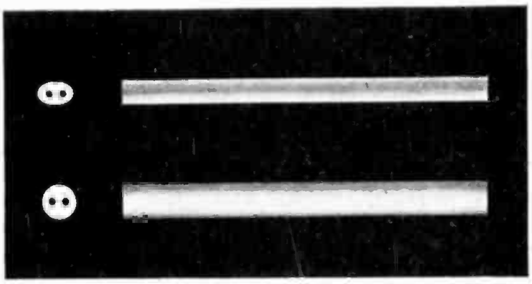

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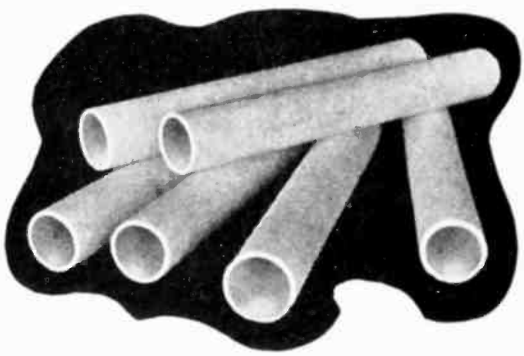

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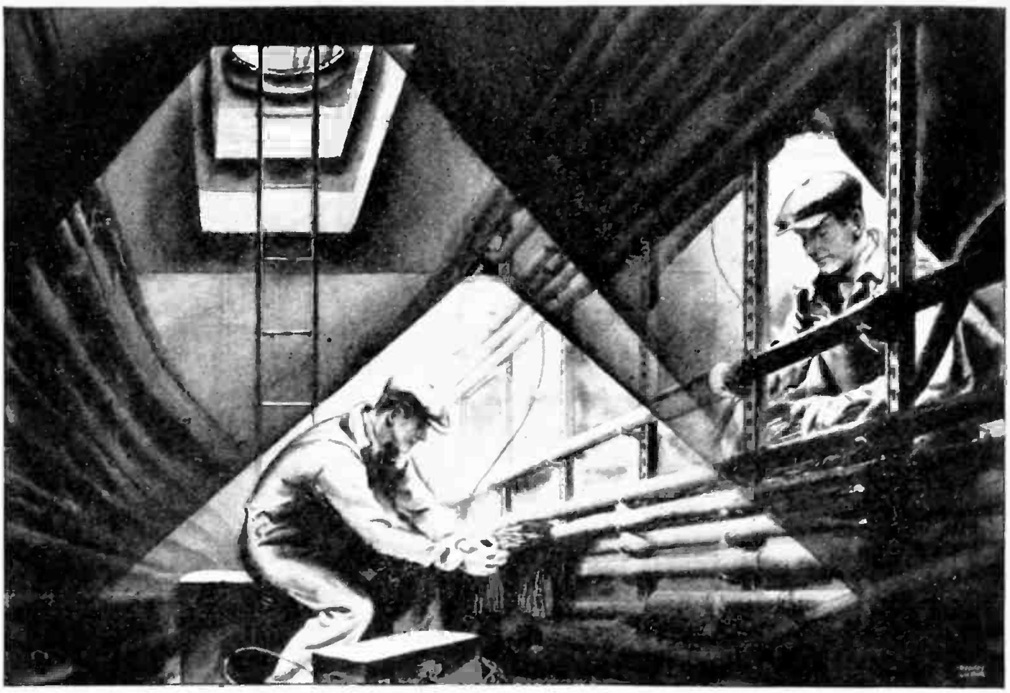

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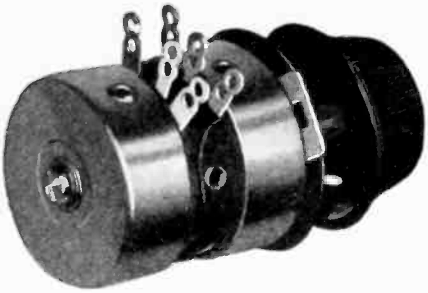

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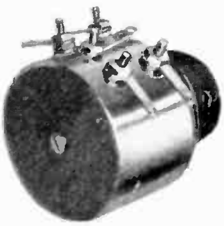

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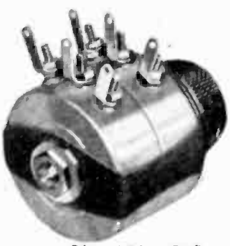

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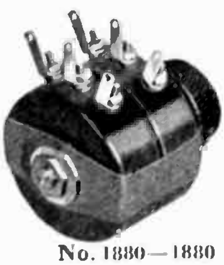

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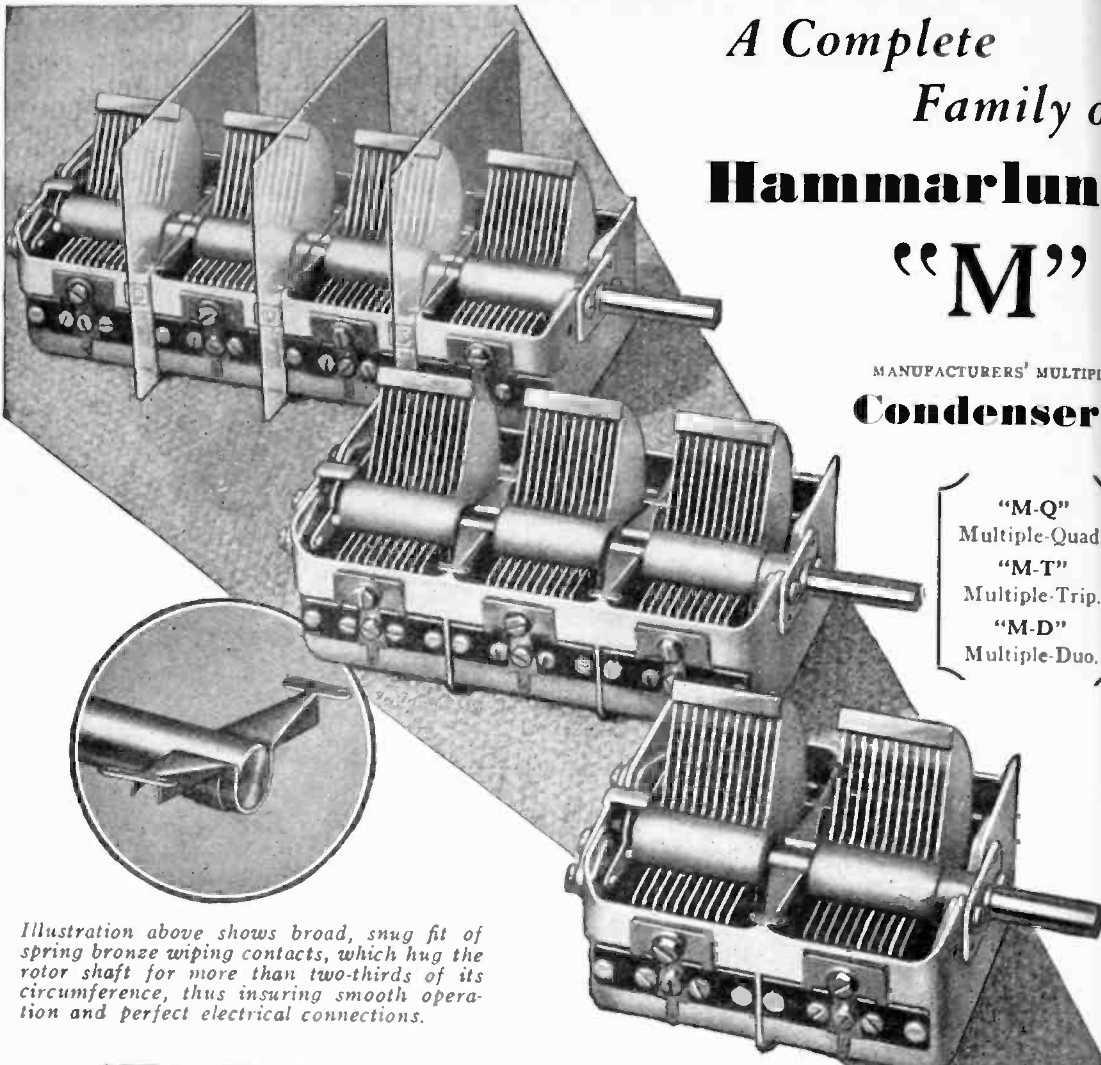

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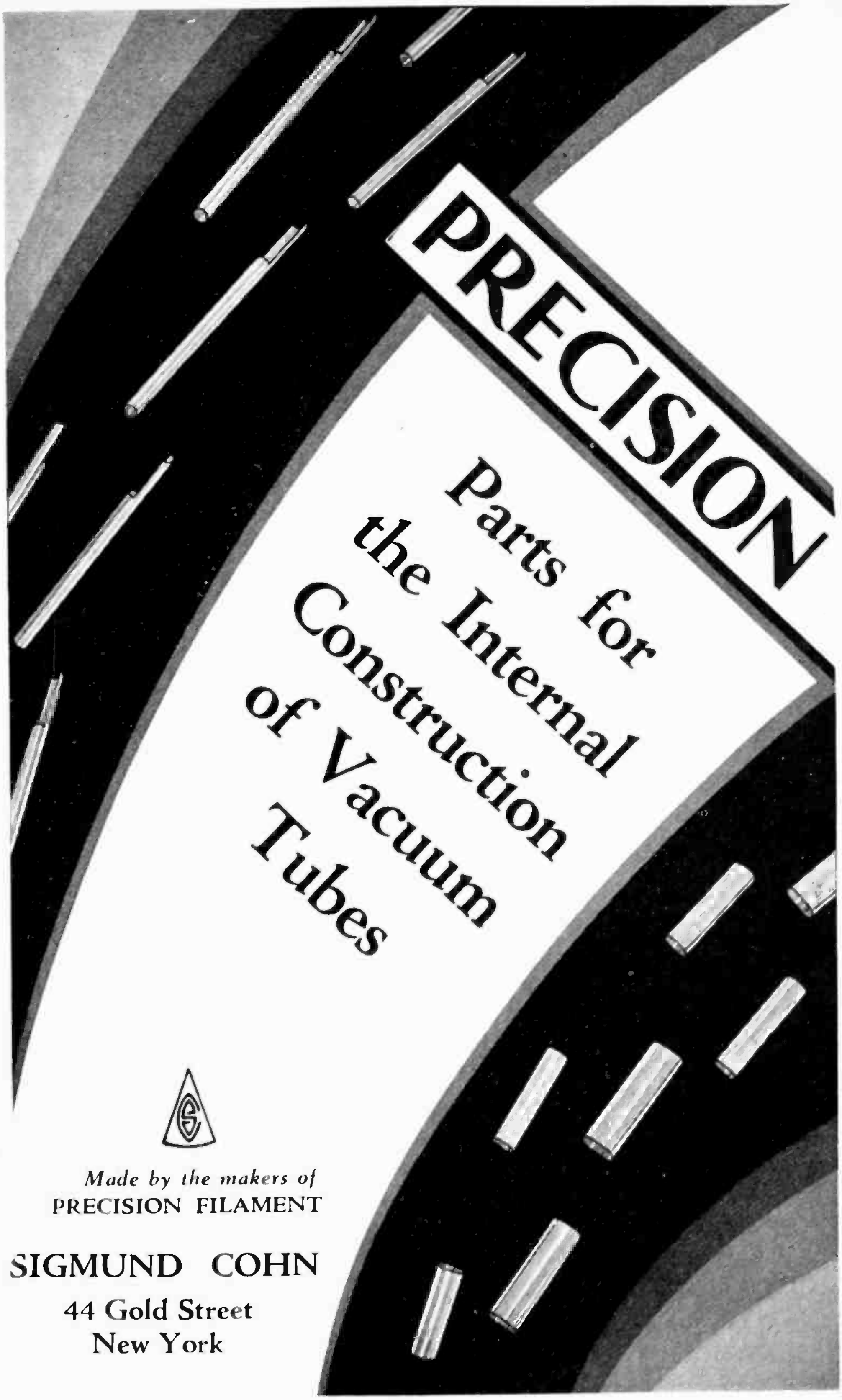
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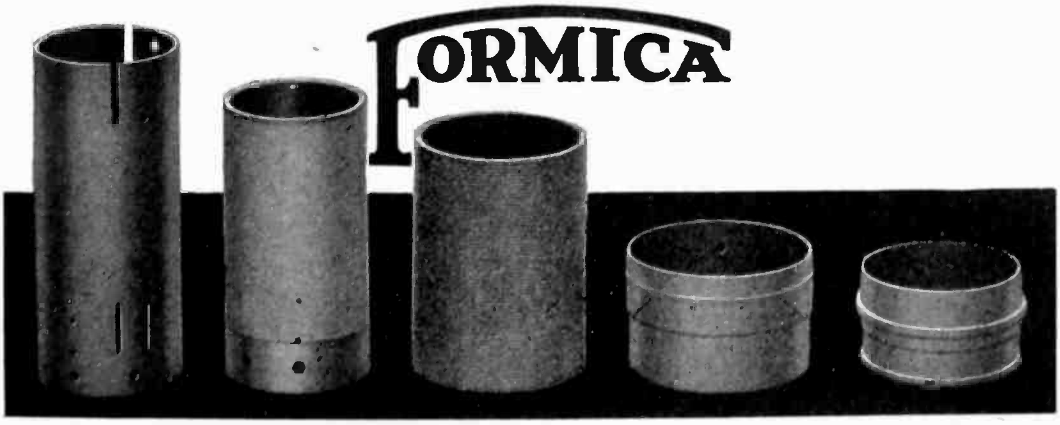

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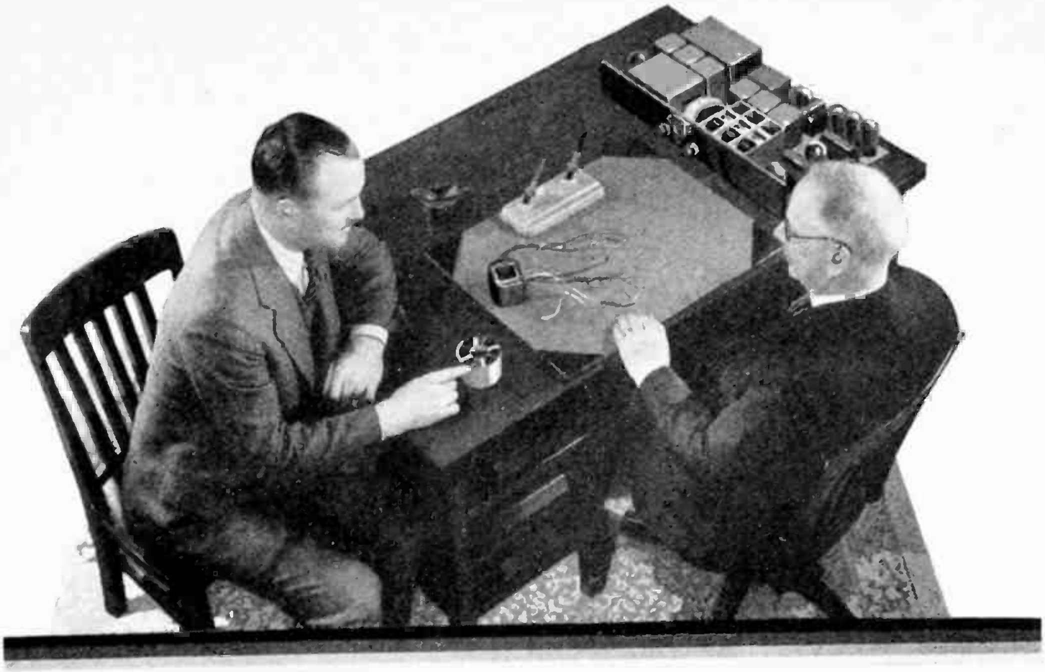

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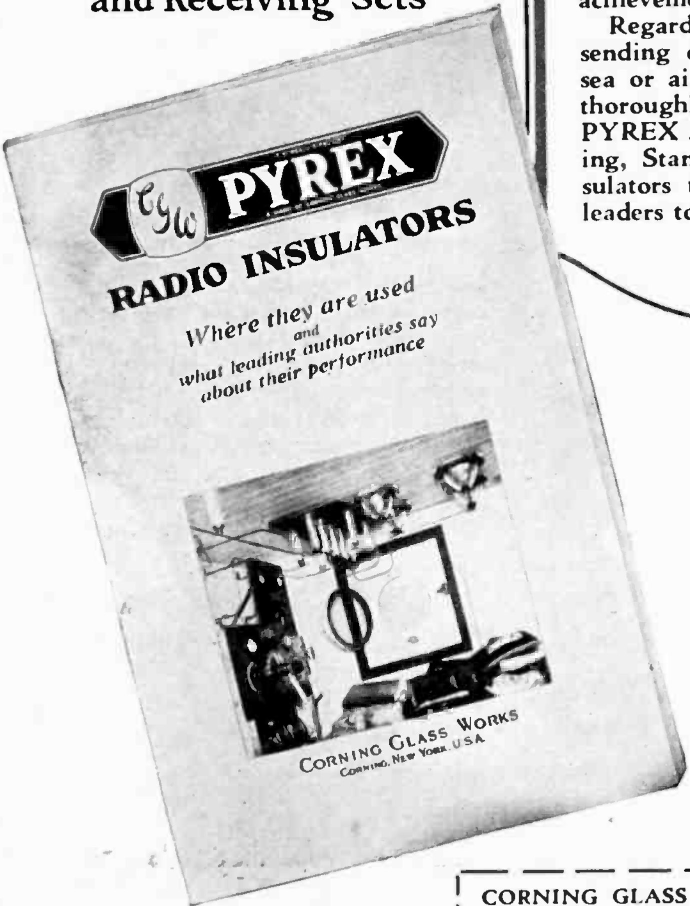

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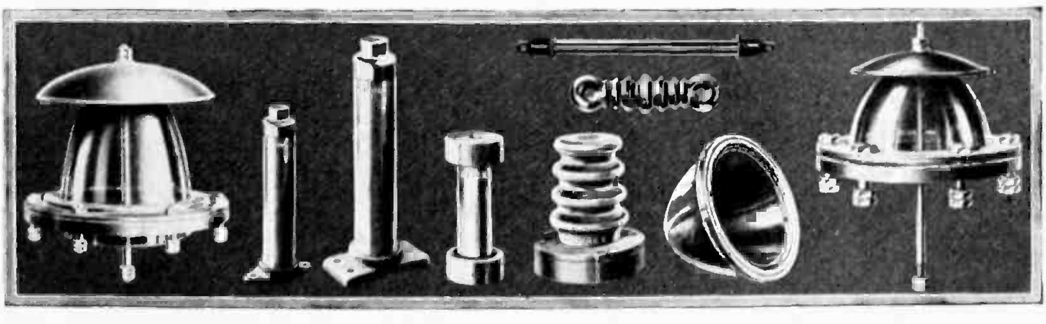

Alphabetical Index to Advertisements

A

Aerovox Wireless Corp..........XXIV
Allen-Bradley Co..................XL
American Tel. & Tel. Co..........XVI
American Transformer Co......XLVI
Amperite Corp................XXVII
Amrad Corp...................XXV
Arcturus Radio Tube Co........XLIII

C

Cardwell, Allen D., Mfg. Corp..XLVII
Central Radio Laboratories..XLVIII
Clarostat Mfg. Co., Inc.........XLII
Cohn, Sigmund................XXIII
Condenser Corp. of America......XXI
Continental Carbon Inc............LI
Cornell Electric Mfg. Co.........LIX
Corning Glass Works.............LV

D

DeForest Radio Co..............XLIV
DeJur-Amsco Corp................LX
Dudlo Manufacturing Co.........LIII

E

Electrad, Inc....................XIX
Electric Specialty Co.............LVI
Engineers Available...............LII

F

Fansteel Products Co., Inc.....XXXII
Formica Insulation Co............XLV
Frost, H. H., Inc................XVII

G

General Radio Co., Outside Back Cover
Grebe, A. H., and Co., Inc..........
..................Inside Back Cover

H

Hammarlund Mfg. Co............XX

I

International Resistance Co....XXVI
I.R.E...........................
...XXXIII, XXXIV, XLIX, L, LIV
Isolantite Co. of America...XXVIII

J

Jewell Electrical Instrument Co......
.........................XXXVIII

L

Lynch Mfg. Co., Inc...............LXII

N

National Carbon Co...............LXI
National Vulcanized Fibre Co.....XLI

O

Operadio Mfg. Co.................XII

P

Pacent Electric Co..............XVIII
Polymet Manufacturing Corp..XXXIX
Powell, R. C. & Co., Inc...........XIII
Professional Engineering Directory
..............................LVII

R

R.C.A. Radiotron Co., Inc.......XXIX
Roller-Smith Co.................XXII

S

Scientific Radio Service...........XXX
Scovill Manufacturing Co.......XXXI
Shakeproof Lock Washer Co..XXXVI

T

Thermal Syndicate, Ltd...........XIV
Thordarson Electric Co.............XI

U

United Scientific Labs..............XV

W

Ward-Leonard Electric Co.....XXXV
Weston Electrical Instrument Corp.
.........................XXXVII

NEW DELUXE "CUB" CONDENSER

Actual Size .1 Mfd. Weight 12½ Grams

Totally enclosed and hermetically sealed against exaggerated temperature and humidity. Light weight, thin aluminum housing makes possible advantageous self-mounting feature.

High Insulation Resistance
Extraordinary Dielectric Strength
Capacities .01 to .25 mfd.

These new DeLuxe "Cub" condensers are compact, light in weight and of attractive appearance. They cut labor operations and cost from 20% to 50% less. *Write for sample.*

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FILTER CONDENSERS
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POWER FACTOR CORRECTION BANKS
UNIFORMLY HIGH INSULATION RESISTANCE
PAPER DIELECTRIC CONDENSERS (All Types)

Write for Sample DeLuxe "Cub" Condenser and Catalog of Complete Line of Cornell Products.

Cornell Electric Mfg. Co., Inc.

LONG ISLAND CITY, NEW YORK

Alphabetical Index to Advertisements

A

Aerovox Wireless Corp..........XXIV
Allen-Bradley Co..................XL
American Tel. & Tel. Co..........XVI
American Transformer Co......XLVI
Amperite Corp................XXVII
Amrad Corp..................XXV
Arcturus Radio Tube Co........XLIII

C

Cardwell, Allen D., Mfg. Corp..XLVII
Central Radio Laboratories..XLVIII
Clarostat Mfg. Co., Inc.........XLII
Cohn, Sigmund................XXIII
Condenser Corp. of America......XXI
Continental Carbon Inc.............LI
Cornell Electric Mfg. Co..........LIX
Corning Glass Works.............LV

D

DeForest Radio Co..............XLIV
DeJur-Amsco Corp...............LX
Dudlo Manufacturing Co.........LIII

E

Electrad, Inc....................XIX
Electric Specialty Co.............LVI
Engineers Available...............LII

F

Fansteel Products Co., Inc.....XXXII
Formica Insulation Co............XLV
Frost, H. H., Inc................XVII

G

General Radio Co., Outside Back Cover
Grebe, A. H., and Co., Inc..........
..................Inside Back Cover

H

Hammarlund Mfg. Co............XX

I

International Resistance Co....XXVI
I.R.E...............................
...XXXIII, XXXIV, XLIX, L, LIV
Isolantite Co. of America...XXVIII

J

Jewell Electrical Instrument Co......
...........................XXXVIII

L

Lynch Mfg. Co., Inc...............LXII

N

National Carbon Co...............LXI
National Vulcanized Fibre Co.....XLI

O

Operadio Mfg. Co................XII

P

Pacent Electric Co..............XVIII
Polymet Manufacturing Corp..XXXIX
Powell, R. C. & Co., Inc...........XIII
Professional Engineering Directory
..............................LVII

R

R.C.A. Radiotron Co., Inc.......XXIX
Roller-Smith Co..................XXII

S

Scientific Radio Service...........XXX
Scovill Manufacturing Co.......XXXI
Shakeproof Lock Washer Co..XXXVI

T

Thermal Syndicate, Ltd...........XIV
Thordarson Electric Co.............XI

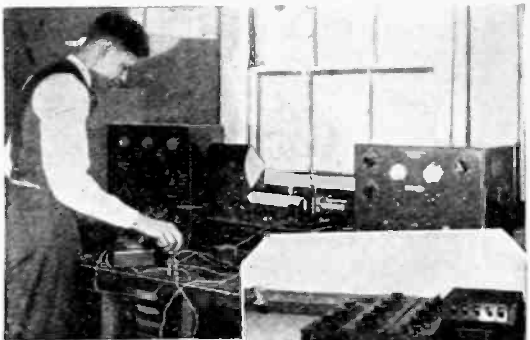
U

United Scientific Labs..............XV

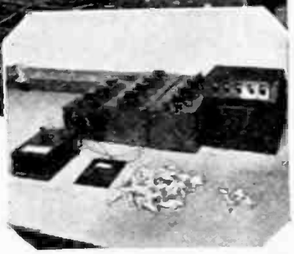
W

Ward-Leonard Electric Co.....XXXV
Weston Electrical Instrument Corp.
...........................XXXVII

—where there can be no compromise



A corner in the laboratory of the National Company at Malden, Mass., where some of the best radio equipment is designed.



A standard bridge in the National Company's laboratory where every resistor is tested before it is used. While the manufacturers' tolerance is an average of 10%, actual tests have shown Lynch Resistors to have a tolerance of about 5%.

with . . QUALITY !

LYNCH RESISTORS

are used in Laboratories of

GENERAL ELECTRIC CO.
WESTINGHOUSE ELECTRIC & MFG. CO.
GRIGSBY-GRUNOW COMPANY
WESTERN ELECTRIC CO.
BELL LABORATORIES

and by

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SCOTT TRANSFORMER CO.
POWERS CINEPHONE CO., INC.
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LYNCH RESISTORS

METALLIZED RESISTORS

Furnished in pigtail or cartridge type, ½ watt, 1 watt and 2 watts; all desirable capacities. Casings are ceramic—caps molded on, tapered for easy mounting. Pigtails are tinned copper, molded into the caps.

Precision Wire Wound Resistors

Especially adaptable for milliameter and voltmeter multipliers. ½ to 1% tolerance. Designed for uses where accurate, non-inductive, high value, low distributed capacity resistors are required. Molded, permanent contacts. Rating of one watt.

MANUFACTURERS: The 1930 production of Lynch Metallized Resistors will use new and heavier type filament which gives Resistors an extra margin of safety, greater current carrying capacity, lower temperature coefficient and makes them more rugged.

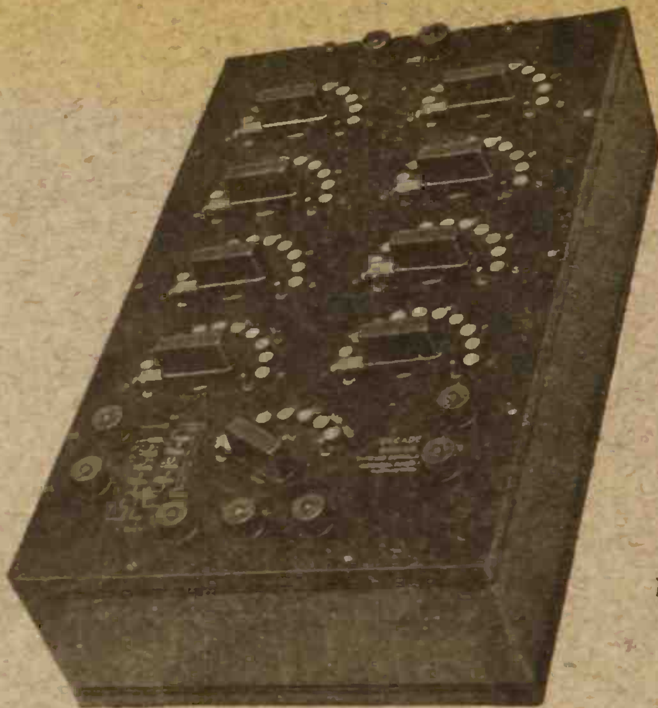
Send us your Resistor specifications and requirements so that samples of this new Resistor can be submitted for test and prices quoted.

LYNCH MANUFACTURING COMPANY, Inc.

1775 Broadway — New York City

Manufacturers of QUALITY RADIO PRODUCTS

A GENERAL-PURPOSE BRIDGE



Type 193
Decade Bridge

Price $115.00

THE TYPE 193 Decade Bridge is a general utility bridge designed for measurement of resistance, capacitance, and inductance at audio frequencies. Standards may be either adjustable or fixed, but the former are much to be preferred.

The dial-switch method of adjusting the ratio arms and the resistance compensation arm greatly facilitates making adjustments. Each decade switch has 11 stops (0 and 10) which further add to the convenience of the instrument.

By a suitable choice of the standards and the ratio arms, the bridge may be made direct reading.

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CAMBRIDGE A, MASSACHUSETTS

British Branch: 40 Buckingham Gate,
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FOR MEASURING ELECTRICAL QUANTITIES AT ALL FREQUENCIES

STANDARD AND SPECIAL ITEMS FOR LABORATORY AND INDUSTRIAL USE

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