VOLUME 19 JULY, 1931 NUMBER 7

Walter J Malone

PROCEEDINGS
of
The Institute of Radio Engineers

W Aaw

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

INSTITUTE SECTIONS

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H. F. Dobbs	ATLANTA	Philip C. Bangs, 23 Kensington Rd., Avondale Estates, Ga.
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D. D. Israel	CINCINNATI	Ralph P. Glover, 5737 Hamilton Ave., Cincinnati, Ohio.
G. B. Hammon	CLEVELAND	P. A. Marsal, National Carbon Co., Inc., P.O. Box No. 400, Cleveland, Ohio.
R. S. Kruse	CONNECTICUT VALLEY	George Grammer, 334 South Quaker Lane, West Hartford, Conn.
Lewis N. Holland	DETROIT	Samuel Firestone, Room 615, 2000 Second Ave., Detroit, Mich.
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Pendleton E. Lehde	NEW ORLEANS	Anton A. Schiele, 1812 Masonic Temple, New Orleans, La.
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L. P. Wheeler	WASHINGTON	Herbert G. Dorsey, U. S. Coast and Geodetic Survey Washington, D. C.

Proceedings of the Institute of Radio Engineers
Volume 19, Number 7 *July, 1931*

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED JUNE 1, 1931

Elected to the Associate Grade

California	Berkeley, 2710 Hillegass Ave.	Power, L. T.
	Oakland, 3227 61st Ave.	Stewart, J. A.
	Santa Monica, 1343 Chelsea St.	Slyfield, C. O.
	Vallejo, 1040 Georgia St.	Rice, A. R.
	Whittier, Route 3, Box No. 139	Wiggins, W. S.
Delaware	Wilmington, 1405 Delaware Ave.	Warfield, T. J.
Dist. of Columbia	Bellevue-Anacostia, U. S. Naval Research Lab.	Trapeur, R. E.
Illinois	Chicago, 21 W. Grand Ave.	Sanderson, G.
	Cicero, 1637 S. 51st St.	Kailus, Peter C.
Indiana	Kokomo, 908 W. Taylor St.	Dawson, F. A.
	South Bend, 718 Sherman Ave.	Work, L. P.
Iowa	Diagonal	McCartney, M.
Louisiana	Monroe, Box No. 893	Mask, W. B.
Maryland	Salisbury, 613 Park Ave.	Perry, J. E.
Massachusetts	East Springfield, 577 Page Blvd.	Taylor, E. H.
Minnesota	Pipestone, 103 E. French St.	Port, C. G.
	Saint Paul, 1941 James St.	Swendson, L. G.
Mississippi	Tupelo, Box No. 482	Stewart, E. B.
Missouri	St. Louis, 6604 Tyler Ave.	Goodberlet, E.
New Jersey	Arlington, 412 Highland Ave.	Lubben, A. G.
	Audubon, 111 Chestnut St.	Feldman, N.
	Audubon, 105 Central Ave.	Pierson, I. S.
	Berkeley Heights, Box No. 26	Romano, M. A.
	Boonton, 113 Schultz St.	Barber, A. W.
	Camden, Plaza Club Hotel	Burnison, J. W.
	Collingswood, 34 Woodlawn Ave.	Bov, R. F.
	Freehold, 25 Park Ave.	Hampton, E. C.
	Montclair, 104 Alexander Ave.	Jacobus, R. A.
	Newark, 250 S. 7th St.	Lampey, J. J.
	Oaklyn, 125 E. Bettlewood Ave.	Byler, H. W.
New York	Binghamton, 3 Carlton St.	Craft, J. A.
	New York City, Trade Dept. British Consulate General, 44 Whitehall St.	Langley, H. E.
	New York City, Hotel Montclair, Rm. 1600	Waring, J. T.
	Port Jefferson, 103 Myrtle Ave.	Bergere, S.
Ohio	Cleveland, Hotel Belmont	Wynne, W. A.
	Dayton, 360 W. 1st St.	Colvin, R. A.
	Dayton, 1615 Roosevelt St.	Mumma, R. E.
	Dayton, 3516 N. Main St.	Peoples, A. G.
	Osborn, R.D. No. 1	Kilheffer, L. D.
Pennsylvania	Lemoyne, 440 Market St.	Smiley, W. A.
	Philadelphia, 933 W. Somerset St.	Bowers, F. L.
	Philadelphia, 3714 N. Carlisle St.	Higgins, H. R.
	Philadelphia, 3834 N. 13th St.	Karker, A. S., Jr.
	Philadelphia, 1239 N. Hollywood St.	Klingelhoeffer, R. H.
	Schuylkill Haven, 336 Dock St.	Eiler, D. L.
	York, R.D. No. 7	Herr, B. R.
Rhode Island	Woonsocket, 371 Nursery Ave.	Heroux, G. A.
Texas	Dallas, 1311 Republic Bank Bldg.	Newton, W. W.
Virginia	Lynchburg, 1017 Knight St.	Henry, N. E.
Alaska	Juneau, Box No. 1498	O'Loughlen, B. W.
Bahama Islands	Nassau, c/o British Government Hospital	Cruikshank, J. M.
Canada	Toronto, Ont., 46 Spencer Ave.	Creswick, E. A.
	Toronto, Ont., 322 Logan Ave.	Fichter, L.
	Toronto, Ont., 433 Delaware Ave.	Ford, F. W. R.
	Toronto, Ont., Bell Telephone Co. of Canada	Geiger, D. G.
	Toronto, Ont., 95 Westwood Ave.	Girdler, H. J.
	Toronto, Ont., 1424 Queen St. W.	McIntyre, R. G.
Egypt	Cairo, Marconi's Wireless Tel. Co. of Egypt	De Barro, F.
England	Douglas, Manchester, 13 Park Rd.	Horne, T.
	Enfield, Middlesex, 8 St. Georges Rd. Forty Hill	Bate, J. R. T.
	Hull, Yorks, "Lynton," 17 Silverdale Rd.	Rapson, E. T. A.
	Ilford, Essex, 351 Wanstead Park Rd.	Randall, M. C.
	London, N.W. 9, 259 The Broadway, West Hendon	Barnard, R. M.
	London, N.W. 2, Radio Section, P.O. Research Station, Dollis Hill	Hawking, W.
	London, 143 Sunny Gardens Rd., Hendon N.W. 4	Starr, A. T.
	Piccadilly, Manchester, Broadcasting House, British Broadcasting Corporation	Lycett, E. L.
	Portsmouth, Hampshire, 34 Mafeking Rd.	Wills, S.
	Rugby, 22 Holbrook Ave.	McDonald, D.

England (cont.)	Slough, Bucks, 12 First Crescent, Cloucester Ave.	Stoddart, J. A.
	Westcliff-on-Sea, 161 Hainault Ave.	Bradfield, G.
	Worksop, Notts, Launt House, Carlton Rd.	Webster, R.
	York, St. Sampson's Square	Shackleton, S. M.
Guam	Agana	Sutton, R. E.
Holland	Amsterdam, Riouwstraat 3	Bruchiss, L.
Philippine Islands	Cavite, 4 Cisneros St.	Minoza, I. P.
Scotland	Glasgow, Cathart, 48 Kingsacre Rd.	Barwise, H.
	Glasgow, E.I., 56 Drumover Dr.	Maceachern, H.
	Glasgow, S2, c/o Livingstone, 35 Langside Rd.	Wilson, T. A.
	Greenock, 25 Stront Crescent	Foster, H. W.
	Greenock, 20 Crawford St.	Young, A. D.
South Australia	Adelaide, 148 Watson Ave., Toorak	Baxter, E. W.

Elected to the Junior Grade

Illinois	Chicago, 1055 N. Leamington Ave.	Liebhart, G. R.
New York	Ozone Park, L. I., 86-11 103rd Ave.	Bulzoni, A. L.
Canada	Toronto, Ont., 699 Sammon Ave.	Cooper, A. S.
England	London S.E. 10, 185 Trafalgar Rd., E. Greenwich	Drinkwater, E. W.
Scotland	Dumfries, Townhead Manse, 14 Moffat Rd.	Thomson, I. W. S.
	Gourock, 24 John St.	Balfour, W.
	Kion, Ailsa View, Stewart St.	Fraser, E. S.

APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before July 31, 1931. Final action on these applications will be taken on August 5, 1931.

For Election to the Associate Grade

California	Bakersfield, 1010½ L St.	LeMert, R. D.
	Long Beach, 1901 Golden Ave.	McDowell, L. W.
	Long Beach, 2017 Florida St.	Strong, F. G.
	Los Angeles, 1547 W. 46th St.	Findley, T. J.
	Los Angeles, 4316 S. Hoover St.	Lippincott, E. W.
	Los Angeles, 3922 Cumberland Ave.	Mosier, J. A.
	Los Angeles, 1843 Echo Park Ave.	Perry, R. M.
	Pasadena, 121 State St.	Blatterman, H. L.
	San Francisco, c/o Heintz & Kaufman, 311 California St.	Barnette, A. F.
	San Francisco, 735 Larkin St.	Kramer, J. F.
	San Jose, 1335 Magnolia Ave.	Williams, D. F.
Colorado	Yuma, P. O. Box No. 307	Beehler, E. G.
Dist. of Columbia	Washington, 3519-13th St. N. W., Apt. No. 301	Kelley, J. L.
	Washington, War Dept., Message Center	Stoner, F. E.
Illinois	Champaign, 805 Neil St.	Mann, G. R.
	Chicago, 5409 Blackstone Ave.	Bell, J. R.
	Chicago, 4843 S. Union Ave.	Gallagher, D. F.
	Chicago, 1507 Melrose St.	Isenberg, R.
	Chicago, 6117 Darchester Ave.	Johnson, L. E.
	Chicago, 1630 Sedgwick St.	Karpuk, A. S.
	Chicago, 1130 N. La Salle St.	Laycock, R. F.
	Chicago, 2032 W. Roosevelt Rd.	Passow, E. B.
	Chicago, 5007 Barry Ave.	Thrun, A. G.
	Newton	Sims, L. H.
Indiana	Anderson, Box No. 999	McKee, A. L.
	Fort Wayne, Steinite Mfg. Co.	Abelson, L. S.
	Indianapolis, 36 N. Irvington Ave.	Gutheil, C. M.
Kansas	Topeka, 1451 Byron Ave.	Beasley, W. A.
Louisiana	New Orleans, 5347 Coliseum St.	Christy, E. H.
Maine	Bangor, 127 Hammond St.	Gustin, R. W.
Massachusetts	Cambridge, 888 Massachusetts Ave., Suite 54	Krumming, A.
	Cambridge A, Mass. Inst. Tech.	Tucker, C. E.
	Springfield, 26 Palm St.	Bacon, F. C.
	Springfield, 19 Calvin St.	Martin, D.
Michigan	Grand Rapids, 407 Cherry St., Apt. No. 212	Darrow, J. A., Jr.
Montana	Bozeman, 108 E. Story St.	McKay, W. M.
Missouri	Kansas City, 1538 E. 50th St. Terrace	Brennan, J. W.
	St. Louis, 2110 Withnell Ave.	Ramey, P. T.
New Jersey	Camden, 333 Front St.	Hoch, A. L.
	Jersey City, 41 Virginia Ave.	Norton, J. E.
	Jersey City, 68 Summit Ave.	Sponseller, J. F.
	Short Hills, 207 Wellington Rd.	Harkins, P. B.
	Westmont, 25 Emerald Ave.	Weber, A. K.
New York	Bayside, c/o Hazeltine Corp., 210th St. and 26th Ave.	Whitman, V. E.
	Brooklyn, 340 Vernon Ave.	Korb, A.
	New York City, 15 E. 26th St.	Block, W. S., Jr.
	New York City, 596 Riverside Dr.	Moises, G. S.
	New York City, c/o Kenyon Trans. Co., 122 Cypress Ave.	Mitchell, I. A.
	New York City, 60 E. 42nd St.	Moscowitz, D. J.
	New York City, 1090 Simpson St.	Sadick, A. I.
	Rego Park, 6212 Saunders St., Apt. No. J4	MacDonald, W. W.
	Richmond Hill, 130-06 Liberty Ave.	Palasky, H.
	Riverhead, c/o RCA Comm. Inc.	George, R. W.
	Rochester, 4 Charlotte Ave.	Newton, W. E.
	Rochester, 80 S. Plymouth Ave.	Peterson, D. H.
	Rochester, 158½ Park Ave.	Prickett, N. R.
	Schenectady, 2226 Watt St.	Towlson, H. G.
	Yonkers, 33 Hudson St.	Rudove, A. J.
Ohio	Columbus, 175 N. Front St.	Morgan, B. A.
Oklahoma	Oklahoma City, 322 E. 15th St. N.	Horton, J. R.
Oregon	Medford, KMED, Sparta Bldg.	Rush, F. M.
Pennsylvania	Philadelphia, 1853 Daly St.	Palm, F. J.
	Philadelphia, 1430 N. 16th St.	Segal, I. A.
Virginia	Lynchburg, 4101 Fort Ave.	Orth, J. T.
Washington	Mount Vernon, 1539 S. 3rd St.	Hannaford, H. W.
	Seattle, 523 Kulien Bldg., 3131 Western Ave.	Shepard, W. D.

W. Virginia	Clarksburg, 731 Locust Ave	Koerner, R. S.
Australia	Clovelly, NSW, 35 Winchester Rd	Falson, A. K.
Belgium	Calmpthout, Heide F. 80	Van Meel, G. M.
Canada	Hamilton, Ont., 174 Main St. W	Bennett, H.
	Montreal, P. Q., Northern Electric Co., Ltd. 637 Craig St. W.	Curran, C. C.
	Montreal, P. Q., 3515 Durocher, Apt. 52	Weir, C. V. F.
	Regina, Sask., 423 Federal Bldg	Megill, W. J.
	Saskatoon, Sask. 101 Grain Bldg	Michalenka, A.
	St. Catherine's, 63 Pleasant Ave	Palen, E. L.
	Toronto, Ont., 243 Pape Ave	Ainsworth, H.
	Toronto, Ont., 357 Kingston Rd	Foote, G. E.
England	Chelmsford, Essex, 28 Van Diemans Rd	Stead, C.
	Liverpool, 83 Broad Green Rd., Broad Green	Anderson, G.
	London N. 1., 12 Thornhill Bridge Place	Roberts, J. A.
	London W. 5., 27 Grange Park, Ealing	Wardle, P. G.
	North Wembley, Middlesex, Oak Lodge, Nathan Rd	Hay, A. E.
	Swindon, Wilts, 27 North St	Love, H. J.
	Whitechurch, Cardiff, 36 Lon Isa, Rhiwbina	Jones, D. E.
France	Paris, Rue du Laos 7	Fernandez, J. D. P.
	Paris, 15eme, 10 Rue de Langeac	Turpin, R.
Italy	Milan, Viale Abrussi 2F	Guarineri, G. M.
Japan	Kawasaki, c/o The Tokyo Electric Co	Kato, K.
	Haramura Asagun, Hiroshima Prefecture	Kinoshita, S.
	Kawasaki, c/o The Tokyo Electric Co	Miyaji, T.
	Tokyo, 1229 Sekigahara Ohimachi	Shinohara, K.
Porto Rico	Ponce, 6 Cristobal Colon St., P. O. Box 749	Rodriguez, J., Jr.
South Africa	Durban, P. O. Box 1219	Levine, L. E.

For Election to the Junior grade

Illinois	Chicago, 1130 N. LaSalle St	Tickus, L. J.
N. Dakota	Grand Forks, 406 Oak St	Dettman, R. A.
Ohio	Dayton, 1126 Wisconsin Blvd	Marquette, D. L.
Texas	New Braunfels, P. O. Box No. 384	Rompel, W. A.
	Stamford, 406 E. Campbell St., Box No. 529	Billingsley, J. B.
Canada	Kinistino, Sask	Barber, O. B.

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(Terms expire January 1, 1932, except as otherwise noted)

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LLOYD ESPENSCHIED
Member Board of Direction 1931

Lloyd Espenschied, a member of the Board of Direction from 1915 to 1926 and also in 1931 was born in St. Louis, Mo., on April 27, 1889.

He was graduated from the Pratt Institute in 1909 after taking a course in industrial and electrical engineering. During the summers of 1907 and 1908 he was a shipboard radio operator. Upon graduation he became connected with the Telefunken Wireless Telegraph Company of America and in 1910 joined the technical staff of the American Telephone and Telegraph Company.

Mr. Espenschied has taken an active part in Bell System radio developments, including the first long-distance telephone experiments in 1915 and the more recent developments of ship-to-shore radiotelephony and overseas radiotelephone services. At the present time he is a high-frequency transmission engineer in immediate charge of radio development.

In addition to his radio work he has had a responsible part in the development of wire communication systems, particularly in high-frequency carrier current telephony and telegraphy. He has contributed much on the technical side of communications as indicated by many patents issued in his name, as well as in the capacity of an engineering executive. He has participated in a number of national and international conferences dealing with radio communications.

Mr. Espenschied is a charter member of the Institute and transferred to the grade of Fellow in 1924. He is a Fellow of the American Institute of Electrical Engineers and has contributed technical papers to the publications of both societies.

Radio Transmissions of Standard Frequency, July, August, and September, 1931

The Bureau of Standards announces a new schedule of radio transmissions of standard frequencies. This service may be used by broadcast and other stations in adjusting their transmitters to exact frequency, and by the public in calibrating frequency standards and transmitting and receiving apparatus. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to about 1000 miles from Washington, and some of them at all points in the United States. The time schedules are different from those used in transmissions prior to this July.

There are two classes of transmissions provided: one, transmission of the highest accuracy at 5000 kc for two hours afternoon and two hours evening on three Tuesdays in each month; the other, transmissions of a number of frequencies in two-hour periods in the afternoon and evening, one Tuesday a month. The transmissions are by continuous-wave radiotelegraphy. The 5000-kc transmissions consist mainly of a continuous cw transmission, giving a continuous whistle in the receiving phones. The first five minutes of this transmission consist of the general call (CQ de WWV) and announcement of the frequency. The frequency and the call letters of the station (WWV) are given every ten minutes thereafter. The transmissions of the other type are also by continuous-wave radiotelegraphy. A complete frequency transmission includes a "general call," "standard frequency signal," and "announcements." The general call is given at the beginning of each 18-minute period and continues for about two minutes. This includes a statement of the frequency. The standard frequency signal is a series of very long dashes with the call letters (WWV) intervening; this signal continues for about 8 minutes. The announcements follow, and contain a statement of the frequency being transmitted and of the next frequency to be transmitted. There is then a 6-minute interval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 280, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequencies are received (or even only a single one), persons can obtain as complete a frequency meter calibration as desired by the methods of generator harmonics.

The 5000-kc transmissions are from a transmitter of 1 kilowatt power; they occur every Tuesday except the first in each month. The other transmissions are from a transmitter of 1/2 kilowatt power; they are given on the first Tuesday of every month both afternoon and evening.

5000-Kilocycle Transmissions
2:00 to 4:00 P.M., and 10:00 P.M. to 12:00 Midnight, Eastern Standard Time

July	August	September
14	11	8
21	18	15
28	25	22
		29

Multifrequency Transmissions

Eastern Standard Time		Frequencies in Kilocycles		
		July 7	August 4	September 1
2:00 P.M.	10:00 P.M.	1600	3600	6400
2:18	10:18	1800	4000	7000
2:36	10:36	2000	4400	7600
2:54	10:54	2400	4800	8200
3:12	11:12	2800	5200	8800
3:30	11:30	3200	5800	9400
3:48	11:48	3600	6400	10000

The frequencies in the 5000-kilocycle transmissions are piezo controlled, and are accurate to much better than a part in a million. The frequencies in the multifrequency transmissions are manually controlled, and are accurate to a part in a hundred thousand.

Since the start of the 5000-kc transmissions the Bureau of Standards has been receiving reports regarding the reception of these transmissions and their use for frequency measurements from nearly all parts of the United States, including the Pacific coast and Alaska. The Bureau is desirous of receiving more reports on these transmissions, especially because radio transmission phenomena change with the season of the year. The data thus far obtained cover the first six months of 1931, and give information regarding approximate field intensity, fading, and the suitability of the transmissions for frequency measurements.

It is suggested that in reporting upon the field intensity of these transmissions, the following designations be used where field intensity measurement apparatus is not at hand: (1) hardly perceptible, unreadable; (2) weak, readable now and then; (3) fairly good, readable with difficulty; (4) good, readable; (5) very good, perfectly readable.

A statement as to whether fading is present or not is desired, and if so, its characteristics, such as whether slow or rapid, and time between peaks of signal intensity. Statements as to the type of receiving set used in reporting on the transmissions and the type of antenna used are likewise desired. The Bureau would also appreciate reports on the use of the transmissions for purposes of frequency measurement or control.

Reports on the reception of the transmissions should be addressed to Bureau of Standards, Washington, D. C.

Committee on Sections

A meeting of the Committee on Sections was held at the office of the Institute at 7 P.M. on Friday, May 15, and was attended by C. W. Horn, chairman; Austin Bailey, D. H. Gage, B. E. Shackelford, and H. P. Westman, secretary.

A number of items regarding the financial status of Sections of the Institute were considered and a tentative draft of a proposed standard financial form to be used by Sections in submitting financial statements at the end of each year was prepared.

An agenda was prepared for the annual meeting of the Committee on Sections to be held during the convention in Chicago.

Institute Meetings

Atlanta Section

A meeting of the Atlanta Section was held at the Atlanta Athletic Club on May 22, H. F. Dobbs, chairman, presiding. A paper by Henry Reid on the "Multi-Mu tube and its Importance to Broadcast Receivers" was presented. At the conclusion of the presentation of the paper a discussion was held and entered into by Messrs. Bangs, Dobbs, Dougherty, and Wills.

Several of the members present then discussed the problems of power interference.

The meeting was attended by fourteen members and guests, all of whom were present at an informal dinner which preceded it.

Boston Section

The May 15 meeting of the Boston Section was held at Cruft Laboratory, Harvard University, Cambridge, G. W. Pierce, chairman, presiding.

A paper on "Application of Frequencies above 30,000 Kilocycles to Communication Problems" by Messrs. Beverage, Peterson, and Hansell of RCA Communications, Inc., was presented by H. H. Beverage.

The 5000-kc transmissions are from a transmitter of 1 kilowatt power; they occur every Tuesday except the first in each month. The other transmissions are from a transmitter of 1/2 kilowatt power; they are given on the first Tuesday of every month both afternoon and evening.

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Boston Section

The May 15 meeting of the Boston Section was held at Cruft Laboratory, Harvard University, Cambridge, G. W. Pierce, chairman, presiding.

A paper on "Application of Frequencies above 30,000 Kilocycles to Communication Problems" by Messrs. Beverage, Peterson, and Hansell of RCA Communications, Inc., was presented by H. H. Beverage.

The meeting was attended by one hundred and four members and guests, a number of whom were present at the informal dinner which preceded it.

Buffalo-Niagara Section

The May meeting of the Buffalo-Niagara Section was held on the 13th of the month at Edmund Hayes Hall, University of Buffalo, S. W. Brown, chairman, presiding.

The paper of the evening by Walter R. Jones, an engineer of the Sylvania Products Company was on "Recent Developments of Vacuum Tubes."

The paper covered a number of types of pentodes and tetrodes which were described in detail and a number of empirical curves and data obtained from measurements, presented. The informal discussion which followed the presentation of the paper was entered into by a number of the nineteen members and guests in attendance.

Chicago Section

The April meeting of the Chicago Section was held on the 23rd in the auditorium of the Engineers Building, Byron B. Minnium, chairman, presiding. The paper of the evening by David Grimes of the Radio Corporation of America was on "Superheterodyne Receiver Design."

The author discussed in detail the superheterodyne design problems such as image frequency ratios, reradiation, correct intermediate frequency, application of the pentode, selectivity, etc.

The discussion which followed the paper was entered into by Messrs. Arnold, Krantz, Landon, Million, Minnium, Polydoroff, and several others of the one hundred and ninety-five members and guests in attendance.

Cincinnati Section

The nineteenth meeting of the Cincinnati Section was held at the Hotel Alms, Dorman D. Israel, chairman, presiding.

The paper of the evening "The Nature of Radiant Energy" was presented by D. A. Wells of the Physics Department of the University of Cincinnati.

Dr. Wells traced briefly the historical and theoretical development of the electromagnetic theory of radiation, from Newton, Maxwell, and Hertz to the present day. The radiation spectrum was divided into the following frequency bands: radio, heat or infra-red, visible, ultra-violet, X-rays, gamma rays, and cosmic rays. It was stated that the difference between these various groups is characterized essentially by differing orders of frequency. Each class of radiation (with the exception of the cosmic) was demonstrated with the following generating and receiving apparatus.

Type of Radiation	*Generating Means*	*Receiving and Detecting Means*
Radio	$3\frac{1}{2}$-meter v.t. osc.	dipole antenna and lamp
Heat	radiant heater	reflector, thermocouple and galvanometer
Visible	arc	prism, filters, and screen
Ultra-violet	mercury-vapor arc	fluorescent substances
X-rays	X-ray tube	excitation of fluoroscope screen
Gamma rays	radium	electroscope

The paper was discussed by Messrs. Austin, Boyle, Kilgour, and Nichols of the forty-three members and guests in attendance.

Connecticut Valley Section

The May meeting of the Connecticut Valley Section was held on the 7th of the month at the Hotel Charles, Springfield, Mass., R. S. Kruse, chairman, presiding.

A paper on "Radio in the Tropics" was presented by H. W. Holt of the Radio Engineering Department of the Westinghouse Electric and Manufacturing Company. The paper covered some of Mr. Holt's experiences in the jungle along the Magdalena River in Colombia, South America, where he spent nearly a year installing commercial radio apparatus. Some of the difficulties attendant upon the installation of radio apparatus in the tropics were discussed. Approximately one hundred photographs taken by the author were exhibited.

The meeting was attended by thirty-one members and guests.

Los Angeles Section

The May 19 meeting of the Los Angeles Section was held at the Engineers Club in Los Angeles, T. E. Nikirk, chairman, presiding.

The Los Angeles Section of the American Institute of Electrical Engineers was the guest of the Institute Section at this meeting.

A paper on "Mechanical and Installation Problems of the New 50-kw RCA Transmitter at KFI" and "An Electrical and Circuit Description of the New 50-kw RCA Transmitter" were the two papers presented by J. J. Farrell and G. W. Fyler both of the General Electric Company. The papers related specifically to the new transmitter now being installed for the new broadcast station KFI at Buena Park, California which is some twenty miles southeast of Los Angeles. Mr.

Farrell described the problems of design, experimental work, and testing of such a transmitter prior to its acceptance as a commercial unit. He considered also the difficulties encountered in the problems of materials, mechanical design, and arrangements that entered into his work as installation engineer of the new station. Mr. Fyler, with the aid of slides, discussed the electrical design and circuit arrangement of the new transmitter, explaining the construction of the power supply and rectifier systems, the crystal oscillator frequency controls, the various stages of power amplification, and the antenna system.

A 100-kw tube was on display at the meeting along with several smaller amplifier and rectifier tubes which are a part of the new station.

Seventy-five members and guests attended the meeting.

Philadelphia Section

The May meeting of the Philadelphia Section was held on the 13th at the Engineers Club, D. O. Whelan, vice chairman, presiding. The paper of the evening on "Technique of Loud Speaker Sound Measurements" was presented by Stuart Ballantine of the Boonton Research Corporation.

The author described various pieces of apparatus used for the purpose of determining frequency—sound pressure curves. These included a new logarithmic optical recorder permitting the automatic registration of frequency response characteristics. A number of slides were projected to illustrate the paper which evoked considerable discussion by the members and guests in attendance.

Nominations for the election of officers at the annual meeting in June were made.

Rochester Section

A meeting of the Rochester Section was held at the Sagamore Hotel on April 16, Virgil M. Graham, presiding.

A paper on "Recent Developments in Vacuum Tubes" was presented by E. W. Ritter of the RCA Radiotron Company. Mr. Ritter traced tube development from other days to the present giving the reasons for the various types which have been produced. He touched upon the design of tubes in general, describing briefly the advantage of the pear-shaped intermediate size bulb for police radio sets and airplanes where space is limited. He stated all the desirable characteristics of the larger bulb have been retained and no disadvantages introduced.

The so-called variable-mu tube was described and the conical grid construction was explained. This tube finds its greatest value in avoiding cross-talk and tube cut-off when listening on a broadcast receiver to one of two powerful signals operating on frequencies close together.

Mr. Ritter discussed the pentode and what might be expected of it. He described the construction and discussed its operation characteristic from the standpoint of the ability of the tube to overcome the dip in the plate-current curve which is a characteristic of the 227 tube.

The paper was discussed by Messrs. Grimes, Manson, Wise, and several others of the ninety-seven members and guests in attendance.

On April 18 the annual dinner meeting of the Rochester Engineering Society and its affiliated societies was held at the Rochester Chamber of Commerce, Virgil Palmer, President of the Rochester Engineering Society, presiding.

The speaker of the evening was General James G. Harbord who is chairman of the Board of Directors of the Radio Corporation of America.

General Harbord's subject "In Years to Come" covered a considerable number of predictions as to the future of the musical instrument field. He vividly pictured electrical instruments which will compete with great symphony orchestras and be easily and simply manipulated with comparatively little previous training on the part of the artist.

Personal Mention

Francis H. Engel, formerly in the Vacuum Tube Department of the Radio Corporation of America of New York City is now with the Hygrade Lamp Company of Salem, Mass.

Thomas L. Gottier has joined the Engineering Department of the United Research Corporation of Long Island City previously being with the RCA Victor Company at Camden.

Formerly chief engineer of Station WFAA in Dallas, Texas, Conrad F. Harington has become general manager of station KFUL in Galveston, Texas.

Robert Holmes has left Loftus, Ltd., of Liverpool, England, to become a radio and research development engineer for Lissen and Company, Ltd., of Middlesex, England.

R. H. Langley, previously director of engineering of the Crosley Radio Corporation, is now doing consulting work at 165 Broadway, New York City.

Previously in the Engineering Department of Radio Industries of Canada, Ltd., Winnipeg, D. J. H. Leitch has become engineer and secretary-treasurer of the Canadian Sound Engineering Company, Ltd., of Winnipeg.

A. Roy Macartney, formerly with the Honolulu Broadcasting Company, Ltd., has joined the engineering staff of the Mutual Telephone Company in Honolulu.

John F. Morris has become studio engineer for the National Broadcasting Company, Inc., leaving the Great Lakes Broadcasting Company where he was control room engineer.

James P. O'Donohue has been advanced to the rank of chief engineer of the Postal Telegraph Cable Company in New York City.

E. K. Oxner has become chief engineer of the High-Frequency Laboratories of Chicago.

L. S. Hillegas-Baird, formerly with the Leslie F. Muter Division of Steinite is now connected with the High-Frequency Laboratories of Chicago.

Formerly with the RCA Victor Company of Massachusetts, J. P. Putnam has joined the staff of the Television Laboratories of Boston.

Previously a vacuum tube engineer for the General Electric Company, Jack D. Ryder has joined the electrical force of the Bailey Meter Company of Cleveland, Ohio.

Richard F. Shea has joined the engineering staff of the Pilot Radio and Tube Manufacturing Company of Lawrence, Mass., having been previously connected with the Atwater Kent Manufacturing Company.

A. J. Spooner, formerly chief engineer and secretary-treasurer of the Oklahoma Broadcasting Co., Inc., is now chief radio engineer of the KGFF Broadcasting Company of Shawnee, Okla.

A. C. Stelzer has joined the Engineering Department of the Electrical Test Equipment Division of the General Motors Radio Corporation of Dayton, Ohio, having previously been in the Test Methods and Engineering Division of the RCA Victor Company.

E. M. Strout has left Station KSTP of St. Paul, Minn., to become chief engineer of the Waterloo Broadcasting Company of Waterloo, Iowa.

Minoru Taguchi formerly in the Radio Engineering Department of Nihon Musen has left to join the engineering staff of the Japan Wireless Telegraph Company of Tokyo.

Allen C. Wooldridge has joined the engineering staff of Wired Radio Company, having formerly been connected with R.C.A. Communications, Inc.

Part II

TECHNICAL PAPERS

Proceedings of the Institute of Radio Engineers
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NOTES ON LOUD SPEAKER RESPONSE MEASUREMENTS AND SOME TYPICAL RESPONSE CURVES*

By

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Summary—*The difficulties encountered in loud speaker measurements are briefly reviewed and a description of the acoustic features of a particular indoor measuring system is given. Outdoor testing arrangements are described whereby double as well as single radiating loud speakers are measured with negligible ground reflection error. It is pointed out that the over-all electrical fidelity curve of a radio receiver is an inadequate performance index; the electro-acoustic fidelity embracing the frequency response of the loud speaker is suggested as more informative. The interpretation of loud speaker response curves in terms of what one may expect to hear is discussed.*

Reponse curves shown and discussed include the following:

(1) Effect of type of cone corrugation.

(2) Comparison of radio receiver cabinet with flat baffle.

(3) Curves showing that the face dimensions of a box baffle are of equal importance with the path length from front to rear of cone in determining low-frequency response.

(4) The effect of extreme length in a box baffle.

(5) Measurements of the same loud speaker in cabinets of different size.

(6) Over-all electro-acoustic response curves of radio receivers.

(7) Loud speakers with improved high-frequency response.

(8) Some examples of outdoor measurements.

PROGRESS in electrical measurements of the modulation frequency characteristics of radio receivers has been so rapid that we now find production line measurements made thousands of times daily that were not performed successfully in laboratories four or five years ago. Acoustic measurements, however, have not progressed as rapidly as have electrical measurements, chiefly because of the nature of sound itself. Loud speaker response measurements have been discussed in several published articles, references to some of which will be made in the course of this paper. The present writer's chief aim is to present and discuss results of measurements rather than to review measuring methods. It is believed, however, that in the present state of the art a brief discussion of the major acoustic difficulty of indoor measurement should be included, not only for the sake of completeness, but to add emphasis to the precautions which must be observed in the attainment of significant results. A particular outdoor measuring arrangement also will be described.

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Acoustic Difficulties of Indoor Measurement

The problem in loud speaker response measurements is to determine, over the frequency range, the relation between the acoustic power that the loud speaker transmits to a specified position or region when it is connected to the supply source for which it was designed, and the maximum electrical power which it is possible to draw from the supply source. In order to obtain results independent of acoustic surroundings, the measurements should preferably be made outdoors under rather special conditions, as will be described later. However, owing to interference by noise, bad weather, etc., it is necessary in practice to make most of the measurements indoors under conditions which are designed to imitate as closely as possible those which obtain in free space. A calibrated condenser microphone is used for picking up the sound from the loud speaker, which is driven by an oscillator of suitable output power capacity.

The principal acoustic difficulty in making indoor loud speaker measurements is due to reflection of sound at the bounding surfaces of the room. The disturbing effect of this is the creation of a standing wave system which results in wide variations in the steady state sound pressure in different parts of the room, and this "sound pattern" shifts about in a complicated manner as the frequency is changed. Therefore, a microphone in any fixed position would indicate large changes of sound pressure with frequency even though the loud speaker were capable of generating a uniform sound pressure over the frequency range in free space. The use of a large room lined with sound absorbing material reduces this effect but, with a reasonable quantity of any known material and with the size rooms ordinarily available, it is still disturbingly present at a normal listening distance from the loud speaker. Reflection at low frequencies is particularly difficult to eliminate because of the inefficiency of the best sound absorbers in this range. For example one-inch-thick felt having an absorption coefficient of 63 per cent at 1000 cycles will absorb only about 12 per cent of the incident sound at 64 cycles.

One way of overcoming the difficulties just discussed would be to take a large number of separate observations at different points in the room at each frequency and average them. The equivalent of this slow and laborious procedure, however, may be better accomplished by the method described by Bostwick,[1] wherein the microphone is rotated over a circular path inclined to the axis of the loud speaker and is at the same time constantly turned so that it always faces the plane of the

[1] L. G. Bostwick, "Acoustic considerations involved in steady state loud speaker measurements," *Bell. Sys. Tech. Jour.*, January, 1929.

speaker. The diameter of this path should, for satisfactory averaging, be greater than and at least equal to a half wavelength at the lowest frequency to be measured. A slow-acting meter of the thermocouple type is chosen to indicate the output of the microphone and gives a fairly steady average reading as the latter sweeps through the varying sound pressures. As the microphone terminal voltages are proportional to the pressures on its diaphragm and the thermocouple meter deflections obey the square law, this reading is proportional to the average of the squares of the pressures around the microphone path. This latter average is proportional to the sound energy density in the region traversed by the microphone.

In spite of the acoustic difficulties which have been mentioned, measurements made in a suitable room by the method above described afford much valuable information, particularly of a comparative nature. The effect of design changes having a relatively small influence upon performance may readily be evaluated, as the stability of the system in such that results may be repeated over long periods of time with an average deviation of plus or minus one decibel.

Outdoor Measurements

As mentioned before, to obtain the greatest accuracy, especially at low frequencies, it is necessary to make response measurements outdoors. Such measurements are especially valuable in calibrating a room used for measurements, particularly at frequencies below the averaging ability of the rotating microphone. Free space measurements are also absolutely essential in determining the directional radiation characteristics of loud speakers. Outdoor measurements have been described by Bostwick[1] and by Malter,[2] and several interesting arrangements have been suggested by Kellogg.[3] Many of the methods so far disclosed, however, have been strictly suitable only for loud speakers which radiate in one direction, such as the horn and the directional baffle types, because the arrangements were such that disturbing reflections from the ground or from the supporting structures could take place when measuring double radiating speakers. As virtually all loud speakers in commercial radio receivers radiate from both front and rear, some method of making accurate free space measurements of their response characteristics is of considerable interest.

[1] *Loc. cit.*

[2] Louis Malter, "Loud speakers and theatre sound reproduction," *Jour. S. M. P. E.*, June, 1930.

[3] Edward W. Kellogg, "Loud speaker sound pressure measurements," *Jour. Acous. Soc. Amer.*, October, 1930.

Fig. 1 shows the arrangement which has been set up in a vacant lot beside the Stromberg-Carlson factory at Rochester. The tops of the four poles are connected by steel cables clamped together where they cross at the center of the span, the loud speaker under test being suspended from this point and securely guyed as shown. The microphone with its associated preliminary amplifier is shown suspended from one

Fig. 1—Outdoor loud speaker measuring set-up.

of the cables, a heavy iron bar being interposed in the suspension to prevent the possible transmission of mechanical vibrations from the loud speaker to the microphone via the tightly stretched steel cables. The batteries for the microphone amplifier are located in the box on the ground and connections to the measuring set-up are made over lead covered pairs laid on the ground and terminating in low impedances. The measuring set is located near a window in the factory about 350 feet away where a clear view of the outdoor structure and its surroundings may be had. This is a convenient arrangement as the operator can check for residual noise readings when there is occasional heavy traffic on the adjoining street and refrain from measuring if necessary.

The distance to the nearest building is about 300 feet and the poles are of sufficient height to bring the axis of the loud speaker at least 35 feet above the ground. Assuming spherical radiation and a measuring distance of 10 feet on the axis and employing the optical analogy of an image loud speaker, the ratio of the reflected to the direct sound pressure at the microphone position is 1 to 7. As most practical loud speakers have somewhat directional radiation characteristics, this ratio will in most cases be more favorable. It was late last fall when these measuring arrangements were completed and it was possible to obtain but few data before bad weather set in. Some distinguishable but practically negligible effects of ground reflection may be seen in the response curves, examples of which will be shown later.

Over-all Electro-Acoustic Fidelity of Radio Receivers

The sole physical function of a broadcast system, including the radio receiver, is the faithful transfer of *sounds* from place to place. The part of the entire radio receiver in this, aside from selection of the proper signal, is strictly that of an electro-acoustic transducer; that is, it takes electric power from the antenna and converts it into acoustic power at the loud speaker. The efficiency and fidelity with which this function is accomplished constitutes a useful and significant performance index; indeed, it is the only objective criterion by which the so-called tone quality of the receiver may be judged.

Measurements of the electrical fidelity of a receiver are very useful in development work but, except in their limiting features, offer little information when considered alone about the over-all performance of the receiver. It is a well-known fact that the variations with frequency of the sound output of most loud speakers are large compared with the departures from flatness of the fidelity curves of their associated receiver chassis but, in spite of this, we frequently find magazine writers, advertising men, and even engineers comparing the fidelity of receivers solely on the basis of their electrical characteristics. In the engineering departments of many manufacturers there is no doubt that much effort is spent in securing a good-looking electrical fidelity curve in a receiver chassis, which is then wedded to a loud speaker with an unknown and often poor response characteristic. This inconsistency is due, of course, to the lack of reliable loud speaker response measurements, and it is hoped that it will soon be abolished by the more widespread application of such knowledge of the subject as is available.

Over-all receiver measurements which show the *shape* of the electro-acoustic curve may be made by combining the loud speaker response measuring apparatus with the standard I.R.E. fidelity measuring set-up. The measuring technique appears to be somewhat cumbersome, however, and in our own work we have so far preferred to combine the results of separate determinations of the electrical fidelity of the receiver and the response of the loud speaker, mounted in the correct cabinet. We have defined over-all response as the ratio of the sound pressure produced by the loud speaker under specified acoustic conditions to the modulated radio-frequency voltage applied to the standard artificial antenna under standard I.R.E. test conditions. All these data may be derived from the regular fidelity and sensitivity measurements of the receiver and from the response curve of the loud speaker. The results may be conveniently expressed in db referred to a zero level of one bar per microvolt. (One bar equals one dyne per square centimeter.)

This synthetic method of obtaining the over-all curve has its advantages and its disadvantages. The separate measurements are simpler in practice than the combined ones, and the results give not only the shape of the fidelity curve but the electro-acoustic sensitivity of the receiver as well. Disadvantages are that the over-all data are not obtained by a single direct measurement and that the effect of possible acoustic coupling between the loud speaker and vacuum tubes or condenser plates is not measured. This latter condition, however, may readily be detected by a separate test.

A few remarks on equalization may be pertinent in connection with this subject of over-all receiver characteristics. The type of equalization referred to consists in the shaping of the frequency characteristic of some portions of the system to compensate for defects in some other portion which it may be impracticable to overcome otherwise. This was frowned upon a few years ago when radio receivers and loud speakers were sold separately, on the principle borrowed from wire telephony and from broadcast practice that each portion of the system must figuratively stand on its own feet. This reasoning should not apply, however, to the case of the modern radio receiver where all of the elements contributing to the over-all frequency characteristic are inseparably associated, and it is felt that proper *fixed* equalization in such cases is not only justified but highly desirable.

The Interpretation of Loud Speaker Response Curves

The interpretation of response curves in terms of what one may expect to hear when reproducing speech and music under average listening conditions involves many physiological and psychological factors concerning few of which definite data are available. Because of the difficulties involved, there appears to be no substitute in the present state of the art for the actual experience gained in comparing the results of response measurements with critical listening tests.

It is also difficult to pass on to others the knowledge thus acquired because of the inadequacy of the spoken or written word to describe definitely such complicated sensations as are involved in audition. Engineers engaged in this sort of work usually become able with experience to assign approximately the frequency range covered and also the approximate location and magnitude of the most salient features of the response curve after listening to a loud speaker under familiar acoustical conditions and, when discussing loud speaker performance, are inclined to describe their conception of what the response curve would look like rather than the auditory sensations which are evoked. There are, however, some descriptive terms which have come into common use and which may be related to the response curve. For instance, reproduction is said to be "dull" or lacking in "presence" when the higher frequencies are suppressed. On the other hand, the exaggeration of frequencies in the neighborhood of 2000 cycles may produce a "hollow" quality which is quite unnatural but which may give good speech articulation. Extreme peaks in this region or higher in the frequency scale usually give a "hard," "harsh," or "shrill" effect, particularly when the low frequencies are relatively deficient. Extremely narrow and sharp peaks in the higher frequency region may produce what, for want of a better term, may be called a "sing" which may color the whole reproduction. This effect is the natural consequence of a lightly damped resonant system which is continually being shock-excited and which contributes its characteristic transients to the speaker output. Peaks in the upper part of the frequency spectrum also have the effect of increasing the "hissing" or "swishing" types of background noise associated with radio and phonograph reproduction. Among those who have not had the opportunity of comparing listening tests with the results of measurements, peaks in the region around 3000 cycles may create the erroneous impression that a loud speaker has an extended high-frequency range when, in reality, its response may fall off sharply above, say, 3500 cycles.

Peaks in the lower frequency region in the neighborhood of 100–200 cycles produce the all too familiar "boomy" or "tubby" effect, particularly when associated with the rather common condition of lack of the higher frequencies. Extreme peaks in this region are responsible for a distressing "thump" sometimes heard when reproducing piano, kettle drums and the lower plucked strings. A loud speaker with a large peak at the lower end of the frequency range, another at the high end and with an extended depression in the middle range produces a peculiar effect. Critical listening tells one that both high and low frequencies are being well reproduced, but that there is an obvious deficiency somewhere. Lack of "fullness" is fairly descriptive of this tone quality. Loud speakers having an extended series of high peaks in the lower and middle ranges were found to produce, on speech, a confused effect somewhat similar to that experienced in listening to normal speech in a very reverberant room. The response curve of such a speaker will be shown later.

In connection with the interpretation of response curves it is interesting to note that the ear apparently is able to discriminate to a considerable extent between effects due to the dynamical characteristics of a loud speaker and some of those due to the listening room itself. A response curve taken in an ordinary room at a fixed position with respect to a loud speaker bears little resemblance to one taken by the rotating microphone method in a heavily damped measuring room. The irregularities are so numerous and of such magnitude as almost completely to obscure those disclosed by the regular response measurement, and are subject to decided change with any substantial shift in the microphone position. In spite of these facts, however, one may listen to the loud speaker from almost any normal position in the room without noticing much change in the characteristic quality of the reproduction. Furthermore, it appears that with practice, one is able to predict the outstanding features of the regularly taken response curve equally well anywhere within the normal range of listening positions. This seems to justify the validity and usefulness of response measurements made under the special conditions necessitated by acoustic difficulties, as opposed to those made under the more nearly normal listening conditions suggested by some.

Experimental Results

The response curves of Figs. 2 to 8 were based upon indoor measurements made in a heavily damped room $15 \times 22 \times 11$ feet high. The absorbent lining consisted of 1-inch hair felt, the walls, floor, and ceiling being completely covered. Additional short strips of felt were suspended from the ceiling and longer, twisted strips were hung around the walls.[4] The reverberation time at 512 cycles was 0.1 second as calculated by the Eyring formula.[5] The orbit of the rotating microphone was inclined at an angle of 45 degrees to the axis of the loud speaker and was 8 feet in diameter, thus setting the theoretical low-frequency limit for good averaging at about 70 cycles. The center of the microphone

[4] Graham and Olney, "Engineering control of production," Proc. I. R. E., August, 1930.

[5] Carl F. Eyring, "Reverberation time in 'dead' rooms," *Jour. Acous. Soc. Amer.*, January, 1930.

orbit was located 10 feet distant from the loud speaker and on the axis of the latter, and 4 feet 6 inches from one end of the room. This was about as far from a wall as it was feasible to locate the rotator center on account of the size of the room and the necessity for maintaining a reasonable measuring distance and at the same time providing sufficient space for setting up large horns. About 10 frequency points per octave were taken. The reference level was 1 volt, 1 ohm, and 1 bar as proposed by Bostwick.[1] The condenser microphone was mounted in a 6-inch spherical housing and corrections were applied both for diffrac-

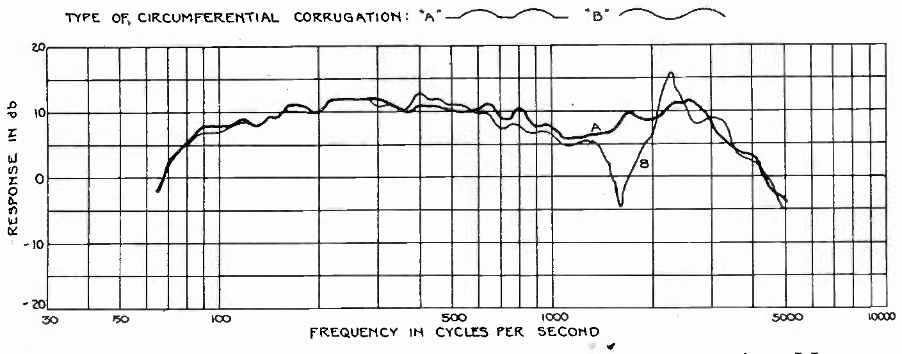

Fig. 2—Response of 8-inch cones with different types of corrugation. Measured in 53 × 53 inch flat baffle and in the same field structure. Points below 65 cycles not shown because of measuring conditions.

tion of sound around the microphone[6] and for resonance in the cavity in front of its diaphragm,[7] assuming normal incidence of the waves. A low-pass filter always was included in the measuring circuit so that only the fundamental output of the loud speaker under test was indicated.

In the outdoor measurements whose results are given in the curves of Figs. 9 and 10, the loud speaker was located as shown in Fig. 1, with the microphone placed 10 feet distant and on the sound axis. The measuring apparatus and the microphone corrections were the same as for the indoor measurements.

The measuring conditions having been stated, we may now proceed with the presentation and discussion of the results. In Fig. 2 is shown the importance of the form of corrugations employed for controlling the resonances which occur in cones at the higher frequencies where they fail to act as rigid pistons. The two cones were of practically identical construction and material and the corrugations were of the same number and spacing, the only difference being in their shape. On a comparative listening test loud speaker *A* was judged to give more natural

[1] *Loc. cit.*

[6] Stuart Ballantine, "Effect of diffraction around the microphone in sound measurements," *Phys. Rev.*, 32, 988, 1929; Proc. I. R. E., July, 1930.

[7] W. West, "The pressures on the diaphragm of a condenser transmitter in a simple sound field," *Jour. I. E. E.* (London), April, 1930.

reproduction than loud speaker *B*, the latter appearing to introduce considerable transient distortion in the upper frequency range as manifested by a pronounced "crackly" effect on loud music. Without reliable response measurements, the location of the cause of such a fault and its correction would be difficult and uncertain matters.

Fig. 3 shows a comparison between the response of a loud speaker in a 53×53 inch flat baffle and in a console-type radio cabinet in which attention has been paid to the acoustic features. Curve *B*, taken with the cabinet several feet away from a wall shows by its comparative smoothness in the lower frequency region, a freedom from pronounced

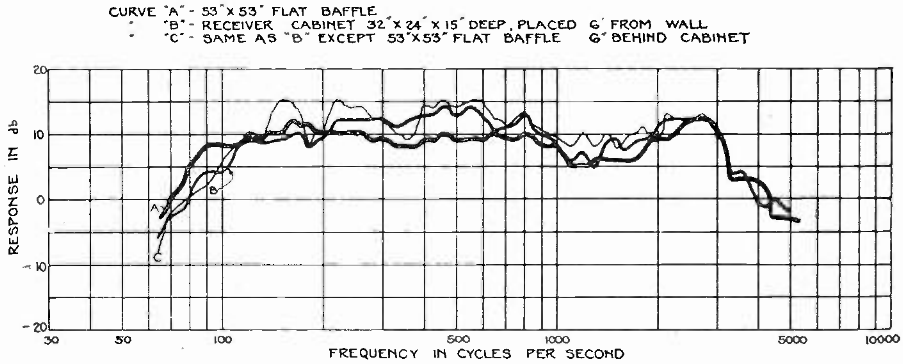

Fig. 3—Response of electrodynamic cone speaker in flat baffle and in commercial receiver cabinet. Points below 65 cycles not shown because of measuring conditions.

resonances in the cavity behind the loud speaker or in the walls of the cabinet itself. Curve *C* shows the detrimental effect imposed by the condition of ordinary use, namely, the rear of the cabinet placed close to a wall. The peak at 150 cycles is undoubtedly due to resonance in the now more completely enclosed space behind the loud speaker. The bottom of this cabinet was left entirely open in order to minimize this effect, which is decidedly more pronounced in cabinets with closed bottoms.

The curves of Fig. 4 are interesting in view of the popular idea, expressed in numerous magazine articles and advertising statements, that the limiting effect of a baffle in the reproduction of the lower frequencies depends only on the length of circulation path from front to rear of the loud speaker diaphragm. Three baffles were constructed to the dimensions given in the figure, all three having the shortest distance from front to rear of the diaphragm the same (48 inches) but having different face areas. The same loud speaker was measured in each of these baffles. It will be noticed that, as the face area is decreased, the low-frequency response falls off markedly and, as the depth of the box is increased, the effects of cavity resonance become evident in the peaks which appear in the curves. In the case where the depth of the box exceeds its side dimension an extended series of peaks is produced, as shown in curve *C*.

The effect of carrying the depth of a box baffle to extremes as compared with the face dimensions is shown in Fig. 5 where the baffle takes

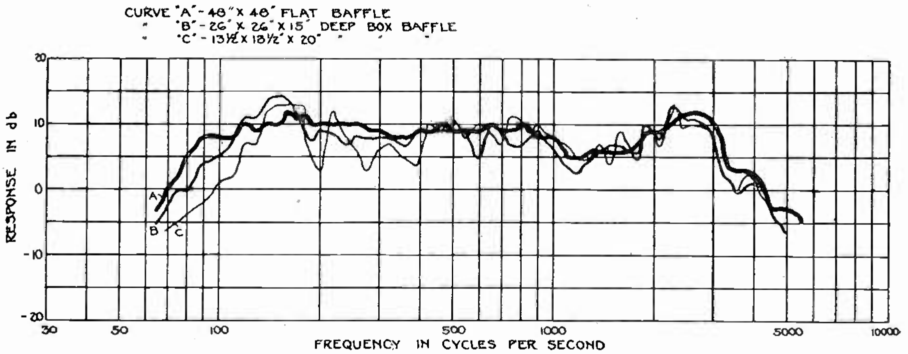

Fig. 4—Response of electrodynamic cone speaker in three baffles having same circulation path length. Points below 65 cycles not shown because of measuring conditions.

the form of an organ pipe. The response curve of the same loud speaker in a flat baffle is shown for comparison. The large variations in response caused by air column resonance within the pipe are to be noted,

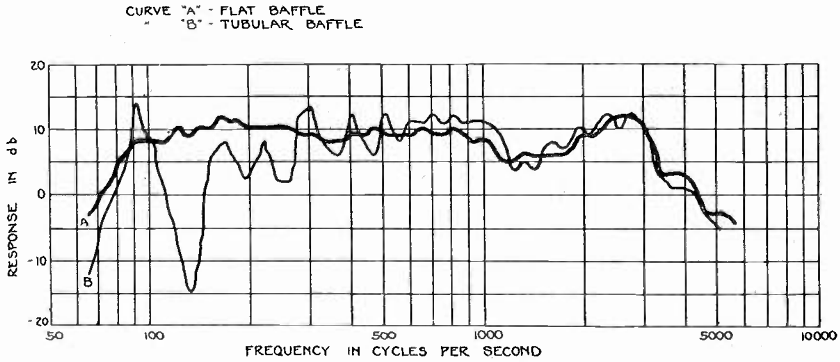

Fig. 5—Response of electrodynamic cone speaker in a 53 × 53 inch flat baffle and in a 60×13×13 inch tubular baffle. Points below 65 cycles not shown because of measuring conditions.

particularly the change of 29 db between 90 cycles, the fundamental note of the pipe, and 130 cycles. It is interesting to observe that most of the peaks fall at frequencies somewhat higher than those corresponding to harmonics of the fundamental tone, in view of the fact that in pipes which are wide (from the organ builder's point of view) in comparison with their lengths, the overtones are always sharp. This is the loud speaker referred to before, which gave the effect of excessive room reverberation in reproducing speech.

In Fig. 6 is shown the effect of cabinet size and construction upon the response of a high-grade loud speaker. Curve *A* is that of the speaker in a substantially constructed console cabinet without a bottom, it being the same one shown in Fig. 3, curve *C*. Curves *B* and *C* (Fig. 6) were made with the loud speaker housed in progressively smaller open-backed boxes built of 1/2-inch lumber and having the speaker opening located near the top of the front face to simulate the smaller type cabinets now so common. The loss of low frequencies resulting from restricted baffle area and the peaks caused by acoustic resonance in the boxes and by vibration of the sides may readily be noted.

Curve *D* (Fig. 6) is of the same loud speaker mounted in a "midget" receiver cabinet with sides and top made of 1/8-inch bent plywood and

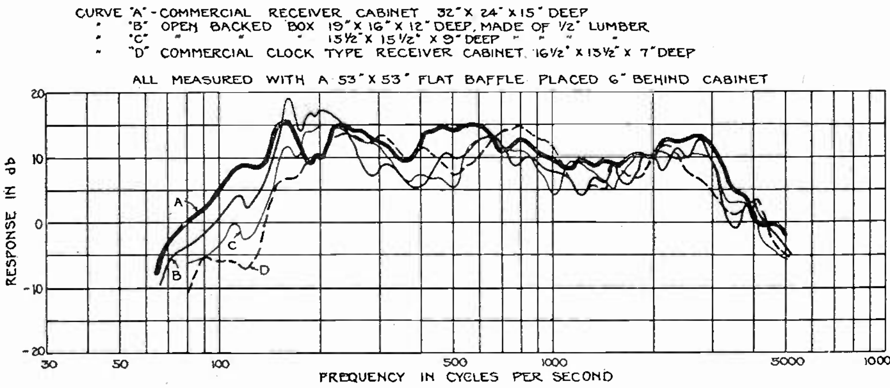

Fig. 6—Response of electrodynamic cone speaker in different size cabinets. Points below 65 cycles not shown because of measuring conditions.

having a front panel 1/4-inch thick. The extreme loss of low frequencies as compared with the console cabinet of curve *A* is strikingly evident. This small cabinet was measured with the grille cloth removed, as the cloth which is regularly mounted in it, was of such close weave that it produced an average measured loss of about 10 db. The boxes of curves *B* and *C*, Fig. 6, had no grille cloth over the speaker opening. The receiver cabinet of curve *A* was measured with its grille cloth in place but the weave was so open that a hardly measurable acoustic loss resulted.

It sometimes is suggested that the low-frequency response of these small receivers may be made equivalent to that of full size consoles by equalization in the audio amplifier. Considering the approximately 10-db difference between curves *A* and *D* in the region under discussion, this means that it would be necessary for the output tubes in the smaller receiver to deliver 10 times the power of those in the larger one for equal sound output in the low-frequency range. It also means that the diaphragm of the loud speaker in the smaller cabinet would have to be driven at more than three times the amplitude of the other. As both tubes and loud speakers are being worked close to their maximum power handling capacity in this frequency range in even the larger receivers, the impracticability of this suggestion with present apparatus is evident. It should be borne in mind that the same highly developed loud speaker was employed in all the above cases, the object being to demonstrate only the effect of the cabinet.

In an over-all comparison of a high-grade, full size receiver with a "midget" set, the latter is at a greater disadvantage than is shown in the curves or indicated in the previous statements because it usually is

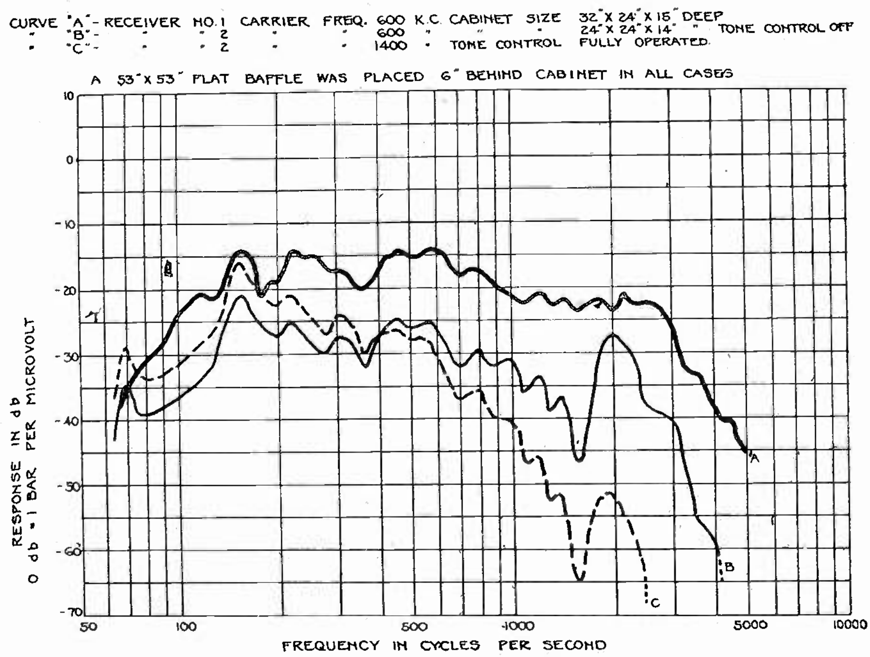

Fig. 7—Over-all response curves of two radio receivers. Points below 65 cycles not shown because of measuring conditions.

fitted with a much less efficient loud speaker of small size and has an audio amplifier of lower power handling capacity. The smaller size of the loud speaker diaphragm also increases the amplitude of motion necessary to generate low-frequency sounds and makes even more impracticable their satisfactory reproduction by the suggested equalization.

In Fig. 7 are shown the over-all response curves of two radio receivers in cabinets of the dimensions given. Curve *A* is of a receiver in a full size console cabinet. Curves *B* and *C* are of another receiver in a small size console cabinet. Note the stronger low-frequency response of the receiver with the larger cabinet. The larger cabinet also is fitted with a better loud speaker, as is indicated by the smoother response curve, particularly in the higher frequency region. Both the loud speaker and the electrical circuits of the receiver of curve *B* contribute to the sharp cut-off above 3000 cycles that occurs even when the tone control, that is provided on this receiver, is set to the position giving the maximum frequency range. Note the enormous frequency distortion and the limited frequency range disclosed by curve *C*, taken with the tone control set to the position giving maximum low-frequency emphasis. The natural character of voices and musical selections is completely lost with this setting, as the higher fundamental tones and a great proportion of the overtones are obliterated and, in what remains, the lower frequency components are greatly exaggerated with respect to the higher ones.

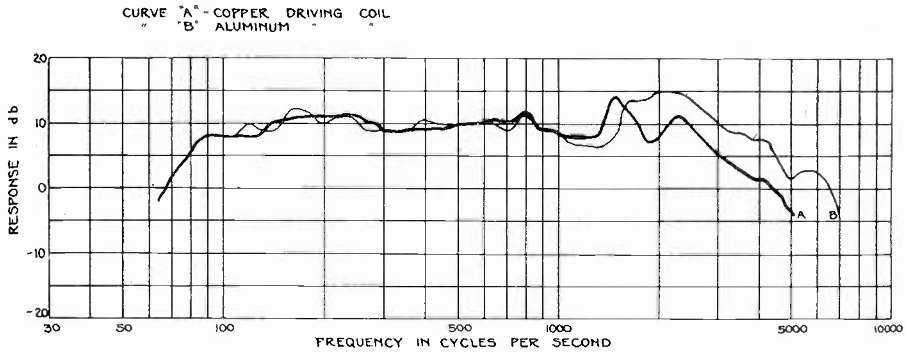

Fig. 8—Response of same cone with copper and with aluminum driving coils. Measured in 53 by 53 inch flat baffle. Points below 65 cycles not shown because of measuring conditions.

In most receivers which are sufficiently selective to meet present-day conditions, the high-frequency response is limited more by the electrical circuits of the receiver than by the loud speaker itself. Radio loud speakers having considerably better high-frequency response than those ordinarily available may be had by the use of lighter driving coils.

Fig. 8 shows the response curve of an experimental loud speaker in which the usual copper driving coil was replaced by an aluminum one of the same size. The curve of the same loud speaker with the copper coil is shown for comparison. The same cone was used in both cases. It will be noted that there is no substantial change due to the lighter coil except over that portion of the frequency range wherein the cone breaks down into flexural wave motion. This is to be expected because the reduction in the calculated mechanical driving point impedance, considering the diaphragm rigid and including its acoustic load in an infinite baffle amounts, at 100 cycles, only to about 10 per cent, and changes the approximately 87-degree phase angle by less than 1 degree. At frequencies above 1000 cycles, however, where wave motion takes place, it is reasonable to expect that less than the entire mass of the cone is in active vibration and also that the radiation resistance per unit area is close to its ultimate value. Under these circumstances the reduction in driving coil mass obviously would be more effective in increasing the efficiency of the loud speaker. It appears also from the change in shape of the upper portion of the curve that the substitution of the lighter coil affected the modes of vibration of the cone. The aluminum coil loud speaker gave substantially improved quality of reproduction even when operated from a fairly selective receiver, although, of course, it also raised the level of hissing types of background noise.

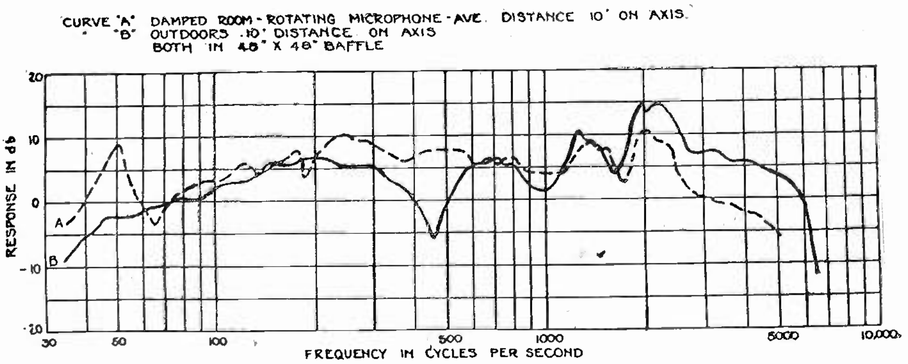

Fig. 9—Comparison of indoor and outdoor response measurements on electrodynamic cone speaker.

In Fig. 9 is shown a comparison between indoor and outdoor measurements made on the same loud speaker under conditions described earlier in this paper. The discrepancies between the two curves will bear some explanation. The peak at 50 cycles in the indoor curve is due to the large dimensions of the standing wave pattern in the room as compared with the size of the region swept over by the microphone, thus preventing a true average from being obtained. This peak was present in the original indoor curves of all of the loud speakers shown in this paper whose response extended sufficiently low. *In the published curves, points below about 65 cycles were omitted, when obtained at all, because of this effect peculiar to the measuring condition.*

The next major discrepancy is the depression in the outdoor curve reaching a response minimum at 460 cycles. This minimum occurred at the same frequency when the measuring distance was increased to 15 feet. It is caused by interference at the microphone position between the sound coming from the front of the loud speaker and that diffracted around the baffle from the rear of the diaphragm. The waves start from the front and rear of the diaphragm in phase opposition and if the distances which they traverse in reaching the microphone differ by an integral number of wavelengths, the pressures at the microphone will oppose and a response minimum will be indicated. When the distances differ by an integral number plus one-half wavelengths, the pressures will add and a response maximum will be produced.

In mathematical form, the condition for

$$\text{minimum pressure is } \frac{d_1 - d_2}{\lambda} = N, \text{ an integer,} \tag{1}$$

$$\text{and for maximum pressure is } \frac{d_1 - d_2}{\lambda} = N + \tfrac{1}{2} \tag{2}$$

where d_1 and d_2 are the distances and λ is the wave length. These relations are often more convenient when expressed in the following form obtained by substituting c/f for λ in the above equations, c being the wave velocity, and f the frequency:

$$f = N\left(\frac{c}{d_1 - d_2}\right) \text{ at a pressure minimum,} \tag{3}$$

$$\text{and } f = N + \tfrac{1}{2}\left(\frac{c}{d_1 - d_2}\right) \text{ at a pressure maximum.} \tag{4}$$

The lowest frequency at which this effect can occur is obviously that obtained by taking N equal to unity. Theoretically there should be a succession of minima and maxima starting at this frequency and continuing throughout the range of the loud speaker. In the practical case, however, the radiation becomes so directional at the higher frequencies that the effect of the sound from the rear is negligible at the microphone position. In curve B of Fig. 9 the first maximum which should theoretically fall at 690 cycles may be indicated by the rise centering around 700 cycles. This is more or less masked by the succeeding depression which is caused by an antiresonance in the diaphragm itself. None of the succeeding theoretical minima and maxima may be detected with any certainty.

It is interesting to compute the distance traveled by the wave from the rear of the diaphragm by solving (3) for d_1. Taking f equal to 460 cycles, N as 1, c as 1130 feet per second, and d_2 as 10 feet, d_1 comes out 12.5 feet. The shortest possible distance over which the sound from the rear could reach the microphone by bending sharply around the edge of the baffle is 11.8 feet. Considering that the microphone position was uncertain in any direction within probably an inch or two and that the flow lines of the sound may not bend sharply around the baffle edges, the agreement between the above values for d_1 may be taken as good evidence that the cause of the minimum at 460 cycles has been correctly assigned. The reason that the frequency at which this depression occurred did not shift appreciably when the measuring distance was increased is that the *difference* in the lengths of the two paths was changed only slightly thereby.

The greater high-frequency response indicated by the outdoor curve is easily explained. It is well known that the radiation at high frequencies from diaphragms of the size commonly employed in baffle-type speakers is quite directional, being concentrated in a beam along the axis of the cone. In the outdoor measurement, the microphone was

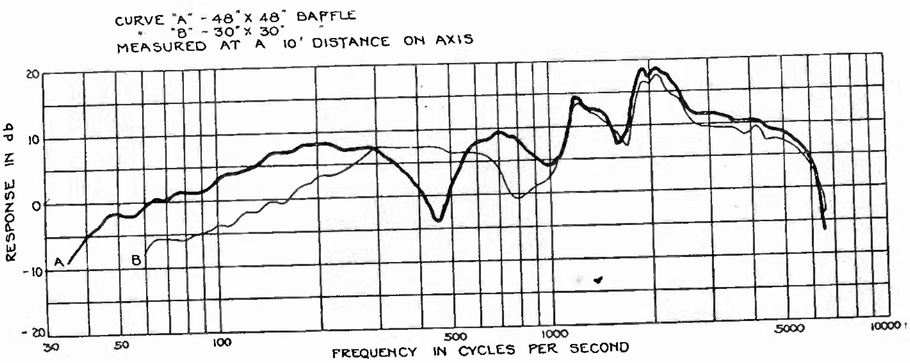

Fig. 10—Outdoor response curves of electrodynamic cone speaker in different size flat baffles.

located directly in the path of this beam while in the indoor test the rotating microphone subtended a comparatively wide solid angle of the radiation and never actually crossed the beam. *The results of the two methods therefore are not directly comparable.*

In Fig. 10 is shown a comparison between outdoor measurements at the same microphone position on a loud speaker mounted in a 48×48 inch and in a 30×30 inch flat baffle. Note the marked reduction of low-frequency response with the smaller baffle. The measurements in both cases were carried down to as low frequencies as the limitations of the testing apparatus would permit. In the case of curve *B*, the first minimum due to interference between the sound coming from the two sides of the speaker occurs at 800 cycles. The distance traveled by the back wave, calculated in the same manner as for the previous case of the larger baffle, is 11.8 feet compared with the shortest possible path of approximately 11.0 feet, the difference in these values being about the same in magnitude and direction as for the larger baffle.

The humps in both curves below 200 cycles appear to be caused by reflection from the earth and by their smallness indicate that this disturbing factor has been reduced to a practically insignificant magnitude. Considering the loud speaker as a single point source of sound and assuming a similar image loud speaker located 70 feet below it in free space to represent the effect of reflection by the earth, the first theoretical interference maximum would occur at about 19 cycles and other maxima would appear at this frequency interval throughout the range where earth reflection was effective. It is interesting to observe that in most cases these minor humps follow approximately the above frequency spacing. The measurements were made only to the nearest decibel which may account for the discrepancies. The agreement between curves A and B above 1000 cycles is thought satisfactory for this class of measurement, considering a certain amount of uncertainty in the relative location of the loud speaker and microphone and the fact that a period of seven days elapsed between the two tests.

Any *single* free space response measurement of a loud speaker which does not radiate uniformly at all frequencies, made with a fixed relation between loud speaker and microphone, is of more or less limited value. An interesting possibility in connection with the outdoor set-up described and one which, in a single response curve, would furnish a maximum of useful information regarding the performance of a loud speaker, is the taking of a free space response average over a region representative of a range of listening positions, somewhat as is done indoors by means of the rotating microphone. In the outdoor set-up, a convenient and practicable arrangement might be to place the microphone in a fixed position on the sound axis and then to oscillate the loud speaker slowly through a suitable angle about its vertical axis. The mechanical arrangements involved should not be as difficult or as cumbersome as those required in the indoor set-up and by the addition of a scale and pointer which could be read from the ground and the provision of proper controls for the driving motor, would also be convenient for taking directional distribution curves. The thermocouple and long-period meter would, of course, be used as a response indicator in the oscillating measurements.

Conclusion

It is thought that acoustic measurements have now been developed to the point where sufficient information is available to permit significant and useful loud speaker tests to be made. While the difficulties surrounding this type of measurement are numerous and there is considerable danger of misleading results unless proper precautions are observed, improvement in methods is bound to come as more workers accumulate experience in this field. The experimental curves presented in this paper are indicative of how valuable such measurements may be to engineers.

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HIGH AUDIO POWER FROM RELATIVELY SMALL TUBES*

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Introduction

SINCE the early use of a headphone at the base of a horn for a loud speaker, the demand for higher output power from an audio system has steadily increased. This demand for higher power has been met mainly by the development of tubes for higher plate dissipation and lower plate resistance. As a result of such development, the 245-type tube is the most commonly used vacuum tube for output systems. Other output tubes are very closely related to the 245 tube in so far as amplification constant and plate resistance are concerned and differ principally in rated plate dissipation and plate voltage for various output levels.

These low plate resistance tubes developed primarily for audio output systems are intended to operate as class "A" amplifiers. To increase the output of such systems, it is necessary to resort to higher plate dissipation and usually to higher plate voltages, both of which are expensive to incorporate in the construction of a vacuum tube.

The purpose of this paper is to present a method by which audio outputs five to ten times the usual output of a tube of a given size may be obtained with the same plate voltage, lower average plate dissipation, and no serious effects on the tube. The above results are obtained by using the tubes in such a manner that advantage is taken of the essential features of the class "B" amplifier.

An amplifier of this type[1] was developed for use as a source of high audio power for plate modulation of a broadcast station and may be applied to any system requiring a relatively high audio output from a minimum of equipment.

A discussion of the class "A" and class "B" amplifiers with diagrams illustrating the operation of each class will be given to set forth some of the essential operating features of these amplifiers.

Class "A" Amplifiers

The class "A" amplifier as indicated above, is used almost exclusively for audio output systems in which the output is aperiodic. This

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[1] Loy E. Barton, "A plate modulation transformer for broadcasting stations", *University of Arkansas Engineering Experiment Station Bulletin* No. 8.

type of amplifier is also used in radio-frequency amplifiers in which the output of plate circuit is tuned for selecting the desired frequency.

The diagram in Fig. 1 represents the instantaneous plate and grid voltages and plate current for a condition of essentially maximum

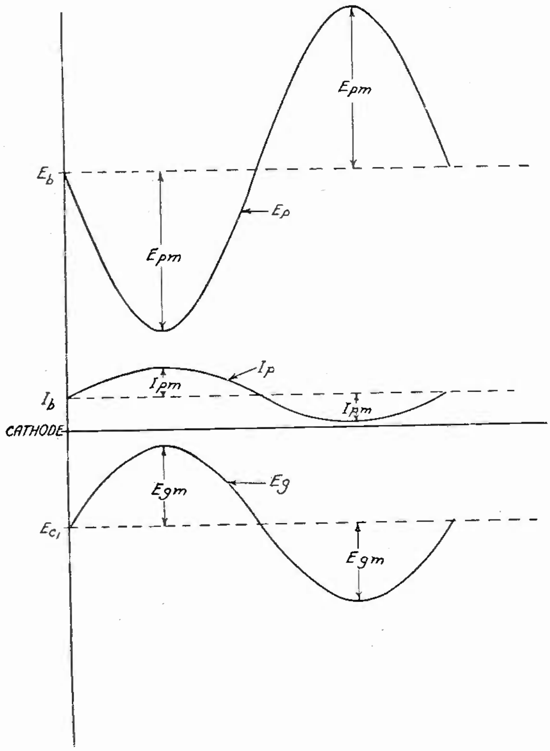

Fig. 1—Diagram of instantaneous voltages and currents for a class "A" amplifier.

output power of a class "A" amplifier into a tuned or untuned circuit, with a resistance load expressed in terms of the values indicated on the diagram as:

$$\frac{E_{pm}}{I_{pm}} = R_p \text{ or load resistance.}$$

In which

$$E_{pm} = \text{maximum a-c plate voltage} \tag{1}$$

$$I_{pm} = \text{maximum a-c plate current.}$$

The plate voltage is represented by the broken line E_b, the plate current, by the broken line, I_b, and the grid voltage (normally negative) by the broken line E_{c1}.

The power in the above case for sinusoidal waves is limited by the minimum plate current and may be calculated as follows when I_{pm} is somewhat less than I_b.

$$0.707E_{pm} \times 0.707I_{pm} = \text{power output} \qquad (2)$$

$$E_b \times I_b = \text{power input.} \qquad (3)$$

The plate circuit efficiency from (2) and (3) is given by the following relation:

$$\frac{0.5\, E_{pm} I_{pm}}{E_b I_b} = \text{efficiency.} \qquad (4)$$

If E_{pm} approaches E_b as a limit and I_{pm} approaches I_b as a limit, the efficiency is 50 per cent for sine wave outputs.

The above expression for power output is independent of the required grid swings and internal plate resistance of the tube as long as a particular grid swing will cause the same maximum pulsating plate current through a resistance equal to the calculated value of R_p. In order that I_b equal a constant for tubes differing only in plate resistance and amplification factor, E_{c1} must be nearer zero for high plate resistance tubes and the grid potential may need to swing considerably positive to obtain a constant a-c plate current maximum I_{pm}. It is obvious that power will be necessary to maintain an undistorted voltage wave on the grid when it swings positive.

Load resistance curves for two tubes as class "A" amplifiers, differing only in plate resistance and amplification constant, are given in Figs. 2 and 3. These load curves are used in preference to the usual plate voltage plate current family of curves for output calculations because certain features of the class "A" amplifiers are revealed that are not commonly recognized and this type of curve is particularly adapted to output calculations for the class "B" audio amplifier. These curves are drawn for the same load resistance in a circuit as shown in the figures and the plate current is adjusted to 20 ma in the two cases by the proper grid voltage with 400 volts on the plate. The voltage drop in the load resistance R_p due to the plate current is compensated for by increasing the plate voltage supply. Referring to Fig. 1, the plate voltage represented by E_b is 400 volts and the plate current represented by I_b is 20 ma. Therefore, the plate dissipation for each of the two tubes is the same.

The grid potential is varied about the voltage line required to obtain 20 ma with 400 volts on the plate and the corresponding plate current is plotted. The load resistance R_p is adjusted to such a value that the maximum increase in plate current over the straight portion of the curve is equal to approximately the maximum plate current decrease for equal grid swings about the starting grid potential. It will be noted that in each case the grid must be driven positive if the above plate current condition is obtained. If the grid cannot be driven positive, the grid voltage for the 20 ma plate current for the given plate dissipation would have to be more negative so that a grid swing to zero will give the above plate current change. Such a bias will require a tube of the same type as the two tubes chosen but with a lower plate resistance and lower amplification constant.

It will be noted in the above cases, that the maximum plate current change above and below the 20-ma line is approximately 18 ma before the curve bends appreciably at either end. Therefore, the a-c power dissipated in R_p, which is 15,000 ohms for all cases, is obtained from (1) and (2).

$$0.5\ I_{pm}^2 R_p = 2.43 \text{ watts.} \tag{5}$$

In which $I_{pm} = 18$ ma

$$R_p = 15{,}000 \text{ ohms.}$$

The plate dissipation at no signal and input power with signal is:

$$E_b\ I_b = 8 \text{ watts.}$$

Therefore, the plate circuit efficiency for the above case is:

$$\frac{2.43}{8} = 30.3 \text{ per cent.}$$

In Fig. 2, the grid swings to a peak of 10 volts positive with a peak grid current of 3 ma,which represents a minimum instantaneous input resistance of approximately 15,000 ohms. This input resistance is based upon the peak grid current for the required voltage swing from the starting grid potential. Therefore, the peak instantaneous input power required of the driver is the product of grid voltage swing and the maximum grid current. Since the resistance of the input circuit over most of the cycle is very high and the resistance represented by the grid current curve is approximately 3000 ohms, the driver resistance must be sufficiently low to deliver the above peak power over a part of the input cycle without seriously affecting the wave shape of the input voltage. It will be noted that at 18-ma increase in plate current for this tube, the curve is still straight so that a power slightly above the 2.43 watts may be obtained if a slightly higher resistance is used.

In Fig. 3, the grid swings to a peak of 20 volts positive with a peak grid current of approximately 7.5 ma, which represents a minimum instantaneous input resistance of about 3000 ohms as calculated above, and the resistance from the grid current curve is approximately 2900 ohms. The peak input power is about the same as required by the tube

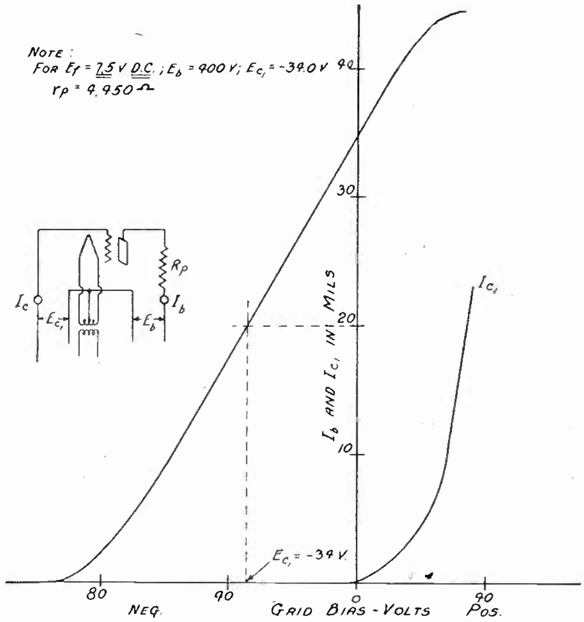

Fig. 2—Load characteristic of the UX-210 as a class "A" amplifier.

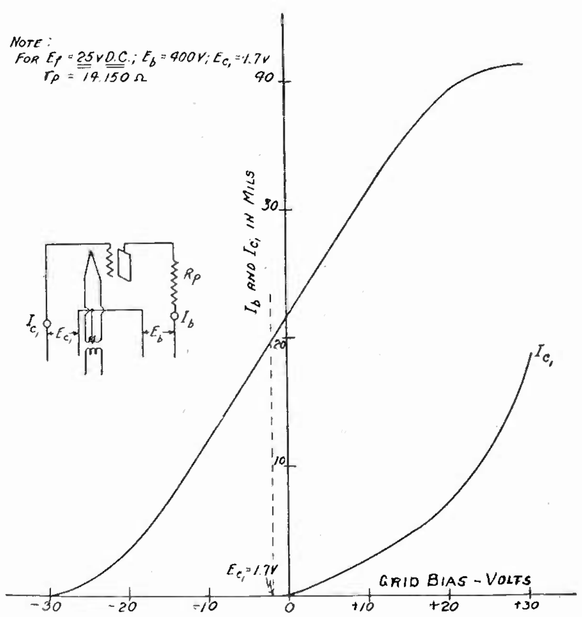

Fig. 3—Load characteristics of the UX-841 as a class "A" amplifier.

in Fig. 2, but because of the fact that the power is required over a larger part of the input cycle, it is somewhat easier to supply an undistorted wave to the grid.

If the tubes represented by Figs. 2 and 3 are operated in a push-pull amplifier connected as indicated in Fig. 4, the output transformer primary impedance from plate to plate is 30,000 ohms, the power output will be doubled and the problem of driving the grids becomes less difficult because the input resistance is more nearly constant over the entire cycle. This statement applies particularly to two output tubes operating at low bias so that one grid or the other requires grid current at all times. This flow of grid current through one side or the other of the input transformer secondary reflects to the primary of this transformer an essentially constant impedance into which the driver tube works.

It will be noted from Figs. 2 and 3 that the slope of the grid-current curve is nearly the same for the two types of tubes. This

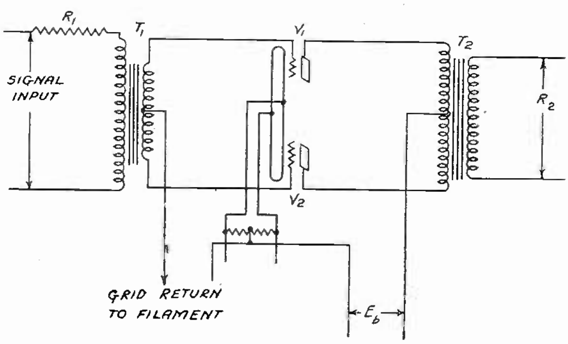

Fig. 4—Circuit for push-pull class "A" amplifier or for class "B" audio amplifier.

means that the effective series resistance of the source driving the amplifier must be considerably below the minimum value of input resistance if it varies considerably over the cycle as indicated in Fig. 2. Since the grid voltage swing for the tube in Fig. 2 is great, a driver capable of supplying considerable power must be used. Therefore, the higher amplification constant tubes are much more desirable if the grids are to be driven positive.

Since the input systems for the above tubes operating in a push-pull manner must be low impedance, a lower load resistance may be used which in turn, permits higher plate-current swings at the expense of added distortion, but with no greater plate dissipation. This increased plate current permits a much greater power output with a permissable increase in distortion.

From the above discussion, it is evident that for an undistorted input wave, the output load resistance for maximum power output is a function of the plate current and plate voltage and is essentially independent of plate resistance and amplification constant provided the grid swing is not limited. For example, in the above cases, the maximum plate current increase was 18 ma through a resistance of 15,000 ohms. The increased drop in the resistor is the product of these values which is 270 volts and the voltage at the plate for the increased current is only 130 volts. Consequently, a class "A" amplifier power output is limited essentially by plate dissipation and plate current change so that the value of load resistance is determined by the minimum plate voltage required at the maximum positive swing of the grid to permit a plate current of approximately 190 per cent of normal plate current.

The plate resistance was measured for each of the tubes under the conditions of bias and plate voltage as indicated above and is noted in each figure. Although the maximum output power into a definite load resistance is about the same for both the tubes, the regulation of the amplifier output under varying load resistances is a function of the load resistance and the internal plate resistance of the tube. Therefore, if a load resistance for optimum power output is used and the excitation is constant, the output voltage will be nearly constant for very low resistance tubes for small changes in load resistance. If, however, the plate resistance in the above case is very high with respect to the same load resistance, the output current will be nearly constant for small changes in load resistance.

Class "B" Amplifiers

A diagram similar to Fig. 1 is shown in Fig. 5 for the class "B" amplifier. The grid potential E_{c1} for this type of amplifier is such that the plate current without signal is essentially zero as represented by the broken line I_b. No plate current flows during the negative swing of the signal but the plate current is essentially proportional to the instantaneous value of the grid voltage on the positive swings. Therefore, the output voltage is essentially proportional to the input voltage during the positive half of the input cycle and the peak plate current is limited only by the emission, the load resistance, and plate voltage.

The load resistance for this amplifier is calculated as with the class "A" amplifier which is:

$$\frac{E_{pm}}{I_{pm}} = R_p \text{ load resistance.}$$

The power output for a condition as represented is:

$$\frac{0.707\, E_{pm} \times 0.707\, I_{pm}}{2} = \text{power output.} \tag{6}$$

The efficiency of this amplifier is given by:

$$\frac{\dfrac{0.707\,E_{pm} \times 0.707\,I_{pm}}{2}}{\dfrac{0.637\,I_{pm}\,E_b}{2}} = \text{efficiency.} \qquad (7)$$

If E_{pm} approaches E_b as a limit, then the efficiency becomes 78.5 per cent for half-sine-wave outputs.

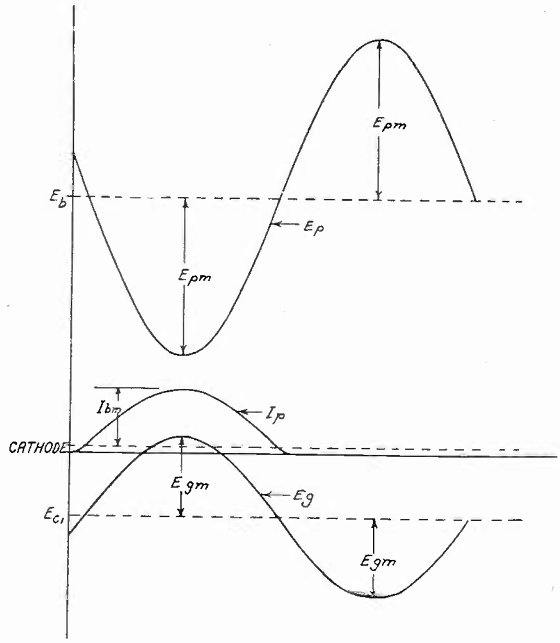

Fig. 5—Diagram of instantaneous voltages and currents for a class "B" amplifier.

If the output is tuned to the frequency of the input, the output voltage or E_b will be nearly sinusoidal but if the output is not tuned, only the half wave during plate current flow will be sinusoidal. As stated above, the maximum plate current I_{pm} is only limited by emission, load resistance, and plate voltage. Therefore, the value of the above expression for power may be several times the output power of a tube limited in plate current swing when operating as a class "A" amplifier. This type of amplifier with tuned radio-frequency output is commonly used as the output amplifier for radiophone transmitters and has a plate circuit efficiency on peak outputs of 50 to 70 per cent.

Class "B" Audio Amplifiers

The class "B" amplifier as explained above, is a tuned radio-frequency system and is not useful as represented for an audio output amplifier. However, if advantage is taken of the fact that the output during one-half of a cycle into an untuned load is essentially sinusoidal for a sinusoidal input, another similar tube may be used in such a manner that an undistorted output may be obtained. The circuit for such an amplifier is shown in Fig. 4.

A typical push-pull output circuit is used for the class "B" audio amplifier with certain deviations from the commonly used class "A" push-pull amplifier. The grid return to the filament in Fig. 4 is connected through the correct potential to limit the plate current of the tubes V_1 and V_2 to essentially zero. The impedance ratio of the input transformer T_1 is such that the reflected resistance on the primary, when the grids of V_1 and V_2 are driven positive, is high compared to the internal resistance of the driving source represented by R_1 as explained for class "A" amplifiers when the grids are driven positive. This relation of the internal resistance of the driver compared with the minimum reflected resistance of the grids of V_1 and V_2 reduces the distortion of the voltage applied to the grids of these tubes.

When the grid of one tube swings in a positive direction from its d-c value, the plate current for that tube increases with the swing and flows through one-half of the output transformer primary. The output voltage for this half cycle bears a linear relation to the input voltage. At the beginning of the next half cycle, the above tube becomes idle because its grid becomes more negative and the other tube functions in a manner similar to the first tube, except the plate current flows in the other half of the output transformer primary and the output voltage is 180 degrees out of phase with the first half wave. These two output waves then will form a wave similar to the input wave with no distortion if the plate current and grid voltage have a linear relation. This relation of plate current and grid voltage is not quite linear as will be seen from the grid voltage plate current curves, but by proper precautions, the distortion can be reduced to a point that is not objectionable.

It should be noted that the input transformer delivers current to the grids of the output tubes from only one side of the secondary at any particular instant which must be considered in the design of this transformer. It should also be noted that the entire output power must be transferred from only one side of the output transformer primary during each half cycle. Therefore, the load impedance the tube is working into is calculated as if only one tube is supplying the total power from one side of transformer primary but in calculating plate dissipation, each tube functions for one-half the time so that the total plate loss is divided between the two tubes.

The plate current input to the output tubes resembles a true full-wave rectified current, the frequency of which is double the frequency of the signal. Therefore, the power input to the plates of the output tubes is:

$$
\begin{aligned}
0.637\, I_{pm} E_b &= \text{power input} \\
I_{pm} &= \text{peak plate current} \\
0.637\, I_{pm} &= I_b \text{ or average plate current.}
\end{aligned}
\qquad (8)
$$

in which

The output power from (6) is:

$$\frac{I_{pm}{}^2 R_p}{2} = \text{power output for two tubes} \qquad (9)$$

and,

$$\frac{I_{pm}{}^2 R_p}{1.274\, I_{pm} E_b} = \text{efficiency.} \qquad (10)$$

Load resistance curves similar to the load curves for class "A" amplifiers were taken for several tubes adapted to class "B" audio amplification. As was noted above, the plate current is not limited as in the class "A" amplifier so that a load resistance for maximum power output is such a value that the limit of emission is approached. The minimum instantaneous plate voltage and the allowable plate dissipation are also factors which determine the load resistance. No allowance for plate voltage drop due to the load resistance was made in taking data for the load curves because the plate current for zero signal is nearly zero.

The peak plate current at a point on the desired load curve which is within the permissible distortion limit is used to calculate the output power of two tubes operating as a class "B" audio amplifier. This value of peak plate current and the corresponding load resistance are used in (9) to obtain the power output of two tubes operating as a class "B" audio amplifier. The input plate power is calculated by using (8) and the plate circuit efficiency by (10).

The input resistance of the amplifier is more or less indefinite as with the class "A" amplifier with the grids driven positive, but a minimum value can be obtained from the grid current curves to indicate the input series resistance that may be permitted for a given distortion. The minimum input resistance is calculated from the maximum grid current and the voltage swing required for this current, however, the slope of the grid current curve must also be considered because if this resistance represented by the slope of the curve is low compared with the effective grid series resistance, considerable distortion will result This is especially true if the operating bias is considerably negative as it would be for the medium and low amplification factor tubes. If the grid current curve has a negative resistance, a condition for oscillation will occur over a small portion of the input cycle if the effective series input resistance is large compared with the negative resistance. A low negative input resistance produces an effect similar to a rattle and is not a harmonic of the input signal.

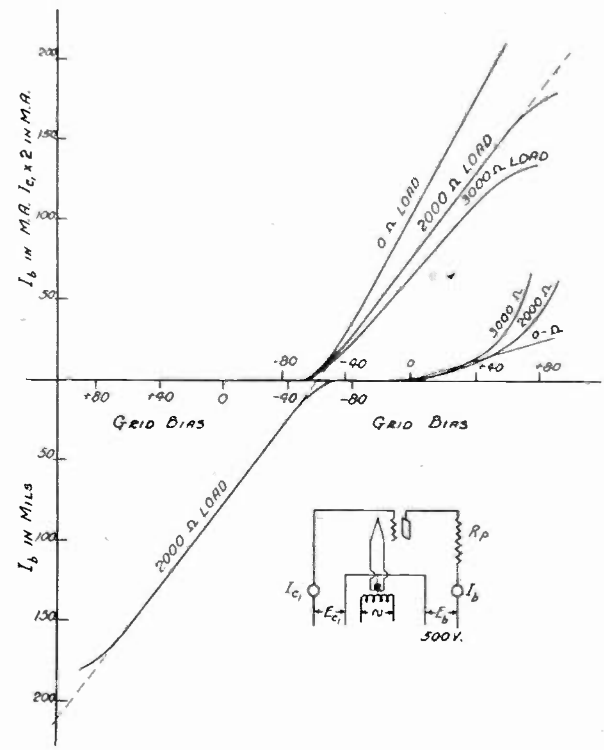

Fig. 6—Load characteristics of the UX-210 as class "B" audio amplifier.

The tubes used for the curves in Figs. 6 and 7 are the same tubes used respectively for the curves of Figs. 2 and 3. These identical tubes were used for the class "A" amplifier load curves and the class "B" audio-amplifier curves so that a direct comparison could be made in the operation of these amplifiers.

The standard UX-210 tube as represented in Fig. 6 has a peak plate current of approximately 170 ma without serious deviation from a line through the straight portion of the curve for the 2000-ohm load curve. Another load curve similar to the upper curve is drawn below the zero plate current line and with a reversed grid voltage scale. This curve represents another tube connected in a push-pull manner with the tube above the zero plate current line and is placed in such a position that a line drawn through the straight portion of the upper curve also passes through the straight portion of the lower curve. The point through which this line passes on the zero plate current axis is the proper bias to use for minimum distortion if the two tubes are similar. The above load is the load per tube during the time it operates so that the plate to plate impedance of the output transformer is 8000 ohms.

The above values of plate current and load resistance are used in (9) to calculate power output.

$$0.5\ I_{pm}^2\ R_p = 28.9 \text{ watts output}$$

Plate power input as calculated from (8) is:

$$0.637\ I_{pm}\ E_b = 54.15 \text{ watts input.}$$

The plate circuit efficiency is:

$$28.9/54.15 = 53.4 \text{ per cent.}$$

The plate loss for full power output is 25.25 watts for the two tubes and is essentially constant to an output power of about 30 per cent of the maximum below which it begins to decrease rapidly. This variation in plate loss with respect to the power output is obvious from (10) which indicates that the plate circuit efficiency is a direct function of the peak plate current.

The above calculations are based upon no distortion of the input sine-wave signal when the grid swings positive. Since the input resistance is a variable it is only possible to reduce the distortion to a practical minimum by the proper design of the driver. The peak grid swing for 170 ma in a 2000-ohm load is 75 volts positive from a bias of 58 volts or a total swing of 133 volts. The peak grid current is 15 ma so that the minimum input resistance is approximately 8800 ohms, and the peak instantaneous power is 2 watts which must be supplied without affecting the input wave seriously. The resistance as calculated from the slope of the grid current curve is approximately 5000 ohms.

The above information indicates that if only a small amount of distortion is permitted, the series input resistance to each grid should be approximately 20 per cent of the minimum input resistance or about 1000 ohms because the difference in the above resistances is great. If the effective input series resistance is the above value and two tubes having a plate resistance of 2000 ohms operating in push-pull are used to drive the UX-210 tubes, the resistance in series with the primary is 4000 ohms. Therefore, the impedance ratio from total primary of the input transformer T_1, Fig. 4, to each grid is 4 to 1 or a turn ratio of 2 to 1. The required grid voltage swing is 133 so that a primary voltage peak of 266 is required. A study of the UX-245 characteristics at normal voltages indicates that two of these tubes in push-pull will successfully drive the UX-210 tubes as class "B" audio amplifiers without serious distortion.

The same calculations as above may be made from the curves in Fig. 7 for the UX-841 tube which is a tube similar to the UX-210 except for amplification constant and plate resistance. So far as the power output is concerned, it can be seen that the output of two of these tubes is approximately the same as for the UX-210 tubes and the efficiency is essentially the same. However, the bias required for

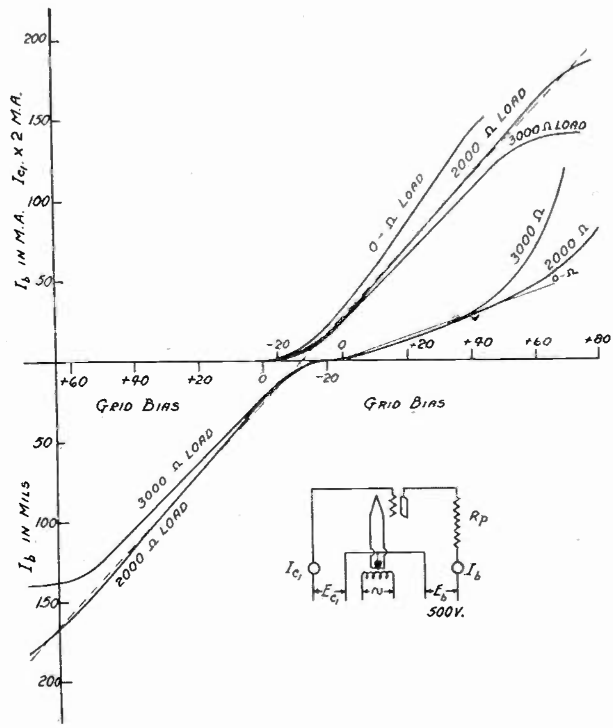

Fig. 7—Load characteristics of the UX-841 as a class "B" audio amplifier.

these tubes is approximately 13.5 volts and the peak positive grid swing is 65 volts with a peak grid current of 25 ma. The peak swing from normal bias in this case is 78.5 volts so that the minimum input resistance per grid is 3140 ohms and the slope of the grid current curve represents 2600 ohms. Since the above resistances are near the same value, the input series resistance may be as high as in the case of the UX-210 without increased distortion. If full advantage is taken of the above 255 volts peak swing on the primary of the driver transformer, an impedance ratio to each grid may be approximately 12 to 1 or an equivalent series input resistance of 330 ohms. This resistance is about

12 per cent of the minimum input resistance as compared to 20 per cent for the UX-210 tubes. Therefore, the UX-841 tubes are much easier to drive as class "B" audio amplifiers than the UX-210 tubes although the power output is essentially the same in the two cases.

The load curves in Fig. 8 for the UX-112A indicate what may be expected of two of these tubes as class "B" audio amplifiers. The curves indicate that the output power for the 2000-ohm load is about

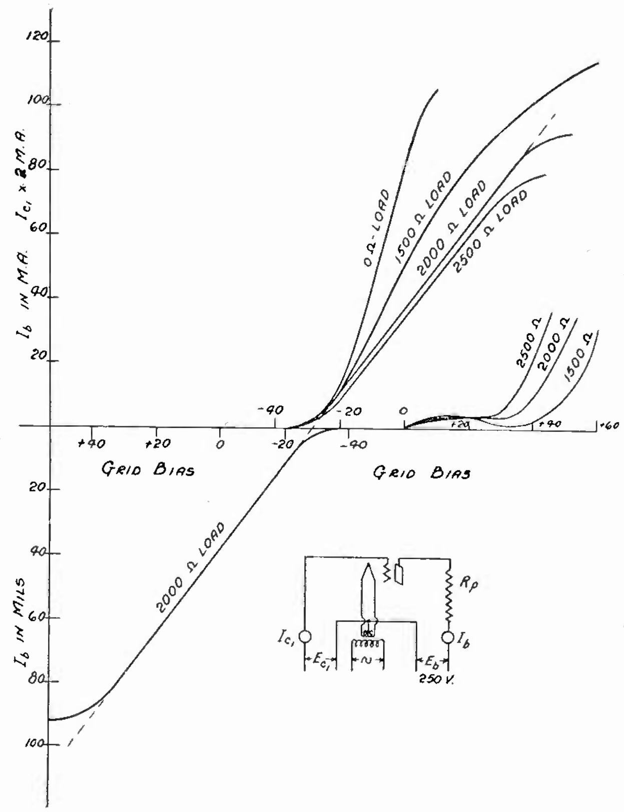

Fig. 8—Load characteristics of the UX-112A as a class "B" audio amplifier.

6 watts. If the plate voltage is raised to 300 volts, the optimum load resistance is about 2500 ohms and an output of 10 watts may be expected. The bias for 250 volts on the plates is about 28 volts and the positive grid swing is approximately 35 volts. The same precautions must be taken in driving these tubes as are taken with the UX-210 because of the flow of grid current on only a part of the input cycle. The grid current curves indicate a slight negative resistance at some points, but the resistance is so high that no difficulty is encountered.

One UY-227 operated at maximum rated voltages will drive the UX-112A tubes as class "B" audio amplifier without serious distortion although it is preferable to use two UY-227 tubes either in push-pull or parallel.

From a standpoint of high audio power from small tubes as a class "B" audio amplifier, the RCA-230 is perhaps the most interesting, especially as adapted to battery operated radio receivers. A study of the curves of Fig. 9 indicates that an output of approximately one watt can be obtained from these tubes with a maximum average plate

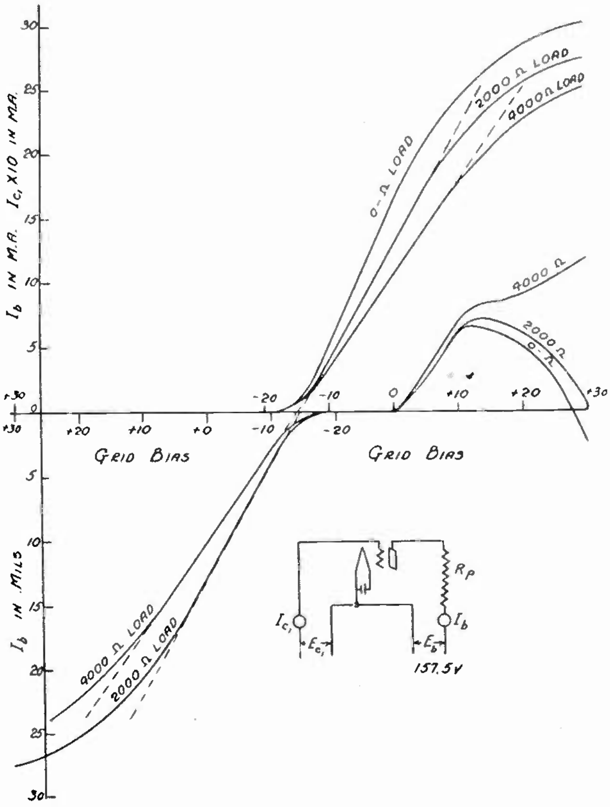

Fig. 9—Load characteristics of the RCA 230 as a class "B" audio amplifier.

battery drain of 14 ma at 157.5 volts, the other 22-volt block of a 180-volt battery being used for bias. The above plate current is the value for maximum output of a sine-wave signal. However, because of pauses, distorted musical waves, and extreme variation in amplitudes of radio signals, the average battery drain to the output tubes is probably 6 to 8 ma for a radio signal at full volume. A well designed battery operated set using two RCA-230 tubes as the output amplifier would have an average plate battery drain of 12 to 15 ma on full volume and a filament drain of approximately 0.5 ampere at two volts.

The grid current curves in Fig. 9 indicate negative resistance at certain load resistances and instantaneous grid voltages but the minimum resistance represented by the curves, positive or negative, is approximately 20,000 ohms. Consequently, an RCA-230 tube operating at 157.5 volts on the plate and 2-ma plate current, will successfully drive the output amplifier. The coupling transformer should have an impedance ratio of approximately 3 to 1 from the primary to each grid of the output tubes. The load resistance on each tube should be

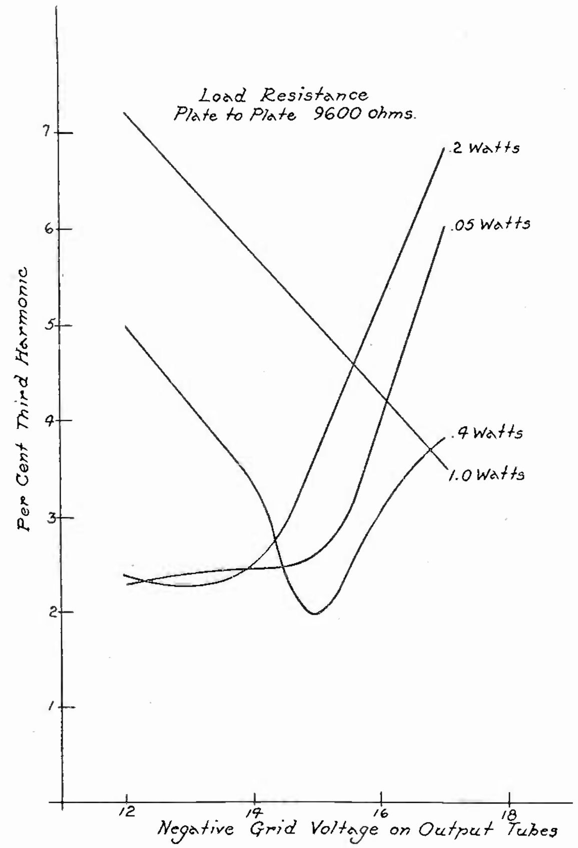

Fig. 10—Third harmonic distortion of two RCA 230 tubes as a class "B" audio amplifier for various power levels and bias voltages.

4000 ohms minimum or 16,000 ohms from plate to plate of the output tubes. Slightly more power can be obtained from the output stage by using a load resistance of 2500 ohms but at the expense of increased plate battery drain and excessive plate current peaks in the output tubes. The output tubes may be damaged by abnormally high volume if distortion is disregarded when a load resistance for maximum power output is used. If the higher load resistance of about 4000 ohms per plate is used, it is improbable that the output tubes can be driven with another RCA-230 tube to destructive plate current peaks.

Several curves of per cent harmonic distortion for various conditions of operation are given in Figs. 10 and 11 for two RCA-230 tubes as class "B" audio amplifiers and driven with an RCA-230 as indicated above. The input signal had less than 0.2 per cent second or third harmonics and the harmonics were measured by means of an RCA Victor voltage analyzer. Harmonics above the third are not plotted because they were below one per cent except in cases where the third was greater than 5 per cent. The curves in Fig. 10 indicate that the optimum bias is very nearly the bias obtained from the curves of Fig. 9. These curves also indicate that the bias may vary plus or minus one

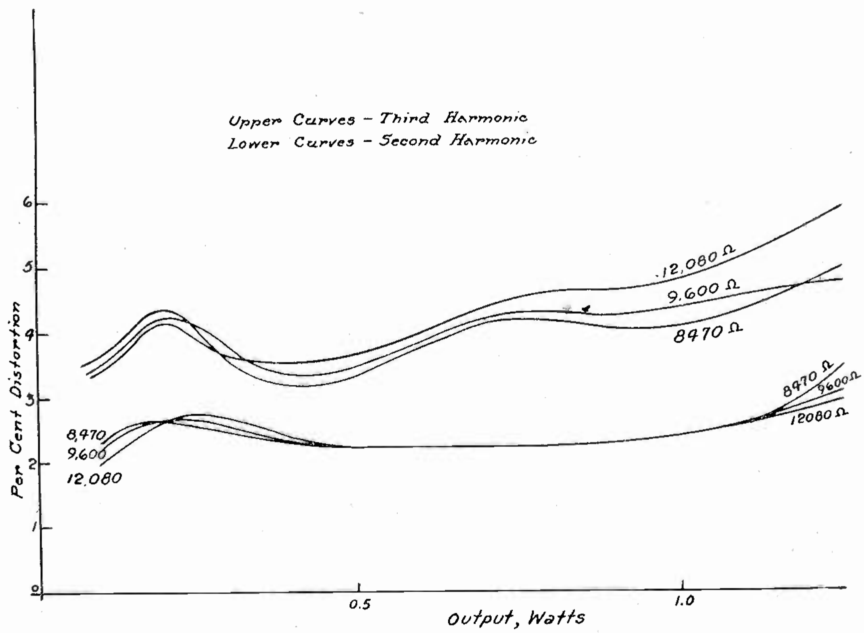

Fig. 11—Distortion of two RCA 230 tubes as a class "B" audio amplifier as a function of power output for various load resistances.

volt about the optimum value without serious distortion. The second harmonic curve corresponding to the curves in Fig. 10 are not shown but were below 3 per cent at 12 volts bias and about 2 per cent at 17 volts bias.

The above representative curves on a few of the common tubes may be used as guides in making estimates of the power output possibilities of other tubes. None of the present low plate resistance output tubes are desirable as class "B" audio amplifiers because of the required grid swing and the relatively high power required to swing the grids positive. It is of interest to know that 150 to 200 watts audio power may be obtained by using two UV-203A tubes or two UV-211 tubes with 1000 volts on the plate. It is also interesting to know that the RCA Victor Company has successfully constructed an audio amplifier of the class "B" type which delivers approximately 2500 watts. This amplifier uses four UV-849 tubes with 2800 volts on the plates and the output stage is driven by two UV-845 tubes in push-pull. The entire amplifier has not more than 4 per cent harmonic distortion and has a good frequency characteristic. The amplifier is entirely a-c operated and the plate voltage is supplied from a 4.5 kva high voltage transformer.

Conclusions

The class "A" amplifier is particularly adapted to use in cases where a single source of voltage is available for grid and plate supplies as in the case of a radio receiver or where poor regulation of the plate supply will cause difficulty. If provision is made to drive the grids of such an amplifier considerably positive greater peak plate current may be obtained for considerable increase in power output at the expense of relatively small increase in distortion. If the grids are driven positive for the peak plate current, the load resistance for maximum power output is a function of plate dissipation and in some cases, emission, but is independent of internal plate resistance of the tube.

Tests to date indicate the output voltage of a class "A" audio amplifier for a given signal input is a function of the relation of load resistance and the internal resistance of the output tubes. Consequently, if high plate resistance tubes are used compared to the load resistance, the output current is essentially constant for a wide range of load resistances for a given signal input. The output power is also nearly constant over a wide range of load resistances if the grid is driven to maximum output for the various loads.

The class "B" audio amplifier is very similar to the class "A" amplifier in regard to regulation of output and fairly constant output power for wide variation of load resistances except that this effect is more pronounced because the load resistance of the class "B" audio amplifier is much less than the optimum load resistance for the class "A" amplifier. However, the power output of the class "B" audio amplifier is usually limited only by emission or plate dissipation on peak signals. The plate current is nearly proportional to signal voltage so that the power taken from the plate supply source varies from a small value to full plate power required for maximum output. The problem of obtaining a plate supply with the necessary regulation is one which must be considered in connection with class "B" audio amplifiers unless batteries are used. As is obvious, this problem involves a rectifier supply which will carry the pulsating load with good regulation.

Another problem which has been met in developing this type of amplifier is to provide for a well regulated bias supply unless the supply source is a battery. This is difficult because of the relatively high grid current taken by the grids of the output tubes. This problem probably will be met commercially by using tubes that operate at very low bias, as indicated in Fig. 7. The bias is, at the present time, obtained preferably from a source separate from the plate supply to the amplifier and is not made automatic in the usual sense of the word. The above problems do not exist when a battery is used to supply the voltages which makes the class "B" audio output amplifier very practical for battery operated sets.

The problem of driving the amplifier can be successfully solved if some distortion is permitted. At the present stage of development, it seems that class "B" audio amplifiers inherently have a negligible amount of distortion because the power output must come from one tube at a time and the transfer of the load from one tube to the other cannot be made without a limited amount of disturbance or suitable compensation. However, as indicated above, the distortion is not serious if proper precautions are taken.

The low cost of the class "B" audio amplifier, the conservation of plate supply power for relatively high outputs and the ability of a tube to deliver 5 to 10 times its usual audio output, are factors worth considering where relatively high power is desired at a minimum cost.

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NOTES ON RADIO TRANSMISSION*

By
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Summary—*Considerable data on radio transmission have been obtained the past few years in connection with the establishment and operation of various radiotelephone services by the Bell System. It is the purpose of these notes to present certain aspects of some of these data which may be of interest in the development of a general physical picture of radio transmission and in indicating the effects of disturbances accompanying storms in the earth's magnetic field.*

The general results which are arrived at are:

1. Neglecting short time fading, the maximum field strength received at a given point for frequencies up to at least 4 megacycles are in general agreement with those calculated by the inverse-distance law and the minimum field strength (over-water transmission) are in approximate agreement with those calculated by the Austin-Cohen formula.

2. There appears to be a daylight absorption band in the neighborhood of 40 kilocycles (North Atlantic transmission) which reduces minimum daytime fields in that vicinity below the minimum limit given above.

3. The effect of solar disturbances is to increase the absorption to "sky wave" transmission throughout the entire radio-frequency spectrum generally and to reduce or eliminate the 40-kilocycle absorption band thereby increasing daylight fields for transmission on frequencies in that vicinity.

1. General Picture

Short-Wave Transmission (70 Meters to 17 Meters; 4.5 Megacycles to 17 Megacycles)

THE various daytime observations made at Forked River, New Jersey, on 4-megacycle transmission from the S. S. Leviathan at various distances from Ambrose Channel Lightship are shown in Fig. 1. The fields in general decrease with increasing distance, and at a distance approaching 300 nautical miles the spread of the values increases with the minimum fields following the original trend. The values for mighttime transmission are shown in a similar way in Fig. 2. Values were considered to be nighttime if all or any part of the transmission path were in darkness, and therefore include some points which really come in the transition period.

Preliminary approximations had indicated that the trend of the daytime values followed roughly those calculated by the Austin-Cohen formula and that the maximum nighttime values followed roughly those expressed by the inverse-distance law. Furthermore, the maxi-

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mum daytime values approached the inverse-distance values and the minimum nighttime values approached the Austin-Cohen values.

As data on transmission from other ships and with other frequencies appeared to bear out the above assumptions, the curves representing the inverse-distance and Austin-Cohen limits were fitted to the data of

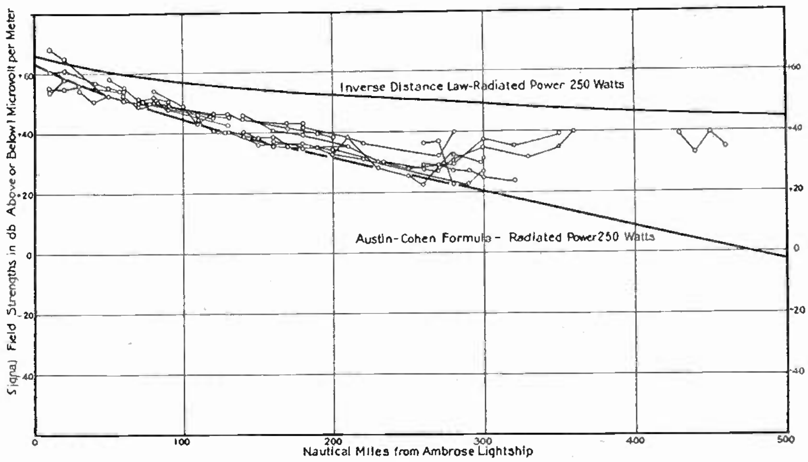

Fig. 1—Variation of signal field strengths with distance. Transmission from S. S. Leviathan. Measured at Forked River, N. J. Day values only. October to December, 1929, and April to August, 1930. Frequency 4 megacycles.

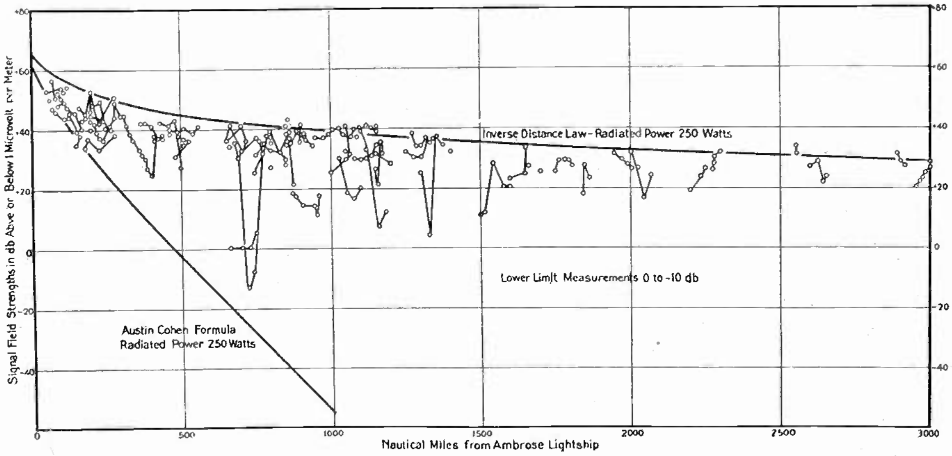

Fig. 2—Variation of signal field strengths with distance. Transmission from S. S. Leviathan. Measured at Fork River, N. J. Night values only. October to December, 1929, and April to August, 1930. Frequency 4 megacycles.

Figs. 1 and 2. The radiated power was then computed. The value of 250 watts as given is not out of line with that estimated from transmitter power and antennas.

It should be noted that the transmission to Forked River from points less than 50 to 100 miles from Ambrose involves transmission partly over land so that the lower fields observed from these points may be neglected for the general study.

The inverse-distance equation for at least the distance involved in these notes may be given as

$$E = \frac{120\pi HI}{\lambda D}$$

which when combined with the equation for radiated power

$$P_{\text{watts}} = \frac{1580 H^2 I^2}{\lambda^2}$$

becomes

$$E_{\mu v/m} = \sqrt{P_{kw}}\,\frac{3 \cdot 10^5}{D_{km}}\,.$$

The fact that the maximum fields noted are not materially greater than the inverse-distance values might be explained on the basis that these values obtain at times of "no loss" in the transmission medium, assuming spherical expansion of the wave energy.

The Austin-Cohen factor $e^{-0.0015D/\sqrt{\lambda}}$, by which the inverse-distance values are multiplied to represent the minimum values, will be remembered to be an empirical one based on measurements made on transmission up to 4000 kilometers on wavelengths from 1000 to 4000 meters.

Although the upper limit of field strength seems to be that postulated by the inverse-distance law and the minimum fields appear to be those calculated by the Austin-Cohen formula, the actual fields received at any particular time depend, of course, on the state of the transmitting medium at that moment. The transmitting properties of the medium change from hour to hour due to varying exposures to sunlight and related phenomena resulting in diurnal characteristics of radio transmission which vary with the frequency of transmission. Superimposed upon this diurnal variation are such effects as accompany solar disturbances.

For a given frequency, the envelope of maximum values of a mass plot of numerous measurements made at various times of day and at various seasons is a curve dropping off approximately inversely with distance. This is true up to the maximum distances covered by the measurements, that is, up to 3000 to 5000 miles. The envelope departs from the inverse-distance curve at the relatively short distances so that apparently the minimum distances at which inverse-distance values may be obtained appear to be roughly as follows:

Frequency	Minimum Distance for Inverse-Distance Values to Obtain
4 mc	100 Nautical Miles
8 "	400 " "
13 "	600 " "
17 "	700 " "

No distinction between seasonal variations is made in the above figures nor between day and night values.

The variation of maximum *daylight* field strengths with distance as measured during a *particular* season (summer, 1930) in connection with

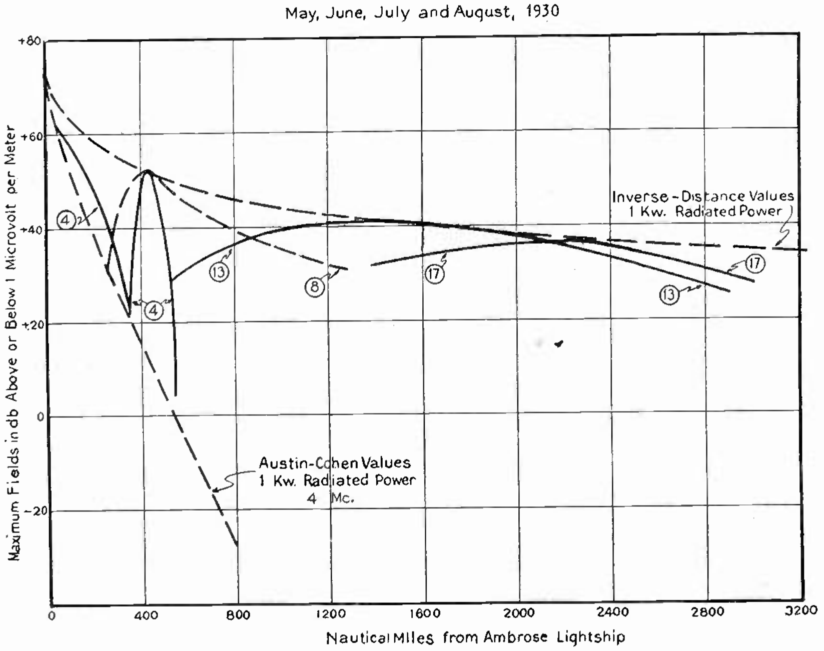

Fig. 3—Variation in maximum daylight field strengths with distance. Corrected to 1-kw radiated power. Figures denote approximate frequency in megacycles. Fields from S. S. Homeric, Leviathan, Majestic, and Olympic. Measured at Forked River, N. J. May, June, July, and August, 1930.

the ship-shore radiotelephone service is shown in Fig. 3. The appearance of the 4-megacycle "sky wave" is quite evident. It is of interest to note that the distance for maximum 4-megacycle field strengths is the same as for 8 megacycles. The distances corresponding to the maximum fields for the various frequencies are approximately as follows:

Frequency	Distance for Maximum Fields
4 mc	450 Nautical Miles
8 "	450 " "
13 "	1500 " "
17 "	2300 " "

It should be kept in mind that the values of signal field strengths expressed by the Austin-Cohen formula apply only to over-water transmission. Although a general study on the over-land attenuation at the higher frequencies has not been made, we do have some information on transmission on 4 megacycles which is the frequency used in ship-shore telephone service for near-by transmission. In a recent paper on the ship-shore radio telephone system[1] it was stated that "moving either the transmitting or receiving terminal a mile back from the coast line increases the attenuation some 8db at 4.5 megacycles." The investigation† referred to also showed that the attenuation over wet marsh could be assumed to be equivalent to that of sea water—at least for the short distances measured. Fig. 4 shows the results of more

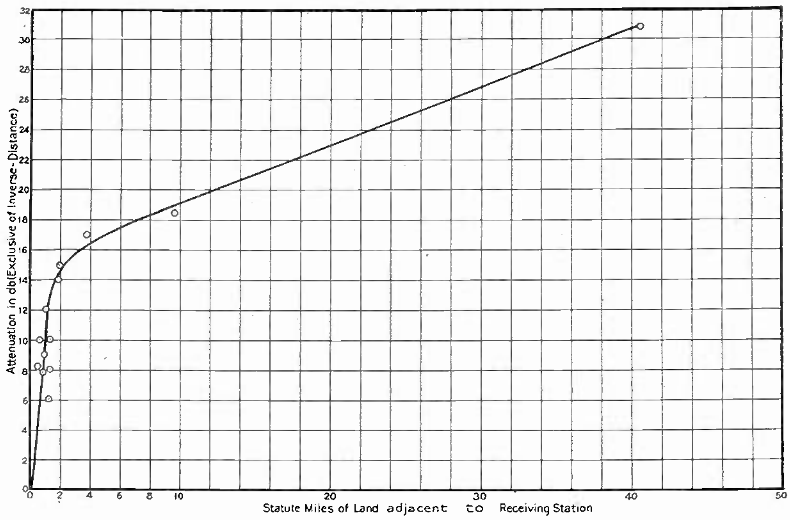

Fig. 4—Over-land transmission on 4 mc from ships at sea. Measurements made at Forked River and vicinity, 1930.

recent tests on transmissions from the ships with the field strengths measured at various distances inland from the New Jersey coast. It will be noted that the loss is of the order of 9 db for the first mile, 14 db for the first two miles, and 17 db for the first five miles after which the attenuation increases at the rate of close to 0.38 db per mile. This attenuation of 0.38 db per mile is about five times the attenuation in db for sea water at this frequency.

800-Kilocycle (380-Meter) Transmission

Fig. 5[2] shows the results of measurements made a number of years ago by H. T. Friis on transmission on 800,000 cycles which fall very nicely between the two limits inverse-distance and Austin-Cohen values.

† Discussed in paper being prepared for publication by R. A. Heising.
[1] W. Wilson and L. Espenschied, "Radio telephone service to ships," *Jour. A.I.E.E.*, 1930.
[2] H. W. Nichols and L. Espenschied, "Radio extension of the telephone system to ships at sea," Proc. I.R.E., **11**, June, 1923.

Long-Wave Transmission (5000 to 17,000 Meters; 60 Kilocycles to 17 Kilocycles)

In a previous analysis of long-wave transmission data[3] an attempt was made to obtain idealized transmission curves on the various frequencies. For the night fields the maximum values were used and found to approach the inverse-distance values as a limit. For the daytime, however, average values were used. In view of the results on the higher

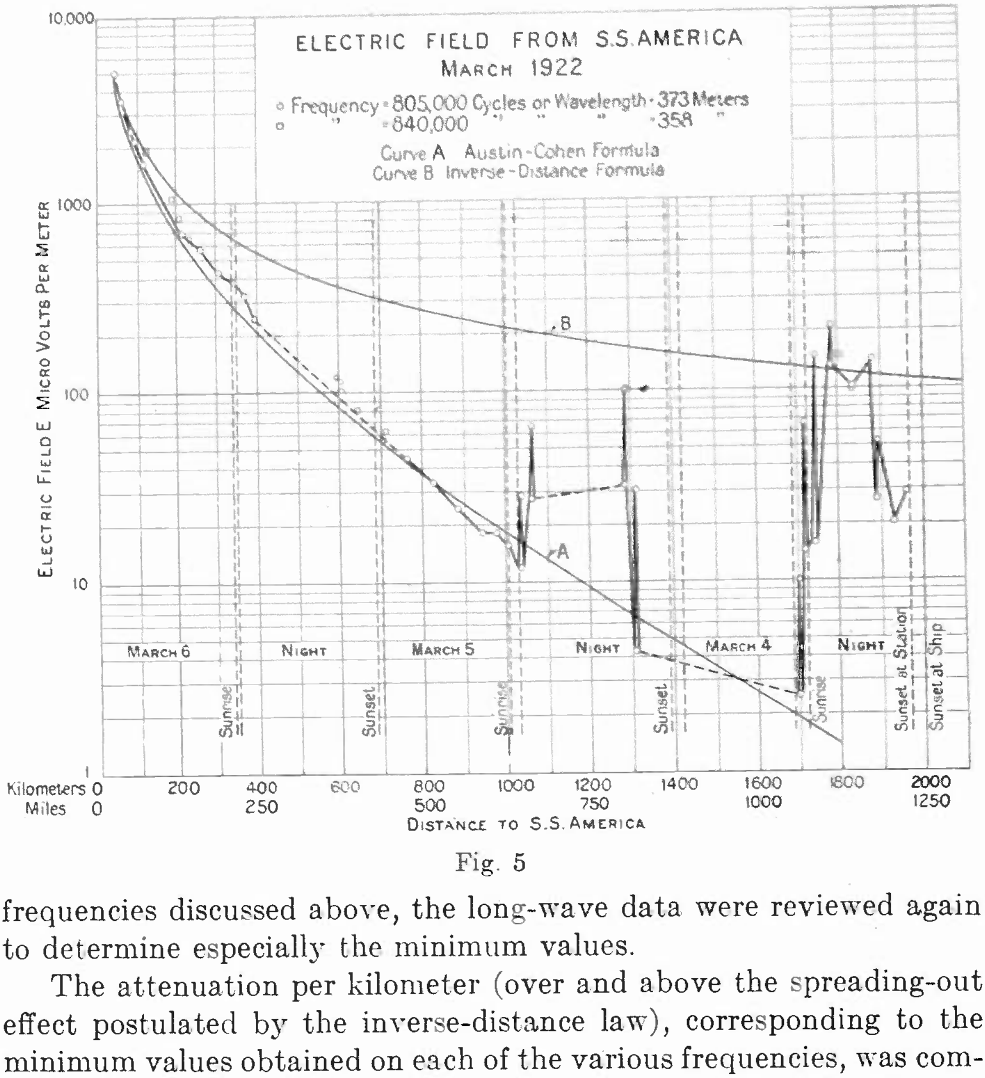

Fig. 5

frequencies discussed above, the long-wave data were reviewed again to determine especially the minimum values.

The attenuation per kilometer (over and above the spreading-out effect postulated by the inverse-distance law), corresponding to the minimum values obtained on each of the various frequencies, was computed from the data. These, together with the curve representing the attenuation in db per mile computed from the Austin-Cohen factor $e^{-0.0015D/\sqrt{\lambda}}$ are shown in Fig. 6. The basis for each of the points is given below:

[3] C. N. Anderson, "Correlation of transatlantic radio transmission with other factors affected by solar activity," Proc. I.R.E., 16, March, 1928.

Point "a" represents the 4-megacycle transmission as discussed above in connection with Figs. 1 and 2.

Point "b" represents the 800-kilocycle transmission discussed in connection with Fig. 5.

Point "c" represents the value of attenuation obtained from the 57–60-kilocycle transmission from Rocky Point, Long Island, to New Southgate, England, and Cupar, Scotland, the past eight years. This represents considerable data and it is well established that the attenuation per kilometer for minimum values is above that given by the Austin-Cohen factor.

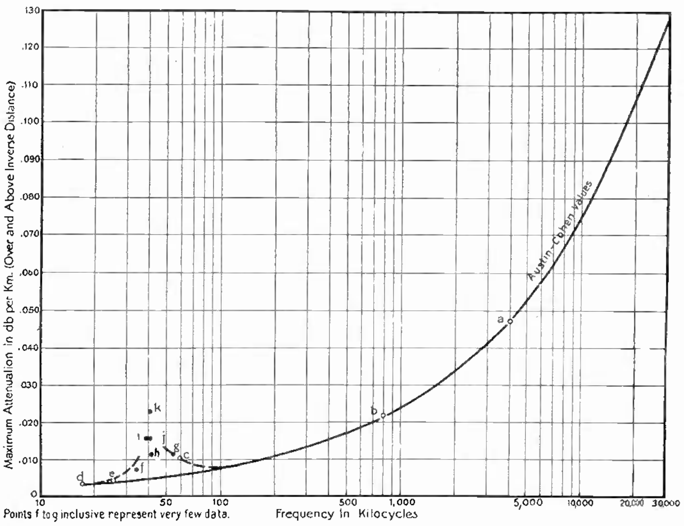

Fig. 6—Variation in maximum attenuation with frequency. Solid line represents Austin-Cohen factor. Dotted line represents 40-kc daylight absorption band.

Point "d" represents the value for transmission on 17,130 cycles from Rocky Point, Long Island, to New Southgate, England, 1923 to 1925. The point is reasonably reliable. The attenuation per kilometer for minimum fields is only very slightly greater than that given by the Austin-Cohen factor.

Point "e" represents the value for transmission on 25,700 cycles from Marion, Massachusetts, to New Southgate, England, 1923 to 1925 and is probably equally reliable with point *"d"*.

Considering these five points, only point *"c"* lies definitely off the Austin-Cohen curve. This led to a consideration of other data between 25 kilocycles and 60 kilocycles. Some measurements showing unexpectedly low fields at 40 kilocycles (Fig. 7) had been previously reported on.[3] These data, together with other miscellaneous data, are represented by the following points:

[3] *Loc. cit.*

Point "f" represents very scattered data obtained at Green Harbor, Massachusetts, in 1923, on transmission from GBL (Leafield, England) on 34,130 cycles. About as much as can be said for this point is that the attenuation is at least as great as indicated.

Point "g" represents measurements on transmission from GKB (Northolt, England) on 54,500 cycles during the latter part of 1923 at Riverhead, Long Island. Although there is considerable data during the period, the minimum values are approximately at the lower measurement limit and hence apt to be unreliable although attenuation is at least as great as that indicated.

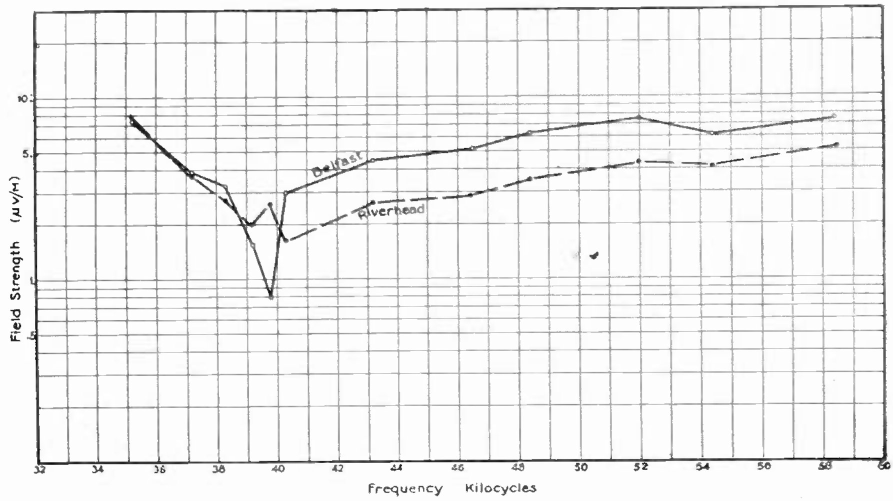

Fig. 7—Variation of signal field with frequency. Transmission from Northolt, Eng. (GKB) Daylight signal fields corrected to 2-kw radiated power.

Points "h" and "i" are determined from a relatively small number of measurements at Belfast, Maine, of GKB (Northolt, England) on 43 kilocycles and 39 kilocycles, respectively. Because of small amount of data, points are not reliable.

Points "j" and "k" are results of measurements made July 25 and March 25, 1925, respectively at Chedzoy, England, on transmission from GKB (Northolt) on 43 kilocycles. The distance between the Chedzoy and Northolt is only 135 kilometers, but as data for each point cover only a period of a day, they are probably unreliable.

It will be noted that, although practically all the latter points are in themselves unreliable, collectively they tend to indicate abnormal transmission in the neighborhood of 40 kilocycles. (The error of the different points is likely to be in the direction of giving too little attenuation.)

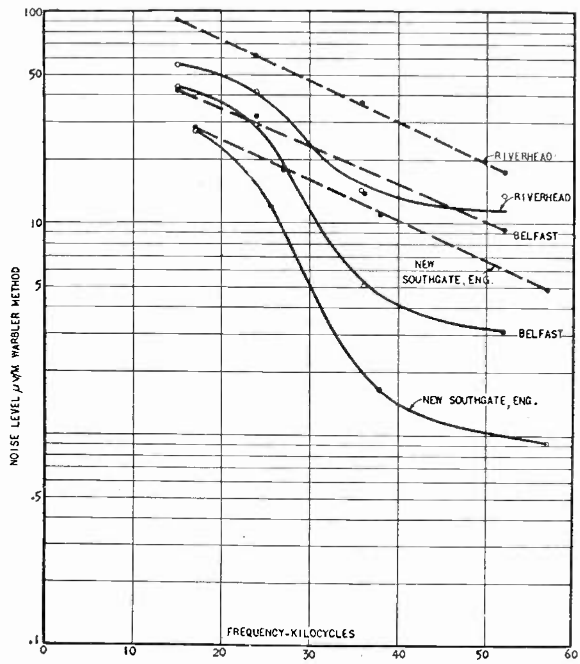

Fig. 8—Frequency distribution of noise. New Southgate, Eng., Belfast, Maine, Riverhead, L. I. Nighttime – – – Daytime ———. August to December, 1924.

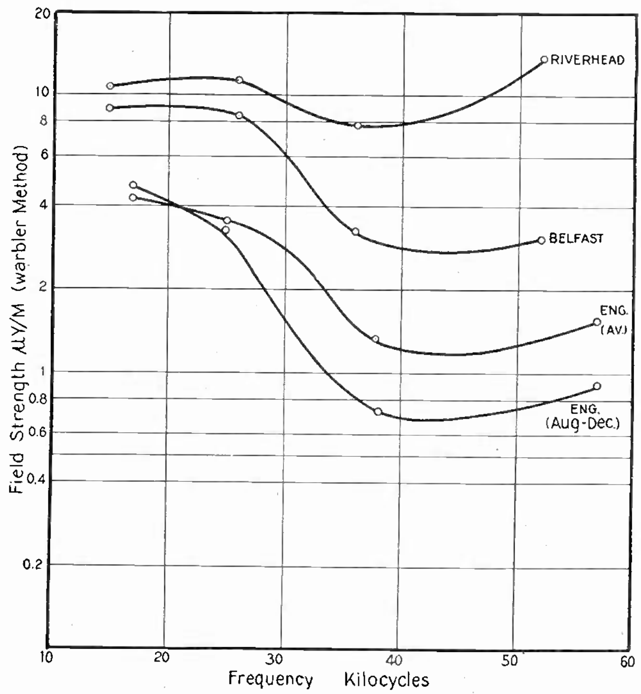

Fig. 9—Frequency distribution of noise. Corrected to equal powers radiated assuming nighttime distribution to be proportional to $\sqrt{P}$ radiated.

It is of interest to note what the effect would be if this absorption band did not exist. The 17-kilocycle fields would increase only very slightly, the 25.7-kilocycle fields would increase slightly more, the 57–60-kilocycle fields would increase considerably more and the 40–43-kilocycle fields would increase most of all. For the 57-kilocycle transmission, the minimum fields such as occur during the "sunset-dip" would be of the same order as the normal average of daylight fields. This is exactly what happens during periods of solar disturbances.

It may be of interest to note that this absorption band also appears to affect static noise. It is reasonable to expect this in view of our picture of static on long waves being generated largely in the tropics and hence being received over a relatively long transmission path. Fig. 8[4] represents some data on the frequency distribution of noise for daytime and nighttime. Fig. 9 shows roughly the differential, as it were, between nighttime and daytime noise as derived from Fig. 8 and indicates greater attenuation in the vicinity of 40 kilocycles than at frequencies 20 kilocycles on either side.

It might be pointed out that evidence indicating an absorption band in the vicinity of 1500 kilocycles, for transmission to transatlantic distances,[5] pertains to the higher "sky wave" fields. No data are available bearing upon the question of whether this absorption band also exists in the "ground wave" transmission at this frequency. This explains why no indication is made of the 1500-kilocycle absorption band in Fig. 6.

2. Effect of Solar Disturbances

Long-Wave Radio Transmission

Some of the early data indicating the effect of solar disturbances on long-wave radio transmission are shown in Fig. 10[4]. A generalized picture of the effect upon transmission during the sunset period is given in Fig. 11[3]. The results shown in both of these figures vary with the intensity of the disturbance and the time elapsed after the occurrence of the disturbance.

The difference in the magnitude of the effect accompanying a disturbance in the earth's magnetic field upon the 43-kilocycle daylight signal fields as compared with transmission on 57 kilocycles is shown in Fig. 12. The two diurnal curves for each frequency show the field

[3] *Loc. cit.*

[4] L. Espenschied, C. N. Anderson, and A. Bailey, "Transatlantic radio telephone transmission," Proc. I.R.E., January, 1926; *Bell Sys. Tech. Jour.*, July, 1925.

[5] H. W. Nichols and J. C. Schelleng, "Propagation of electric waves over the earth," *Bell Sys. Tech. Jour.*, April, 1925.

strengths the week-end before and the week-end after the disturbance. It is to be noted that whereas the increase of the daylight field strengths on 57 kilocycles after the disturbance was only of the order of 25 per cent, the increase on 43 kilocycles was of the order of 100 per cent (400 per cent considering minimum fields).

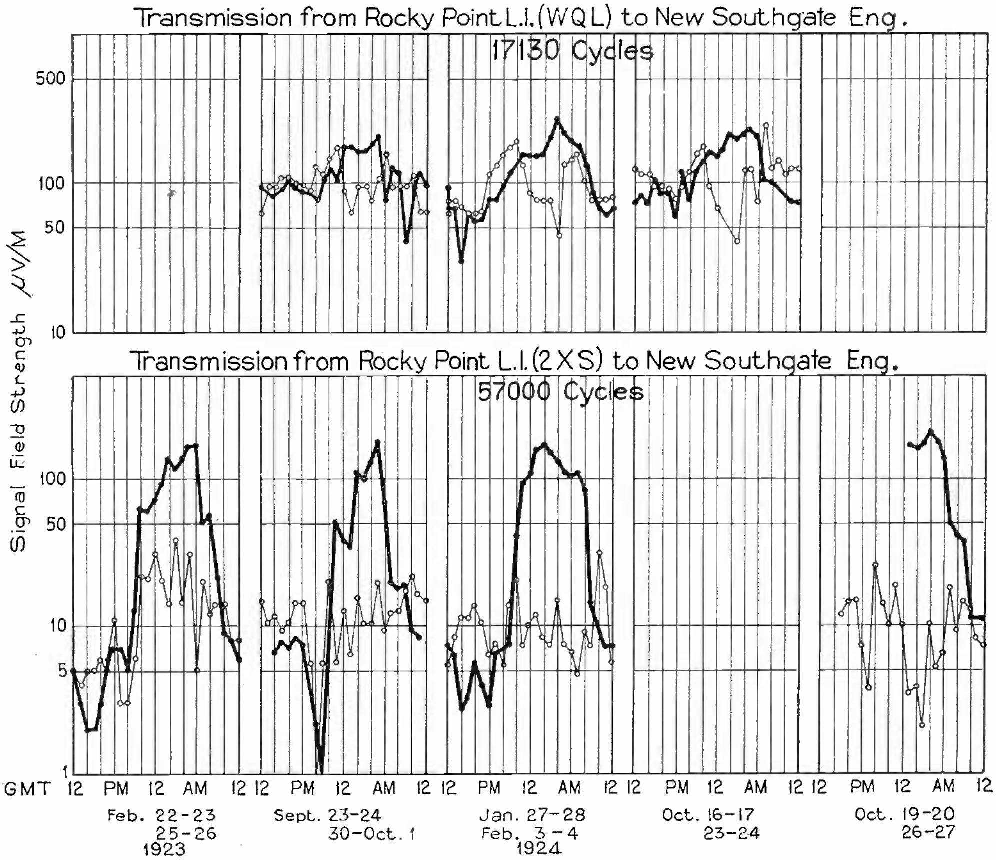

Fig. 10—Effect of solar disturbances on radio transmission. Heavy line indicates more or less normal transmission week-end before disturbance occurred. Abnormal transmission on following week-end indicated by light line.

From Figs. 10, 11, and 12 several effects of solar disturbances are evident:

(a) Little or no increase in daylight signal fields on 17-kilocycle transmission; slight increase in 25-kilocycle fields, considerable increase in 57-kilocycle fields, and still greater increase in the 43-kilocycle field strengths.

(b) Reduction (or elimination depending upon intensity of disturbance) of "sunset-dip."

(c) Decreased night field strengths to values sometimes as low as or less than average daylight values. Minimum night fields are either greater than or of same order as the Austin-Cohen values.

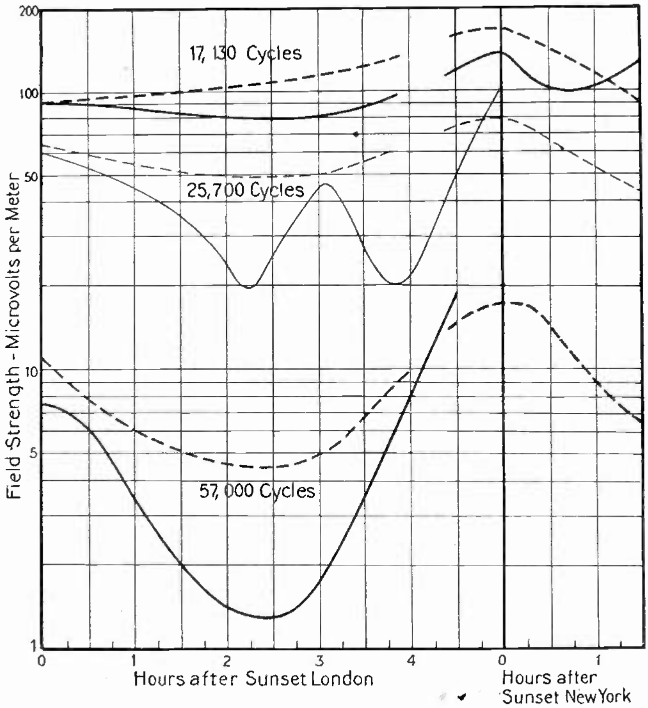

Fig. 11—Major sunset effects on radio transmission. Transmission from Rocky Point, L. I., (57 kc and 17.1 kc) and Marion, Mass., (25.7 kc) to New Southgate, England. Solid line indicates undisturbed and dashed line disturbed conditions.

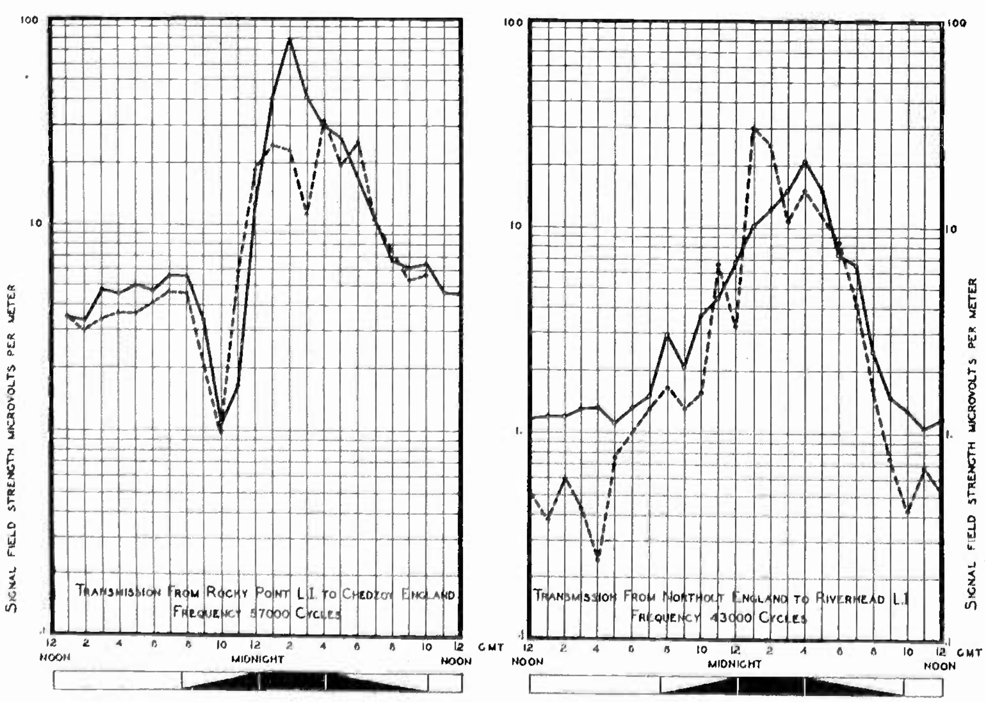

Fig. 12—Effect of solar disturbance on radio transmission. Diurnal variation of signal field strength on 57,000 cycles and 43,000 cycles showing greater increase of daylight signal field strengths on 43,000 cycles than on 57,000 cycles. Dotted line indicates before the disturbance. Solid line indicates after the disturbance. Date of disturbance, May 5, 1925.

These results indicate the reduction or elimination of the 40-kilocycle daylight absorption band. The decreased values obtained at nighttime (presumably "sky wave" transmission) are similar to the low fields experienced on short waves (also "sky wave" transmission) during such disturbed periods.

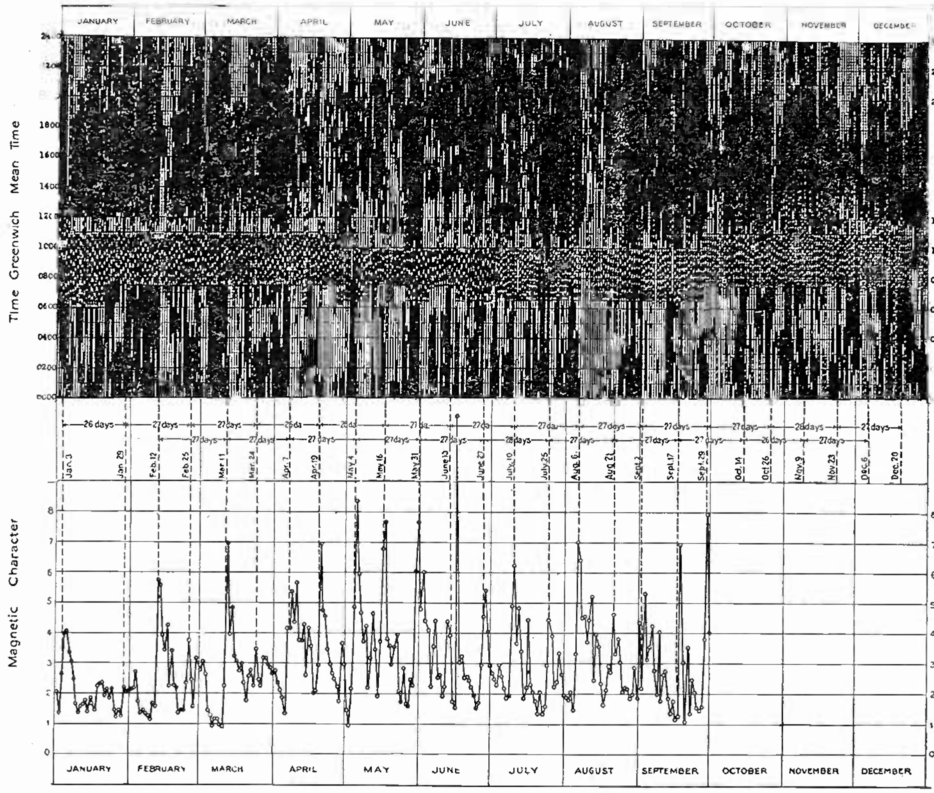

Fig. 13—Transmission on New York–London short-wave circuits, 1930. Solid portion indicates time during which one or more two-way short-wave radiotelephone circuits were serviceable. Dashed portion indicates no data and blank portion indicates time when no circuit was serviceable.

Short-Wave Transmission

The low fields experienced in short-wave transmission during solar disturbances are apparently due to increases in attenuation rather than a change in the optimum distance for that frequency.[6] As the high night fields on long waves decrease during such storms to values equal to or approaching those represented by the Austin-Cohen factor, a similar situation may obtain on short waves. As the Austin-Cohen values for short waves at distances beyond a few hundred miles are so far below any possibility of measurement, it is doubtful whether this

[6] C. N. Anderson, "Notes on effect of solar disturbances on transatlantic radio transmission," Proc. I.R.E., 17, September, 1929.

point will ever be definitely established. It is known that even with storms not exceptionally severe it has been impossible even to hear a heterodyne beat note of the transatlantic signal using effectively large transmitting powers, large receiving arrays and high gain receivers which means that the fields are less than the order of a few thousandths of a microvolt per meter.

The effect of solar disturbances on the utility of short-wave radio circuits is shown in Fig. 13 which indicates the time during which at least one out of the three short-wave circuits between the United States and England gave serviceable transmission, when no circuit gave

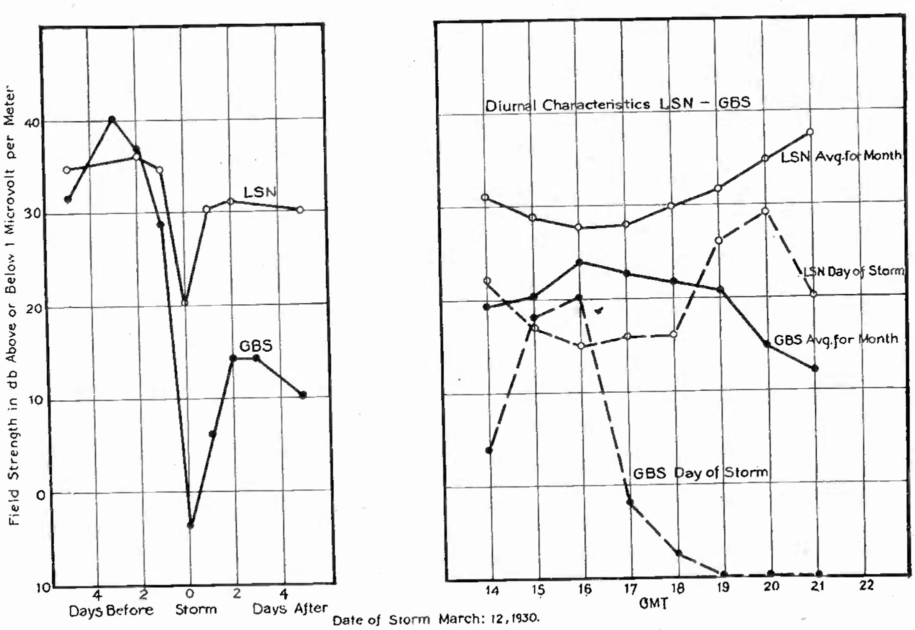

Fig. 14—Comparison of the variation in field strengths before and after a solar disturbance. Transmission from Rugby, England, to Netcong, N. J., and from Buenos Aires, Argentina, to Netcong, N. J. (GBS) Frequency 18.34 megacycles (16 m) Frequency 21.02 megacycles (14 m) (LSN).

serviceable transmission, and the no data periods when the circuits were not in operation. It is of interest to note the different series of disturbances separated by the 27-day period corresponding to the time of rotation of the sun as viewed from the earth. Such daily magnetic character figures as are available[3] are shown below.

The difference in the effect of solar disturbances upon transmission from the United States to England as compared with transmission from the United States to Argentina has been very marked. The difference in both the immediate effect and in the recovery is shown in Fig. 14. Without any doubt the effect upon transmission across the

[3] *Loc. cit.*

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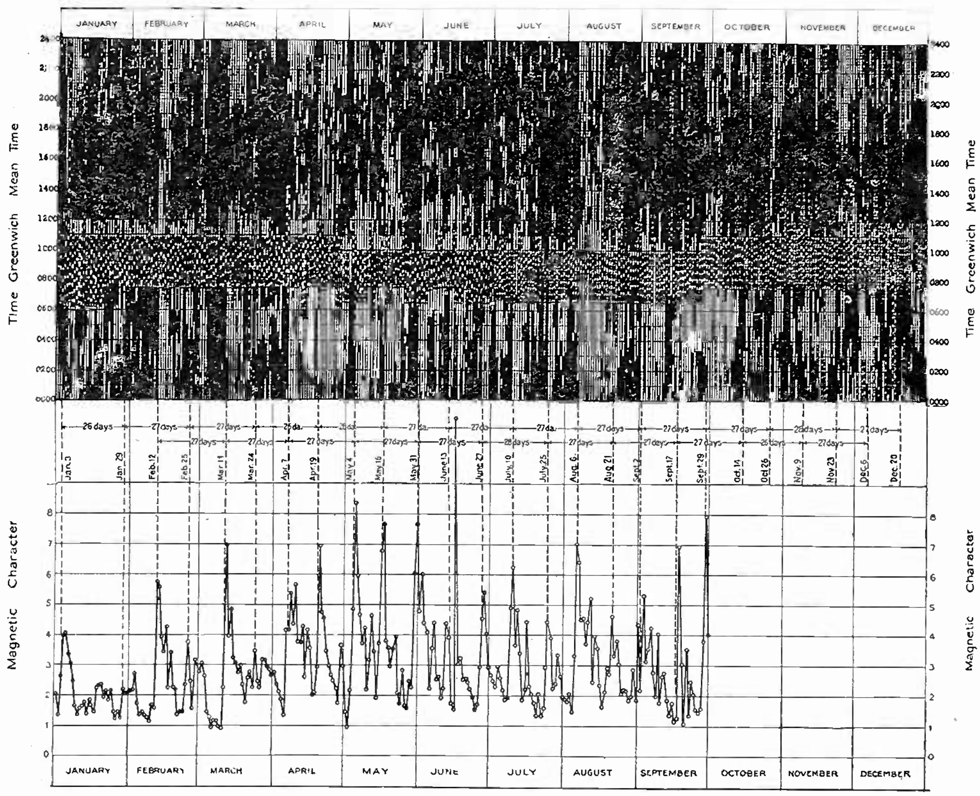

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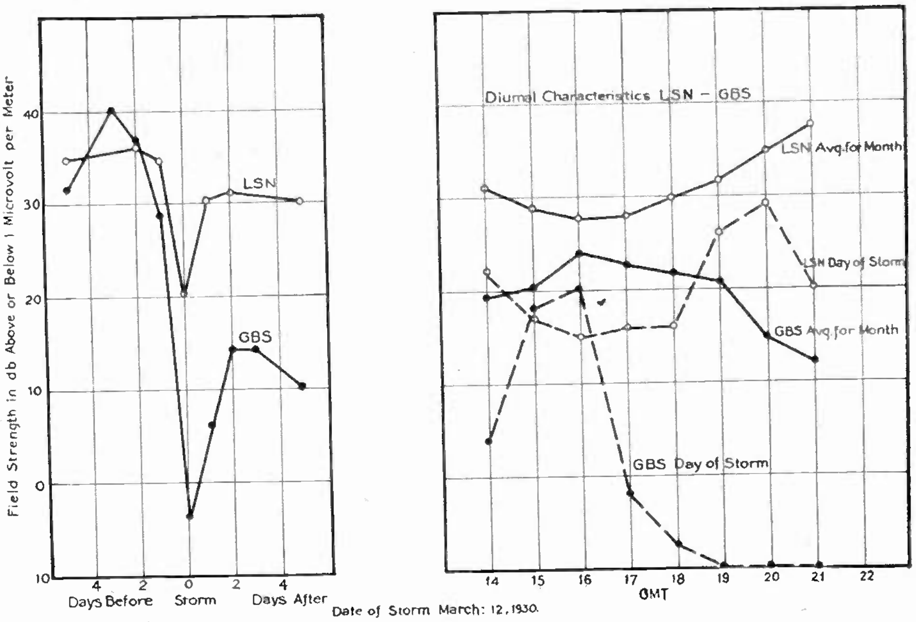

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[3] *Loc. cit.*

North Atlantic is by far the more serious. The difference in exposure of the two transmission paths to the effects of solar disturbances is suggested by Fig. 15 which shows the relation of the transmission paths to the distribution of aurora. It is evident that whereas the New York to London path passes close to the region of maximum auroral frequency, the New York-Buenos Aires path is largely out of the region of visible aurora.

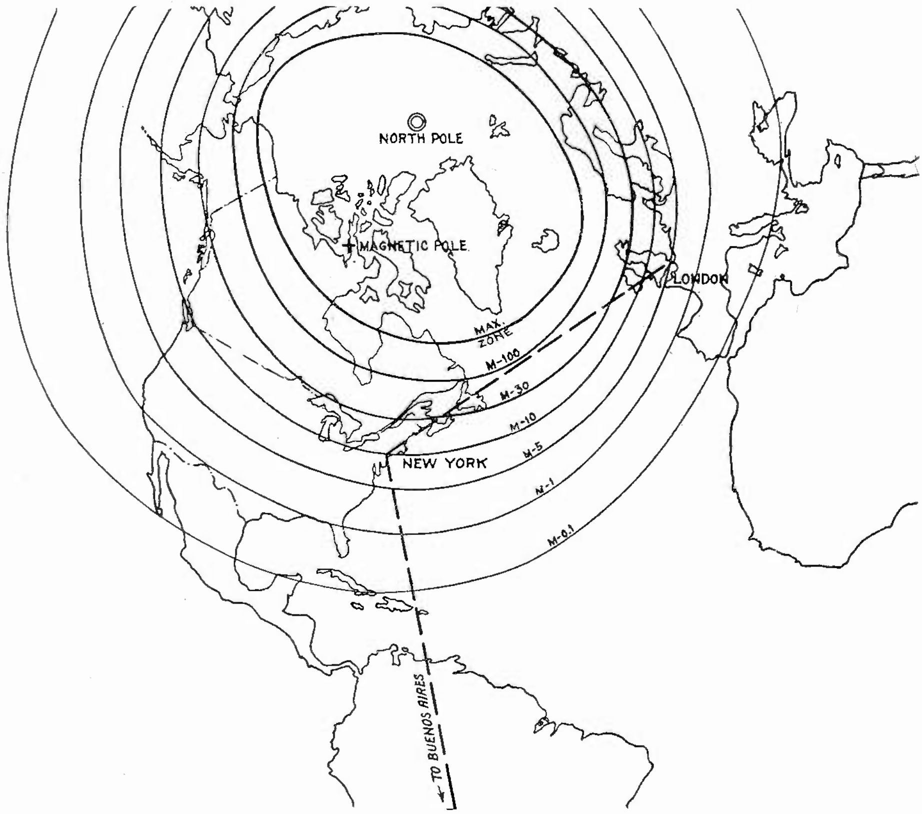

Fig. 15—Radio transmission and solar disturbances. Isochasms or lines of equal auroral—frequency (after H. Fritz). Relation of isochasms to radiotelephone transmission paths.

A comparison of how radio transmission varies from year to year is given in Fig. 16. In the lower part of the figure is given also the variation in magnetic activity and sun spots. It is evident that the correlation between these various phenomena is not a close one. Witness for example, the low sun spot numbers in March, April, and May, 1930, and at the same time the high magnetic activity (highest since the beginning of 1923), the relatively high long-wave fields and the low short-wave fields. Radio transmission and magnetic activity seem to vary together in this particular case but in other cases again do not seem to be so closely related. In correlating radio transmission with activity in the earth's field it should be kept in mind that the former is more sensitive to solar emission and requires a longer period of recovery than the latter. Similarly in correlating radio transmission and magnetic activity with sun spot numbers better results would be ob-

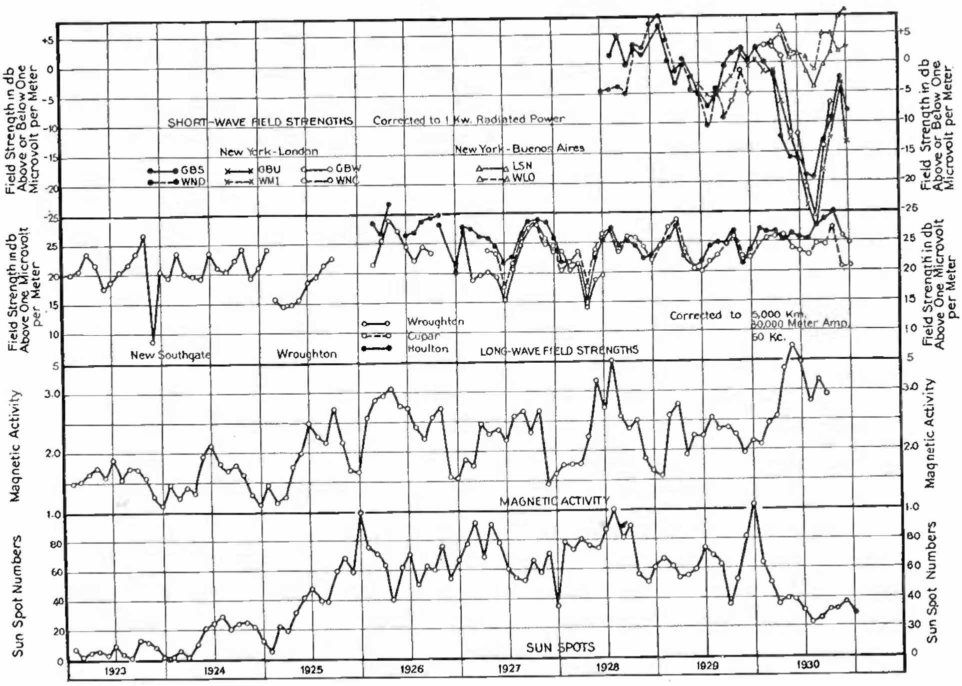

Fig. 16—Radio transmission and solar activity.

tained if differentiation were made between sun spots whose emission may affect the earth and those which are so far from the center of the sun's disk as to have apparently little or no effect upon terrestrial phenomena.

Proceedings of the Institute of Radio Engineers
Volume 19, Number 7 *July, 1931*

A NOTE ON THE RELATION OF METEOR SHOWERS AND RADIO RECEPTION*

By

Greenleaf W. Pickard

(RCA Victor Company of Massachusetts, Boston, Mass.)

Summary—*A comparison of reception measures with meteor showers over five-year periods indicates an accompanying increase of night fields and a decrease of day fields. The highest correlation appeared between monthly means of meteor hour rates and night fields. No associated disturbance of terrestrial magnetism was found.*

IT WAS first suggested by Nagaoka that meteors may disturb the Kennelly-Heaviside Layer and so affect radio transmission. The mechanism he postulates is set forth in the following quotation from his recent paper[1]:

> "The great commotion caused by the meteors in the ionized layer calls forth vortices in it, and tends to disperse the dusts about the paths. The speed of passage is generally sufficient to cause ionization in the air, and as the atmosphere is already highly ionized, the fine particles form nuclei for collecting the ions from the surrounding. The condition of the ionized layer is thus temporarily disturbed, and apparently introduces discontinuity in the transmission of the radio waves which impinge on the layer. *The irregular and diffuse reflections from the ionized layer are the general result.* The effect is somewhat analogous to the reflection of light from a mirror with numerous scratches made in a disorderly confused manner. If the paths of meteors be straight and horizontal, the effect would not be complicated, but as it is inclined at more than 20 degrees and actually to a certain degree curved, the diffraction caused by long discontinuity and the scattering of waves by dust particles generally tend to weaken the intensity of radio waves, and especially to disturb the directional transmission."

The writer has been interested in the possibility of meteoric effects since early in 1910 when the earth passed through or near the tail of the Halley Comet. At that time a series of audibility meter measurements were made at Amesbury, Massachusetts, on Station DF at Manhattan Beach, New York, but without finding any change which could be attributed to the comet. As a matter of fact, the earth's approach to this comet was not then accompanied by any meteoric shower, and in any event audibility meter readings on an intermittently operated spark station would hardly have been competent to

* Decimal classification: R113.5 Original manuscript received by the Institute April 27, 1931. Presented before the U.R.S.I. in Washington, D. C., May 1, 1931.

[1] Hantaro Nagaoka, "Possibility of the radio transmission being disturbed by meteoric showers," *Proc. Imp. Acad.*, Tokyo, 5, No. 6, 1929.

show any small effect. But in the summer of 1921, while making directional observations at Seabrook Beach, New Hampshire, on a number of low-frequency European stations[2] an interesting coincidence was observed between a sound in a radio receiving circuit and the passage of a brilliant meteor. The author has hitherto refrained from publishing this incident, partly because standing alone it had little weight, partly also for the reason that 1921 amplifiers had the habit of intermittently self-generating all sorts of weird noises, but principally because until now there has been no good setting for this solitaire observation.

At 1.11 on the morning of August 29, 1921 the writer was concluding a series of directional measurements on the French station LY at Bordeaux, then operating at 12.8 kc. The apparatus employed at Seabrook consisted of a 20-turn coil aerial, 3 meters on a side, a 4-stage resistance coupled amplifier, with external heterodyne oscillator, detector, and two audio stages. The coil aerial was set nearly on null for LY—N74 degrees E—and there was little sound in the phones, save a very light crackle of static, and a faint 800-cycle note from the Bordeaux Station. Glancing up and eastward from the degree scale of the direction finder, the author saw a bright yellow-green meteor, of approximately -2 or -3 magnitude, with an origin somewhere between Cassopeia and Perseus, traversing a path of over 30 degrees. Coincident with the passage of this meteor, a sharp hiss was heard in the phones, lasting for a little over a second, and, as nearly as could be judged beginning and ending sharply with the start and finish of the meteor trail.

Having now over five years of systematic night field measurements of WBBM at his disposal, the writer decided to examine these statistically for a possible correlation with meteor showers. While the results so far are of a preliminary character, and do not yet amount to a conclusive proof, nevertheless they strongly indicate a relation.

According to Olivier[3] there are nine principal meteor showers each year, which are tabulated below:

TABLE I

Name	Duration in Days	Date of Max.	Hour Rate of all Meteors on this date
Quadrantids	3	Jan. 2	28
Lyrids	4	Apr. 20	7
Eta Aquarids	8	May 2–4	7+
Delta Aquarids	3	Jul. 28	27
Perseids	35±	Aug. 11–12	69
Orionids	14 +	Oct. 19–23	21±
Leonids	3−	Nov. 14	21
Andromedes	2	Nov. 24	16
Geminids	14+	Dec. 11–13	23

[2] "The direction and intensity of waves from European stations," Proc. I.R.E., **10**, June, 1922.

[3] Bulletin No. 8, The American Meteor Society.

The hour rates given above are means of a number of years; the showers in any one year may diverge widely from these values. Inasmuch as the WBBM measurements are in the period 1926–1930, for which no published meteor data are yet available, the author has been forced to use the general rates of the table in dealing with night reception, instead of the actual hour rates of the individual showers in this period. In this comparison of WBBM with meteors, there was first reduced each of the nightly values to a ratio of a 27-day average to a 5-month mean. This, at the expense of a considerable reduction in

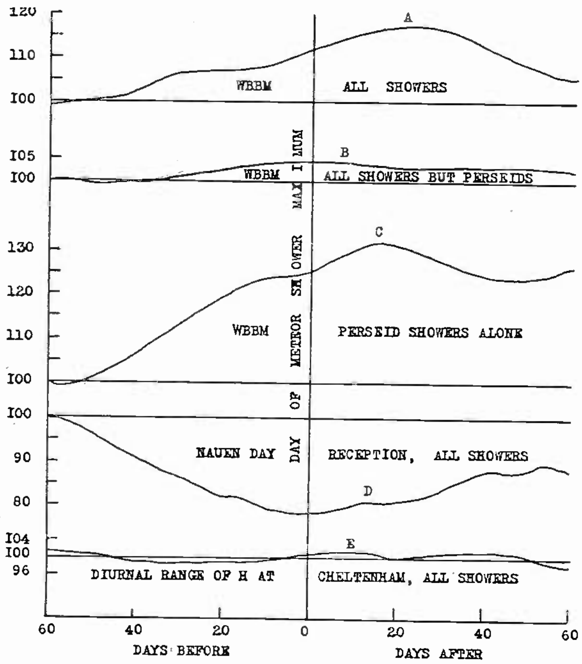

Fig. 1

amplitude, substantially eliminates the prominent solar rotation period, the lunar period of 29.5 days and also any seasonal effect. Means were then made of these nightly 27 day/5 month ratios for 60 days before and after the central shower dates, weighting the reception values according to the hour rates of each shower. The result, for five years of meteor showers and WBBM reception, is shown in Fig. 1, curve *A*. It is evident that reception reaches a maximum about 25 days after the central shower date, with an amplitude of about 17 per cent. And even taking into account the smoothing effect of the 27-day average in the reception values, it appears that the effect is not sharply defined, but is spread over some two months. As the Perseid or August shower is normally the most important, it should form a major contribution to the effect, and in curve *B* the result of taking out the five

Perseid showers is shown. The amplitude is now reduced to about 5 per cent, and its maximum now coincides approximately with the central shower date.

In curve C the effect of the Perseid showers alone is taken. As this results from only five showers, its detail is not significant, but it does indicate the dominant effect of the August showers, with a maximum after the central shower date.

Olivier has published an excellent tabulation of the principal showers up to the end of 1925, with their individual hour rates and actual maximum dates.[4] Although no systematic measurements of night reception are available for this period, Austin's day field measurements of Nauen were taken for comparison, for the six-year period

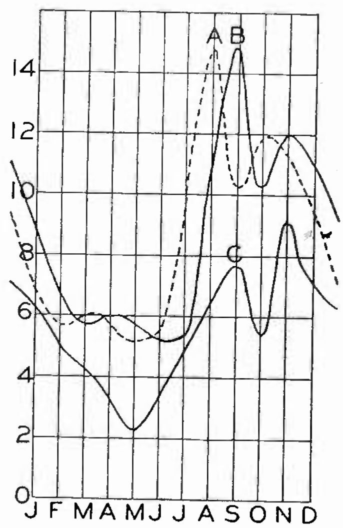

Fig. 2

1920–1925, with the result shown in curve D. Agreeably to the general rule that a disturbing cause affects night and day reception inversely, Nauen reception shows a marked depression, centering on the shower date. Finally, in curve E of Fig. 1, is shown a comparison of the diurnal range of H, as measured at Cheltenham, Maryland, with the meteor showers of 1920–1925. It would not seem from this that meteors caused any appreciable disturbance of terrestrial magnetism.

There is another comparison which can be made between meteors and radio reception, which disregards individual showers, and takes only average monthly hour rates. Olivier has given a tabulation[5] of monthly hour rates by six authorities, a mean of which there was plotted as broken-line curve A, Fig. 2. Below this in curve C are

[4] Publications of the Leander McCormick Observatory.

[5] Charles P. Olivier, "Meteors," Williams & Wilkins Co., Baltimore, Maryland, 1925, page 182.

plotted WBBM monthly means for 1926–1930. A relationship is obvious; in fact the correlation between these curves is $r=+0.67 \pm 0.11$. But it is also apparent that the curves are somewhat out of phase, reception lagging (as also indicated by *A*, Fig. 1) about a month behind meteors. In curve *B* in full line the author has advanced the meteor curve one month, which obviously improves matters, and the correlation now becomes $r=+0.87 \pm 0.05$.

The writer believes the results above establish a *prima facie* case for a relation between meteors and radio reception; certainly the subject warrants further study. It becomes increasingly clear that the vagaries of radio transmission cannot be attributed to a single cause, but merely arise out of the complex situation in which they occur. So far, marked correlations have been found with solar activity, terrestrial magnetism, lunar-hour angle, meteor showers, and minor correlations with temperature and barometric pressure.

It is perhaps too early to speculate on the mechanism by which meteor showers affect radio transmission. At least it would seem that Nagaoka's hypothesis does not fit the above facts, for if meteors merely "scratched the mirror," all radio transmission should be adversely affected, instead of which we find an increase of night fields. It seems more likely that meteors act as an ionizing agent at such a level and in such a manner as to improve night transmission and depress day fields.

The author wishes to express his indebtedness to Professors Charles P. Olivier and Willard J. Fisher for the meteor data employed, and to the U. S. Coast & Geodetic Survey for the magnetic measurements.

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THE COÖPERATION COMMITTEE PROGRAM FOR 1930—1931*

AS WAS mentioned in the report of last year, the French Government radio authorities began transmitting daily Ursigrams from the Eiffel Tower (FLE) in Paris, on December 1, 1928. This service has since been regularly maintained and extended to corresponding daily emissions from FLY Lafayette Station near Bordeaux (15.7 kc/s or $\lambda = 19.1$ km) and FYB Paris-Pontoise (10,581 kc/s or $\lambda = 28.35$ m). Both emissions are made at 20.30 G.M.T.[1] and in clear uncoded) French language.

Commencing with August 1, 1930, there has been a regular daily Ursigram emitted in code, from the U. S. Navy Radio Station NAA, Washington, D. C., under the auspices of "Science Service", Washington. The transmission is at present at 22.00 G.M.T. on frequencies 12.045 and 4.015 mc/s. This American Ursigram service has been brought into effect by the coöperative effort of a number of American scientific organizations, especially the U. S. Coast and Geodetic Survey, the Smithsonian Institution, the Carnegie Institution, the U. S. Army, the U. S. Navy, Science Service, and the American Section of the U.R.S.I. The data for the daily bulletin are forwarded to Science Service, Washington, from distant parts of the United States, largely with the aid of the Army radio net. The Ursigram is compiled by Science Service and delivered to the Navy Department, which issues it from NAA. These Ursigrams are ordinarily received in western Europe, in Canada, South America, and over the Atlantic Ocean broadly. The data are also given in mimeographed bulletins sent by Science Service to interested parties.

The French Ursigram relates to variations in:

(1) the earth's magnetic field observed at Val Joyeux near Paris;

(2) the gradient of atmospheric electric potential observed at Val Joyeux near Paris; and

(3) Solar surface activity, sun spots, faculae, etc., observed at Meudon near Paris.

The American Ursigram relates to variations in:

(1) MAG, the earth's magnetic field, as observed at Tucson, Ariz.;

* Decimal classification: R005. Original manuscript received by the Institute, May 15, 1931. Report of the Coöperation Committee, presented by the Chairman, Dr. A. E. Kennelly, at a meeting of the American Section of the International Scientific Radio Union, Washington, D. C., May 1, 1931.

[1] Greenwich Mean Time, i.e., time which begins with zero at midnight, sometimes called Greenwich Civil Time.

(2) SUN,[2] groups and number of sun spots, as observed at Mount Wilson, California.

(3) SOL,[2] the solar thermal constant, as observed at Montezuma, Chile.

The American Ursigrams are in code groups mainly of five numerals. They commence with the word *URSI* and end with the word *Scienservc.* The code is explained in a Science Service Bulletin, and also in the PROCEEDINGS I.R.E., *18*, 1469; September, 1930. They are compiled weekly by Science Service and a limited number are mailed to applicants, together with corresponding compilations of French Ursigrams for the same week. It is possible that the code material may in the near future be supplemented by plain language comments emphasizing any related events of particular interest not covered in the code message. These comments are to be considered as additions to, and in no sense substitutes for, the coded reports. The added labor involved in transmitting such information is usually not large. It may be desirable to include more complete data in regard to solar activity (such as flocculi, magnetic indexes, faculae, etc.) in the Ursigram data also, perhaps weekly.

The possibility of including data as to major earthquake disturbances should also be considered.

It has been the aim of the Liaison Committee of the *URSI* to distribute information concerning these Ursigram services to many interested persons and groups. It has similarly been the aim of the American Section Committee on Coöperation to aid in the same task.

PUBLICATION OF AMERICAN URSIGRAMS

Commencing with the December, 1930, number, *Terrestrial Magnetism and Atmospheric Electricity,* an International Quarterly Journal (conducted by Louis A. Bauer and J. A. Fleming and published under the auspices of the Johns Hopkins Press, Baltimore, Maryland, U. S. A.) has printed a quarterly table of the contents of American Ursigrams (MAG, SUN, and SOL), in decoded form, from material collected at the Department of Terrestrial Magnetism, Carnegie Institution of Washington. The first publication appears on pages 252–253 of No. 4, Vol. **35**, (December 1930) covering the quarter August 1–October 31, 1930. A preliminary announcement of the code to be employed appeared on pages 184–185 of the same Journal, in No. 3, Vol. **35** for September, 1930.

[2] It is proposed to change these designations to SOL and RAD, respectively, to avoid conflict with some proposals of the French Committee.

Proposed British Ursigram

A weekly coded Ursigram from the British Committee has been under contemplation, for emission from French Ursigram radio transmitting stations, covering measurements of the apparent height of the ionized atmospheric layer or layers, at specified times, places, and frequencies, and also apparent positions of major sources of atmospheric electrical disturbances. It is hoped that this program may be put into effect, as soon as arrangements for making, collecting, and transmitting the data shall have been completed.

It seemed desirable to the URSI Liaison Committee that cosmic data collected in Europe should be emitted, if possible, through the French ursigram stations only, so as to avoid the difficulties and confusion incident to the reception of a number of Ursigrams issuing from different European radio stations at different hours of the day.

Projected Changes and Extensions in the French Ursigrams

It is expected that the French Ursigrams will be changed, in the near future, from plain French language to code, the transmitting times, stations, and frequencies remaining the same as at present. The plan is to emit in code at 21.15 G.M.T.

(1) a meteorological bulletin (BAR)
(2) a magnetic field bulletin (MAG)
(3) an electric field bulletin (ELC)
(4) a solar bulletin (SOL)

These may also soon be followed by a coded British bulletin, as already mentioned, and by a coded bulletin on auroral observation emanating from the URSI Norwegian committee.

Projected Extensions in the American Ursigrams

It is proposed to add to the daily American Ursigram a coded bulletin (KHL) relating to measurements of the apparent height of the ionized layer or layers, at specified times, locations, and frequencies. It is also proposed to add a coded bulletin concerning auroral observations (AUR) at College, Alaska, in approximate latitude (using data from the Alaska Agricultural College and School of Mines) 64.8 degrees N., longitude 147.7 degrees W., within 2 degrees of the arctic circle. The KHL bulletin may be inaugurated on an initial weekly basis about June 1, using data from the Bureau of Standards, and to be given more frequently whenever the supply of data permits.

The code used for the KHL data is similar to that used for the other data. It includes a three-letter identification and two five-digit groups. It can be expanded indefinitely, if data become available from a multiplicity of sources by adding more five-digit groups. Following the three-letter identification, the first five-digit group gives the place and the radio frequency, and the second five-digit group gives the day of week, time of day, and effective height of the layer. An illustration would look like

KHL 31348 61513

The three-letter identification stands for Kennelly-Heaviside layer. In the first five-digit group, the letter given first indicates the place (the code for these places to be agreed upon and published; *e.g.*, 3 for Washington). The other four digits in this group give the frequency, in tens of kilocycles per second. In the second five-digit mean group, the first digit gives the day of the week, precisely as is done for "Sol," "Mag," and "Sun." The second and third digits give the nearest hour in Greenwich Mean Time. The fourth and fifth digits give the effective height of the layer, in tens of kilometers.

Thus, the above illustration states that the effective layer height at Washington, for a frequency of 13,480 kc/s, on Friday at 15 Greenwich Civil Time, was 130 kilometers.

Whenever there are observational data on two heights (E and F layers of Appleton), a complete numerical group should be given for each.

This code permits the radio frequency to be stated with sufficient precision, over the whole range of radio frequencies. It also gives the time of day and the layer height with sufficient precision. The layer height given is the virtual layer height ordinarily given directly by echo-signal calculations.

Any unused character will be transmitted as an X, as is already the practice in the present daily Ursigrams. For example, a frequency of 830 kc would be transmitted as XX83, the time 5 would be transmitted as X5, and a layer height of 80 kilometers would be transmitted as X8.

Meetings of the U. S. National Comittee on Coöperation

During the past year, meetings have been held in Washington on April 28 and November 20, 1930, at which discussions were devoted to the coöperation program, in general, and the Ursigram program, in particular. The list of committee personnel is the following: A. E. Kennelly, (chairman); E. F. W. Alexanderson, L. W. Austin, Watson Davis, J. H. Dellinger, J. A. Fleming, N. H. Heck, S. C. Hooper, W. J. Humphreys, G. W. Kenrick, (secretary), H. Mitchell, W. J. Peters, A. H. Taylor, K. B. Warner, and L. E. Whittemore.

Proceedings of the Institute of Radio Engineers
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TEMPERATURE RATING OF ENGINE DRIVEN AIRCRAFT RADIO GENERATORS*

By

C. B. Mirick[1] and H. Wilkie[2]

([1]Naval Research Laboratory, Bellevue, Anacostia, D. C.,
[2]Department Signal Office, Fort Shafter, Honolulu, T. H.)

Summary—*Previously described methods of temperature measurement and computation are applied to engine driven aircraft radio generators in flight. Observed and computed heating curves are shown from which an emission constant for this type of machine has been derived.*

A STUDY of temperature rise in wind driven aircraft radio generators is reported in an earlier issue of these transactions.[1] It is here desired to take up an application of the same methods to engine driven aircraft radio generators. So far this analysis has been confined to direct-current double voltage generators designed for mounting in a vertical position directly behind the upper cylinder of the radial air-cooled engine. The gear ratio of drive is such that the generator turns one and one-half times engine speed.

In modern planes, the engine speed may vary over quite a wide range, which entails some kind of voltage regulation. This is usually accomplished by means of a vibrating reed type of regulator which varies the field current at a rate depending upon the speed of the engine. By this means the voltage is held almost constant over a two-to-one speed variation. With such an arrangement, the losses in the generator, and therefore the heating, should vary considerably with speed. For purpose of test the speed was kept constant as possible at what was considered a normal cruising speed. During flight tests, the same speed was maintained as in ground tests. Other conditions of flight, such as altitude and ambient temperature varied quite widely, though some effort was made to keep them constant. All of the flight tests were made in a Vought Corsair plane equipped with a Pratt and Whitney Wasp engine.

The generators on which these tests were made, are shown in the photograph while their weights, losses, and efficiencies, are given in Tables I and II. Figs. 1, 2, and 3 show an observed temperature rise in the No. 1 generator with regulator, as tested in still air, at 3/4 load, full load, and 1-1/4 load respectively. Figs. 7, 8, and 9 show similar tests of the No. 2 generator. The *b* curve on each of those curve sheets

* Decimal classification: 621.313.33. Original manuscript received by the Institute, March 13, 1931.
[1] Proc. I. R. E., 17, December, 1929.

TABLE I

Engine Driven Generator No. 1

Weight	20.5 lbs.
Diameter	4.75 inches
Length	9.5 inches
Rating	
6 amperes at 15 volts	90 watts
0.36 amperes at 450 volts	162 watts
Total	252 watts

Speed during test. 2500 r.p.m.
Altitude during flight test

4	2600 feet
5	4000–6000 feet
6	6000 feet

		Losses	Over-all Efficiency
Total Losses and Efficiency	at 75 per cent load	126.5 watts	59.6 per cent
	at 100 per cent load	147.6 watts	63. per cent
	at 125 per cent load	164.5 watts	65. per cent

TABLE II

Engine Driven Generator No. 2

Weight	35 pounds
Diameter	5.625 inches
Length	11 inches
Rating	
10 amperes at 14.6 volts	146 watts
0.36 amperes at 1000 volts	360 watts
Total	506 watts

Speed during test. 2500 r.p.m.
Altitude during flight test

10	2000 feet
11	1800 feet
12	1800 feet

		Losses	Over-all Efficiency
Total Losses and Efficiency	at 75 per cent load	207. watts	65 per cent
	at 100 per cent load	265. watts	65.6 per cent
	at 125 per cent load	275. watts	70 per cent

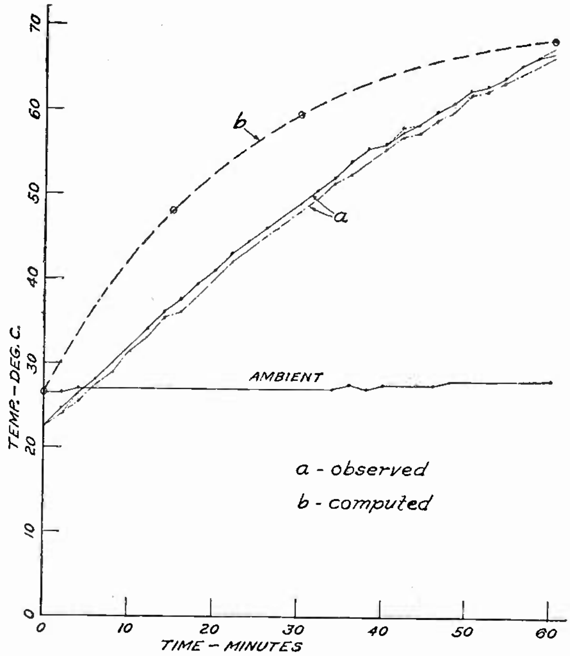

Fig. 1—Generator No. 1 in still air, 75 per cent load.

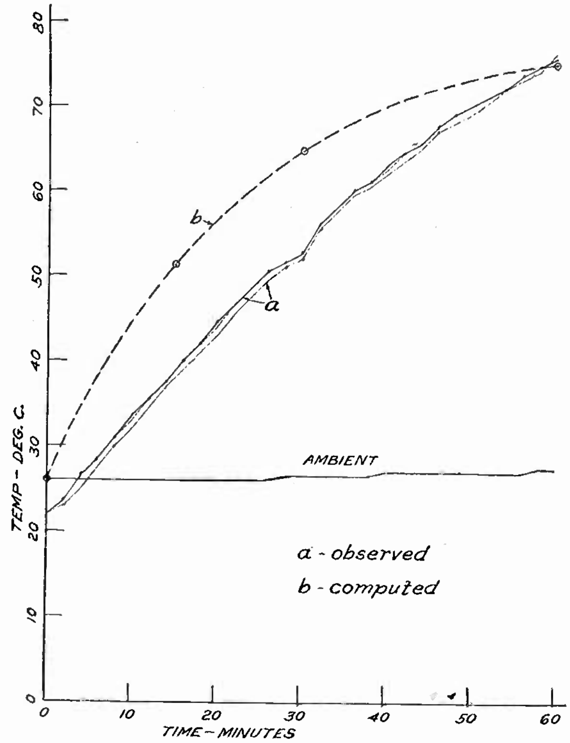

Fig. 2—Generator No. 1 in still air, 100 per cent load.

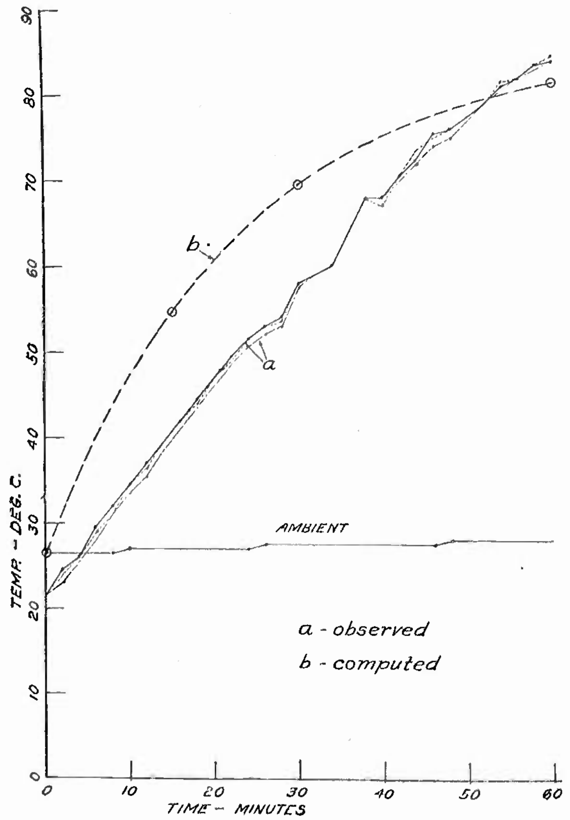

Fig. 3—Generator No. 1 in still air, 125 per cent load.

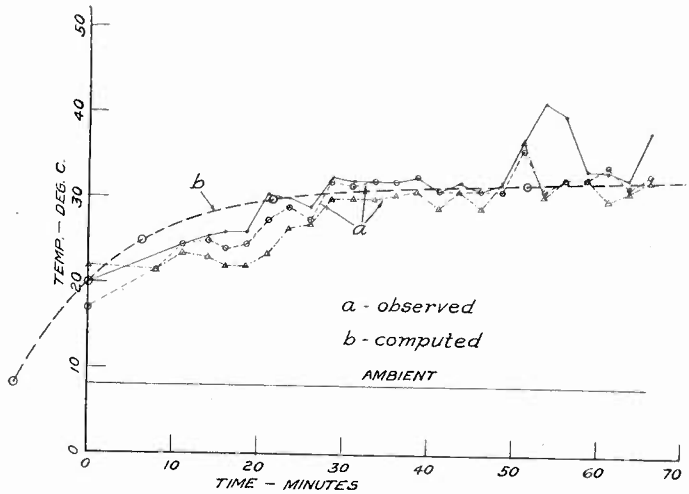

Fig. 4—Generator No. 1 in flight, 75 per cent load, altitude 2600 feet.

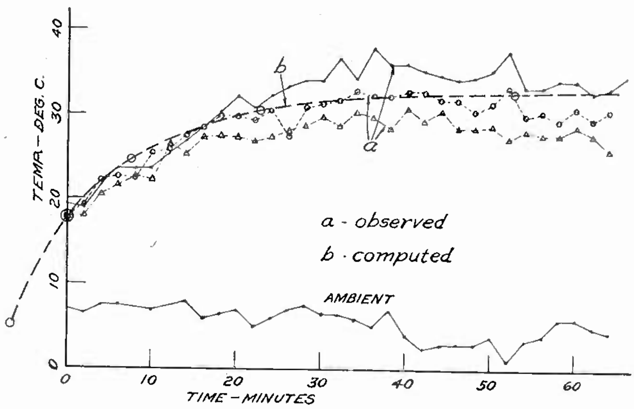

Fig. 5—Generator No. 1 in flight, 100 per cent load, altitude 4000 to 6000 feet.

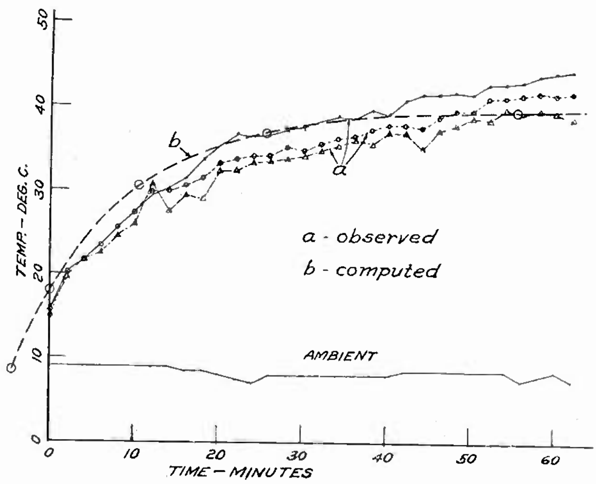

Fig. 6—Generator No. 1 in flight, 125 per cent load, altitude 6000 feet.

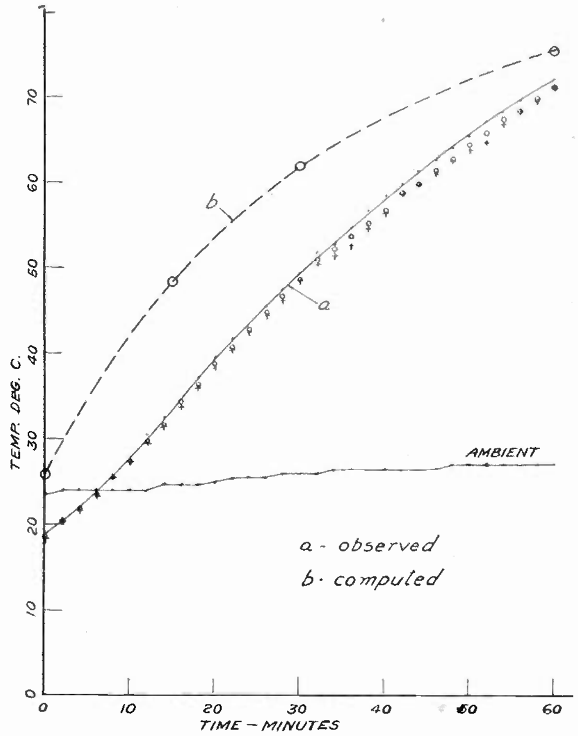

Fig. 7—Generator No. 2 in still air, 75 per cent load.

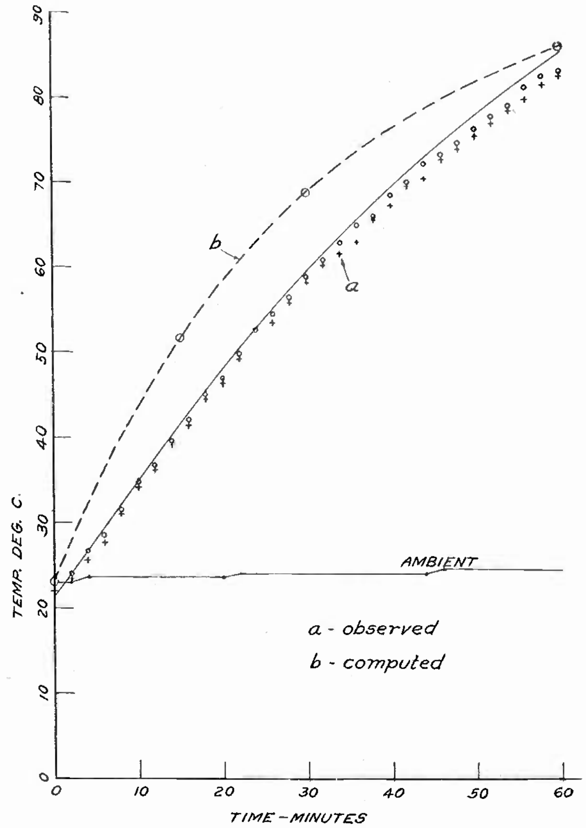

Fig. 8—Generator No. 2 in still air, 100 per cent load.

is computed from the dimensions of the generator and the measured losses using the method outlined in the earlier paper. The area used in computations was that of a cylinder having a diameter equal to the mean diameter of the generator, and a length equal to the over-all measurements including the bearings. The emissions constant of 0.033 was used for still air. Figs. 4, 5, and 6 show observed heating curves for the No. 1 generator operated during flight, but with load and speed

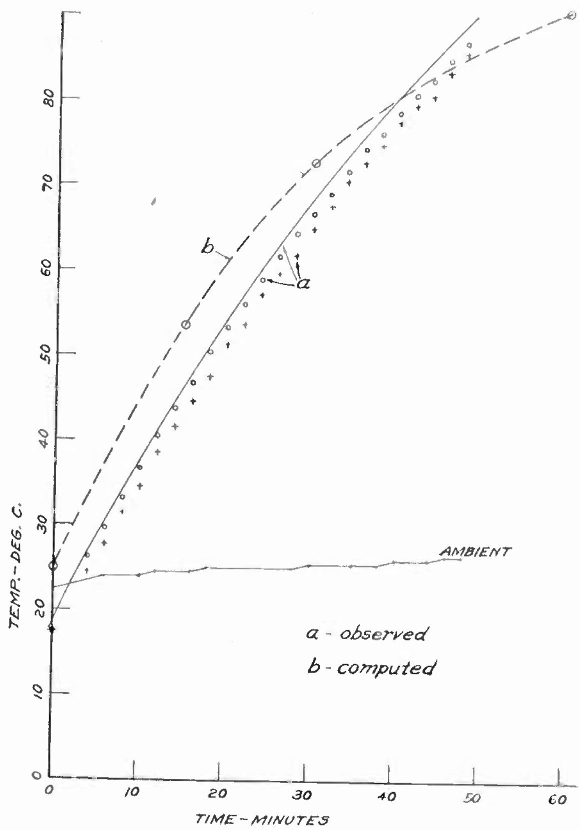

Fig. 9—Generator No. 2 in still air, 125 per cent load.

conditions corresponding to those prevailing during the preceding ground test. Figs. 10, 11, and 12 are similar records of the No. 2 generator. Based on these six flight measurements a new constant of emission for engine driven generators in flight was determined as 0.0671. Using this constant the *b* curve for the flight test sheets was computed.

It may be noted that the observed heating curves for the ground tests are quite different in shape from the computed heating curves, whereas previous tests of other generators of the wind driven type show a much closer agreement. This is believed to be the result of thermal lag and temperature gradients within the generator. As originally pointed out, such conditions are to be expected, though why this type of machine reveals a wider variation from the normal type of heating curve from those previously measured is not entirely clear. It may be noted

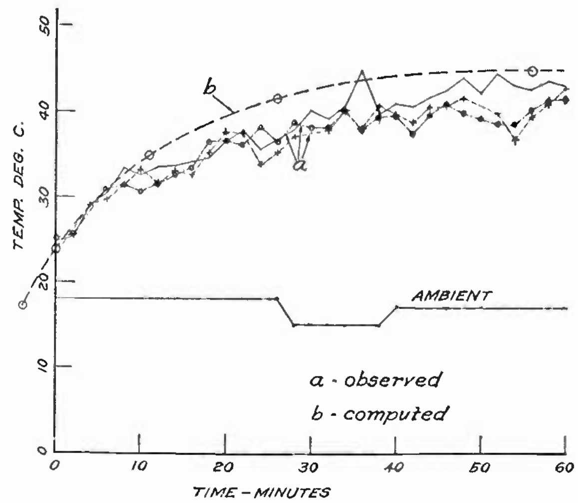

Fig. 10—Generator No. 2 in flight, 75 per cent load, altitude 2000 feet.

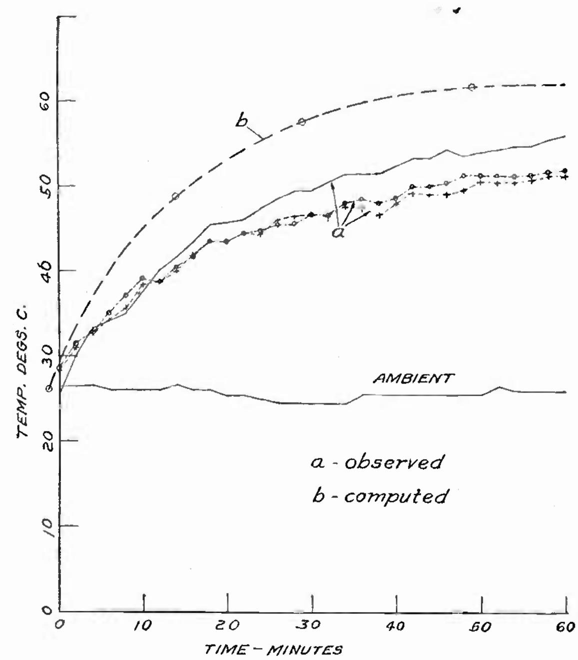

Fig. 11—Generator No. 2 in flight, 100 per cent load, altitude 1800 feet.

that the observed curves for these ground tests start below the recorded ambient temperature in each case. This is due to the fact that the runs were always started in the morning and the room warmed up before the generator did. No doubt this condition effects the general shape of the curve.

In studying the sheets showing flight tests, it is seen that the initial generator temperature is always considerably above the ambient temperature as observed by thermometer on a wing strut. This occurred in spite of the fact that the generator was always flown with open field circuit until the temperature settled down, before commencing a test. This usually required about 10 minutes. It was at first reasoned that the generator, mounted as it is, directly behind the engine cylinder and in the direct blast of the propeller, would be in a constant stream of warm air, therefore the true ambient temperature would be that at which the observations commenced. However, the curves computed on this as-

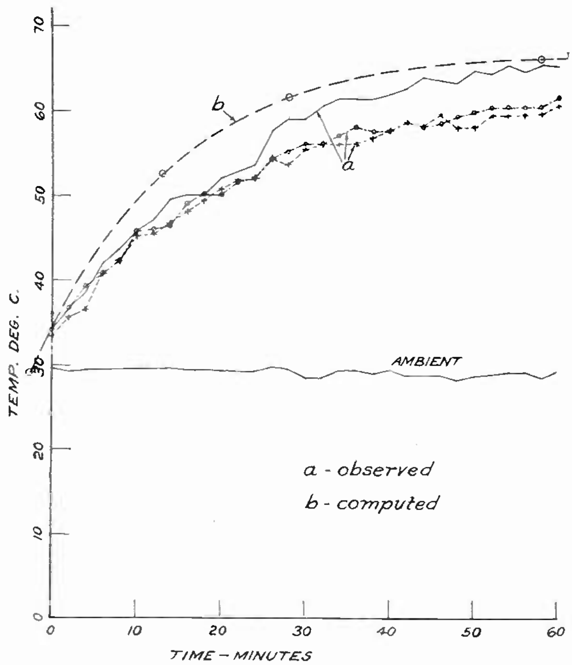

Fig. 12—Generator No. 2 in flight, 125 per cent load, altitude 1800 feet.

sumption were an entirely different shape from the observed curves and no emission constant could be found which seemed to fit.

It was then concluded that the cooling was controlled by the general temperature of air, but that the generator had been subjected to a certain amount of preheating acquired during flight and the previous period of warming up the motor on the ground. By using the temperature rise at the first, observation in the equation for the heating curve and solving for the time, a corrected initial time was obtained from which the computed curve was laid off. This is merely equivalent to an assumption that the preheating and consequent emission took place at the same rate as that which occurred when the generator was operating under load.

Using this procedure, the agreement between the observed and computed curves is found reasonable. It is therefore concluded that an emission constant of 0.067 is fairly close for engine driven generators of this type at least. From this and previous work, it may be stated in round numbers that the temperature rise in engine driven generators

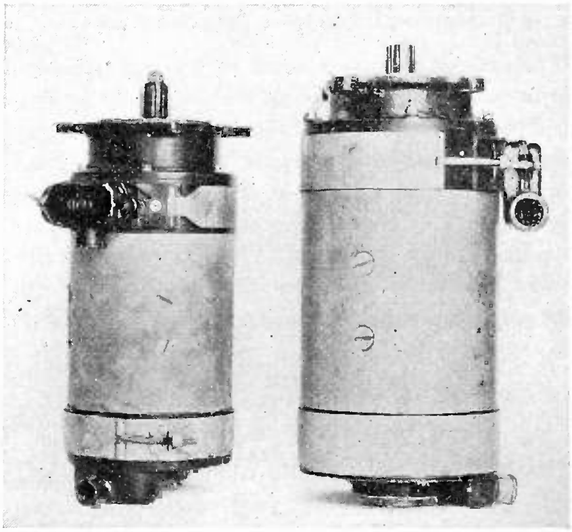

Fig. 13—Engine driven aircraft radio generators used in this study.

in flight is two-thirds of what it would be in still air, while for a wind driven generator in flight it is about one-third. The rapidly accruing flight record of these engine driven generators in service without any evidence of overheating, is taken as confirmation of this conclusion. It is desired to acknowledge the assistance of Assistant Radio Inspector A. H. Johnson who helped with the installations and made some of the flight tests.

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THEORETICAL AND PRACTICAL ASPECTS OF DIRECTIONAL TRANSMITTING SYSTEMS*

By

E. J. Sterba

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Summary—*This paper discusses some of the more important principles involved in the development of the directional transmitting antennas at present employed in the Bell System short-wave facilities. The theoretical performance of directive arrays is presented by means of various curves which have been obtained by integrations based upon Poynting's theorem. The details of the mathematical derivations are omitted for the sake of brevity, but the general procedure and the resulting formulas have been placed in an appendix. Various practical problems encountered in the development are described. These include antenna tuning procedure, transmission line adjustments, and sleet melting facilities.*

I. Introduction

During the past ten years considerable attention has been given to directive antennas. Many of these studies have appeared as papers in various technical publications. This paper and a companion paper by G. C. Southworth[1] discuss some of the problems encountered in the development and design of transmitting antennas for the short-wave transoceanic facilities of the Bell System.

Fundamentally, the advantage to be derived from a directive transmitting antenna is the power economy effected in laying down a required field strength in a chosen direction. The improvement is brought about by interference between waves emitted from the several sources. Reinforcement, usually complete, occurs in at least one direction and reduced signal in one or more directions. The net result is a reduction of radiated power required to maintain a given signal in the direction of reinforcement.

For example, suppose (n) like antennas each bearing a current (I) are distributed over an area at intervals comparable to a wavelength. Suppose that the configuration and phasing is such that in some direction complete reinforcement occurs, that is, the field strength from the combination is (n) times that from one antenna alone or (nI) units. The total power radiated by the combination if there were no couplings between the elements could be expressed as:

$$P_d = nRI^2 \tag{1}$$

* Decimal classification: R125. Original manuscript received by the Institute, October 8, 1930.

[1] G. C. Southworth, Proc. I.R.E., 18, 1502, 1930.

in which (R) is the radiation resistance of each antenna. In order to obtain (nI) units of field strength from one antenna alone it is necessary to increase the current in the one antenna (n) times and the power radiated becomes;

$$P_0 = R(nI)^2. \tag{2}$$

The improvement of the array over one antenna may be obtained from the ratio of powers:

$$\frac{P_0}{P_d} = n. \tag{3}$$

That is, the array requires but $(1/n)$ of the power consumed by one antenna in laying down a required field strength.

On the other hand, if it were assumed that the power input to the array and to the single antenna were equal, it could be shown in a similar manner that the field intensity from the array in the direction of complete reinforcement would be $\sqrt{n}$ times that from the one antenna alone.

In practice, it is generally inconvenient to employ a configuration of antenna elements in which interelement couplings are of negligible magnitude. In fact, there are element configurations and element current phasings for which the power required to lay down a given field intensity in the chosen direction is reduced by more than the factor $(1/n)$. Any attempt at a rigorous computation of the power radiated by a directive array must take into account the interaction of the currents in the individual elements. This may be done directly[2] by computing the electromotive forces (in phase with the currents in each element) which are induced into each element by currents flowing in all the other elements, and then proceeding by the method already outlined. On the other hand, an indirect method for taking account of interelement coupling is involved in the process of assuming currents in the individual elements, computing the intensity about the array and obtaining the power radiated by means of Poynting's theorem. Since the results so obtained may be interpreted as the power required to maintain the currents assumed in the individual antenna elements these will, therefore, include the effects of interelement couplings. The latter method will be employed in this paper.

A convenient numeric for stating the effectiveness of a directive array is obtained from the ratio of power consumed by one antenna to

[2] A scheme of this kind was disclosed to the writer several years ago by J. C. Schelleng of the Bell Telephone Laboratories, who has employed this method as a check upon several of the results to be found in this paper. A similar scheme was recently described by A. A. Pistolkors, Proc. I.R.E., 17, 562, 1929.

the power consumed by an array of antennas when equal field intensities are produced in the chosen direction. If the reference antenna is one of the array elements alone the numeral is the improvement over one of the array elements alone. In order to compare different directive arrays it is necessary to set up an arbitrary standard. It has been found convenient to use a vertical one-half wave Hertz antenna as a standard of comparison. As used the midpoint is somewhat more than one-quarter wavelength above ground. The improvement factor for a directive antenna is defined as the ratio of powers radiated by a standard comparison antenna to that radiated by the directive array when equal intensities are produced in the chosen direction. When the improvement factor is expressed in decibels it is called the "gain" of the array.

Many of the computations in this paper will be made for arrays of short elements grounded to a perfect metallic reflector, and will be expressed as decibels improvement over one of the array elements similarly grounded. Identical results would be obtained if an array of short elements far removed from all reflecting surfaces were assumed, providing that the result were expressed as improvement over one of the array elements far removed from reflecting surfaces. This assumption greatly simplifies the computations. Extensive calculations may be made without great effort and it is possible to observe closely the mechanism by which antenna gain is obtained. The earth at the shorter wavelengths is far from being a perfect metallic reflector. It is desirable in many cases to omit the effects of reflection from the earth rather than to assume the reflection laws of a perfect metallic reflector. Where reflection from the earth has been ignored and the results stated as the gain of the array over one of the array elements, the computations may be applied to arrays of either vertical or horizontal elements, it being assumed that the single element is parallel to the elements in the array.

In order to employ Poynting's theorem for computations of power radiated by an array it is first necessary to compute the interference pattern about the array. In general, this is done by adding in correct time and space phase the field from the several antenna elements at a point in an arbitrary direction and remote from the system. This subject has been treated excellently in a number of publications and will not be taken up in this paper. For the same reason no diagrams of directive patterns will be given. The reader interested in these diagrams will find extensive collections in papers by R. M. Foster[3] and G. C. Southworth.[1]

[3] R. M. Foster, *Bell Sys. Tech. Jour.*, 5, 292, 1926.
[1] *Loc. cit.*

The general procedure and the resulting equations for antenna gain as computed from Poynting's theorem have been placed in an appendix to this paper.

II. Theoretical Study of Antenna Gain

This section of the paper is a theoretical study of the mechanism by which gain is obtained from directive arrays. It is confined chiefly to the study of antenna elements in rectangular array. This does not imply that other antenna configurations are unworkable, but that

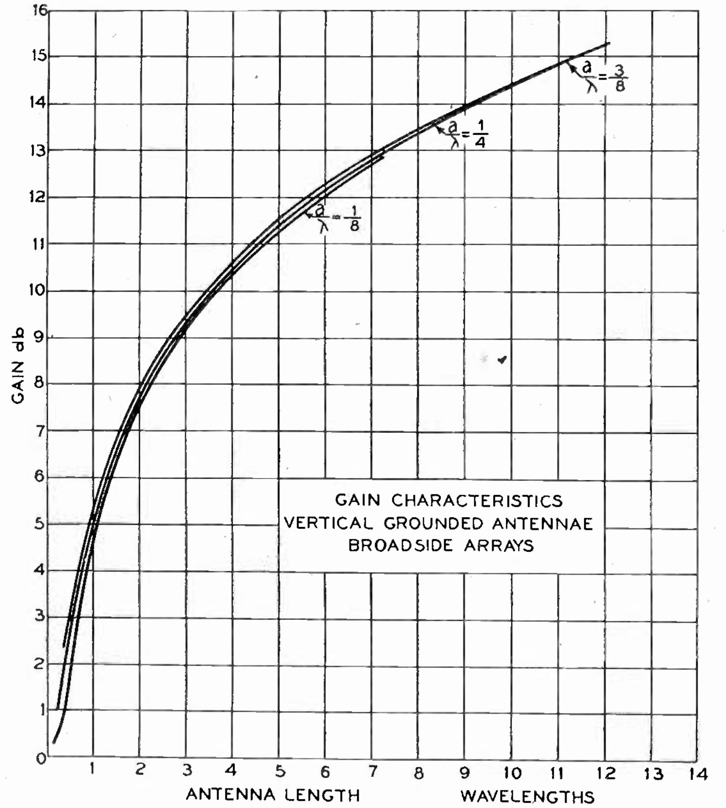

Fig. 1—The curves give the improvement expressed in decibels of a number of short antenna elements in broadside array over one element as a function of array length and for a number of element spacings.

rectangular arrays are very practical from the standpoint of feeding and supporting structures.

A rectangular configuration of antennas may be viewed as several rows and columns of elements. A row of parallel elements may be called a coplanar array. Besides extending in two horizontal dimensions the array may extend in the vertical direction. Thus, a coplanar array may comprise several tiers of elements stacked one above the other. Tiers of elements may be viewed as colinear arrays.

Computations are made, first, for the more simple configurations and then for the combinations of these simple arrangements which form a large array. In order to keep the problem within hand some additional assumptions are necessary. The utility of the results is not seriously affected by assuming that the elements are spaced at equidistant intervals, that the elements are excited by currents of equal magnitude, and that the phase of the element currents progresses in uniform intervals from element to element of the array.

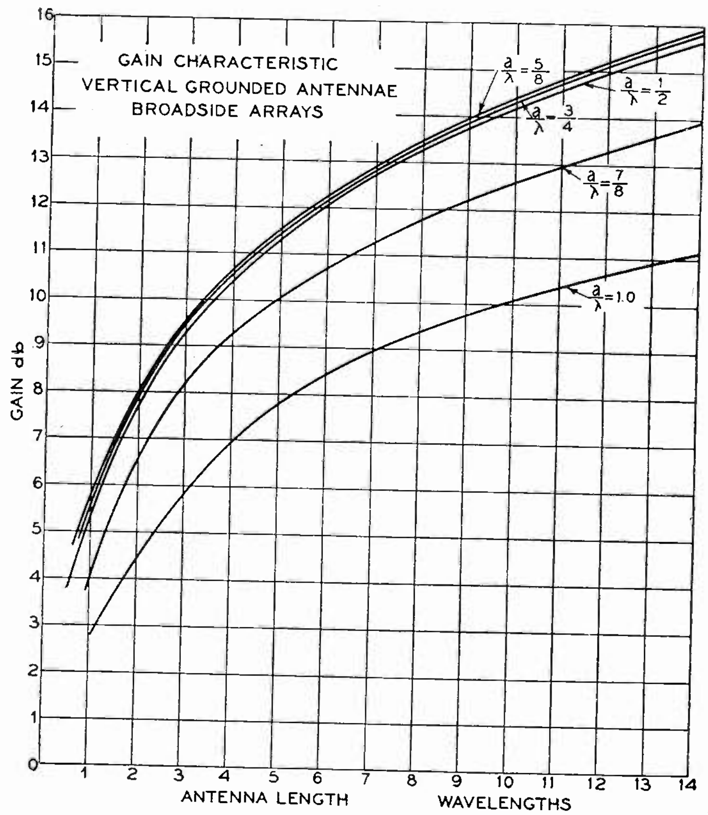

Fig. 2—The curves are similar to Fig. 1 and give cases of wide element spacing.

Coplanar arrays may be divided into two general types: broadside arrays and endfire arrays, so called because the preponderance of radiation is normal to the plane and in the plane, respectively, of the array.

There is a preponderance of radiation normal to the plane of an array when all the elements bear currents of equal magnitude and phase. The advantages to be derived from such an array expressed in decibels improvement over one of the array elements for the case of equal power inputs may be obtained from the curves of Figs. 1, 2, and 3. The first two figures plot decibels improvement of the array over one of the array elements as a function of array length and for several element spacings. Two sheets were employed owing to the proximity of the curves. It is of interest to note that for element spacings less than three-quarters wavelength the improvement factor is a function chiefly of array length. Another important observation is that for array lengths longer than about two wavelengths an improvement of 3.0 db is obtained when the array length is doubled.

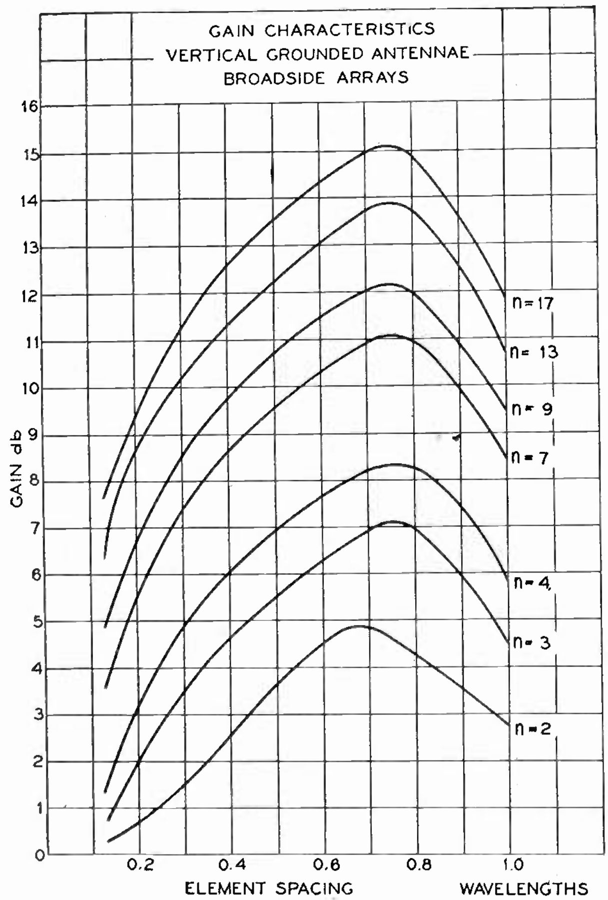

Fig. 3—The data of Figs. 1 and 2 are replotted to show the optimum element spacing relation.

The foregoing computations are replotted upon Fig. 3 as a function of element spacing and the number of elements. When plotted in this manner an optimum spacing is disclosed which, as the number of elements is increased, approaches approximately three-quarters wavelength. In view of the fact that the improvement factor is a function chiefly of over-all array length, optimum spacing is usually not of great utility except where for economical reasons only a few elements may be erected.

If in a coplanar array the current phasing in fractions of a period is made numerically equal to the element spacing expressed in fractions of a wavelength, an endfire array is formed. Computations for endfire arrays appear as the curves of Fig. 4. The improvement factor expressed as decibels improvement over one element of the array is plotted as a function of array length for a number of element spacings. As may be determined by comparing Figs. 1 and 4 there is no great

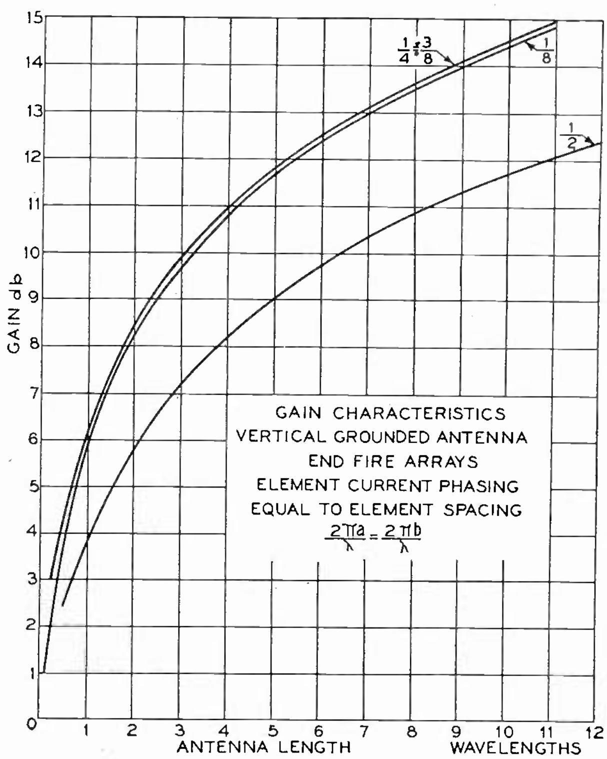

Fig. 4—The curves give the improvement expressed in decibels of a number of elements in endfire array as a function of array length and for a number of element spacings and phasings.

difference in the improvement factor for broadside and endfire arrays providing that the element spacing is less than three-eighths wavelength.

There is one well-known property of endfire arrays which is very useful. If the element phasing is an odd number of quarter periods and the element spacing is an odd number of quarter wavelengths the preponderance of radiation is in one direction. Arrays employing this one principle may be constructed. The more common scheme, however, is a combination of unidirectional endfire systems and broadside systems, that is, rectangular arrays.

Equations for the improvement factor of rectangular arrays appear in the appendix. Computations for only one particular array have been placed in the paper. These are the curves of Fig. 5. The computations were made for the following conditions. The array is assumed to comprise (q) broadside rows of (n) elements each. Both the row and element spacings are assumed to be three-quarters wavelength. The endfire phasing, that is, the column phasing, is assumed to be three-quarters period.

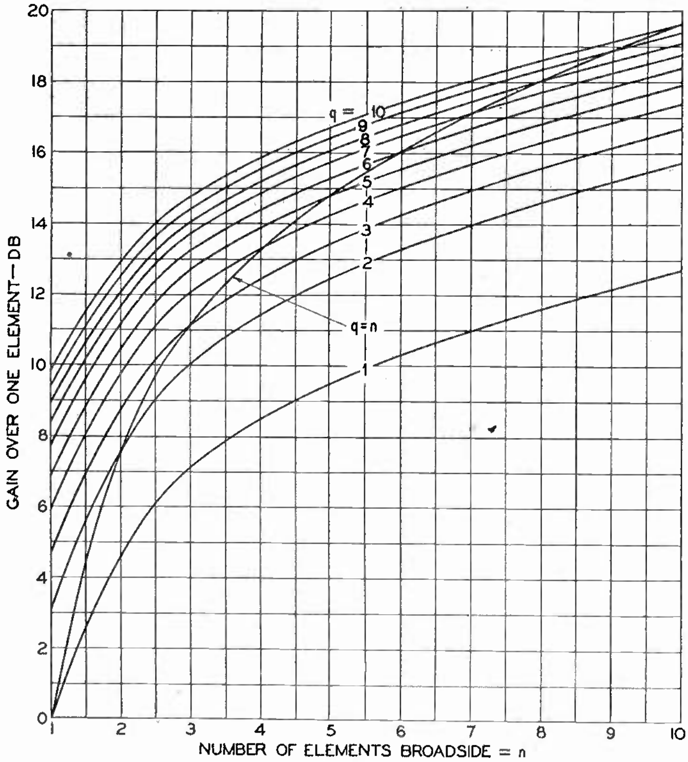

Fig. 5—The curves give the improvement expressed in decibels for one case of rectangular arrays as a function of the number of elements (n) placed in broadside rows and the number of elements (q) placed in endfire columns. Both row and column spacing is assumed to be three-quarters wavelength. The endfire phasing is assumed to be three-quarters period.

The first point of interest is that exactly 3.0 db is acquired by employing the second row so as to form unidirectional endfire columns. Upon examining the equations it is found that an apparent gain of 3.0 db (an improvement factor of 2.0) is obtained over one broadside row whenever the column phasing is an odd number of quarter periods. Of course, the gain is 3.0 db only for those combinations which give complete reinforcement, that is, the column spacing must also be an odd number of quarter wavelengths. The explanation is simple. If two elements in a column bear currents ninety degrees out of phase, the increase in power required to maintain unit current in one element due to the coupling between elements is exactly offset by the decrease in power required to maintain current in the other. The total power required to drive both elements is, therefore, exactly twice that for one element and if the element spacing is such that complete reinforcement is obtained in but one direction, twice the power is radiated in that direction or a gain of 3.0 db is acquired.

The improvement from three rows is approximately 1.0 db more than two rows. Succeeding rows add less to the gain of the array. In

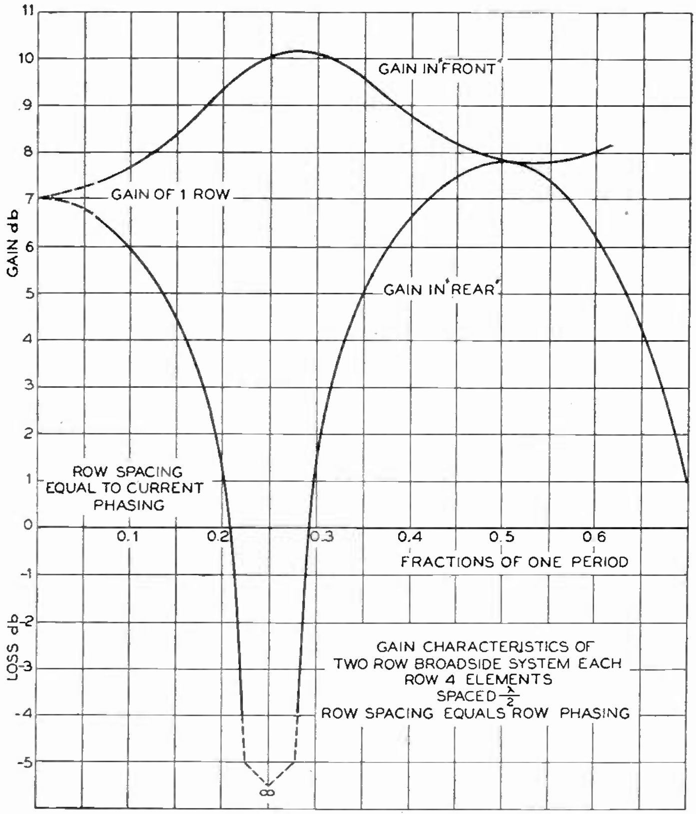

Fig. 6—The curves give the improvements expressed in decibels both in the preferred direction and the opposite null as a function of the numeric: row phasing expressed in fractions of a period equal to row spacing expressed in fractions of a wavelength for a short two-row system.

order to obtain marked improvement the length of the columns must be of the same order as the length of the rows.

The two-row unidirectional system is of practical interest because it is the ideal case of the familiar exciter-parasitic reflector system. For this reason some additional computations have been made. The assumptions are that the array comprises two broadside rows, that each row is four elements spaced at one-half wave intervals and that the row spacing and row phasing are variable. The computations have been made for the smallest unit which is typical of a large array.

In Fig. 6 the gain of the above two-row system is plotted as a function of row phasing for the particular case when the row phasing expressed in fractions of a period is made numerically equal to the row spacing expressed in fractions of a wavelength. Both the gain in "front," the preferred direction, and to the "rear," the opposite null, are shown on the figure. The results are corrected for incomplete reinforcement. The null to the "rear" occurs for the values: 1/4 wavelength, 1/4 period. The sides of the curve in this region are very steep

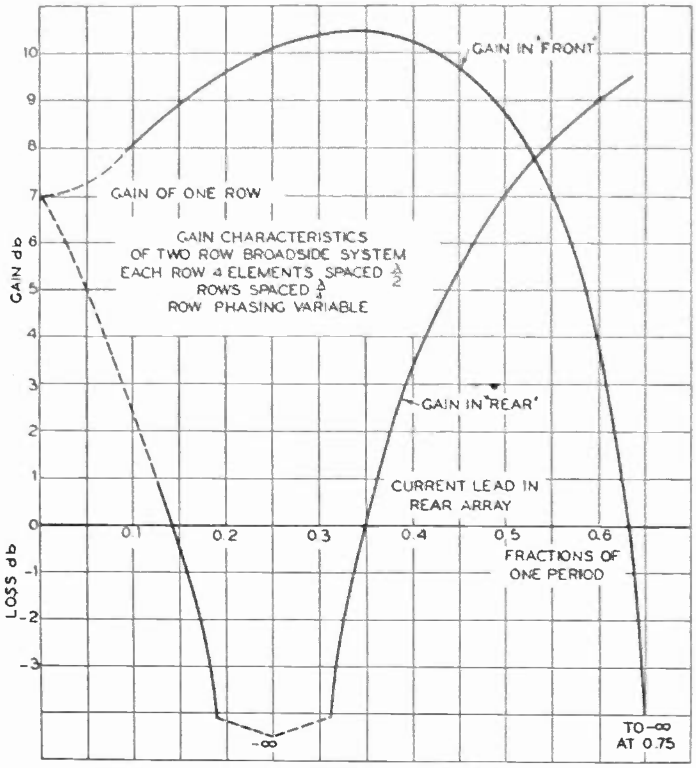

Fig. 7—This is a special case of Fig. 6 in which the row spacing is fixed at one-quarter wavelength and the row phasing is the variable. The dependence of undirectional transmission upon the phase of the current in the second row is of particular interest.

which means that the adjustment is critical. Maximum gain in "front" does not occur for this adjustment but at a somewhat higher value. The increase in gain in passing to the optimum adjustment is of negligible value.

In order to examine the effects of row phasing in more detail the data of Fig. 7 have been computed. The two-row system of Fig. 6 is assumed. It is also assumed that the row spacing is fixed at one-quarter wavelength and that row phasing is the variable. It is useful to observe that the phase adjustment is not very critical if an absolute null in the "rear" is not required and that the gain in "front" is even less critical.

It is for this reason that a parasitic reflector may be slightly detuned to obtain a desirable null without measurably affecting the gain in the preferred direction.

When the practical limitations of length and breadth of a directive system are reached the system may be extended in a vertical direction. That is, several units may be placed in tiers, one above the other, so as to obtain additional signal improvement. In order to study the gain to be acquired by placing antenna elements in tiers it is first desirable to examine some simple cases in detail and then to pass to the more complex systems.

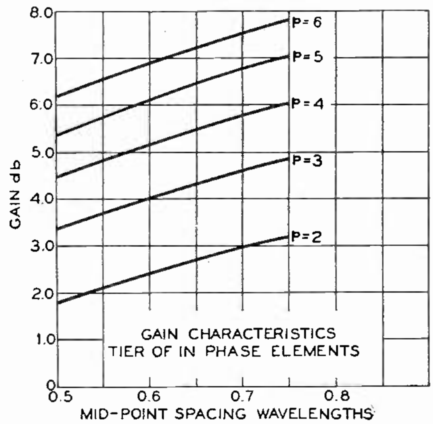

Fig. 8—The curve gives the improvement expressed in decibels of short in-phase elements in colinear array over one element as a function of mid-point spacing and for a number of elements. For the case of several tiers of broadside arrays the results given here are low by approximately 0.6 db.

The improvement of (p) short colinear in-phase elements expressed as decibels improvement over one element is plotted on Fig. 8 as a function of midpoint spacing. The equations appear in the appendix. Computations have been made only for midpoint spacings of practical importance. For large spacings the improvement curves oscillate and approach ten times the logarithm to the base ten of the number of tiers.

The improvement incurred in placing several tiers of broadside arrays one above the other cannot be obtained directly by adding the tier improvement, expressed in decibels to the row improvement expressed in the same units. There are couplings between elements in such a system occurring in those directions for which diagonals may be drawn which modify the net improvement. For brevity computations are given for one practical system. It is a broadside row of (n) elements spaced at one-half wavelength intervals, (p) tiers spaced at one-half wavelength intervals between midpoints and a similar curtain forming

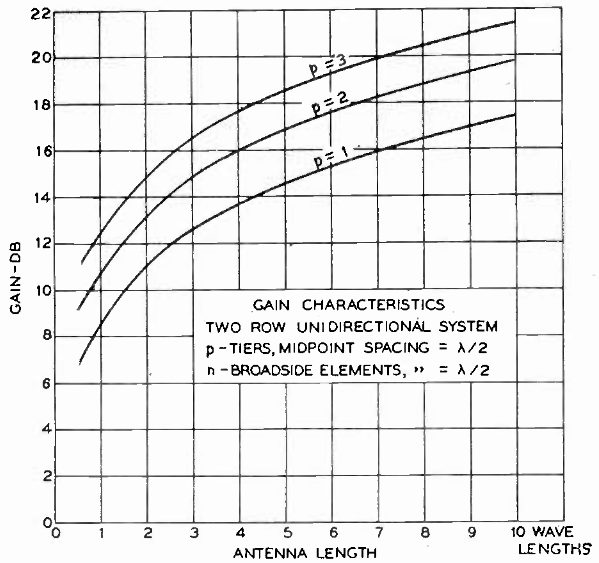

Fig. 9—The curves give the improvement expressed in decibels of a unidirectional, two-row, broadside system as a function of broadside length and for a number of tiers. Short elements far removed from reflecting surfaces are assumed.

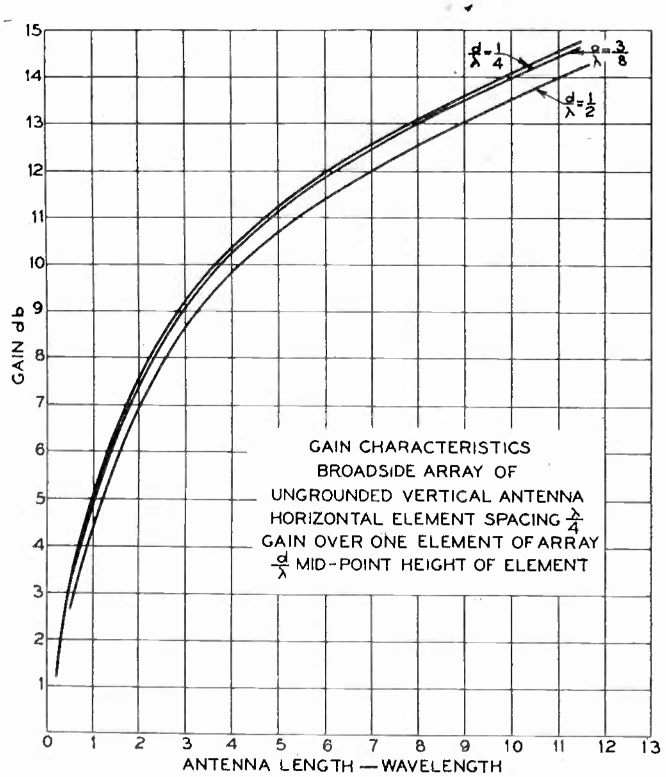

Fig. 10—The curves give the improvement expressed in decibels over one of the array elements of an ungrounded broadside row of short elements above a perfect metallic reflector for several midpoint heights. The element spacing is assumed to be one-quarter wavelength.

a second row one-quarter wavelength away and phased at one-quarter period with respect to the first row. The computations appear in Fig. 9. The equations are in the appendix. It is interesting and useful to observe that substantially the same result would have been obtained if to the improvement factor for the broadside row had been added the tier factor of Fig. 8, the quantity 3.0 db for the second row and a cor-

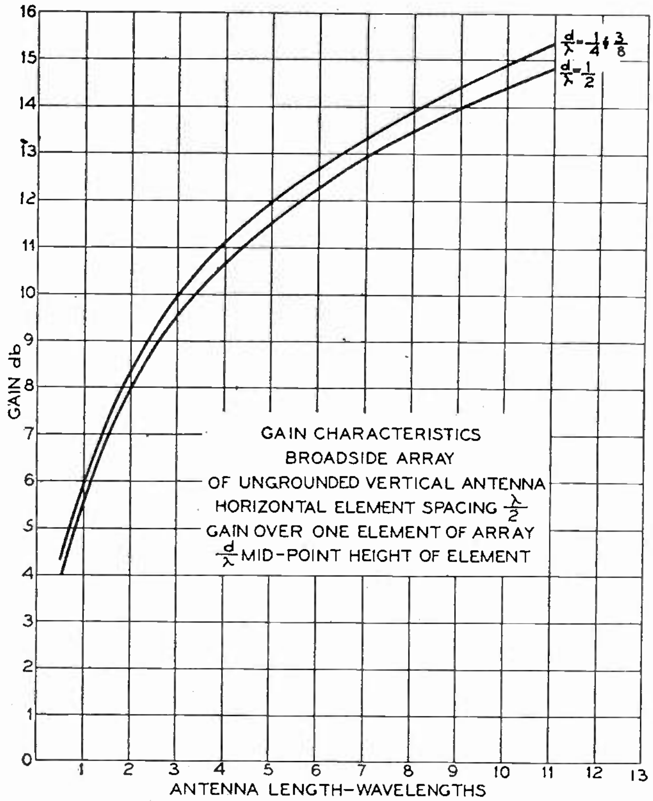

Fig. 11—The curves give the improvement expressed in decibels over one of the array elements of an ungrounded broadside row of short elements above a perfect metallic reflector for several midpoint heights. The element spacing is assumed to be one-half wavelength.

rection factor of 0.6 db for all broadside lengths longer than one wavelength.

So far in the paper the reflecting properties of the earth have been ignored. Computations of antenna gain which involve the dielectric properties of the earth are difficult. Some interesting information, however, may be obtained by assuming that the earth is a perfect metallic reflector. On this basis the curves of Figs. 10 and 11 have been calculated. The curves plot the improvement of a broadside array over one of its elements for a number of element midpoint heights above ground and for two horizontal element spacings. The results so obtained are very little different from those of Fig. 1 for the case of short grounded elements or short elements in space. The reason is that the results have been expressed as the gain of the array over one of its own elements. The results indicate that those couplings between antenna elements which involve the earth do not affect to a marked degree the broadside gain of an array.

An attempt to reduce the improvement factor to the gain over a standard comparison antenna is not so readily computed. If the earth is assumed to be a perfect metallic reflector the data of Fig. 12 are readily obtained. The figure plots the gain of a vertical antenna element of variable midpoint height over the standard comparison antenna. The curve indicates in particular that the midpoint height of a vertical element should be from 0.4 to 0.5 wavelength above a low resistance ground.

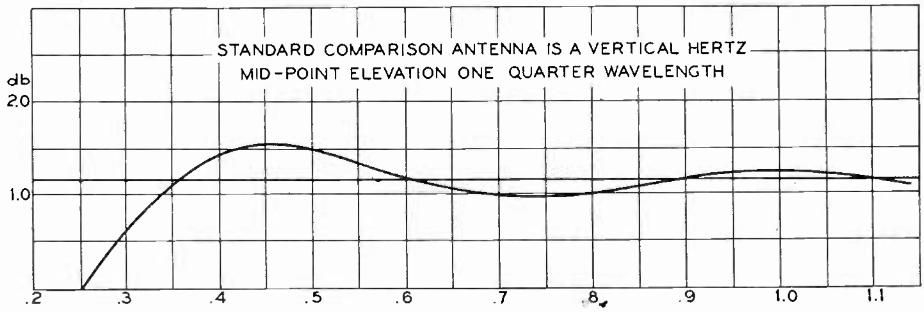

Fig. 12—The curve gives the improvement of one element over a standard comparison antenna as a function of midpoint antenna height for well conducting ground.

III. Limitations of Antenna Gain

Under the conditions assumed in the foregoing computations there appear to be no limitations to the gain which may be realized from directive systems. The computations show that if the number of elements in an array are increased by a factor of two so as to increase one dimension by the same factor a gain of approximately 3.0 db is to be expected. There are, however, some very real limitations to the total gain obtainable.

The cost and annual charge are limiting factors. From the standpoint of annual charge it is more economical, after the dimensions have reached certain values, to increase the signal strength by augmenting the radio power equipment rather than enlarging the antenna. This economical limit is to a large extent a function of the wavelength.

The sharpness of the directive pattern may limit the antenna gain. Owing to a lack of homogeneity in the transmitting medium changes in direction along the path of a ray may occur. When the angular width of the pattern is less than these deviations difficulties in the form of fading and weak signals are to be expected. Both the horizontal and vertical directivities are affected. The phenomena are too variable to be described by any simple statement.

Such deviations are most evident when the directive pattern in the vertical plane is narrow. Under certain conditions the theoretical gain expected from some combination of units such as several tiers of elements is never realized. In fact an appreciable loss over one unit may be found. At times the major lobe should be directed at an appreciable angle from the horizontal. Quantitative information on this subject is meager. In the 16-meter region vertical directivities such as those obtained from three-tier-high antennas are very useful. On the other hand, the utility of more than two tiers in the 30-meter region is very questionable.

Directional fluctuations in azimuth reduce the gain when this is obtained by excessive horizontal directivity. The limit is variable and not definitely known. For a broadside antenna there is some evidence that little is gained by extending the antenna length beyond eight or ten wavelengths.

Assuming the most useful radiation to be at low angles, the foregoing considerations imply that the maximum to be expected from a directive array is a gain of 20–22 db. When the angle is so high that the standard comparison antenna is ineffective this limit may be too low. Other reservations may be required by the neglect of the actual properties of the earth.

An example of antenna construction which approaches the apparent limit of useful gain may be of interest. It is a 16.35-meter broadside array of one-half wave elements in three tiers and 8.5 wavelengths long. It was erected and tested at the Deal, N. J., field laboratory. From Fig. 9 the gain of the antenna is 20.6 db. To this should be added 0.6 db obtained from Fig. 12 for 0.3 wavelength height of lowest tier. The latter correction is questionable because the soil at Deal is far from being a perfect metallic reflector. The gain also may be obtained from Figs. 8 and 11.

Gain for broadside of 8.5 wavelengths	14.2 db
Gain for three tiers	3.4
Correction for tier gain (see section II)	0.6
Gain for reflector	3.0
	21.2 db

The gain of the antenna was measured in New Southgate, England. The results when corrected for transmission line loss were found to be 21.7 db. The data appear as the curves of Fig. 13. The field intensities plotted on this figure are expressed as decibels above one microvolt per meter. The variations of the gain about the average result are largely observational errors due to fading.

It is well known that the dielectric properties of the soil have a marked influence upon vertical directivity and antenna gain. An antenna developed upon one site might, conceivably, not be suitable at

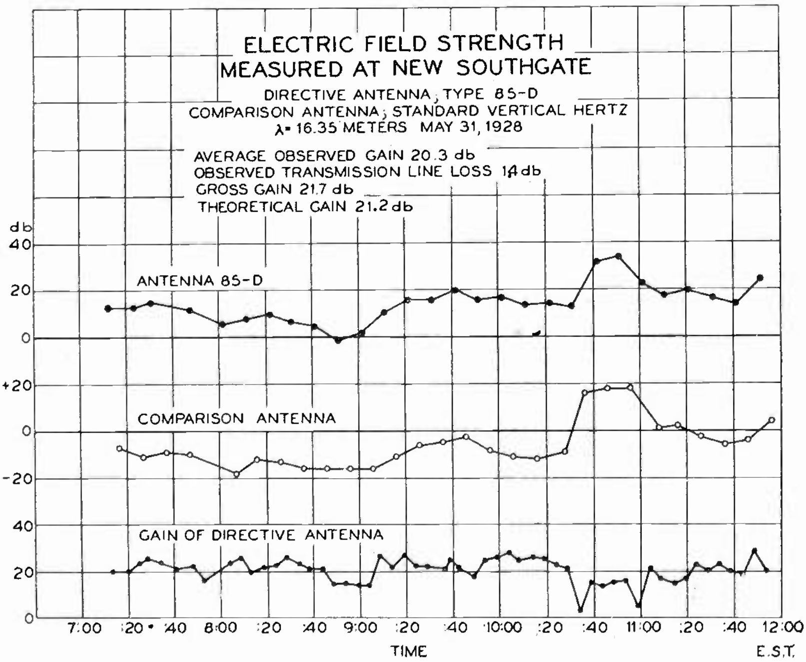

Fig. 13—The figure plots a typical measurement of antenna gain. The directive array is a three-tier-high broadside antenna 8.5 wavelengths long employing a parasitic reflector. A standard comparison antenna was employed.

another. At some sites and at some frequencies the conduction currents exceed the displacement currents and at others the opposite is true. The characteristic phenomena for these two cases are radically different. Under conditions for which the conductivity is large, the calculations made in the preceding section apply substantially without correction. On the other hand, when the displacement currents predominate the optical properties of the soil are very different from those of metallic reflection. Although the foregoing equations may give a fair approximation of gain over a comparison antenna located upon the same site, these may be found very much in disagreement if the comparison antenna is erected upon soil of radically different dielectric properties.

Some idea of what is to be expected when the displacement currents in the soil predominate may be obtained from Fig. 14. The curves depict the "space" characteristic in the vertical plane for vertical

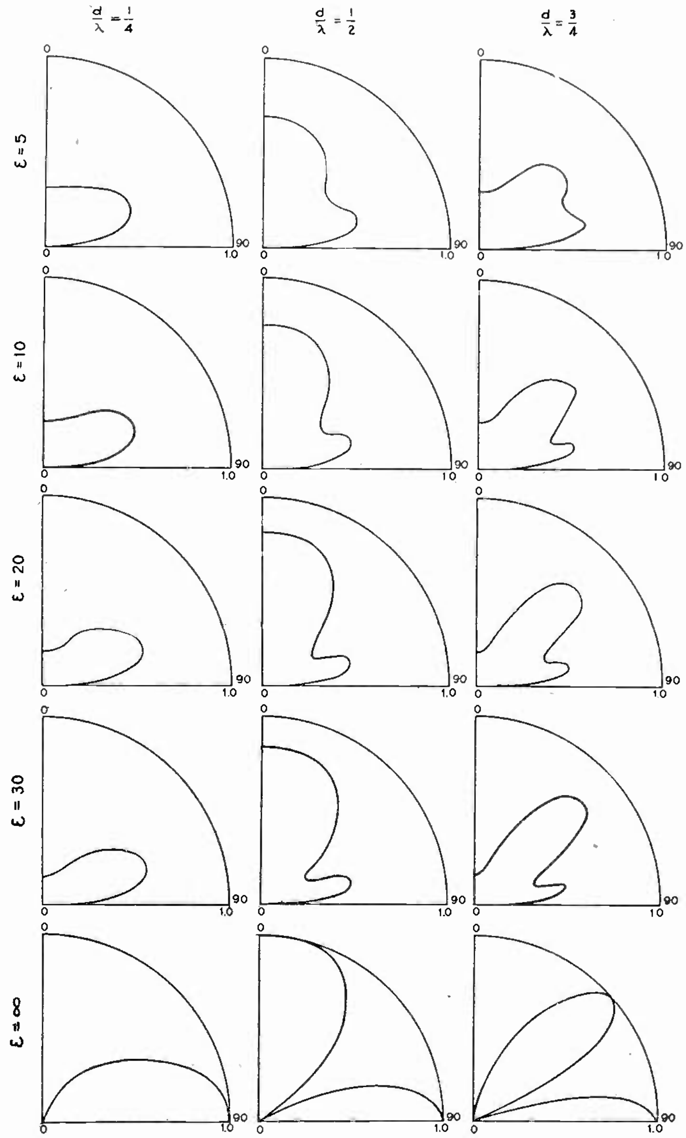

Fig. 14—The curves give the directive diagram in the vertical plane about a spherical source of radiation for several source heights above a perfect dielectric and for several values of dielectric constant. The lowest row is the case of perfect metallic reflection.

polarization. By "space" characteristic is meant the diagram which would be obtained if the antenna were replaced by a spherical radiator. In order to obtain the total characteristic these should be multiplied point for point by the diagram of the element grouping in the vertical plane. The reader interested in a more complete treatment is referred to a paper "Asymptotic Dipole Radiation Formulas."[4]

It is interesting to observe that the antenna for which the above experimental gain measurements were given was erected upon soil for which the dielectric constant is approximately 30 and the resistivity 600,000 ohms-cms. The close check between calculated and experimental values is probably a coincidence.

IV. Measurements of Antenna Gain

It may be well to mention some of the precautions which should be observed in measuring the gain of a directive antenna. Such measurements usually are made by observing the field intensity at a remote point in the chosen direction from the directive array and from a comparison antenna. It has been found convenient to employ a vertical one-half wave Hertz as the standard of comparison. As used, the midpoint is usually one-quarter wavelength above ground. It is desirable to supply equal power to both antennas in order to eliminate corrections and errors in reducing the data to a common power level.

Considerable care must be exercised to eliminate all stray coupling between the directive antenna and the comparison antenna. Suppose, for example, that the gain of an antenna from which an improvement of 20 db is possible is being determined by experiment. At the time in question suppose that the signal intensity from the comparison antenna is being measured. A stray coupling between the directive and the comparison antennas 20 db down will introduce considerable error because the current induced in the directive array will then be of the same amplitude as the current in the comparison antenna. If at the point of observation the field from the comparison antenna and that due to the induced currents in the directive antenna are in phase the field strength, which to the observer appears to be only that from the comparison antenna, will be 6.0 db too high and the gain so measured will be low by the same amount. If, on the other hand, the fields are opposite in phase or nearly so the apparent gain of the antenna will be absurdly large.

Some additional precautions should be observed when gain measurements are made at a point so distant from the antenna that the fading of signals is involved. Variations of field strength with time may be minimized by observations over a period of (say) two minutes with each antenna. With the additional intervals required for switching the

[4] W. H. Wise, *Bell Sys. Tech. Jour.*, **8**, 662, 1929. Also see M. J. O. Strutt, *Ann. d. Phys.*, **1**, 721, 1929; T. L. Eckersley, *Jour. I.E.E.* (London), **65**, 600, 1927.

complete cycles may then cover a period of approximately ten minutes. This cycle of observation is repeated many times. With the fading usually encountered a significant result can hardly be obtained in less than an hour. A few observations are of little value.

V. Practical Aspects of Antenna Construction

So far in this paper the treatment of directive arrays has been chiefly from a theoretical standpoint. The rest of the paper will be confined to some observations upon antenna construction, tuning, and operation.

It is out of the scope of this paper to review all the schemes which have been employed in the construction of directive arrays. A few

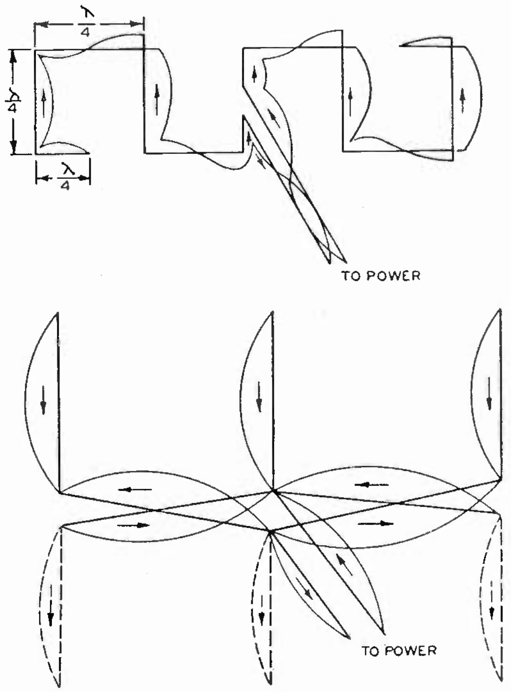

Fig. 15—The upper figure depicts one scheme for obtaining a broadside array and two-wire transmission line from two bent conductors. The lower figure is a practical scheme for employing two-wire lines as combined feed lines and phase shifters between several elements.

broadside schemes[5,6,7] which may be familiar to the reader are depicted upon Figs. 15 and 16. The general scheme in most of these is that a wire may be bent and excited so as to bear standing waves thus forming a number of elements some of which function as radiators and others as phase shifters. Since at intervals of about one-quarter wave-

[5] C. S. Franklin, English Patents Nos. 310,451 and 311,449.
[6] M. Latour, English Patent No. 319,324.
[7] H. Chireix, *L'Onde Elec.*, 7, 169, 1928.

length measured from one end of the wire points of real impedance occur and since at intervals of about one-half wavelength a reversal of phase occurs, the elements are generally multiples of these dimensions.

It would appear from the figures that a long conductor may be bent into a directive array with but little thought as to what effect the geometrical shape has upon the current distribution in the array. It has been found by experiment that this condition is not always true. The

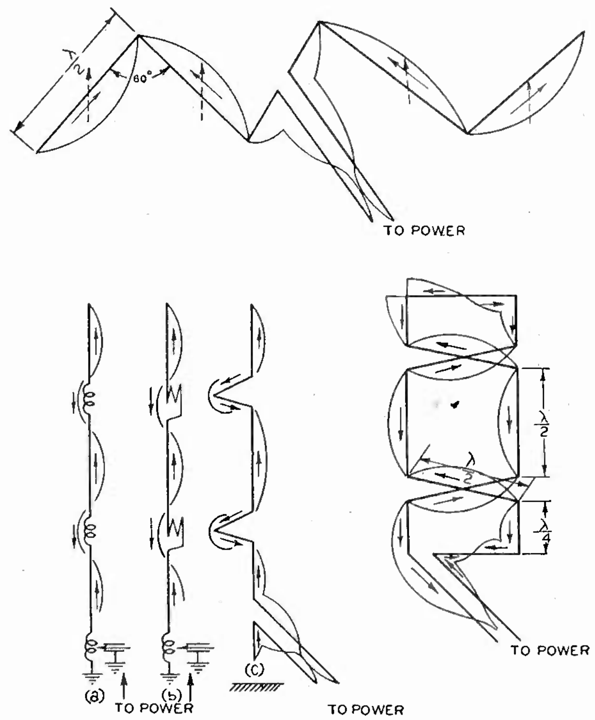

Fig. 16—The upper figure depicts another scheme for utilizing two conductors to form a two-wire transmission line and broadside array. The lower left-hand figures are several schemes by which tiers of antenna elements are obtained. The lower right-hand figure is the scheme employed in the Lawrenceville antenna system for obtaining tiers of elements and closed loops for sleet melting currents.

manner in which the waves stand upon the wire is affected to a marked degree by the proximity of the elements and by reflections from corners, fittings, and other irregularities. The reader is probably familiar with the fact that metallic fittings and harness possess a loading effect upon the conductor which makes up an antenna element. This effect produces a reduction in the wire length from a one-half wave element. Another effect becomes evident when a long conductor is bent into a number of radiating and interconnecting elements. The couplings between elements and reflections from the corners produce a loading effect of opposite character to that ascribed to the fittings. Often the latter effect is pronounced so that it dominates the shortening effect of the fittings. This may not be a general rule for all types of arrays. Nevertheless, this effect has been found in arrays of such varied configuration that in setting up a new design involving bent continuous wires several wavelengths long the one-half wave elements are made from 2 to 5 per cent long as a first approximation of the correct length. Also, it has been observed that these loading effects are not a simple function of the wavelength. The ratio of average element length to the wavelength is very different for two similar arrays one of which is designed to operate at a wavelength appreciably different from the other.

The size of an array curtain built up from a single conductor and excited from one point is limited by the amplitude and phase differences which exist between the currents at that point and at the remote parts of the antenna. Lengths of three to four wavelengths have been employed with success in transmitting antennas. Where the phase changing elements function as two-wire transmission lines (lower diagram of Fig. 15) greater lengths have been employed.

The means for connecting the source of radio-frequency power to the antenna is closely related to antenna design. It is generally desirable and often necessary where several antennas are associated with one transmitter to erect the antennas at some distance from the transmitter. Two kinds of lines have been commonly employed for this purpose, the balanced open-wire type and the concentric tube type. Owing to the large surface area and to the shielding effect upon the inner conductor which eliminates radiation losses the concentric type is the more efficient of the two. The efficiency is obtained at a high construction cost. On the other hand the cost of open-wire lines is small because standard power line material may be employed, maintenance is relatively unimportant and connections to the curtains may be made in a simple manner.

In the construction of open-wire lines certain precautions must be observed. The lengths of the two conductors should be equal. Dissymmetry of this nature is apt to arise at corners. It has been found that one satisfactory method for turning a corner is to employ four insulators and correct the line lengths at the loops between insulators. The irregularity due to the loops does not appear to be as detrimental as inequalities in line lengths. Corners, insulator tie wires, joints, and even the insulators themselves are irregularities. Since the spacing of the line fittings is an appreciable fraction of a wavelength and often several wavelengths, each fitting is a reflection point and these irregularities cannot be treated as continuous loading. It is usually true that the effect of each irregularity is small. The total effect in a long line, however, may be appreciable.

Opportunities for spurious coupling between the antenna and the line should be minimized, particularly near the antenna. The currents so induced flow in the two wires acting as a single conductor. Such currents may affect transmitter operation. In addition, an appreciable loss of power due to radiation and to ground currents is incurred. Since the control of these currents is difficult it is best to avoid as far as possible construction which gives rise to coupling between the line and the antenna. Vertical sections of line near an antenna made up of vertical elements may produce this difficulty. Some control is obtained by circuits which bring these currents to a junction exactly out of phase. For example, suppose that an antenna is made up of two similar sections which are fed by two transmission lines from a common junction. The addition of one-half wavelength of line to one side of the junction produces a 180-degree phase shift in both the normal and in the induced currents. This phase shift may be corrected for the normal currents by reversing the line terminals upon either side of the junction. This reversal does not affect the phase of the induced currents which, therefore, will cancel.

The induced currents also may be shorted to ground by means of a two-wire transmission line one-half wavelength long. The two wires of this shorting device are connected together exactly one-quarter wavelength from the transmission line from which it is desired to drain the induced currents. Thus, a high impedance is obtained for the normal currents on the line and a low impedance to ground for the induced currents. The ground connection is most readily made by means of a counterpoise of one-quarter wavelength long elements.

It is generally known that in order to eliminate reflection losses due to the antenna load at the far end of the line it is necessary that the impedance at the junction of the antenna and the line be real and equal to the surge impedance of the line. Often the antenna curtain dimensions may be chosen such that the junction resistance matches a convenient spacing of line conductor and sometimes the conductor size in the curtain and the line may be alike. More often, however, it is necessary to build up the antenna impedance to that of the line by means of suitable transformers.

Tuned transformers may be employed for matching the impedance of the antenna and the line. In many cases, however, it is convenient to employ a section of transmission line as a transformer. The circuit thus provided is extremely simple and cheap.

If a transmission line of length (l) and surge impedance (Z_0) is connected at the receiving end by a load (Z_r) the impedance at the sending end of the line is given by the familiar relation:

$$Z_s = \frac{Z_0^2 + Z_0 Z_r \coth Pl}{Z_r + Z_0 \coth Pl}. \tag{4}$$

At radio frequencies the propagation constant (P) is practically a pure imaginary and may be written as:

$$P = \frac{j2\pi}{\lambda} \tag{4a}$$

in which $j = \sqrt{-1}$

λ = wavelength in the same units as (l).

If the line is exactly one-half wavelength long the relation simplifies to:

$$Z_s = Z_r. \tag{5}$$

On the other hand if the line is one-quarter wavelength long the relation becomes:

$$Z_s = \frac{Z_0^2}{Z_r}. \tag{6}$$

At radio frequencies the surge impedance is given with sufficient accuracy by the real quantity:

$$Z_0 = 276 \log_{10} \frac{2D}{d} \text{ ohms}, \tag{7}$$

D = wire spacing measured between centers,

d = wire diameter in the same units.

Equation (5) states that a line one-half wavelength long possesses the properties of a one-to-one transformer. If the receiving end load is real the sending end load will be real.

Equation (6) states that a line one-quarter wavelength long may be employed as a step-up or step-down transformer, the ratio for a given case being determined by the surge impedance of the line, that is, by wire diameter and spacing. As in the case of the one-half wave line, if the receiving end load is real the sending end load is real.

The performance of the quarter wavelength line as a transformer may be checked by observing the currents at the receiving end (I_r) and at the sending end (I_s). Employing (6) and assuming that the power measured at the sending and receiving end of the quarter wavelength line is the same:

$$Z_s I_s^2 = \frac{Z_0^2}{Z_r} I_s^2 = Z_r I_r^2 \tag{8}$$

in which it is assumed that all impedances are real. The following useful relations may be obtained from this equation:

$$Z_r = Z_0 \frac{I_s}{I_r} \tag{9}$$

$$Z_s = Z_0 \frac{I_r}{I_s} \cdot \tag{10}$$

It follows that if the currents at any two points one-quarter wavelength apart are equal, the line may be considered as terminated with a load equal to its surge impedance. Practically it is not advisable to depend upon but two readings but to obtain the current distribution in both wires by means of a portable indicating device for a section of line at least one-half wave long. If the data so obtained are plotted as a function of their location along the line the readings when the match is obtained will give practically a line which is parallel to one axis.

Another interesting property of quarter-wave lines is that as the receiving end load approaches zero the sending end load approaches an infinite value. It would be infinite if the propagation constant were purely imaginary. It has already been mentioned that such a loop is very convenient for the purpose of shorting to ground pick-up on the lines from the antenna. It is also very useful for by-passing sleet melting currents and for draining static charges to ground at points where a direct-current circuit is required which is a high impedance to the radio-frequency currents.

In broadside antenna design it is not necessary to match the impedance of every antenna element or panel grouping of elements to surge impedance of the connecting line. It is possible to employ at 15-meters wavelength impedance matches 50 per cent in error for lines one wavelength long without incurring losses of undesired magnitude. This permits grouping a number of elements together before an impedance match is made thereby simplifying both construction and tuning procedure.

The weather has a marked effect upon the performance of an open-wire line. The insulation resistance, velocity of propagation, and surge impedance are affected. Likewise, the tuning of an antenna is affected by the weather. Thus, a line carefully terminated to an antenna will give rise to pronounced standing waves during wet weather and during sleet storms. Since both water and ice are poor conductors the currents are confined to the metallic surface of the wires. In other words, the inductance per unit length of conductor is not appreciably changed. On the other hand, the capacity per unit length is increased. This is true for a film of water on the conductor because the dielectric constant of water is large. The dielectric constant of ice is small at radio frequencies. The radial thickness of ice deposited during a sleet storm, however, may be a large fraction of an inch.

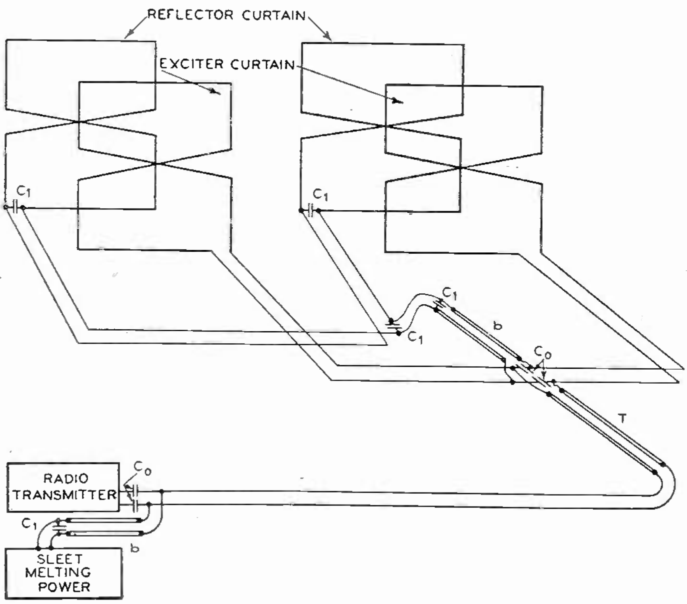

Fig. 17—The figure is a diagram of the sleet melting circuits incorporated in the Lawrenceville antenna system. The condensers are low impedance paths for radio-frequency currents and high impedance paths for the sleet melting currents. The bars marked (*b*) are high impedance paths for radio-frequency currents and low impedance paths for the sleet melting currents. The bars marked (*T*) are impedance transformers.

The effect of films of water upon the wires may be reduced by employing conductors of large diameter. It is, however, often desirable to remove the sleet from the wires because of the dangerous stresses which may be set up in the conductors and in the supporting structure. Since sleet is most readily removed by heating the wires with electric currents a compromise in conductor size is usually necessary.

In order to remove the sleet deposit by heating the wires with electric currents it is desirable that the elements comprising the antenna appear as closed loops to a source of sleet melting power. It is particularly desirable to construct the antenna elements and all lines including those to the transmitter from the same size conductor. All elements and lines may then be placed in series for the sleet melting currents and the entire circuit cleared of sleet in one operation. Circuits considerably greater than unity which is the usual case with a tuned primary. It will, however, be more convenient for the purpose desired here to derive the equations directly.

The following equation derived in Appendix I is obtained for I_2.

$$I_2 = \frac{\mu\, e_g\, j\omega M}{R_2(b + ja) + \omega^2 M^2(1 + j\omega C_0 r_p) + X_2(-a + jb)} \tag{4}$$

where the values of a and b are given in Appendix I.

The values of reactances which will make I_2 a maximum are also derived in Appendix I, and are given in (8), (9), (10), and (11), Appendix I.

The Voltage Amplification

The secondary voltage is approximately

$$e_g{}^2 = \frac{I_2}{j\omega C_2} = -j\omega L_2 I_2 \tag{5}$$

The case of most practical interest will be the voltage amplification $e_g{}^2/e_g{}^1$ when X_2 is varied to make I_2 a maximum. Substituting the value of X_2 given by (9) Appendix I into (4), multiplying by $-j\omega L_2$ and simplifying, we obtain

$$Av = \frac{-\mu\omega M L_2}{\dfrac{\omega^2 M^2(r_p + R_1 + \omega^2 C_0{}^2 r_p{}^2 R_1)}{\sqrt{a^2+b^2}} + R_2\sqrt{a^2+b^2}} \; \frac{a+jb}{\sqrt{a^2+b^2}} \tag{6}$$

$$Av = \frac{-\mu\omega M\omega L_2}{1 + jr_p\omega C_0} \; \frac{1}{\dfrac{w^2M^2R}{Z} + ZR_2} \; \frac{X+jR}{Z} \tag{7}$$

(7) is obtained from (6) by making the following substitution,

$$X + jR = \frac{(a+jb)(1+jr_p\omega C_0)}{1 + r_p{}^2\omega^2 C_0{}^2} \tag{8}$$

where,

$$R = R_1 + \frac{r_p}{1 + \omega^2 C_0{}^2 r_p{}^2}$$

$$X = \omega L_1 - \frac{r_p{}^2\omega C_0}{1 + r_p{}^2\omega^2 C_0{}^2}$$

and,

$$Z = \sqrt{R^2 + X^2}.$$

If the primary is untuned and if $r_p\ (1-\omega^2C_0L_1)$ is large compared with ωL_1, $\omega C_0 r_p R_1$ and R_1, the usual conditions, (6) becomes

$$Av = \frac{\mu\omega M\omega L_2(1-\omega^2C_0L_1)}{\omega^2M^2(1+R_1r_p\omega^2C_0^2)+R_2r_p(1-\omega^2L_1C_0)^2}. \tag{9}$$

If $r_p\omega C_0$ is considerably greater than unity and if the primary is tuned x and x_2 may be considered equal to zero for approximate results. (7) then becomes, if ωL equals $1/\omega C_0$

$$Av = \frac{\mu\omega M\omega^2L^2}{r_p\left[\omega^2M^2+R_2\left(R_1+\dfrac{\omega^2L^2}{r_p}\right)\right]}. \tag{10}$$

If ωM is also adjusted to make I_2 a maximum, the maximum possible value of amplification for a given set of circuit conditions is obtained from (7) by substituting R_2Z^2/R for ω^2M^2.

$$Av\ \max = \frac{1}{2}\,\frac{\mu}{1+jr_p\omega C_0}\cdot\frac{\omega L}{\sqrt{R_2R}}\cdot\frac{x+jR}{Z}. \tag{11}$$

If $r_p\omega C_0$ is small compared with one (11) reduces to the usual equation

$$Av\ \max = \frac{1}{2}\,\frac{\mu}{\sqrt{r_p}}\,\frac{\omega L}{\sqrt{R_2}}. \tag{12}$$

If $r_p\omega C_0$ is large compared with unity (11) becomes

$$Av\ \max = \frac{1}{2}\,\frac{\mu}{\sqrt{r_p}}\,\frac{\omega L_2}{\sqrt{R_2(1+R_1r_p\omega^2C_0^2)}}. \tag{13}$$

If the primary is tuned so that $1/\omega C_0$ equals ωL_1 (13) may also be written

$$Av\ \max = \frac{1}{2}\,\frac{\mu}{r_p}\,\frac{\omega L_2\omega L_1}{\sqrt{R_2\left(R_1+\dfrac{\omega^2L_1^2}{r_p}\right)}}. \tag{14}$$

It is thus seen that the presence of the output capacity reduces the maximum possible amplification. In an untuned circuit $R_1r_p\omega^2C_0^2$ is usually small so the amplification is not reduced very much. In a tuned primary $R_1r_p\omega^2C_0^2$ will be greater than for an untuned primary so that tuning the primary reduces the value of the maximum possible value of amplification. For example, if $R_1r_p\omega^2C_0^2$ equals one, the value of (13) will only be $1/\sqrt{2}$ as much for (12).

waves upon the line from the sending end of the bars toward the power source by means of a portable current indicating device as a check upon the termination. In fact, when some experience has been gained, it is possible by observing only the standing waves on the line to set the current minima and the bar spacing with a few cut-and-try runs.

In a large array where a number of terminated lines are brought to a common junction the leads between the junction and the transformer bars introduce sufficient reactance to preclude a good impedance match. In this case it is desirable to connect the junction and the bars with a one-half wavelength of line, the equivalent of a one-to-one transformer. This loop is also a convenient adjustment for bringing a current node or antinode to the bar terminals, particularly, if the impedance matches ahead of the junction are poor.

The final local test on the antenna is a gain frequency measurement with respect to a comparison antenna to be made at a distance of two or three miles in front of the system. This distance should be sufficiently great so that the difference in attenuation to the point of observation from the directive array and comparison antenna is small.

It is often difficult to obtain a clean-cut directional pattern about an antenna owing to the scattering effect of local obstructions and reradiation from conductors in the vicinity. The use of a comparison antenna is often helpful for the purpose of reducing irregularities in the diagram.

The writer wishes to express his indebtedness to Dr. T. C. Fry of the Bell Telephone Laboratories staff for considerable assistance in performing the integrations indicated in this paper and also to Mr. J. C. Schelleng for valuable advice and independent checks upon the results so obtained.

Appendix I

Gain Equations

The improvement factor for a directive array is by definition the ratio of power (P_0) radiated by one isolated element to the power (P_d) radiated by a number of elements for the case of equal signal intensities in the preferred direction. In the case of complete reinforcement this implies that the total current in the array and in the comparison element be equal.

Employing Poynting's theorem the power radiated by an antenna may be computed from the integral:

$$P = \frac{2c}{4\pi} r_0^2 \int_0^{\pi} \partial\theta \int_0^{\pi/2} E^2 \cos\alpha\partial\alpha \tag{11}$$

in which (E) is the intensity at any remote point defined by the distance (r_0) and the angles (θ) and (α). The meaning of the angles may be found by referring to Fig. 19. If for (E) is written the field about a simple antenna, the integral yields the familiar expression for the power radiated by a simple antenna. The integral is shown evaluated for a hemisphere because a perfect metallic reflector is assumed. For the case of short elements far removed from reflecting surfaces the perfect reflector may be viewed as a plane of symmetry.

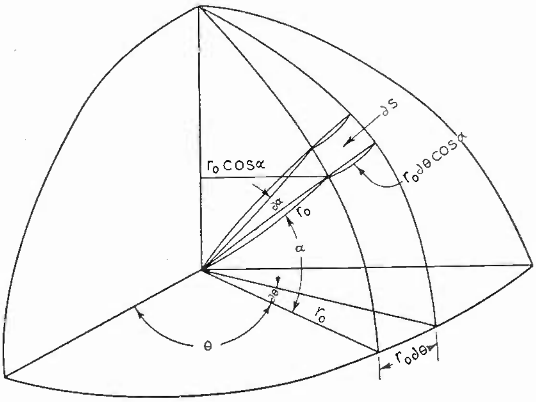

Fig. 19

It is very convenient to assume unit current in the single element and in each of the elements of the directive array for the purpose of computing (E). Upon this basis the improvement factor becomes by definition:

$$\rho = \frac{N^2 P_0}{P_d} \tag{12}$$

in which (N) is the number of radiating elements in the directive array.

The improvement factor for a row of short grounded elements, or for a row of elements far removed from reflecting surfaces expressed in decibels is:

$$db = 10 \log_{10} \rho_1 \tag{13}$$

in which,

$$\frac{1}{\rho_1} = \frac{1}{n} + \frac{3}{n^2} \cdot \sum_{k=1}^{n=1} (n - k) \cos kB \text{ times}$$

$$\left(\frac{\sin ks}{ks} + \frac{\cos ks}{(ks)^2} - \frac{\sin ks}{(ks)^3} \right) \tag{13a}$$

where,

n = number of elements in a row

$s = \dfrac{2\pi a}{\lambda}$, the uniform element spacing,

$B = \dfrac{2\pi b}{\lambda}$, the uniform current phasing.

The improvement factor for a rectangular array of grounded elements, or an array far removed from reflecting surfaces is:

$$db = 10 \log_{10} \rho_2 \tag{14}$$

$$\frac{1}{\rho_2} = \frac{1}{q}\left[\frac{1}{n} + \frac{3}{n^2}\sum_{k=1}^{n-1}(n-k)\cos kB \text{ times} \right.$$

$$\left.\left(\frac{\sin ks}{(ks)} + \frac{\cos ks}{(ks)^2} - \frac{\sin ks}{(ks)^3}\right)\right]$$

$$+ \frac{3}{nq^2}\sum_{l=1}^{q-1}(q-l)\cos lZ \text{ times}$$

$$\left(\frac{\sin lu}{(lu)} + \frac{\cos lu}{(lu)^2} - \frac{\sin lu}{(lu)^3}\right)$$

$$+ \frac{6}{(nq)^2}\sum_{l=1}^{q-1}\sum_{k=1}^{n-1}(q-l)(n-k)\cos lZ \cos kB \text{ times}$$

$$\left(\frac{\sin\sqrt{k^2s^2+l^2u^2}}{\sqrt{k^2s^2+l^2u^2}} + \frac{\cos\sqrt{k^2s^2+l^2u^2}}{(k^2s^2+l^2u^2)} - \frac{\sin\sqrt{k^2s^2+l^2u^2}}{(k^2s^2+l^2u^2)^{3/2}}\right) \tag{14a}$$

in which,

q = the number of rows of (n) elements each,

$u = \dfrac{2\pi y}{\lambda}$, uniform row spacing,

$Z = \dfrac{2\pi z}{\lambda}$, uniform row phasing,

and the notation of the previous equation.

It is of interest to note that for a two-row array and odd quarter phasing ($Z = \pi/2$) between rows the expression reduces to the improvement for one row multiplied by a factor of two, that is, an improvement of 3.0 db over one row.

The improvement factor for a row of elements of which the midpoint height is (d/λ) may be obtained from:

$$db = 10 \log_{10} \frac{\rho_0}{\rho_3} \qquad (15)$$

in which,

$$\rho_0 = 1 - 3\left[\frac{\cos v}{v^2} - \frac{\sin v}{v^3}\right] \qquad (15a)$$

$$\rho_3 = \left[\frac{1}{n} + \frac{3}{n^2} \sum_{k=1}^{n-1}(n-k)\cos kB \text{ times}\right.$$
$$\left.\left(\frac{\sin(ks)}{(ks)} + \frac{\cos(ks)}{(ks)^2} - \frac{\sin(ks)}{(ks)^3}\right)\right]$$

$$- \frac{3}{n}\left[\frac{\cos v}{v^2} - \frac{\sin v}{v^3}\right]$$

$$+ \frac{3}{n^2} \sum_{k=1}^{n-1}(n-k)\cos kB \cdot \frac{2v^2}{k^2s^2 + v^2} \cdot \frac{\sin\sqrt{k^2s^2 + v^2}}{\sqrt{k^2s^2 + v^2}}$$

$$+ \frac{3}{n^2} \sum_{k=1}^{n-1}(n-k)\cos kB \cdot \frac{k^2s^2 - 2v^2}{k^2s^2 + v^2} \text{ times}$$

$$\left(\frac{\sin\sqrt{k^2s^2 + v^2}}{(k^2s^2 + v^2)^{1/2}} + \frac{\cos\sqrt{k^2s^2 + v^2}}{(k^2s^2 + v^2)} - \frac{\sin\sqrt{k^2s^2 + v^2}}{(k^2s^2 + v^2)^{3/2}}\right) \qquad (15b)$$

and,

$$v = \frac{4\pi d}{\lambda}. \qquad (15c)$$

Note that if the gain over a standard comparison antenna is required*

$$P_0 = 1 + \frac{3}{\pi^2} \qquad (16)$$

and that the gain of a comparison antenna height (d/λ) above ground over a standard comparison antenna is

$$db = 10 \log_{10} \frac{1 + \dfrac{3}{\pi^2}}{1 - 3\left[\dfrac{\cos v}{v^2} - \dfrac{\sin v}{v^3}\right]}. \qquad (17)$$

This is the factor plotted in Fig. (12).

* Ideal case of perfect metallic reflector.

The improvement factor for a colinear array of (p) in-phase elements far removed from reflecting surfaces is:

$$db = 10 \log_{10} \rho_4 \tag{18}$$

$$\frac{1}{\rho_4} = \frac{1}{p} - \frac{6}{p^2} \sum_{j=1}^{p-1} (p - j) \left(\frac{\cos (jw)}{(jw)^2} - \frac{\sin (jw)}{(jw)^2} \right) \tag{18a}$$

in which,

$$w = \frac{2\pi g}{\lambda} \text{ is the midpoint spacing term.}$$

The improvement factor for a unidirectional two-row broadside system (p) tiers high far removed from reflecting surfaces is:

$$db = 10 \log_{10} \rho_5 \tag{19}$$

where,

$$\frac{1}{\rho_5} = \frac{1}{2p} \left[\frac{1}{n} + \frac{3}{n^2} \sum_{k=1}^{n-1} (n - k) \left(\frac{\sin (ks)}{(ks)} + \frac{\cos (ks)}{(ks)^2} - \frac{\sin (ks)}{(ks)^3} \right) \right]$$

$$+ \frac{3}{np^2} \sum_{j=1}^{p-1} (p - j) \left[\frac{\sin (jw)}{(jw)^3} - \frac{\cos (jw)}{(jw)^2} \right]$$

$$+ \frac{3}{n^2p^2} \sum_{k=1}^{n-1} \sum_{j=1}^{p-1} (n - k)(p - j) \frac{2(jw)^2}{(ks)^2 + (jw)^2} \frac{\sin \sqrt{(ks)^2 + (jw)^2}}{\sqrt{(ks)^2 + (jw)^2}}$$

$$+ \frac{3}{n^2p^2} \sum_{k=1}^{n-1} \sum_{j=1}^{p-1} (n - k)(p - j) \frac{(ks)^2 - 2(jw)^2}{(ks)^2 + (jw)^2} \text{ times}$$

$$\left(\frac{\sin \sqrt{(ks)^2 + (jw)^2}}{\sqrt{(ks)^2 + (jw)^2}} + \frac{\cos \sqrt{(ks)^2 + (jw)^2}}{(ks)^2 + (jw)^2} \right.$$

$$\left. - \frac{\sin \sqrt{(ks)^2 + (jw)^2}}{((ks)^2 + (jw)^2)^{3/2}} \right). \tag{19a}$$

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MEASUREMENTS ON MAGNETOSTRICTION VIBRATORS*

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THE study of electromechanical vibrating systems through the measurement of motional impedance was first undertaken by Kennelly and Pierce[1] in 1912, in their work on telephone receivers. The method was applied by K. C. Black[2] in 1928 to the study of ferromagnetic rods centrally clamped and lying along the axis of a coil carrying alternating current superposed upon direct current. Such a rod will vibrate mechanically when the axial magnetic field alternates at frequencies close to the elastic natural frequency of the rod. A mathematical analysis of a rod vibrating under the influence of magnetostriction has been given by G. W. Pierce.[3] In the same paper he describes the application of such vibrators to the frequency control of vacuum tube oscillators, in a manner analogous to the use of vibrating quartz crystals.

The work of K. C. Black was done on rods of nickel, stainless steel, and stoic metal (invar steel), and also on nickel tubes. The natural frequencies of the rods he studied lay between 2300 and 3000 cycles per second. The present paper describes similar measurements carried out on rods with natural frequencies near 30,000 cycles per second. The study was made on rods of nichrome, monel metal, stainless steel, stoic metal, and nickel. The dependence of the vibratory characteristics on rod diameter was determined by making measurements on a nickel rod, as its diameter was reduced in successive steps. The rod was annealed before each set of measurements, by heating to redness and cooling in air. Another nickel rod cut from the same bar was bored axially with concentric holes of successively larger diameter, and measurements made at each step. This was done in an endeavor to determine something about the flux distribution across the section of the vibrating rod. The effects of annealing, and of changing the number of turns on the alternating-current coil (in which the rod was supported), were investigated. The effect on the vibratory characteristics of varying the alternating magnetic field on the rod was also measured.

* Decimal classification: 538.11. Original manuscript received by the Institute, March 13, 1931.

[1] Kennelly and Pierce, *Elec. World*, September 14, 1912.

[2] K. C. Black, *Proc. Am. Acad.*, **63**, 49, 1928.

[3] G. W. Pierce, "Magnetostriction oscillators," *Proc. Am. Acad.*, **63**, 1, 1928; Proc. I. R. E., **17**, 42–88; January, 1928.

Summary of Motional Impedance Theory

Let a rod of magnetostrictive material be centrally supported in a coil, and let the coil be placed in the unknown arm of an alternating-current bridge. If we then measure its resistance and reactance as a function of frequency, over a frequency range including the natural frequency of the rod, we obtain curves similar to those of Fig. 1. The

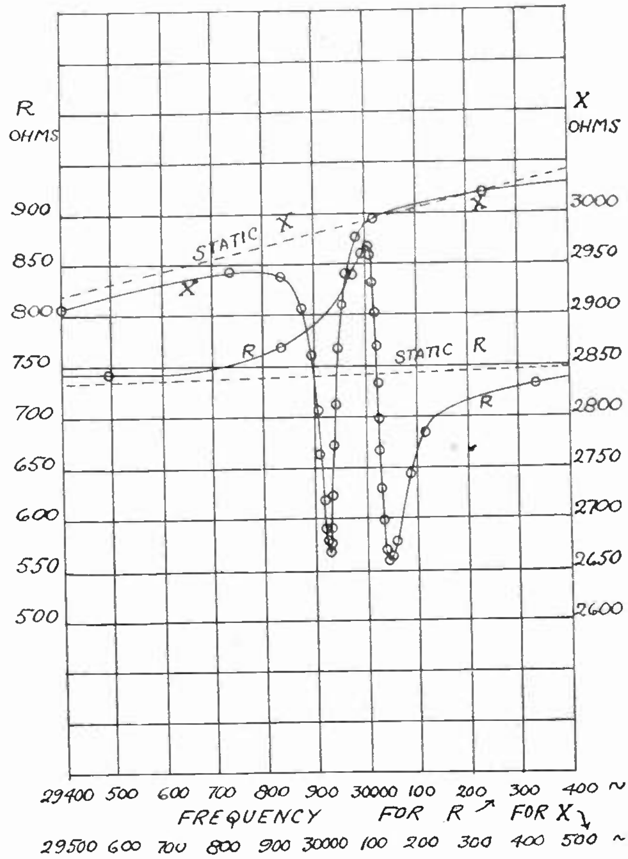

Fig. 1—Resistance R and reactance X plotted against frequency for a nickel rod vibrator. To avoid confusion, the frequency scales of the curves are displaced 100 cycles.

impedance so obtained may be considered as divided into two types: the volts per ampere in the coil due to the resistance and reactance of the rod-coil system, regarded as rigidly clamped, and the volts per ampere in the coil due to the longitudinal vibration of the bar under the action of the periodic mechanical force of magnetostriction. We shall refer to the former as the static impedance, and to the latter as the motional impedance.

The static impedance of such rod-coil systems has been investigated theoretically and experimentally by numerous observers.[4] It has been

[4] Lange and Myers, Proc. I.R.E., 17, 1687, 1929; K. L. Scott, Proc. I.R.E., 18, 1750, 1930.

found to increase uniformly with the square root of the frequency, on account of hysteresis and magnetic skin effect in the bar.

The motional impedance, in which we are here interested, can be separated from the static impedance because the whole motional effect is limited to a small frequency range above and below resonance. Over this range—about a thousand cycles per second for most of the rods measured—the variation of the static impedance is of the order of 1.6 per cent for the cases studied here. We can therefore legitimately represent the static resistance and reactance in the curves of Fig. 1 by

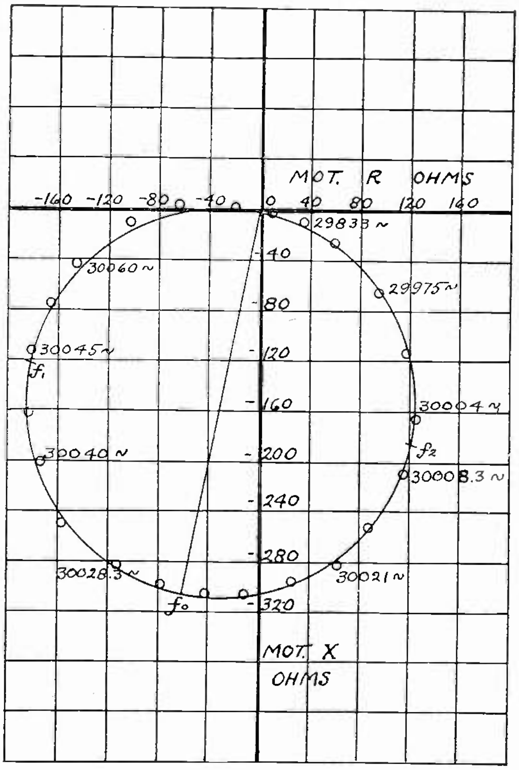

Fig. 2—Motional impedance circle diagram obtained from the curves of Fig. 1.

straight lines (dotted in the figure) asymptotic to the observational curves at frequencies remote from resonance. The motional resistance and reactance for any frequency may then be obtained by subtracting the static resistance and reactance from the total observed resistance and reactance. If we then plot motional reactance against motional resistance, the locus of points for different frequencies will be found to be a circle. (See Fig. 2.)

From the circle diagram thus obtained, it is possible to obtain four quantities which are characteristic of the vibrating system. These are (1) the maximum motional impedance, (2) the resonant frequency of the rod, (3) the decrement of the mechanical vibration, and (4) the angle by which the circle diameter is depressed below the resistance axis. Of these quantities the resonant frequency is the only one which can be obtained more easily by another method.[5]

The maximum motional impedance is the diameter of the motional impedance circle, measured in ohms. It can be shown to be proportional to the velocity of vibration of the ends of the rod. The velocity of vibration is the time derivative of the amplitude of vibration, and hence proportional to it when the time is involved in a factor $e^{j\omega t}$.

The resonant frequency of the rod is the frequency corresponding to maximum motional impedance, or, on the circle diagram, it is the frequency corresponding to the point where the longest radius vector intercepts the circle.

The decrement, Δ, of the mechanical vibration can be obtained from the circle diagram by the relation[6]

$$\Delta = \pi(f_1 - f_2)$$

where f_1 and f_2 are the frequencies corresponding to the two values of motional impedance, one on each side of resonance, at which it has $1/\sqrt{2}$ times its maximum value, the value at resonance.

The fourth quantity obtainable from the circle diagram is the angle through which the diameter through the origin is depressed below the motional resistance axis. The physical significance of this dip angle has not been completely worked out. Kennelly and Pierce, in their original work on telephone receivers, showed that the dip angle was equal to twice the angle of lag of the flux through the diaphragm behind the coil current which induced it. This lag is apparently due to hysteresis and magnetic skin effect in the case of vibrating rods.

Method of Measurement

The method of measurement consisted in placing the coil containing the rod in series with a General Radio Type 222 precision air condenser, as the unknown arm of a resonance bridge. The bridge and its accessory apparatus was carefully shielded, so that no difficulty was experienced in obtaining a sharp balance, at frequencies in the neighborhood of 30,000. The bridge current was furnished by a Western Electric Type 8A variable oscillator, provided with an additional vernier air condenser for fine adjustment of frequency. An amplifier with heterodyne detector was used to determine the condition for bridge balance.

[5] D. S. Muzzey, *Phys. Rev.*, **36**, 935, 1930.
[6] A. E. Kennelly, Electrical Vibration Instruments, Chap. IX, Macmillan, 1923.

The rod was supported in the test coil by means of a pin set transversely through the midpoint of coil and rod. This was found by trial to be the only scheme which did not allow the rod to shake off the support, and did not damp the vibrations. The direct-current magnetic field on the rod was furnished by a 50,000-turn electromagnet with iron

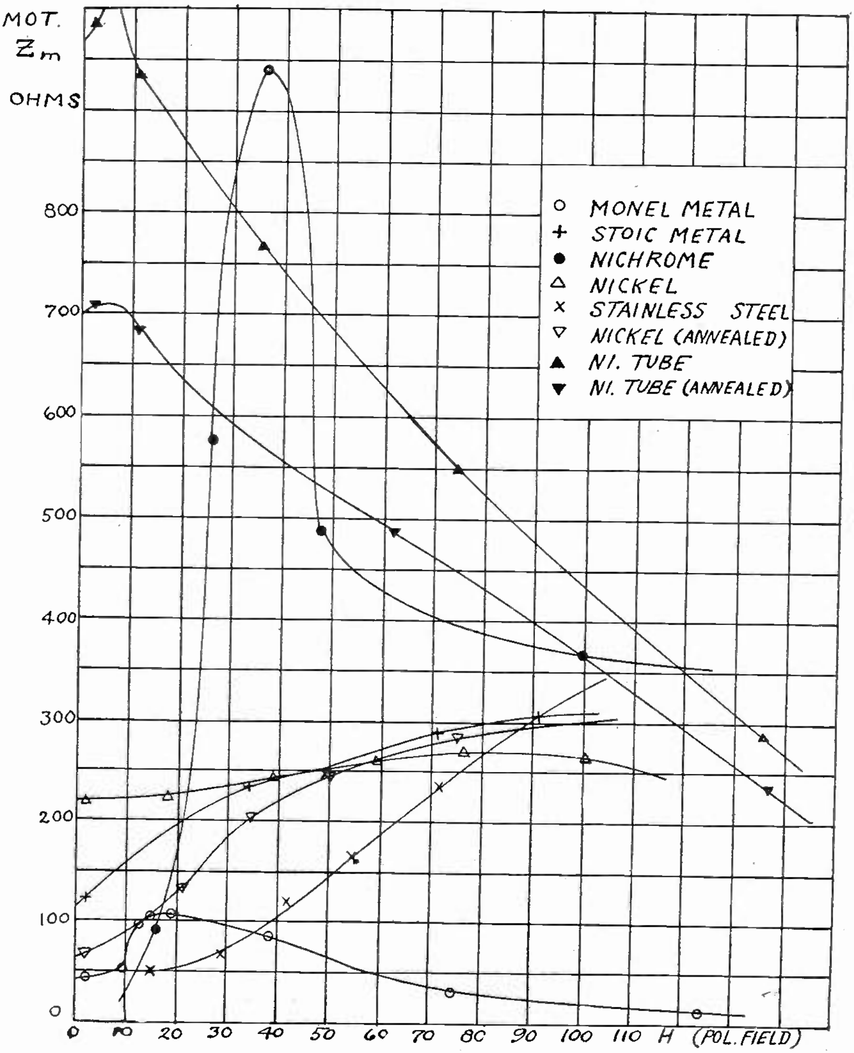

Fig. 3—Maximum motional impedance plotted against polarizing field for eight vibrators.

core. An 11,000-turn choke coil was found adequate to keep the 30,000-cycle current from flowing in the direct-current magnetizing circuit. The magnetic field corresponding to various values of current in the magnetizing coils was measured directly by the ballistic method. Due corrections were made for the demagnetizing effect of the free poles formed at the ends of the test rod.

The frequency of the bridge current was measured by producing an audio-frequency beat note between the third harmonic of the current produced by the oscillator, and the plate current of a magnetostriction oscillator. The frequency of the latter was held fixed by standard rods, at various values which gave convenient beat frequencies. The rods

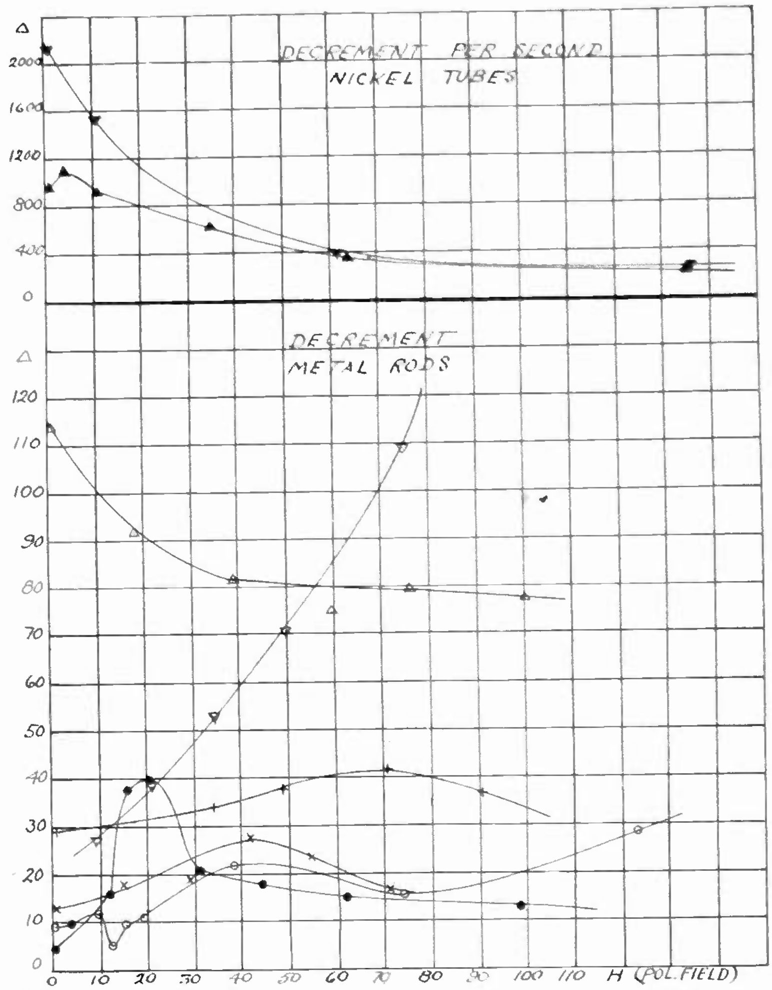

Fig. 4—Mechanical decrement per second plotted against polarizing field for the vibrators of Fig. 3.

were calibrated with reference to a crystal, which was itself known very accurately in terms of the standard second. The beat frequencies so obtained were measured by means of a calibrated audio-frequency meter, similar to the one described by Pierce in the paper previously referred to. With this arrangement it was possible to measure relative frequencies to an accuracy of one part in 100,000, or 0.001 per cent. The absolute accuracy (compared with the standard second) was about one part in 10,000. High accuracy of frequency determination was necessary since in some cases half the circle diagram was traversed in less than five cycles, at 30,000.

Discussion of Results

In this research 101 circle diagrams were obtained, for eighteen different rod specimens. For each specimen several diagrams were obtained, corresponding to various values of magnetic polarization. The results enable us to discuss the effect of polarizing field, diameter of rod, thickness of wall (for tubes), heat treatment, alternating field, and inductance of exciting coil, on the vibratory characteristics of magnetostrictive resonators.

Effect of Polarizing Field

In Figs. 3 and 4 we have plotted maximum motional impedance and decrement per second, as a function of magnetic polarizing field, for eight different vibrators. The observed effects are undoubtedly due to variations in the elastic constants and magnetostriction constants of the metals, as the polarizing field is changed. The behavior of these constants cannot be expressed analytically, so that it is of little help to attempt explanations in terms of them. Our results are thus best expressed in graphical form.

On comparing the motional impedance curves with normal induction curves (not reproduced here, except for monel metal) for the same vibrators, it was evident that in every case where a sharp maximum appears in the curve of maximum motional impedance against polarizing field, it occurs at the value of H which corresponds to the upper bend of the magnetization curve. This is illustrated by Fig. 5(a) and (b) for monel metal.

Our results for the different metals show also that, in general, the resonant frequencies of the vibrators increase with increasing polarizing field. Monel metal is an exception to this, and its resonant frequency (see Fig. 5(c)) is independent of polarizing field except in the vicinity of the upper bend of the magnetization curve. In that region the curve of resonant frequency against polarizing field shows a curious dip, or "wobble." This irregularity is apparently not due to experimental error although its total variation of frequency is eight cycles per second out of 30,000. The resonant frequency of the stainless steel vibrator increased with polarizing field for fields less than 40 gauss, and then began to decrease. With the exceptions just discussed, every specimen studied in this research showed an increase in resonant frequency with increasing polarizing field. This includes the nickel rods of various diameters, and nickel tubes of various thicknesses.

Another conclusion which may be drawn from our results, is that the variation of decrement with polarizing field is related to the variation of resonant frequency. For most of the vibrators measured here the decrement increases with increasing polarizing field, and does so with the same shape of curve as is found for the resonant frequency. The percentage variation of decrement is, however, very much larger

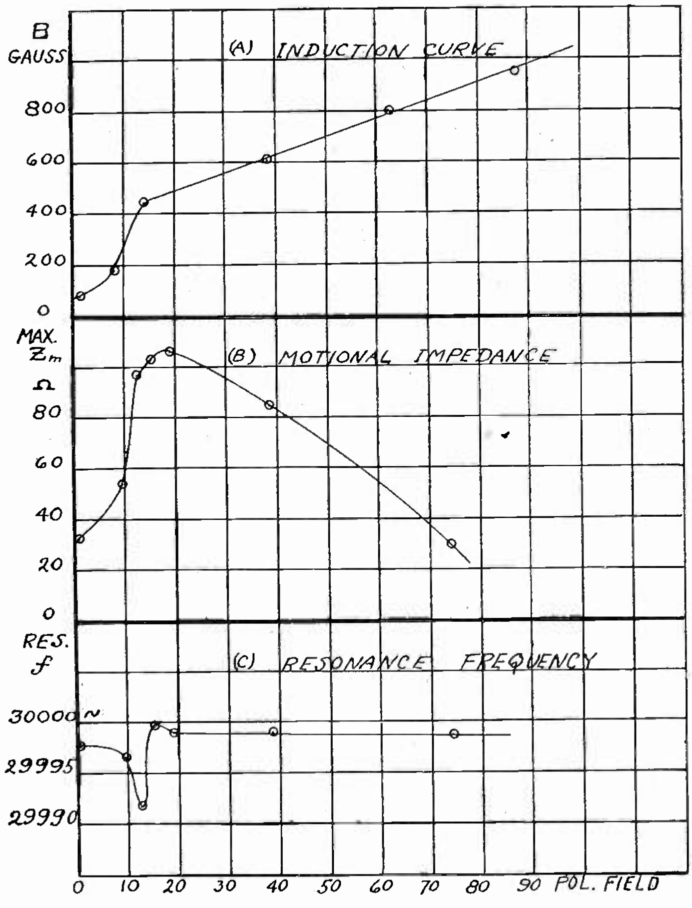

Fig. 5—Typical curves for monel metal, showing relation of magnetization curve to vibratory characteristics.

than the corresponding percentage variation of frequency. For the nichrome and thin nickel tube vibrators the decrement decreases when the resonant frequency increases, and vice versa.

For nichrome, monel metal, stainless steel, and thin nickel tube, the dip angle (twice the angle of lag of rod flux behind the coil current which induces it) decreases as the polarizing field increases. For stoic metal there is very little variation in dip angle, and for nickel rod there is an increase with polarizing field.

Comparison of Metals Studied

For a general comparison of the different vibrators studied in this research, we have compiled a table of their vibratory characteristics. The values of decrement and of maximum motional impedance recorded in the table, are in each case rough averages over a range of polarizing fields from zero to 100 gauss. The complete curves are given in Figs. 3 and 4. The table is merely for rough comparison. Note that a large motional impedance corresponds to a large reaction of the rod vibration on the electrical circuit in which it is placed. A small decrement corresponds to extreme sharpness of resonance of the rod, and hence of the circuit which it controls.

The column marked "Range of Dip Angle" gives the variation in dip angle as the polarizing field varies from zero to 100 gauss. The work of K. C. Black, previously referred to, indicates that there is little or no variation of dip angle with polarizing field, at frequencies near 3000 cycles per second. He also found that the dip angle was close to 90 degrees for all the specimens he studied except the nickel tubes. For them the dip angle was much less. We find that there is a large variation of dip angle with polarizing field, at frequencies near 30,000 cycles per second, for nickel tube, monel metal, nichrome, and stainless steel. A smaller, but still measureable, variation appears in stoic metal, and the nickel rods with various dimensions and heat treatments. Stainless steel shows by far the largest dip angle, for all fields, of any of the metals studied. The thin walled nickel tubes showed the smallest dip angles, as Black found.

For the application of magnetostriction vibrators to the control of frequency in vacuum tube oscillators, it is important to know how much the natural frequency of any given rod is likely to vary with variation of polarizing field. For convenient comparison of the various metals from this standpoint we supply the last column of the table. This tabulates the percentage increase in natural frequency for an increase of polarizing field from zero to 100 gauss.

Effect of Heat Treatment

Owing to limited time and furnace equipment, we were not able to make an extended study of the effects of various heat treatments on the vibratory characteristics of magnetostrictive rods. However, for a thin walled nickel tube and for a nickel rod, we secured data to show the effect of heating the specimen to redness and allowing it to cool in air. The original treatment of the tube is not known, but probably consisted of some type of cold working. The nickel rod was stated by the manufacturer to have been "cold rolled."

The effect of this partial annealing on the thin nickel tube was to reduce the maximum motional impedance (20 per cent to 50 per cent), the resonant frequency (1.7 per cent), the dip angle (5 degrees to 10 degrees), and nearly to double the decrement at small polarizing fields.

The effect of annealing the nickel rod was to reduce the maximum motional impedance (about 100 per cent), the resonant frequency, and the decrement at small fields (from 100 to 20.). The dip angle was increased 15 degrees. For both the rod and the tube, the percentage increase of frequency over the available range of polarizing fields was nearly tripled by annealing. (See the table.)

From these results, we should be inclined to believe that annealing is of doubtful value, as a means of increasing the reaction of a nickel rod employed in frequency stabilization. The advantage, if any, would be in reduction of the decrement. It may be, however, that the effect of annealing is very different in the case of other magnetostrictive materials. It would be interesting to pursue this further.

Effect of Alternating-Current Coil

In the case of nichrome, we tried the effect on the vibratory characteristics of changing the inductance of the alternating-current coil. Circles were taken first with a coil of 1000 turns, and then with a coil of 2000 turns.

We find that doubling the number of turns on the coil approximately doubles the maximum motional impedance, except at small polarizing fields, where the situation is reversed. The other characteristics are slightly increased, but the variation is so slight as to be within the probable error.

Effect of Alternating Field

Six circles, two on stainless steel, and four on the thin nickel tube, were taken to illustrate the effect of varying the alternating field, with the polarizing field kept fixed.

It was found that, for the rod, a variation of 100 per cent in alternating field produced a variation of 1 per cent or less in maximum motional impedance, depression angle, and decrement per second. The variation in resonant frequency was less than the experimental error. In the case of the nickel tube, these variations were about twice as large. For rods of material other than stainless steel the variations of vibratory characteristics with alternating field were no greater, and in most cases less. In all cases an increase in alternating field was found to decrease the maximum motional impedance. K. C. Black found this to be true also at the lower frequencies at which he worked.

Effect of Variations in Rod Diameter

In order to determine the variations in vibratory characteristics as a function of rod diameter, a nickel rod (No. 1) with a diameter which nearly filled the alternating-current coil, was turned down on a lathe to successively smaller diameters. At each diameter circle diagrams were taken, one for each of several polarizing fields. Nickel

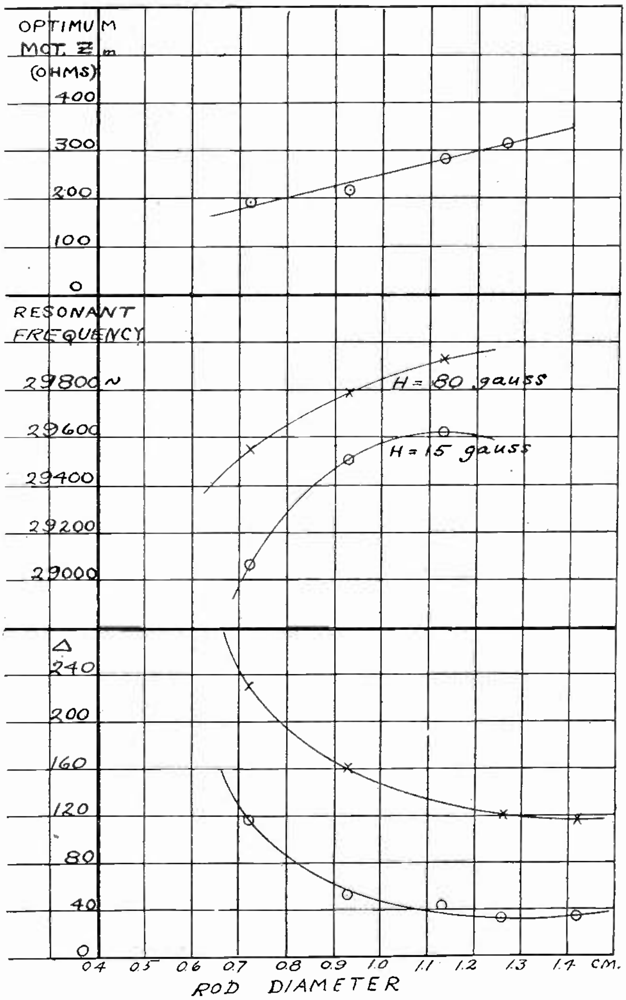

Fig. 6—Variation of vibratory characteristics with diameter of rod.

was chosen because its composition and homogeneity were known with more certainty than for the alloys, and its various magnetic characteristics have been determined by a number of observers.[7]

One new conclusion which may be drawn from the evidence of these nickel rods, is that the variation of resonant frequency with polarizing

[7] For example, K. Honda, Magnetic Properties of Matter, Syokwabo and Co., Tokyo, 1928.

field greatly increases as the diameter of the rod is decreased. This is of importance in applications of magnetostriction to frequency stabilization. We tabulate here the percentage increase of resonant frequency as the polarizing field is increased from zero to 100 gauss, for various values of rod diameter:

Rod Diameter.	Percentage Frequency Increase.
1.415 cm	0.767 per cent
1.266	0.929
1.127	1.10
0.924	1.11
0.739	2.12

In Fig. 6 we have plotted graphs showing the manner in which the largest maximum motional impedance (peak of the curve of maximum motional impedance against polarizing field), the resonant frequency, and the decrement per second vary as a function of rod diameter.

The graph of max., max. motional impedance is nearly a straight line over the range measured. In other words, the largest motional impedance obtainable—regardless of polarizing field—increases linearly with rod diameter.

The second graph illustrates the well-known relation between resonant frequency and rod diameter. This relation is given by Rayleigh's formula for the "free, free" vibration of a longitudinal rod, slightly modified by magnetostriction terms. Muzzey,[5] has given experimental confirmation of this formula. We must note, however, the very considerable influence exerted on this relationship by changes in polarizing field. This effect is much larger in nickel than in monel metal or stainless steel.

The variation of decrement with rod diameter is very striking. The evidence suggests that the decrement varies inversely as the square of the diameter. The decrement is larger at the high polarizing fields, but its variation with diameter is much the same for all polarizing fields.

Effect of Boring Axial Holes in Nickel Rod

In order to work out a mechanism for the vibrations of rods under the influence of magnetostriction, it is important to know whether the driving force is uniform over the rod section. The well-known theory of magnetic skin effect indicates that the alternating flux, which is responsible for the direct magnetostrictive effect, is essentially confined to a thin layer around the periphery of the rod cross section.

According to the theory of magnetic skin effect,[8] the amplitude of

[5] *Loc. cit.*
[8] See, for example, Jeans, Electricity and Magnetism.

the induction density in the rod diminishes to $1/\epsilon$ of its value at the surface, by the time it has penetrated to a depth

$$\delta = \frac{1}{2\pi}\sqrt{\frac{\rho}{\mu f}}$$

where,

ρ = the resistivity of the rod material, in e.m.u.
μ = the permeability of the rod to alternating flux.
f = the frequency of alternation of the flux.

If we compute δ for the materials and frequencies of this research, we find:

$$\text{For nickel: } \frac{0.081}{\sqrt{\mu}} \text{ at } f = 30{,}000 \text{ c.p.s.}$$

$$\text{For nichrome: } \delta = \frac{0.29}{\sqrt{\mu}} \text{ at } f = 30{,}000 \text{ c.p.s.}$$

For the other materials δ lies between these values. We do not know the permeability to alternating flux, for any of the above materials, but it was probably larger than 50 in every case. Thus the greatest effective depth of alternating flux penetration into any of our rods was less than 0.04 cm.

If this magnetic skin effect theory holds for vibrating rods, we must believe that the magnetostrictive forces responsible for the vibration are set up in a surface zone containing in the case of nickel rod (No. 1), for example, less than three per cent of the total cross section of the rod. It is indeed difficult to imagine how elastic deformations in such a thin surface layer could drive the whole rod into such relatively violent vibrations as are observed. We should note that the flux set up by the inverse magnetostrictive effect will also be driven to the surface of the rod by the shielding effect of the eddy currents induced by its variations. Hence we must conclude, either, that the magnetic skin effect theory does not hold in the form given above, when the rod is vibrating, or, that the vibratory motion is confined to the peripheral zone, while the central portion of the rod remains stationary. As we shall see, the latter supposition seems the least reasonable.

To obtain experimental evidence on this question of the vibration mechanism, a nickel rod (No. 2) was bored axially with successively larger concentric holes, and a series of circle diagrams taken for each of the thick walled tubes so obtained. The specimen was annealed before each series of measurements was made.

Some of the results of these measurements are plotted in Fig. 7. We there plot maximum motional impedance, resonant frequency, and decrement per second, against the ratio of the diameter of the axial hole to the diameter of the rod.

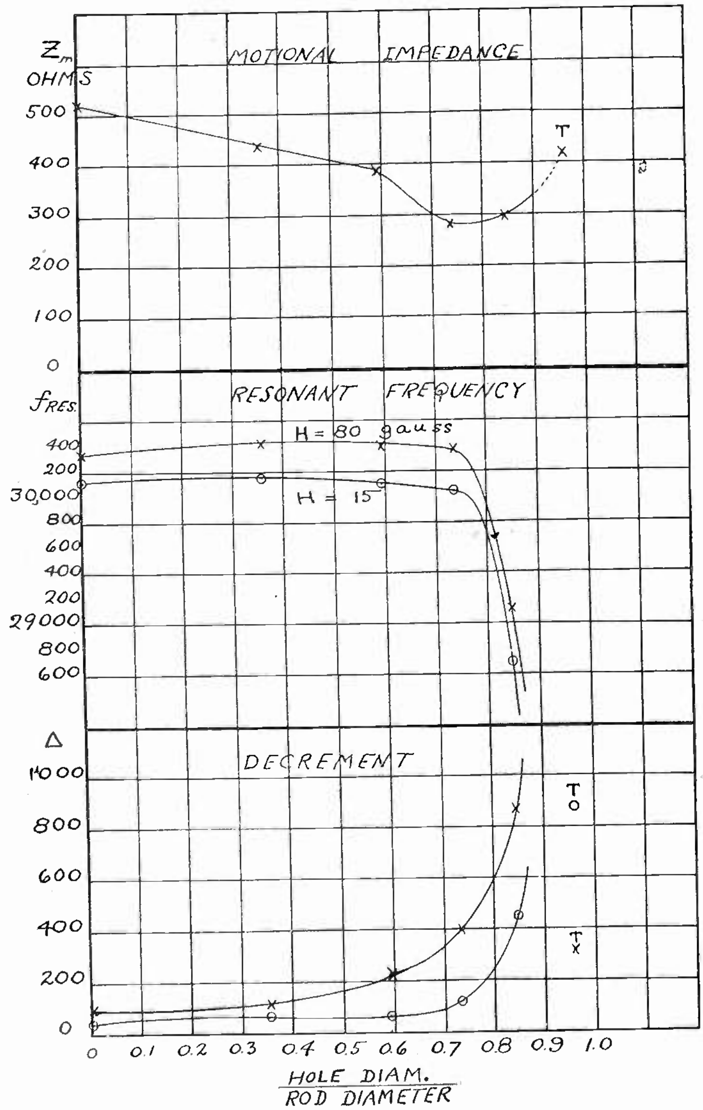

Fig. 7—Variation of vibratory characteristics as the interior of a nickel rod was bored out. The three points marked T were obtained for a very thin walled tube, and not by boring the rod.

We note first that the maximum motional impedance, particularly at large polarizing fields, shows a definite reduction as holes of successively larger diameter are bored in the specimen. This reduction continues until the hole diameter is three-quarters of the rod diameter. Beyond that point the motional impedance increases as the tube wall becomes thinner. It was not practicable to make this tube any thinner, but we found in previous measurements on very thin walled tubes, that they showed by far the largest maximum motional impedance of any of the specimens studied. Hence it is reasonable to suppose that our experimental curve would have continued to rise if we had been able to make further borings.

We notice next that the resonant frequency does not change more than about 0.35 per cent until the hole diameter becomes greater than three-quarters of the rod diameter. As the hole becomes larger than that the natural frequency drops rapidly as far as the measurements could be taken. As the diameter ratio increases from 0.75 to 0.85 the natural frequency decreases 4.3 per cent. We can definitely conclude from this that the frequency equation for vibrating rods does not hold also for tubes, and that the thinner the tube wall, the greater will be the departure of the natural frequency from that given by the theory for rods.

The decrement, or damping factor, shows an increase as larger holes are bored in the rod. This increase is small until the diameter ratio becomes larger than one-half. For larger holes than that the decrement increases exponentially, until it reaches the large values it has been observed to have in thin tubes.

Unfortunately, the experimental evidence described above does not give a definite answer to the question of the mechanism of magnetostrictive vibrations. Further work will be necessary before we can come to a conclusion on the problem, but we may make a few suggestions from our results.

If we take the view that the vibrations are confined to the surface of the rod, while the center remains stationary, it is difficult to see why the amplitude of vibration, and therefore the motional impedance, is as large as it is in solid rods of a substance such as nichrome. It is still more difficult to see why the motional impedance should decrease as the interior of the rod is bored out. This we have found to be the case.

Muzzey has proved that magnetostriction vibrators have the natural frequencies determined by the theory for solid bars. The surface vibration theory makes it necessary to believe that the peripheral zone vibrates with exactly the same natural frequency as the whole bar.

If the surface alone vibrates we should expect a large decrement for solid rods, due to the loading effect of the stationary central portion. We should expect the decrement to decrease on boring out the center. Experimentally we find exactly the opposite.

It is, therefore, unlikely that the vibration is confined to the surface of the rod. We ought to be able to obtain direct evidence on this point by placing a pick-up coil inside a fairly thick walled tube, and measuring the voltage induced in it when the tube is stationary and when it is vibrating. This was done, and it was found that a small, but definitely readable, voltage was induced in the pick-up coil, when, and only when the tube was vibrating. The tube used was of nickel, with a wall 0.11 cm thick. The effective flux penetration on skin effect theory should have been less than 0.01 cm in this case. We have here, then, direct experimental evidence that by some means or other the alternating flux does penetrate into the bar, while vibrating, much farther than magnetic skin effect theory will allow. Further details of this experiment cannot be given until much more work has been done, but we may suggest that a decrease of permeability to alternating flux, due in some way to the vibration of the rod, would account for the observed phenomena.

Engineering Application of These Results

It is very important to know the vibratory characteristics of the rods employed in circuits for the stabilization of frequency. We wish such rods to have small decrement (sharp resonance), large enough motional impedance to react effectively upon the oscillations set up in the circuit, and natural frequencies independent of temperature and intensity of magnetization. It is difficult, of course, to satisfy more than one of these conditions with any given rod. But in order to satisfy any of them, we must know the motional characteristics of rods of different materials. It is possible to guess these in a general way by noting the reaction of the rod vibration on the plate current of a magnetostriction oscillator in which the rod is placed. This method is not quantitative, but much can be learned from it.

For a quantitative comparison of rods we must make motional impedance measurements of the sort described here. We can then choose rods with characteristics as near as possible to the ideal requirements.

Specifically from our results we can say that a solid rod of as large diameter as will fit into the coil, will give vibrations of small decrement and large reactive effect,—in comparison with tubes or with rods of smaller diameter. The variation of natural frequency with polarizing field will also be smaller than in the case of tubes or smaller rods. However, better electrical constants for an oscillator may often be obtained by using a rod of small diameter compared with the coil cross section.

We can also say that monel metal is preferable among the materials studied for its very small decrement, and negligible variation of natural frequency with magnetic state. Its motional impedance is relatively small, but rods of large diameter should stabilize well, none the less. Its chief disadvantage is the necessity for some sort of polarizing arrangement, since its residual magnetism is small.

Stainless steel is recommended as possessing a small decrement, combined with a larger motional impedance than monel metal, and sufficient residual magnetism to make outside polarizing apparatus unnecessary. Additional polarization increases its motional impedance enormously, so that the stabilizing action can be improved by that means.

Nichrome of the composition studied here is not always commercially available, and nichrome of different composition may not vibrate at all. Hence we can recommend for frequency stabilizers only the composition of nichrome which was studied in this research. This was Ni 60.4 per cent, Fe 23.9 per cent, Cr 13.6 per cent, and small amounts of Mn, C, and Si.

The other materials studied have various disadvantages when used for frequency stabilization. Stoic metal possesses a large temperature coefficient of frequency, and a decrement larger than the three materials just discussed. Nickel rods and tubes possess too a large decrements to be of much service in stabilizing frequency. Their large motional impedance makes them, however, good sources of sound.

Further work is now in progress to develop alloys with characteristics of value to engineering applications.

The research described above was carried out under the direction of Professor G. W. Pierce in Cruft Laboratory, Harvard University, Cambridge, Massachusetts.

TABLE OF VIBRATION CHARACTERISTICS

Material	Average Max. Motional Imp.	Average Decrement	Range of Dip Angle	Percentage Frequency Increase with Field
Monel metal	70 ohms	14 $\frac{\text{rad}}{\text{sec}}$	82–105 degrees	0.030 per cent
Nichrome	500	20	60–110 degrees	0.143 per cent
Stainless Steel	120	20	104–140 degrees	0.061 per cent
Stoic Metal (Invar)	250	37	97–101 degrees	0.270 per cent
Nickel Rod (Annealed)	175	65	109–112 degrees	0.770 per cent
Nickel Rod (Cold Rolled)	260	85	94–108 degrees	0.273 per cent
Nickel Tube	700	600	52– 87 degrees	1.53 per cent
Nickel Tube (Annealed)	500	1100	48– 89 degrees	3.04 per cent

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NOTE ON RADIO-FREQUENCY TRANSFORMER COUPLED CIRCUIT THEORY*

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Summary—*Equations considering the effects of output and distributed capacities and primary resistance are developed for radio-frequency transformer-coupled amplifiers using either a tuned or an untuned primary. These equations are transformed to such a form that they may be compared with the well-known equations derived for an untuned primary neglecting the output and distributed capacities. The equations for an untuned primary are verified experimentally.*

It is shown that the amplification obtainable with a tube and a transformer having an untuned primary may be made nearly uniform over a frequency range such as that covered by the broadcast band by adding resistance to the primary to reduce the high-frequency amplification. It is also shown that the addition of primary resistance reduces the selectivity approximately the same percentage as it reduces the amplification. The selectivity of a stage with a tuned primary is found to be approximately the square of the selectivity of a stage with an untuned primary.

THE subject of transformer coupled radio-frequency amplifiers, neglecting the primary resistance, the mutual capacity between windings and the output capacity, which includes the plate to filament capacity, the tube socket capacity, the distributed capacity of the primary winding, and the wiring capacity, has been thoroughly treated and the equations are well known. The effect of mutual capacity was recently treated by Diamond and Stowell.[1]

When impedance coupling is used the mutual capacity is no longer present and the output capacity is part of the tuning capacity so that its effect is of no particular interest except as it affects tuning.

With the advent of high impedance tubes such as the screen-grid tube and the '40 type it seems desirable to study the effects of the output capacity and the primary resistance on the general equations. Both of the above factors under certain conditions will affect the amplification of a stage using an untuned primary. The effect under ordinary conditions may or may not be appreciable with high impedance tubes, but it is negligible with low impedance tubes unless capacity and resistance are intentionally added to the primary to affect the amplification. The reason for the above is that the natural period of the primary and output capacity is considerably lower in frequency for a high impedance tube than it is with a low impedance tube.

* Decimal classification: R141×R382.1. Original manuscript received by the Institute, January 13, 1930. Revised manuscript received by the Institute, August 21, 1930.

[1] Diamond and Stowell, "Note on radio-frequency transformer theory," Proc. I. R. E., **16**, 1194; September, 1928.

Transformers having untuned primaries wound with resistance wire have been used to obtain uniform amplification over the broadcast frequency band. The following analysis is made primarily to study the effects of primary resistance and output capacity on both the amplification and selectivity of a transformer with an untuned primary. The equations developed however, are general and may be applied to the case where both the primary and secondary are tuned.

Fig. 1 shows the circuit studied where C_0 represents the output capacity and the other symbols have their usual significance. It may be easily proved by writing the equations or by applying a general theorem referred to by Smith[2] that C_0 in parallel with the plate resistance r_p and the generator μe_g in series acts as a circuit having a generator

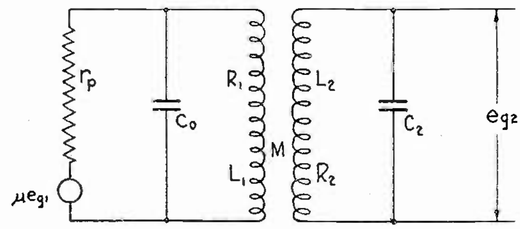

FIG. 1

in series with a resistance and a capacity reactance whose equivalent scalar values are given below.

$$E \text{ series} = \frac{\mu\, e_g}{(1+\omega^2 C_0^2 r_p^2)1/2} \tag{1}$$

$$R \text{ series} = \frac{r_p}{1+\omega^2 C_0^2 r_p^2} \tag{2}$$

$$\text{Series capacity reactance} = \frac{r_p^2 \omega C_0}{1+\omega^2 C^2 r_p^2}. \tag{3}$$

If a tube such as the '24 type of a-c screen-grid tube having an output capacity of 12 $\mu\mu$f and an internal resistance r_p of 400,000 ohms were used the simple analysis, neglecting the output capacity C_0 and the resistance R_1, if appreciable, will no longer apply as the equivalent series resistance will be considerably less than the resistance r_p and the primary reactance will no longer be given by ωL_1.

The circuit might be studied by taking the equation derived neglecting C_0 and replacing μe_g with the vector form of (1), r_p by R_1 plus (2) and adding (3) to ωL_1. Considerable simplicity results when $\omega C_0 r_p$ is

[2] V. G. Smith, "Mathematical study of radio-frequency amplification." PROC. I. R. E., **16**, 533; June, 1927.

of this type are employed in the antenna system at the Lawrenceville, N. J., transoceanic telephone station.

The scheme for clearing an entire antenna and its associated lines of sleet is given in Fig. 17. The antenna depicted is a broadside array with a second parasitic array which functions as a reflector. The antenna elements are arranged in panels to form closed loops and these are brought to common junctions by means of transmission lines. By means of blocking condensers, indicated by (c_0) and the quarter-wave line (b) employed as a large radio-frequency impedance across the line the series circuit for the sleet melting currents and the parallel circuits for the radio-frequency currents are obtained. The condensers (c_1) are low impedance circuits for the radio-frequency currents and serve the purpose of confining the radio-frequency currents to the desired paths. The section of line labelled (T) is the transformer building up the impedance at the junction to the surge impedance of the line.

VI. Tuning Procedure

The following tuning procedure has been successfully employed in setting up several radically different exciter-reflector systems. Although this procedure is not generally applicable to all types of directive arrays it may suggest precautions to be observed in setting up other types.

The first step in the procedure is to bring the parasitic reflector into tune. This is necessary because the reflector introduces reactance into the driven curtain of sufficient magnitude to shift the resonant frequency of the latter. If no previous experience with the particular curtain design has been obtained it is desirable to insert meters in the elements so as to locate and shift the standing waves to the required position on the conductor. It is immaterial for this approximate adjustment whether or not the reflector curtain is driven directly from a power source or parasitically. The standing waves may be shifted either by inserting or cutting out the conductor or by means of one or more adjustable loops of the conductor. The reflector curtain is then excited by the driven curtain and the field strength is observed for a band of frequencies about the required operating frequency at approximately a mile directly to the rear of the antenna. The reflector curtain is said to be in tune at the frequency for which the field intensity is minimum. Usually this dip in field intensity is both sharp and deep. For arrays one wavelength or more long this is approximately the same frequency for which the induced current in the reflector is maximum. The lengths of either the individual reflector elements or, if only a small change is necessary, some one element, are adjusted until the dip in field intensity occurs at the required wavelength.

By means of meter indications the driven curtain is next brought into tune by shifting the current nodes and antinodes into the required position by similar changes of wire length. If but little metallic fittings and harness have been employed it may be found that the average lengths of the antenna elements in the reflector curtain are from two to five per cent more than a one-half wavelength and that the average lengths in the exciter curtain are as much as ten per cent shorter than one-half wavelength.

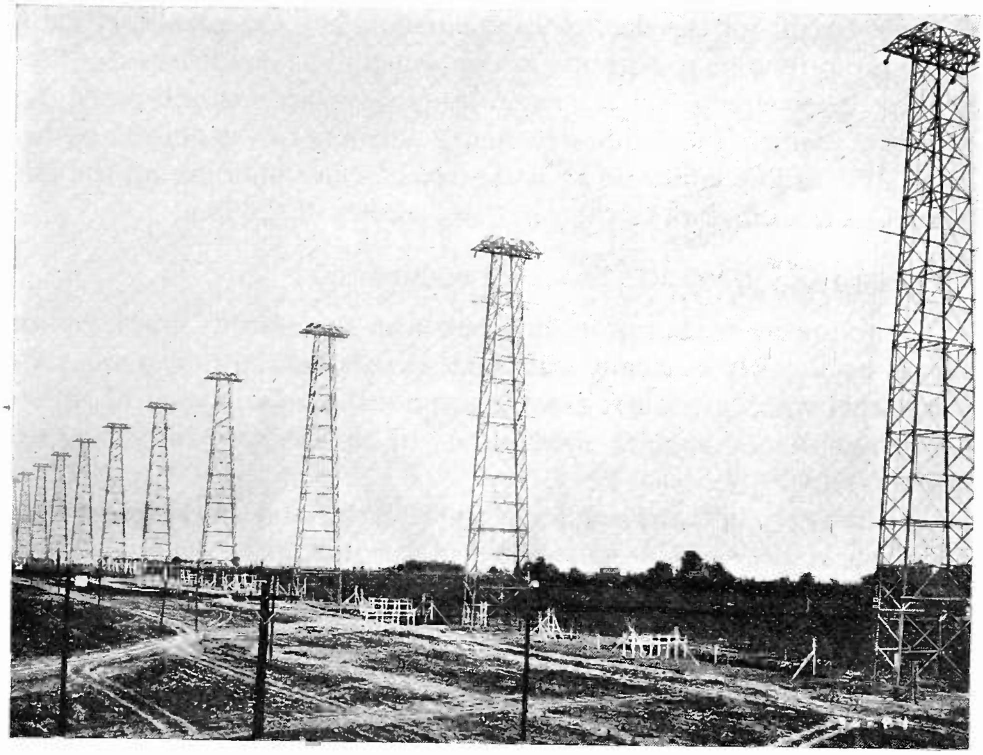

Fig. 18—A view of the Bell System transoceanic telephone station at Lawrenceville, N. J.

The next step in the procedure requires that the impedances for the sections of the driven antenna be built up to the surge impedance of the interconnecting lines. If one-quarter-wave bars are employed as connecting transformers the line lengths must be adjusted so that a current node or antinode appears at the receiving end of the quarter-wave bars. Meters may then be inserted at the extremities of the bars which are spaced at any convenient value and the impedance computed from the surge impedance of the bars and the current ratios as indicated by the meters. The required bar spacing may then be set and checked with meter readings.

It has been found both rapid and convenient to plot the standing

SELECTIVITY

If selectivity is arbitrarily defined as the ratio of the voltage required at any frequency to give the same current as that given by a voltage E at the resonant frequency its value is given by the ratio of the impedance at any frequency to the impedance at the resonant frequency. The impedance at the resonant frequency of the simple circuit[3] is $1(R_2r_p+\omega^2M^2)\omega M$ and at any frequency is

$$(Z_1^2Z_2^2 + \omega^4M^4 - 2\omega^2M^2(X_1X_2 - r_pR_2))^{1/2}/\omega M$$

squaring and dividing we obtain

$$S^2 = 1 + \frac{X_2^2}{\left(R_2 + \dfrac{\omega^2M^2}{r_p}\right)^2}\left(1 + \frac{X_1^2R_2^2}{r_p^2X_2^2} + \frac{X_1^2}{r_p^2} - \frac{2\omega^2M^2X_1}{r_pX_2}\right). \quad (15)$$

If r_p is large compared to X_1 the limiting value of (19) as X_2 increases, becomes

$$S = \frac{X_2}{R_2 + \dfrac{\omega^2M^2}{r_p}} \quad (16)$$

Similarly the impedance found from the approximate equation (9) may be divided by the absolute value of impedance found from (4). If C_0 is small the selectivity becomes

$$S = \frac{X_2}{R_2(1 - \omega^2C_0L_1) + \dfrac{\omega^2M^2}{r_p}\dfrac{(1 + r_p\omega^2C_0R_1)}{(1 - \omega^2L_1C_0)}}. \quad (17)$$

Thus if the impedance at resonance is increased by increasing R_1 the selectivity will be decreased the same percentage as the amplification.

The selectivity of the circuit with a tuned primary may also be found quite easily. The impedance at resonance (10) and the impedance at any frequency are both modified in the same manner by μe_g becoming $\mu e_g/jr_p\omega C_0$ in the equivalent series circuit. The expression for selectivity then becomes the same as that of (15) except that r_p and x_1 are replaced by R and X. In this case x_1^2/R^2 must be considered so that the selectivity considerably off resonance becomes

[3] J. R. Nelson, "Circuit analysis applied to the screen-grid tube," PROC. I. R. E., 17, 334; February, 1929.

$$S = \frac{x_2 x_1}{\left(R_2 + \frac{\omega^2 M^2}{R}\right) R} . \tag{18}$$

It is to be noted that the selectivity given by (18) is approximately the square of that given by (16).

Experimental Verification of the Amplification Equations

Table I gives the physical size and inductances of the windings of the transformers used to verify the equations. The primary and secondary were mounted concentrically with the start of the secondary winding so as to reduce the capacity coupling coefficient to a small value. The resistances of the windings R_1 and R_2 are given in Fig. 2.

TABLE I
Constants of Experimental Transformers

Transformer Winding	Type of Winding	Size of Wire	Inductance Microhenries
Primary 2½-inch tubing	Single Layer Solenoid	No. 36 D.S.C. Copper Wire	300
Secondary 3-inch tubing	Single Layer Solenoid	No. 24 Enameled Wire	240

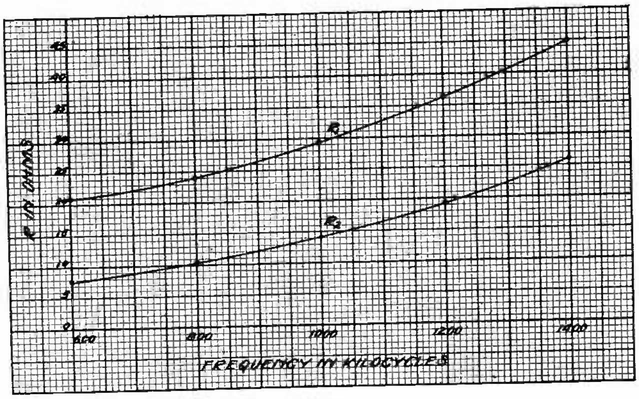

Fig. 2—Coil resistances.
R_1—Primary
R_2—Secondary

The values of the measured amplification of the transformer used with an Eveready Raytheon 240 tube under the conditions specified are given in Fig. 3-A. The values of amplification calculated by the conventional formula are given in Fig. 3-B and the values calculated by (9) are given in Fig. 3-C. The agreement between the measured and the calculated values is much closer when (9) is used than it is when the conventional formula is used.

Fig. 4-A shows the measured amplification curve with 10-$\mu\mu$f output capacity and B of the same figure shows the calculated values.

Fig. 4-D gives the measured values when a resistance of 1000 ohms was placed in series with the primary with 10-$\mu\mu$f output capacity, while C gives the calculated values.

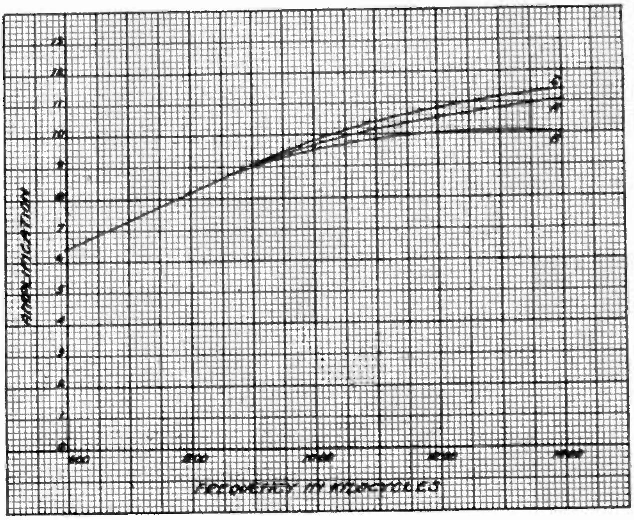

Fig. 3—Voltage amplification with an ER-240 tube.

$g_m = 300 \times 10^{-6}$ mhos
$r_p = 100{,}000$ ohms
$\mu = 30$
$M = 50$ microhenries
$C_0 = 5\ \mu\mu$f
A = Measured
B = Calculated conventional formula
C = Calculated by (9)

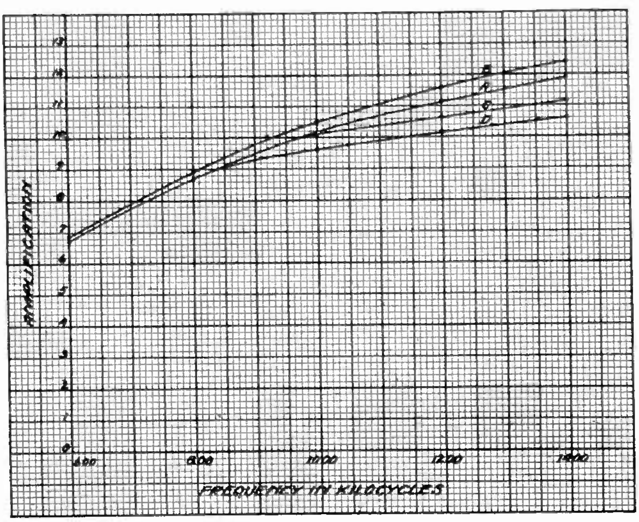

Fig. 4—Voltage amplification with ER-240 tube.

$g_m = 300 \times 10^{-6}$ ohms
$r_p = 100{,}000$ ohms
$\mu = 30$
$M = 50$ microhenries
A = Measured with $C_0 = 10\ \mu\mu$f
B = Calculated with $C_0 = 10\ \mu\mu$f
C = Measured with $C_0 = 10\ \mu\mu$f and $R^1 = 1000$ ohms
D = Calculated with $C_0 = 10\ \mu\mu$f and $R_1 = 1000$ ohms

Conclusions

The output capacity acts with an untuned primary to increase the effective primary inductance and the effective mutual inductance. When the product of $r_p R_1 \omega^2 C_0^2$ is small the amplification will be increased by the output capacity if the mutual inductance is less than the optimum value and will be decreased by the presence of the output capacity if the mutual inductance is greater than its optimum value. The exact analysis shows that the maximum possible amplification obtainable is reduced by the output capacity and primary resistance.

The addition of primary resistance offers a means of controlling the amplification. The high-frequency amplification may be reduced to any value less than that given with a small value of primary resistance without affecting the low-frequency amplification to the same extent. In other words, the addition of primary resistance offers a means of approaching a flat amplification curve as the frequency is varied or a curve having a falling characteristic of the usual interstage transformer. The selectivity will also be decreased by the addition of the primary resistance which fact should be considered in the transformer design.

Appendix I

$$I_0 = jWC_0V_0 \tag{1}$$

$$V_0 = \mu e_{g1} - r_p(I_0 + I_1) \tag{2}$$

$$V_0 = Z_1 I_1 - jwMI_2 \tag{3}$$

$$0 = Z_2 I_2 - jwMI_1 \tag{4}$$

$$\frac{\mu e_{g1}}{1 + jWC_0 r_p} = I_1 Z_1 + \frac{r_p}{1 + jwC_0 r_p} - jWMI_2 \tag{5}$$

$$1_1 = \frac{Z_2 I_2}{jWM} \tag{6}$$

$$I_2 = \frac{\mu e_{g1} jWM}{R^2(b + ja) + W^2M^2(1 + jwC_0 r_p) + X_2(-a + jb)} \tag{7}$$

where,

$$Z_1 = R_1 + jX_1 = R + j(WL_1)$$

$$a = X_1 + wC_0 r_p R_1$$
$$b = r_p + R_1 - wr_0 \times r_p$$

To find when I_2 will be a maximum as any reactance is varied the absolute value of Z_{12} that is, μ_{eg}/I_2 is differentiated with respect to that reactance and the derivative thus found is set equal to zero. Doing this we find the following values of reactances to make I_2 a maximum.

$$W^2M^2 = \left[(R_2^2+X_2^2)\left(X_1^2+R_1^2+\frac{r_p^2}{1+r_p^2W^2C_0^2}1-2WC_0X_1+\frac{2R_1}{r_p}\right)\right]^{1/2} \tag{8}$$

or,

$$W^2M^2 = [(x_2^2 + R_2^2)(x^2 + R^2)]^{1/2} \text{ or } Z_2Z$$

$$X_2 = \frac{\omega^2M^2[(X_1 - \omega C_0r_p^2(1 - \omega C_0X_1)]}{r_p(1 - \omega C_0X_1)^2 + 2R_1r_p + R_1^2 + X_1^2 + r_p^2R_1^2\omega^2C_0^2} \tag{9}$$

or,

$$X_2 = \frac{\omega^2M^2X}{R^2 + x^2}$$

$$X_1 = \frac{WC_0r_p^2}{1 + W^2C_0^2r_p^2} + \frac{X_2W^2M^2}{R_2^2 + X_2^2} \text{ or } X_1 = \frac{W^2M^2X_2}{R_2^2 + X_2^2} \tag{10}$$

$$\omega C_0 = \frac{X_1(R_2^2 + X_2^2 - \omega^2M^2X_2)}{(X_1^2 + R_1^2)(R_2^2 + X_2^2) + \omega^2M^2(\omega^2M^2 + 2R_1R_2 - 2X_1X_2)} \tag{11}$$

where,

$$X = \omega L_1 - \frac{\omega C_0r_p^2}{1 + \omega^2C_0^2r_p^2}$$

$$R = R_1 + \frac{r_p}{1 + r_p^2\omega^2C_0^2}.$$

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MEASURING FREQUENCY CHARACTERISTICS WITH THE PHOTO-AUDIO GENERATOR*

By

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(Communication from the Laboratory of the Reichs-Rundfunk-Gesellschaft, Berlin, Germany)

Summary—*A special alternating-current generator, a photo-audio generator, is used to good advantage as a source of current for measurements of variability with frequency. Compared with the known audio generators (tube generator with or without superposition) hitherto used for the same purpose, the photo-audio generator has the advantage of greater simplicity. The sources of error, as regards sine shape of its potentials and the variability with frequency are so small that they need not be considered in practical measurements.*

The described arrangement is so adjusted that the frequency range in question is continuous, and the result of the measurement is shown as a curve drawn by a recording instrument.

Measurements of quantities dependent on frequency are very important in radio engineering. They are made regularly as a matter of routine and consequently must be as simple as possible. The audio-frequency generator is an essential part used in these measurements. Heretofore such generators were tube transmitters in which either the alternating voltages of audible frequency were produced direct, or by a combination of two high-frequency oscillators which produced true audible frequency by superposition. The devices of the former type are simple to construct except for the oscillation circuit for low oscillations of 30 or 50 cycles. In addition, difficulties are encountered in finding a simple method for the continuous variation of the frequency of the generator within the range from about 30 to 10,000 cycles and at the same time keeping the output voltage constant over the entire frequency range.

The second method, in which the audible frequency is produced by the superposing process, is somewhat more complicated to set up, and there are many more possible sources of error. The fact cannot be doubted that such apparatus has already been used to good advantage, but it is highly desirable to have simpler devices for practical measurements, that can be operated by the layman without difficulty.

As a result of our experience in the transmission of pictures, we were able to develop a special type of audio generator that is very easy to operate. The principle and appearance of such an audio generator

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is shown in Figs. 1 and 1a. "1" is an incandescent lamp whose beam falls on a photo-electric cell "2". If the ray of light is periodically interrupted by a perforated disk "3" driven by the shunt-wound motor "4", the

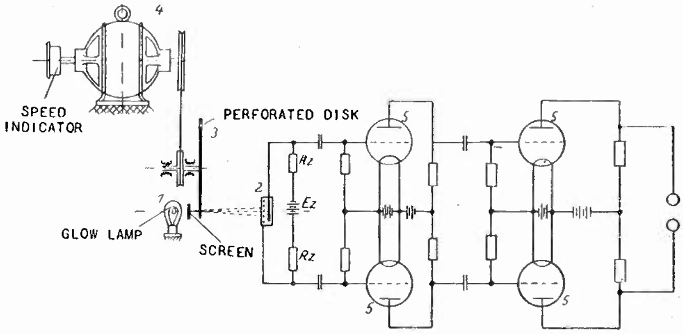

Fig. 1—Diagram of complete circuit.

resistance changes (and with it, the photocell current) in time with the interruptions. These changes in current produce corresponding alternating potentials that are amplified later. In the illustration E_z is a source of photocell current, and "5" the four amplifier tubes. The symmetrical amplifier arrangement is necessary in order to prevent external interference with the comparatively low potential of the photocell.

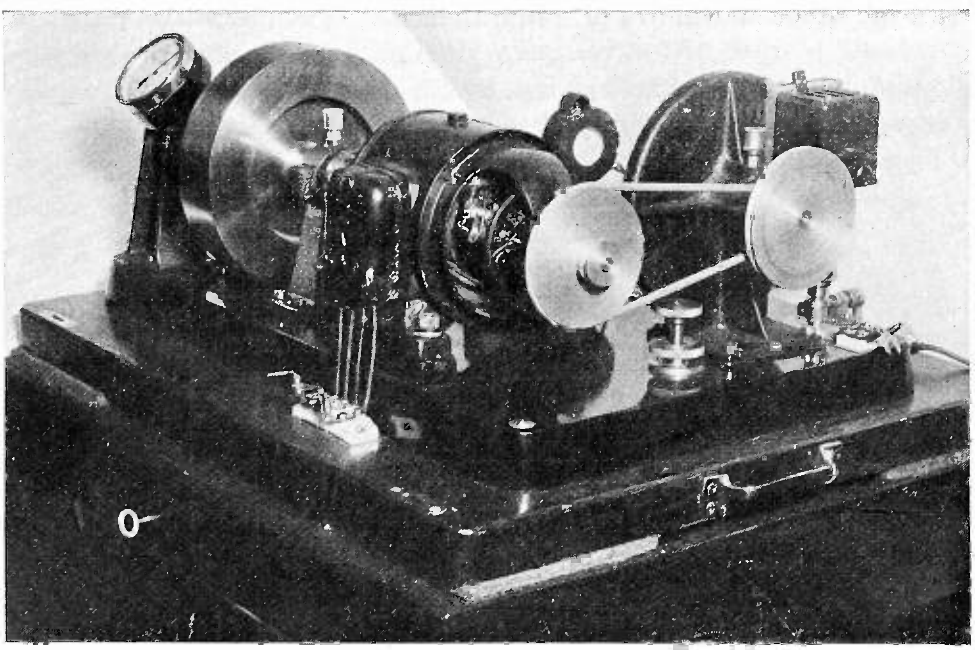

Fig. 1a—Photograph of photo-audio generator.

If the perforated disk has 120 holes, for example, one revolution of the disk a second will give an alternating current of 120 cycles; that is, the frequency range from 30 to 10,000 can be obtained with a disk speed from 15 to 5000 r.p.m. The desired disk speed can be regulated by changing the voltage at the armature of the motor and also by two different ratios for the belt transmission between motor and disk. The frequency can then be determined directly with a tachometer.

The alternating voltage supplied by this generator is constant and independent of the frequency that is produced. It is assumed that the

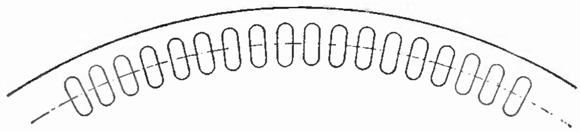

Fig. 2—Shape of holes in rotating plate.

action of the photocell does not depend on time. We shall take up the investigation of this later, but first will discuss the question of the shape of the alternating potential curve. If it is assumed that the change in the resistance of the photocell is proportional to the size of the surface illuminated at each instant, the result of the change in resistance can be calculated as follows, assuming that the holes in the rotating disk have the shape shown in Fig. 2 and that their width is equal to the di-

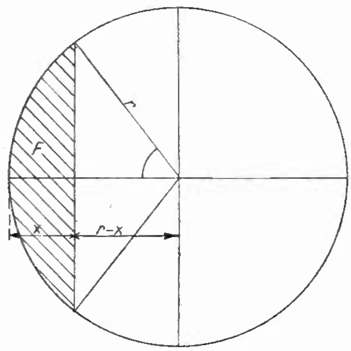

Fig. 3—Graphic illustration of the formula for the illuminated surface of the diaphragm.

ameter of the light beam. Let F be the area of the beam exposed through the slit. Then (see Fig. 3):

$$F = r^2 \operatorname{arc\,cos} \frac{r - x}{r} - \frac{1}{2}\sqrt{r^2 - (r - x)^2} \cdot (r - x).$$

This function is plotted in Fig. 4, and is evidently far from sinusoidal. Measurements with the "distortion-factor†" bridge and observations with an oscillograph have indicated that a very close approximation to sine shape is obtained (probably due to diffraction at the

† The original German text gave this as "Klirrfactor."

boundary lines of the screen and the disk perforations) if the width of the slit on the disk is a trifle larger than the diameter of the screen. The effective value of the resultant harmonics was less than 4 per cent.

There remains to be investigated the extent to which the voltage developed by the photocell in the resistances R_z (Fig. 1) depends on the frequency. Two things must be considered, capacity and ionization of the gas-filled cell.

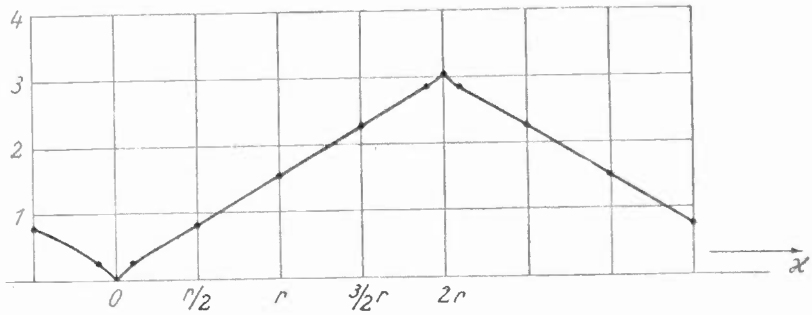

Fig. 4—Curve resulting from Fig. 3.

The capacitive reactance of the cell at the highest frequencies (in this case 10 kilocycles per second) must be large compared with the load resistance R_z if the terminal voltage is not to drop at these frequencies. Suitable tests have shown that the load resistance should not exceed 40,000 ohms. This means that the efficiency of the cell, whose internal ohmic resistance is from 10^7 to 10^8, must be greatly reduced to secure independence of the voltage from the frequency.

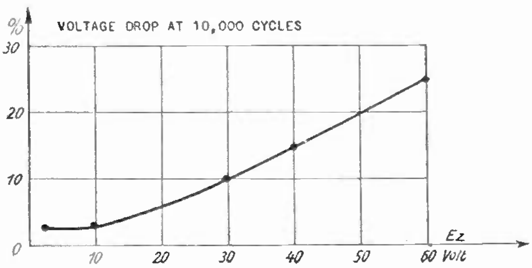

Fig. 5—Drop at 10,000 cycles depending on cell voltage.

The second defect of the cell is due to the gas filling that is necessary in order to obtain the required sensitivity.

Measurements with a tube voltmeter of the alternating voltages developed in the photocell at frequencies from 50 to 10,000 cycles per second show the remarkable fact that there is a falling off of alternating voltages at high frequencies as the d-c potential exceeds 12 volts, the falling off becoming greater the more the potential exceeds 12 volts. This falling off at 10,000 cycles per second is evident in Fig. 5 where the per cent decrease in the a-c photocell voltage at 10,000 cycles per second is compared with that at 50 cycles per second at various d-c potentials. The slight potential decrease of 3 per cent, which is constant up to 12 volts, is caused by the characteristics of the amplifier.

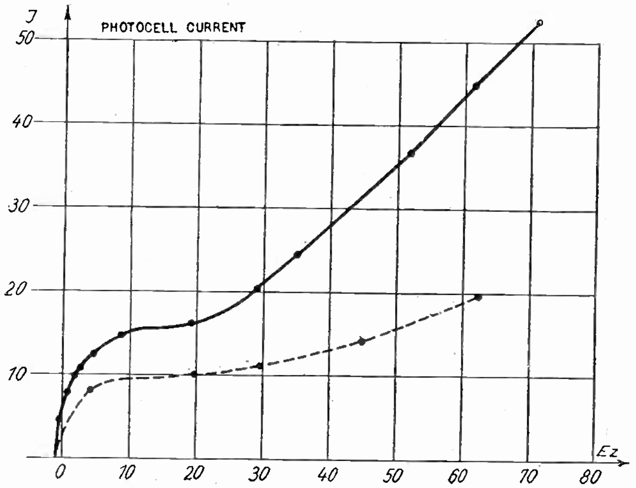

Fig. 6—Cell current versus cell voltage.

Since it was suspected that the function shown in Fig. 5 was due to ionization processes inside the cell, the relation between the cell current and the cell voltage was studied. Measurements were made with the arrangement shown in Fig. 7, where Z is a normal commercial photo-

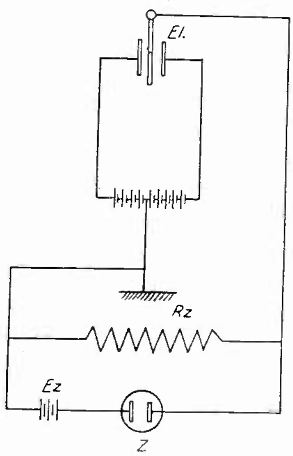

Fig. 7—Arrangement for measuring the cell current.

cell, R_z a resistance of 100,000 ohms, E_l a string electrometer, and E_z the photocell voltage. Fig. 6 shows the current curve on an arbitrary scale and indicates plainly that the cell current I follows an entirely different law above 12 volts from that which holds below 12 volts. The critical point in the curve occurring at about 12 volts leads us to conclude that there is an entirely different physical process in the cell on each side of this potential, in which ionization of the residual gas in the cell plays a part. In every case the position of the critical point in Fig. 6 and that of the bend in Fig. 5, agree very well. On reversing the cell voltage we obtain a perfectly similar function (dash curve, in Fig. 6 on a larger scale) which indicates that the cell grid is also photo active.

The results of other investigations of the effect of different gases on the variability of the cells with frequency are available.[1] They have shown that cells can be made which can be operated with higher direct current voltages up to about 80 volts without affecting the high frequencies.

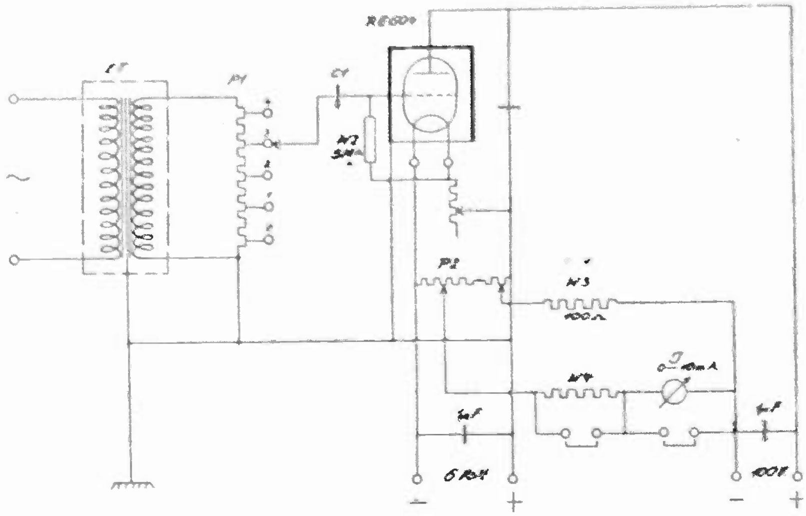

Fig. 8—Connection of the vacuum tube voltmeter.

Just as with any other audio generator this photo-audio generator can be used to apply voltages at individual frequencies to various points of the apparatus being tested, and its frequency curve drawn. The generated frequency is determined by the speed indicator on the motor. However, the following method is much more convenient and more suitable for practical measurements.

The perforated disk motor equipped with a flywheel is brought to the highest speed. The motor is then turned off and allowed to slow down. The frequency of the alternating voltage generated decreases at the same rate as the speed of the motor. This process is carried on in two stages with a change in the transmission ratio between motor and disk between stages, the first including the frequency range from

[1] Fritz Schröter and Günther Lubszynski, "Untersuchungen zur Frage der Trägheit gasgefüllter Photozellen," *Phys. Zeits.*, 31, 897–904, 1930.

10,000 to 500 cycles per second, and the second from 1000 to 50. In order to make a continuous record of such a constantly changing frequency, it is necessary to use a recording device, and for this purpose we use a recording moving-coil milliammeter (0 to 10 milliamperes) with an audion rectifier. The wiring diagram and external appearance of the audion rectifier are shown in Figs. 8 and 9.

As can be seen in Fig. 8, the alternating potential passes through the input transformer $ET(1:2)$, the potentiometer P_1, and through

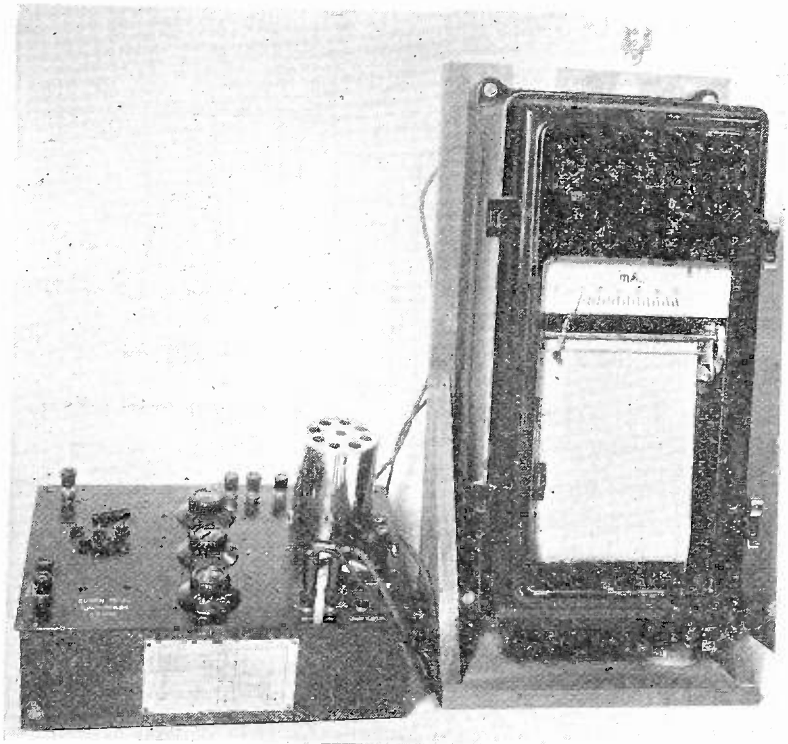

Fig. 9—Photograph of the vacuum tube voltmeter and the registering instrument.

condenser C_1 to the grid of a tube. The leakage of the grid current takes place through resistance W_2. The plate current flowing through resistance W_3 produces a direct-current potential that can be compensated by the potentiometer P_2. The compensation current flows through ammeter I.

For the protection of the instrument a resistance W_4 is inserted, which can be short-circuited after compensation. For input voltage 0, the voltage at resistance W_3 is completely compensated, so that no current flows through the instrument I. The instruments acts when an alternating voltage reaches the grid of the tube.

Fig. 10 shows the calibration curve of the instrument for four different positions of the input potentiometer.

In order to be able to evaluate the curve drawn by the recording instrument, it is well to mark the curve at certain points to correspond to fixed frequencies. If the curve is drawn in the room where the audio generator is located, it is possible to make marks by hand on the slowly moving paper strip (speed 12 cm/min.) in accordance with the revolution counter. This marking also can be done automatically, for example, by short interruptions of the light ray in the photo-audio gen-

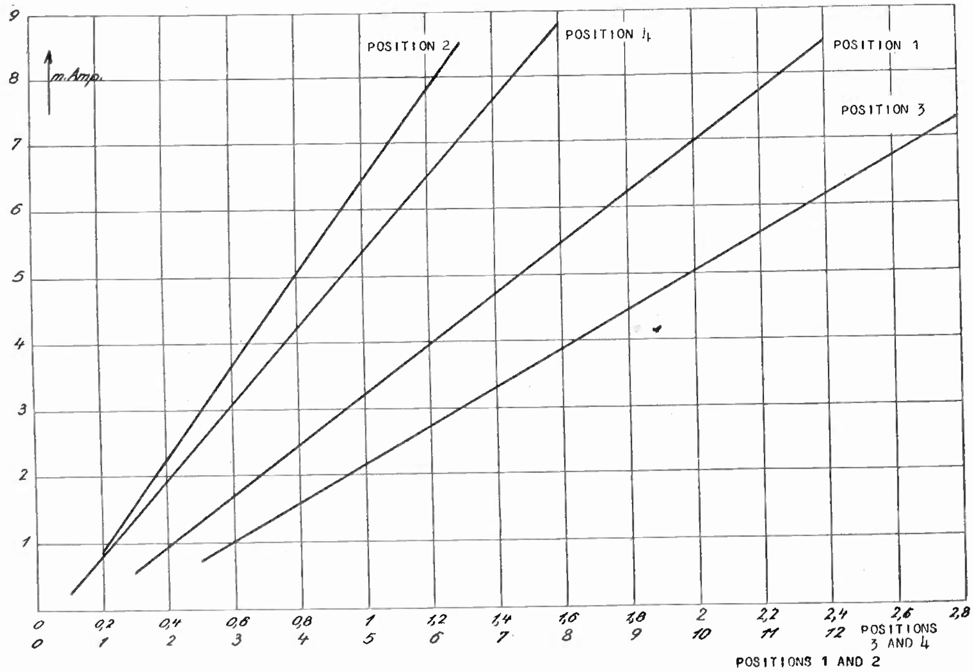

Fig. 10—Calibration curve for equipment shown in Fig. 9.

erator at certain instants. The correct instant for the frequency to be marked can be determined directly by the tachometer at the photo-audio generator. Because of the interruptions of the light ray, the voltage delivered by the photo-audio generator drops for an instant and the curve drawn by the recording instrument shows a corresponding break. Fig. 11 shows such a curve.

Instead of interrupting the light ray, it also is possible to interrupt the output voltage of the generator at the proper instant. The shape of the curve is similar in each case.

With such an arrangement, it also is possible to make measurements if separate rooms are used for the photo-audio generator and the recording instrument. During the measuring itself, the observers who make them in different rooms do not have to be in communication, provided that the time point for beginning the measurement is common to both. This makes it possible in a very convenient manner to make a study of an entire broadcast system from the microphone amplifier through the lines with their amplifiers, and finally through the transmitter itself to a control device at the aerial, to which finally the re-

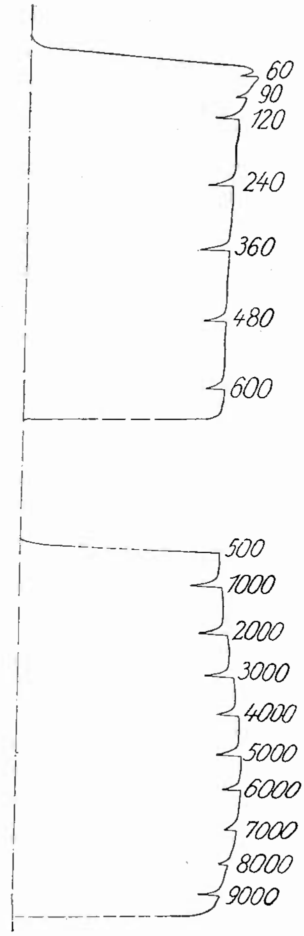

Fig. 11—Frequency curve, taken with photo-audio generator.

cording instrument is connected. Such a measurement can be made in less than 5 minutes, and the curve gives information as to the condition of the entire installation. If the result of the measurement is unsatisfactory, the defective circuit can be traced easily by repeating the measurement connecting the recording instrument to the low-frequency line leading to the transmitter. By comparing the two curves it can be seen whether or not the source of error is in the transmitter. By further measurements in the same manner, faults in the complete line or in parts of the associated amplifiers can be located, connecting the recording instrument at points nearer the microphone amplifier. Obviously it also is possible to obtain the result with one measurement using several recording instruments. Conditions are considerably simplified in practical work. In general, since microphone amplifier and transmitter can be tested first, the trouble is found most frequently in the line between microphone amplifier and transmitter. Generally one or more substitute lines are available, and if the first result is unsatisfactory, the measurement is repeated with new lines and the line giving the best result is used.

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PERFORMANCE OF PIEZO-OSCILLATORS AND THE INFLUENCE OF THE DECREMENT OF QUARTZ ON THE FREQUENCY OSCILLATIONS*

By

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Summary—*In Part I of this paper, the performance of piezo oscillators of the usual Pierce circuits is treated, on the basis of the resonance curves of the quartz (taken experimentally) and with the help of vector diagrams.*

In Part II, the influence of the decrement of the quartz resonator on the oscillation frequency is examined. This study has led to the development of an arrangement which permits the quartz to oscillate in proximity to its frequency of resonance and to reduce thereby the influence of the decrement on frequency to about 1/10 of that usually found.

Part I

It has been known for some time that the frequency of a piezo oscillator of either of the two classic Pierce schemes differs from the frequency of mechanical resonance of the quartz, lying between the frequencies of series resonance, determined by:

$$\omega^2 L' \frac{C' C_1'}{C' + C_1'} = 1, \tag{1-1}$$

(in which L', C', etc., are the constants of the electric circuit equivalent

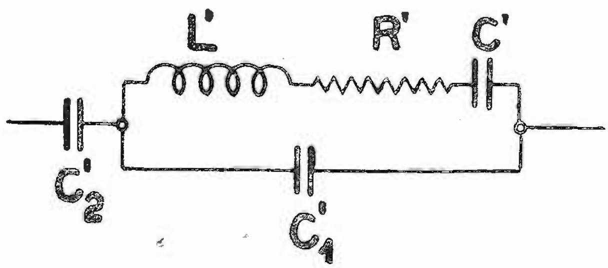

Fig. 1.

of the quartz, Fig. 1) and of the parallel resonance, determined by:

$$\omega^2 L' \frac{C'(C_1' + C_2')}{C' + (C_1' + C_2')} = 1. \tag{1-2}$$

The frequency of mechanical resonance of the quartz plate differs also from these and is given by:

* Decimal classification: R191×537.65. Original manuscript received by the Institute, October 11, 1930. Translation received by the Institute, March 24, 1931.

$$\omega^2 L'C' = 1. \tag{1-3}$$

In his work of November, 1928, Terry[1] who used the equivalent circuit of the quartz as a basis, studied Pierce's schemes analytically. Since the analytical treatment of the problem (notwithstanding the simplification introduced: $K_2 = \infty$, namely, no air gap) leads to rather complex formulas, which are difficult of physical interpretation, it is intended to proceed with the study of piezo oscillators on the basis of the resonance curve of the quartz using principally vector diagrams. This method is less precise and general than the analytical one, although it is more easily applicable.

(2) Accurate measurements have been made on some quartz plates to determine the values of the impedence

$$(Z)_x = \sqrt{\alpha^2 + \beta^2}$$

and of its components α and β as functions of the frequency of the alternating voltage applied to the crystal.

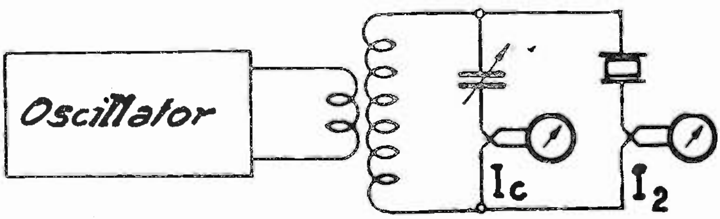

Fig. 2.

The method of measurement employed by the I.E.C. of the Italian Navy at Livorno differs slightly from that used by Dye[2] and that used by Cady[3] to obtain the resonance curve of the quartz. It consists in determining I_c and I_2 (Fig. 2) for each frequency, while C is regulated in a manner to keep I_c a maximum. This method allows the determination of the admittance of the quartz, independently from the e.m.f. introduced in the oscillating circuit by the oscillator, thereby eliminating errors caused by possible variations of the latter. In addition through the difference ΔC between the value C at which I_c reaches its maximum value and value C_0 at which the oscillating circuit becomes resonant at the same frequency, the values of α and β may be determined.

[1] Earle M. Terry, "Frequency of quartz piezo oscillators," Proc. I.R.E., 16, 1486–1506; November, 1928.

[2] D. W. Dye, "The piezo-electric quartz resonator and its equivalent circuit," *Proc. Phys. Soc.* (London), 38, part 5–15, August, 1926.

[3] W. G. Cady, "The piezo-electric resonator, Proc. I. R. E., 10, 83; April, 1922.

(3) *Method of determining α and β.* Applying Kirchhoff's laws to the measuring circuit shown in Fig. 3, (the e.m.f. induced in the coil being

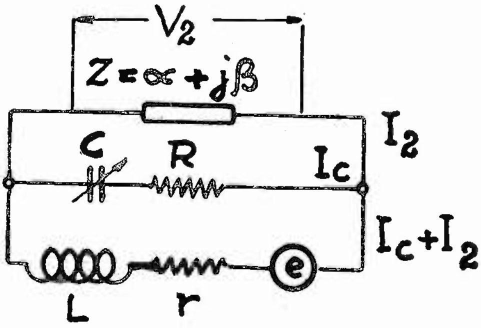

Fig. 3.

designated by e), the following may be written

$$\left.\begin{aligned} e &= \left[(r+R)+j\left(\omega L-\frac{1}{\omega C}\right)\right]I_c+[r+j\omega L]I_2 \\ 0 &= \left[R-j\frac{1}{\omega C}\right]I_c-[\alpha+j\beta]I_2 \end{aligned}\right\} \quad (3\text{-}1)$$

from which the following is obtained:

$$\left|\frac{e}{I_c}\right|^2=(r+R)^2+\left(\omega L-\frac{1}{\omega C}\right)^2+\frac{1}{|Z|^2}\left[(r^2+\omega^2L^2)\left(R^2+\frac{1}{\omega^2C^2}\right.\right.$$
$$\left.\left.+2\alpha R-2\frac{\beta}{\omega C}\right)+2\left(R^2+\frac{1}{\omega^2C^2}\right)(\alpha r+\beta\omega L)\right]. \quad (3\text{-}2)$$

Since $\omega^2L^2/r^2 \geqq 10^3$, (3-2) becomes:

$$\left|\frac{e}{I_c}\right|^2=(r+R)^2+\left(\omega L-\frac{1}{\omega C}\right)^2+\frac{1}{|Z|^2}\left[\omega^2L^2\left(R^2+\frac{1}{\omega^2C^2}\right.\right.$$
$$\left.\left.+2\alpha R-2\beta\frac{1}{\omega C}\right)+2\left(R^2+\frac{1}{\omega^2C^2}\right)(\alpha r+\beta\omega L)\right]. \quad (3\text{-}3)$$

The value of C at which the current I_c reaches its maximum is that which makes

$$\frac{d}{dC}\left|\frac{e}{I_c}\right|^2$$

nil, since e is constant, and, hence, the following results:

$$0=(\omega^2LC-1)-\frac{1}{|Z|^2}[\omega^2L^2+2\alpha r-\beta\omega L(\omega^2LC-2)] \quad (3\text{-}4)$$

Since we have defined C_0 as the value of C at which the oscillating circuit, without the quartz, is in resonance, and since also $\Delta C = C - C_0$, it follows that:

$$\omega^2 LC = 1 + \frac{\Delta C}{C_0} \tag{3-5}$$

and, therefore, (3-4) becomes:

$$0 = \frac{\Delta C}{C_0} - \frac{1}{|Z|^2}\left[\frac{1}{\omega^2 C_0^2} + 2\alpha r - \frac{\beta}{\omega C_0}\left(\frac{\Delta C}{C_0} - 1\right)\right] \tag{3-6}$$

from which, substituting β/α and since

$$|A^2| = \frac{1}{|Z|^2}$$

the following is obtained:

$$0 = \frac{\beta^2}{\alpha^2}(M^2 - N^2) + \frac{\beta}{\alpha}4ArN + (M^2 - 4A^2r^2) \tag{3-7}$$

where,

$$\left.\begin{aligned} M &= \frac{\Delta C}{C_0} - \frac{|A|^2}{\omega^2 C_0^2} \\ N &= \frac{|A|}{\omega C_0}\left(\frac{\Delta C}{C_0} - 1\right) \end{aligned}\right\}. \tag{3-8}$$

The value of β/α becomes:

$$\frac{\beta}{\alpha} = \frac{-|A|rN \pm M\sqrt{N^2 - M^2 + 4|A|^2r^2}}{-(N^2 - M^2)} \tag{3-9}$$

which, assuming $r = 0$, can be written as:

$$\frac{\beta}{\alpha} = \frac{-M}{\pm\sqrt{N^2 - M^2}} \tag{3-9'}$$

In the same manner we obtain from (3-6):

$$\beta = \frac{1}{|A|}\,\frac{-MN \pm 2|A|r\sqrt{(N^2 - M^2) + 4|A|^2r^2}}{N^2 + 4|A|^2r^2} \tag{3-10}$$

$$\alpha = \frac{1}{|A|}\,\frac{2|A|rM \pm N\sqrt{(N^2 - M^2) + 4A^2r^2}}{N^2 + 4|A|^2r^2} \tag{3-11}$$

which, if $r=0$, reduce to:

$$\beta = -\frac{1}{|A|}\frac{M}{N} \tag{3-10'}$$

$$\alpha = \pm\frac{1}{|A|}\frac{\sqrt{N^2 - M^2}}{N}. \tag{3-11'}$$

In general r has a value which can safely be taken as negligible and the simplified equations can be used without serious errors.

From (3-8) it will be seen that $N<O$ if $|\Delta C|<C_0$. In practice it is well to give a low value to the ratio L/C of the oscillating circuit in

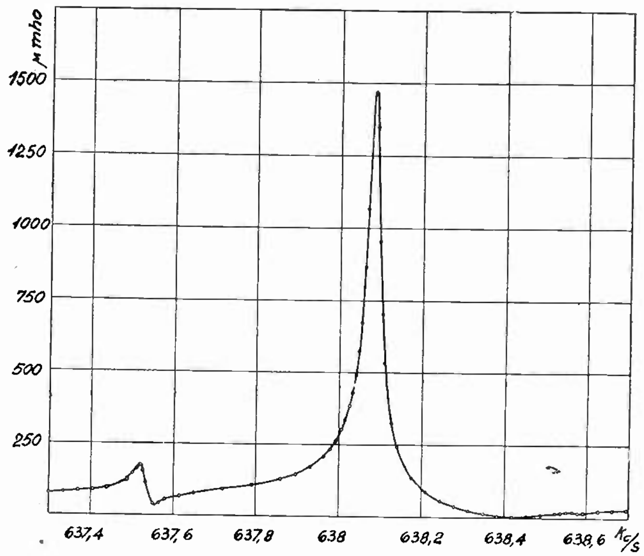

Fig. 4.

order that this condition may be assured. Now, since α cannot be negative it will be seen from (3-11′) that $\sqrt{N^2-M^2}$ is to be taken as negative. Hence, β and β/α will have the sign of M.

(4) In Fig. 4 the admittance $|A|$ measured experimentally is plotted against frequency and in Fig. 5 α, β, and β/α derived from the experimental values of $|A|$ and the experimental values of C_0, ΔC, ω by (3-8), (3-9′), etc. The piece of quartz used for measuring $33\times33\times4.5$ mm was of the type used for secondary standards of frequency by the I.E.C. The maximum and minimum values of $|A|$ correspond respectively to the frequency of the series resonance (1-1) and parallel resonance (1-2) of the crystal.

The curve shown dotted in Fig. 5 gives the values of β/α computed from the electrical constants equivalent to the plate or crystal, obtained by one of the usual methods, from the curve of Fig. 4. This agrees with the experimental curve in a sufficiently satisfactory manner though departing from it noticeably below 638.04 kc/s due to the secondary resonance found at 637.54 kc/s for which, obviously, it has not been possible to allow.

(5) We shall now analyze the behavior of quartz oscillators by means of the well-known method of separate excitation, generally used for triode oscillators (Gutton).

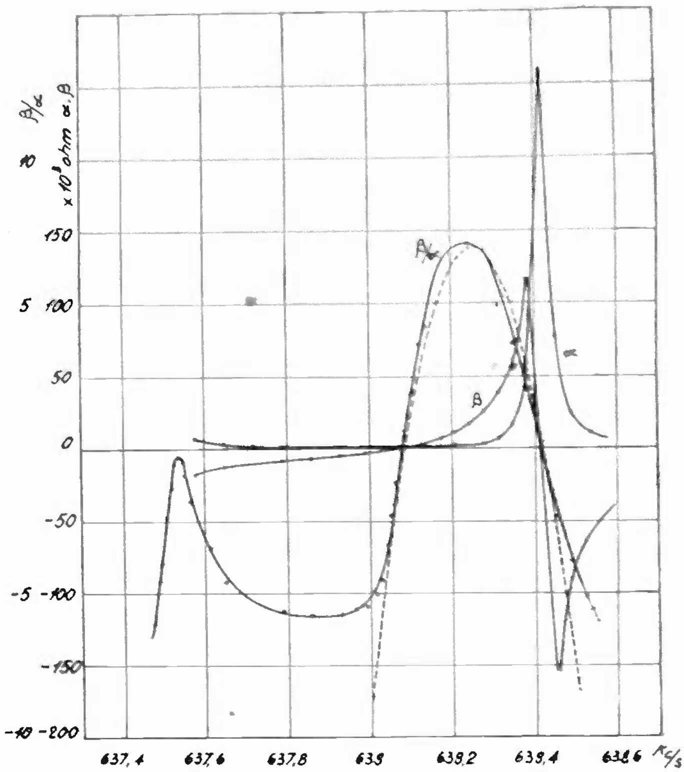

Fig. 5.

Fig. 6 shows the circuit equivalent to a piezo oscillator of one of two fundamental schemes: Z_a is the impedance of the external anode circuit; Z_c is the impedance of the quartz, with the grid-filament capacity, (or grid-plate), in parallel, and C is the grid-plate capacity (or grid-filament) if the circuit with the quartz between grid and filament is considered (or that with the quartz between grid and plate). The alternating voltage on the grid or the control voltage will in the two cases referred to be V_c across Z_c or V_c across C, respectively. E is an alternating voltage which represents the effect of the grid in the anode circuit.

Continuous oscillation is possible at those frequencies for which V_c

is in phase with E and the oscillations will be in equilibrium state in respect to frequency if an increase of frequency in E produces a retardation of the phase of V_c with respect to E, or vice versa.

Since only a qualitative analysis of the behavior is of interest, the electric current in the grid is assumed to be nil and ρ the internal re-

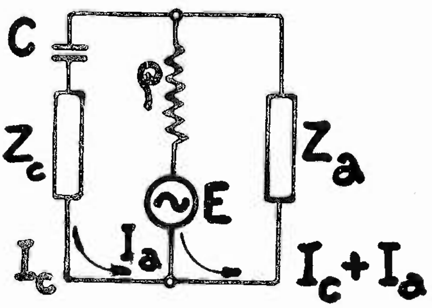

Fig. 6.

sistance of the anode circuit constant. Moreover, we shall assume at the beginning that $I_c \ll I_a$, then $(I_c+I_a)=I_a$.

(6) *Quartz connected between grid and filament*: It will be seen that with this arrangement the quartz will oscillate for values of the capacity of the tuned anode circuit which are less than those needed for

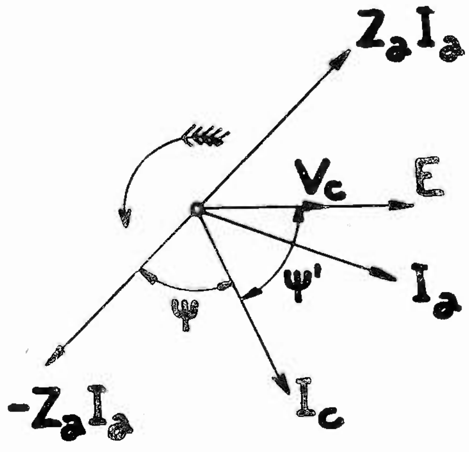

Fig. 7.

resonance, namely, with inductive Z_a. Under these conditions current I_a is retarded with respect to the e.m.f. E by an angle smaller than $\pi/2$. Assuming the signs of the currents to be as shown in Fig. 6, we have to consider the potential $-Z_aI_a$ as active, in relation to current I_c. V_c must be in phase with E and, therefore, it must lead (Fig. 7) $-Z_aI_a$ by an angle supplementary to that which $+Z_aI_a$ makes with E, i.e., by an angle lying between $\pi/2$ and π. This is possible only if Z_c is inductive, while Z_c and the capacity C in series, will be capacitative. In other words, the inductive reactance of Z_c is less than the impedance of the grid-plate capacity. In this case current I_c will lead $-Z_aI_a$ by an angle smaller than $\pi/2$, namely ψ, and V_c in turn will lead I_c by an angle still smaller, namely ψ.'

If Z_c is to be inductive, there is all the more reason why the quartz should have inductive reactance and, therefore, the oscillation frequency is certainly between the frequency of series resonance and that of parallel resonance. In Fig. 8, curves α' and β' are traced as functions of the frequency and are components of the impedance Z_c. The values α' and β' have been computed from those of α and β furnished by the curve in Fig. 5, by means of

$$\begin{aligned}\alpha' &= \alpha\frac{\gamma^2}{\alpha^2 + (\beta - \gamma)^2}\\ \beta' &= -\gamma\frac{\alpha^2 + \beta^2 - \beta\gamma}{\alpha^2 + (\beta - \gamma)^2}.\end{aligned} \tag{6-1}$$

in which $\gamma = 1/\omega C_g$ is the impedance of the grid-filament capacity. 7.2 $\mu\mu$f is taken as the value of that capacity. This was measured for the particular triode used (Philips A415) mounted in a support of small capacity. For the same triode the grid-plate and plate-filament capacities have values 6.3 $\mu\mu$f and 3.8 $\mu\mu$f, respectively. Curve $f(\omega) = 1/\omega c$ (which appears as a straight line parallel to the axis of the abscissa, in view of the short interval of the frequency considered) in reality intersects curve β' at two points. This curve (β') is quite similar to that of β in Fig. 5 and hence these would be two branches of the same curve, of which β' has a value less than $1/\omega C$; however, of the two, only the first is of interest (and is the only one drawn in Fig. 8), for, to the points of the second there correspond values of α', so high, that ψ and ψ' cannot have the value necessary to bring V_c in phase with E. Therefore, the frequency of the oscillation will be slightly less than frequency f_0, corresponding to the first point of intersection between curve β' and curve $1/\omega C$ and the nearer it approaches this, the nearer the angle described between Z_aI_a and E will approach $\pi/2$.

It is, therefore, obvious that $\tan\psi = \dfrac{1}{\alpha'}\left[\dfrac{1}{\omega c} - \beta\right]$ and that $\tan\psi' = \beta'/\alpha'$. In Fig. 8 the values of $\tan\psi$, $\tan\psi'$ and the angle $(\psi+\psi')$ are plotted against frequency. The curve of $(\psi+\psi')$ is of particular interest because it shows just how the state of equilibrium mentioned in section 5 is attained. Whenever the angle between $Z_a I_a$ and E varies sufficiently slowly with the frequency in such a manner that it can be

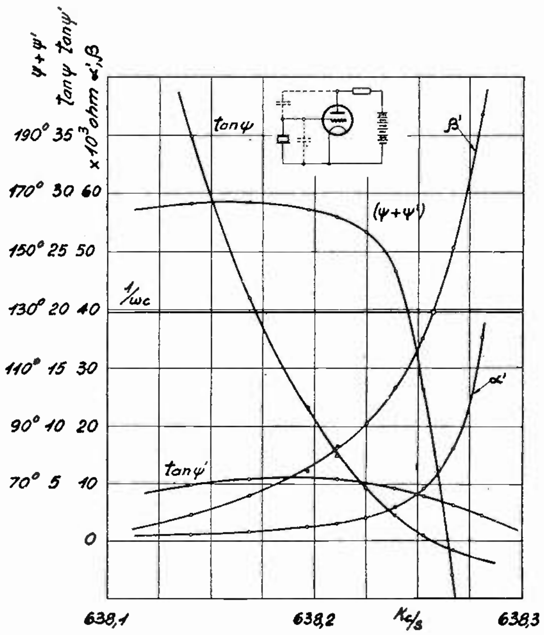

Fig. 8.

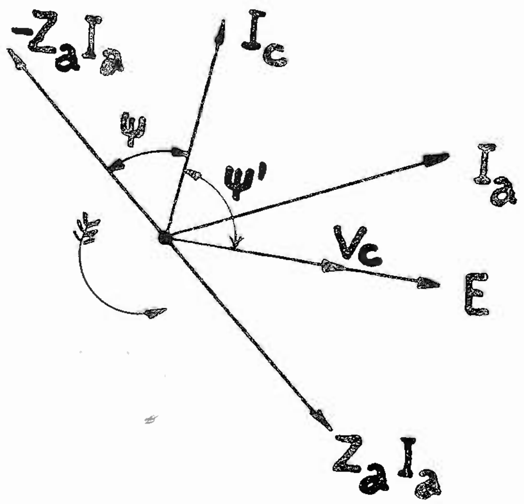

Fig. 9.

held constant in the narrow frequency range considered, the condition of stability is satisfied when the angle $\psi+\psi'$ increases as the frequency diminishes.

(7) *Quartz connected between grid and plate.* In this case the system

oscillates only when the anode circuit impedance is capacitive. Fig. 9 shows, following the procedure in the previous case, that Z_c also must be inductive, but also that it should have a reactance greater than that of the grid-filament capacity. Therefore, in this case, the frequency must be slightly higher than f_0 and the nearer it approaches, the more nearly the angle between $Z_a I_a$ and E will approximate $\pi/2$.

In Fig. 10 curves of α' and β' have been drawn for this case corresponding to those of Fig. 8 for the previous case, from which they differ in that the grid-plate capacity is now in parallel with the quartz (nevertheless, the new frequency f_0 is almost identical with the preceding). Fig. 10 also shows curves for tan ψ, produced by the customary re-

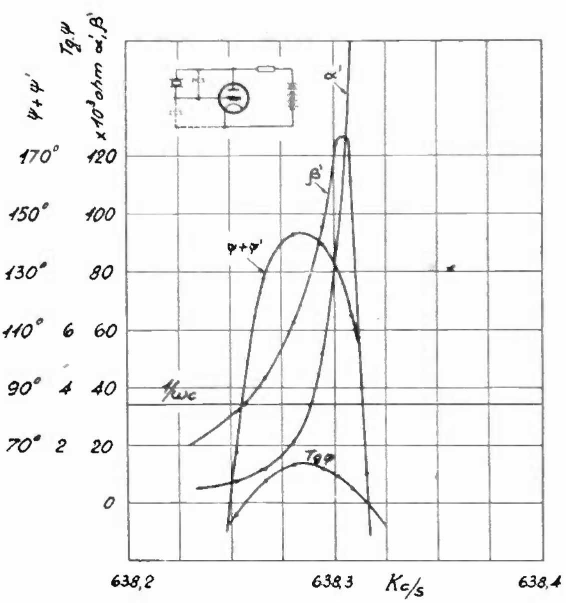

Fig. 10.

lation with inverted sign, and for the angle $(\psi+\psi')$. Angle ψ' is very near $\pi/2$ and is determined by the power factor of the condenser C with sufficient approximation we can consider it equal to $\pi/2$.

(8) The modification of the Pierce scheme investigated by Terry,[1] in which a resistance is inserted in the anode circuit in place of the tuned circuit, (Fig. 11), functions similarly in every respect to the scheme with an oscillating circuit in which the capacity has a higher value than needed for resonance. This becomes possible only if the quartz is inserted between the grid and the plate and the oscillation frequency is slightly higher than f_0. With a simple resistance in the anode circuit, $-Z_a I_a$ would be in opposite phase to E and as the angle $(\psi+\psi')$ is always smaller than π, V_c could not possibly be in phase with

[1] *Loc. cit.*

E. However, it would be necessary to have a capacity in parallel with the resistance (the plate-filament capacity of the triode would suffice, together with that of the resistance), making $Z_a I_a$ not in phase, but slightly behind E, and, therefore, making V_c in place with E.

It has been observed experimentally that by increasing the resistance inserted in the anode circuit, the oscillation frequency approaches f_0; in fact, the angle between $Z_a I_a$ and E increases, for the ratio between

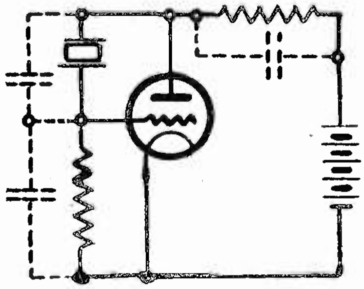

Fig. 11.

the impedance of the capacity and the resistance diminishes, and at the same time the internal resistance of the triode will increase, in view of the fact that the voltage effectively applied to it diminishes.

(9) The simplified hypothesis, $I_c \ll I_a$ is not generally valid in an actual circuit, since I_c is often of the same order of magnitude as I_a. Nevertheless it is logical to assume that this will not alter the mechanism of the behavior of the piezo oscillator.

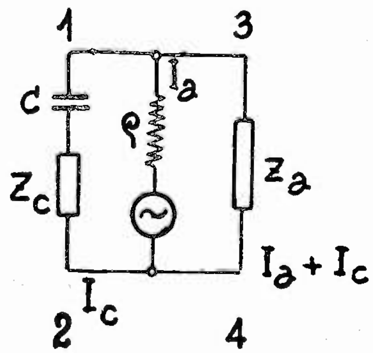

Fig. 12.

It will be remembered that the branch 1-2 of the circuit, including C and Z_c (Fig. 12) always has a reactance of a sign contrary to that of Z_a. Hence, in order that the reactance of Z_a be made enough smaller than that of the branch 1-2 of the circuit and that the resistance of Z_a be smaller than its reactance, the total reactance of the two branches 1-2 and 3-4 in parallel, will have the same sign as that of Z_a and it will, therefore, be greater than this. Moreover, the resulting reactance of the two branches in parallel will be greater than that of Z_a, and it may, therefore, be stated that the nonvalidity of the hypothesis $I_c \ll I_a$ involves a decrease of the angle by which tension V_a at the peaks of Z_a is out of phase with E. Hence, the angle $(\psi+\psi')$ between $-V_a$ and V_c should be increased and the frequency will differ slightly from the value f_0 than calculated (Figs. 8 and 10).

(10) A comparison of the results of the analysis with the experimental behavior of the piezo oscillator is obtained by measuring the frequency of the quartz oscillator with a Philips A415 triode with an inductance in the oscillating circuit having a value of 85 μh.

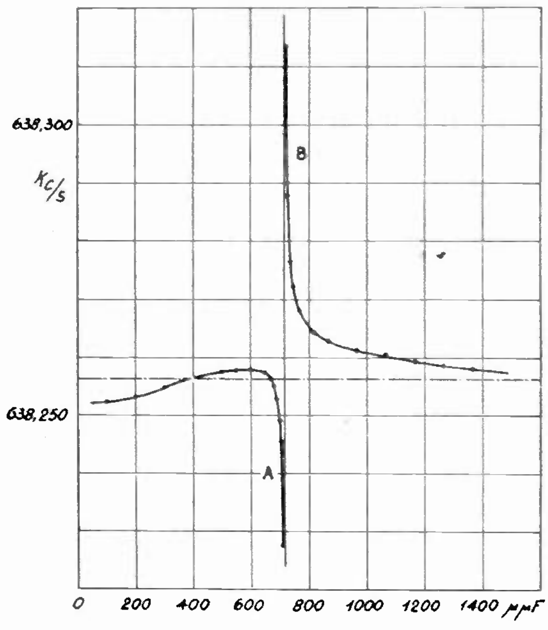

Fig. 13.

The values of oscillation frequencies were traced by a curve in Fig, 13 for the various values of the capacity of the oscillating circuit and by the two schemes, with a crystal inserted between the grid and filament (curve A) and with crystal between grid and plate, (curve B). In the latter case the oscillation frequency will be greater than in the former, in accordance with the theory explained.

The vertical line indicates the value of the capacity at which the oscillating circuit is in resonance.

The horiziontal dash and dot line shows the value of frequency f_0, already determined by the curve of the crystal; the agreement between theory and experiment is noteworthy. Curve A theoretically should be entirely below the horizontal line and should diverge more and more, with increasing capacity of the oscillating circuit, owing to the fact that its reactance increases when it approaches resonance and, therefore, the angle by which V_a is out of phase with E decreases.

Such a partial discrepancy between the theory and the experimental results can only be explained by assuming that the decrement of the quartz increases as the amplitude of the oscillation increases, which supposition is in perfect accord with the results of the measurements of the decrement, according to Watanabe.[4]

(11) The method followed in connection with the study of piezo oscillators permits us to determine the influence which the constants of the circuit and the decrement of the quartz have on the oscillation frequency of the circuit. We shall take up the effect of the decrement more in detail in Part II.

It has already been seen how the frequency varies with the variation of the capacity of the oscillating circuit. A variation of the internal resistance ρ of the triode corresponds to a variation in the contrary sense of the impedance of the anodic oscillating circuit and, therefore, the oscillating frequency should approach f_0 by an increase of ρ and again become farther removed when a decrease occurs; the experimental verification of this deduction has been tried but without success, perhaps because the variation of ρ causes variations in the amplitude of oscillation and, therefore, of the quartz vibration and consequent variations of the decrement which their effects mask. Instead, a variation of grid-plate and grid-filament capacity of the triode produces a variation of this same frequency f_0 and, therefore, variations between them similar to both circuit schemes, as already confirmed by experimental results.

Part II

(12) Among the causes of frequency instability of piezo oscillators one with which we are most concerned is the variation of the decrement of the quartz[5] (due to the adjustment of the plate in the holder), not so much on account of the magnitude of its influence, as on account of its irregularity and uncontrollability.

Consequently, a study of the influence of the decrement on the frequency of oscillation, is of interest in that, by bringing to light the nature of the phenomenon, it may lead to a method of limiting its influence.

[4] Y. Watanabe, "Piezo-electric resonator," Proc. I. R. E., **18**, 695; April, 1930.

[5] F. Vecchiacchi, "Stabilità di frequenza dei campioni piezo-electtrici," *L'Elettrotecnicha*, **15**, 462; June, 1928; *I. E. C. Publications*, No. 43.

The study is essentially analytical, for, a purely experimental research would be well-nigh impossible in this case and, at the most, the experiment would only confirm the analytical results.

(13) The impedence of the electric circuit equivalent to the quartz is expressed by:

$$Z = \frac{1}{\omega C_2'} \frac{\left[1 + \frac{C_0'}{C'} q\right] + \gamma \frac{C_0'}{C'} \phi}{-\frac{C_1'}{C'} \phi + j\left[1 + \frac{C_1'}{C'} q\right]} \tag{13-1}$$

whence,

$$\phi = R'C'\omega \qquad q = (1 - \omega^2 L'C') \qquad C_0' = C_1' + C_2' \tag{13-2}$$

If $Z = \alpha + j\beta$, we shall obtain from (2–1),

$$\left.\begin{aligned} \beta &= \frac{1}{\omega} \frac{C_1' + C_2'}{C_1'C_2'} \frac{\phi^2 + \left[q^2 + q\left(\frac{C'}{C_0'} + \frac{C'}{C_1'}\right) + \frac{C'^2}{C_0'C_1'}\right]}{\phi^2 + \left[\frac{C'}{C_1'} + q\right]^2} \\ \alpha &= \frac{1}{\omega} \frac{C'}{C_1'^2} \frac{\phi}{\phi^2 + \left[\frac{C'}{C_1'} + q\right]^2} \end{aligned}\right\} \tag{13-3}$$

Differentiating the second part of (13–2) we shall obtain, for frequency values not too different from the series resonance frequency, f_s:

$$q = -\left(\frac{C'}{C_0'} + 2\frac{df}{f_s}\right) \tag{13-4}$$

from which we can calculate the value of q for all values of frequency f, as $df = f - f_s$. By means of (13–3) we are now in a position to determine α and β for each frequency value and the various values of ϕ, as functions of the equivalent electrical constants of the plate.

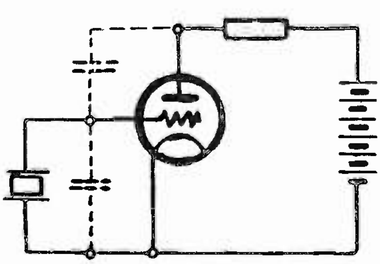

Fig. 14.

By means of (6–1) we shall now calculate α' and β', components of the impedance of the quartz and of the grid-filament capacity in parallel

(Fig. 14). If these values of β' are now plotted in a curve as functions of the frequency for various values of the parameter ϕ, (Fig. 15), the intersections of these curves with $f(\omega)=1/\omega C$ will determine the limiting frequency of oscillation f_0, corresponding to each value of ϕ.

(14) The curves shown in Fig. 15 have been constructed on the basis of electric constant equivalents of the same plate considered previously:

$$L'=2.95 \text{ henry} \qquad C'=0.021\ \mu\mu\text{f}$$

$$C_1'=16\ \mu\mu\text{f} \qquad C_2'=67\ \mu\mu\text{f},$$

giving ϕ various values between 0.3×10^{-4} and 0.8×10^{-4} and assuming a value of 7.2 $\mu\mu$f for the grid-filament capacity.

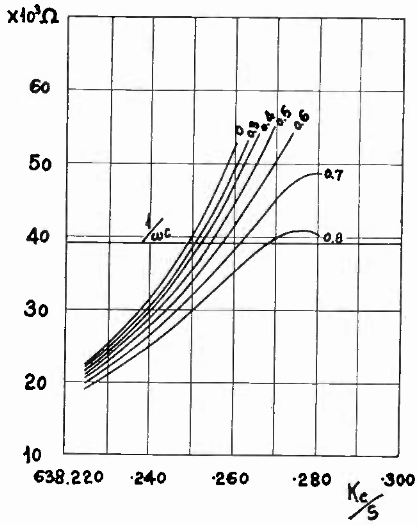

Fig. 15.

In Table I for different values of ϕ, the differences between frequency f_0 and the values this would assume in the ideal case of $\phi=0$, are shown. The plate-grid capacity was taken as 6.4 $\mu\mu$f.

TABLE I

$\phi=1^{-4}\times$......	0.3	0.4	0.5	0.6	0.7	0.8
$\Delta f/f_0=10^{-3}\times$...	2.4	4.7	8.2	12.3	19.1	29.6

With increasing the decrement, the value of f_0 increases. It will be seen in Fig. 15, that an increase of the grid-plate capacity C will slightly decrease the influence of the decrement on the frequency f_0. Vice versa, an increase of the grid-filament capacity C_g is very detrimental, as shown in Table II, on which, as usual the values of $\Delta f/f_0$ for three different values of C_g are shown, namely, 5, 7.2, 10 $\mu\mu$f and for $C = 6.4$ $\mu\mu$f.[6]

TABLE II
Values of $\Delta f/f_0 \times 10^6$ for values of ϕ given in first row

$\phi = 10^{-4}\alpha \times \ldots$	0.3	0.4	0.5	0.6	0.7	0.8
$C_g = 5\mu\mu f$	1.9	2.4	3.9	6.0	8.7	12.2
$C_g = 7.2\mu\mu f$	2.4	4.7	8.2	12.3	19.1	29.6
$C_g = 10\mu\mu f$	3.4	6.6	10.5	16.6	—	—

$\Delta f/f_0 \times 10^6$

The true frequency of the piezo oscillator differs slightly from the value f_0 (sec. 6) and it will vary further, as the impedance of the anodic circuit is made higher or the decrement of the quartz is increased, for a greater difference between the frequency of oscillations and f_0 is necessary to obtain the proper angle $\psi + \psi'$ between the voltage, the peaks of the anodic circuit, $-Z_a I_a$, and the grid voltage V_c. A variation of the quartz decrement will have a twofold effect on the oscillation frequency: the first will be manifest in the variation of limiting frequency f_0, and the second by the difference between the true oscillation frequency and f_0. In the case of a piezo oscillator with the quartz placed between the grid and filament, the above two effects act in opposition owing to the fact that the oscillation frequency is less than f_0, and one will be more effective than the other depending on the impedance values of the anodic circuit and of the internal resistance of the triode. In this manner an increase of the decrement of the quartz may produce an increase of the oscillation frequency if the anodic oscillating circuit is kept at a distance from resonance, and a decrease if it is brought nearer to resonance. If, on the other hand, the quartz is inserted between the plate and the grid, the two effects are in the same direction.

(15) A thorough experimental investigation of these deductions is not possible because, although it may be easy to produce considerable variations of the decrement of the quartz by tapping or disturbing the quartz holder, it is, nevertheless, difficult to determine their magnitudes, especially since only a slight modification in the adjustment of the quartz in the holder suffices to cause important decrement variations. Nevertheless, sufficiently satisfactory experimental verifications have already been obtained on divers points.

In the experiments a quartz holder has been used consisting of two

[6] The values of $\Delta f/f_0$ for $C_g = 10$ $\mu\mu$f and $\phi = (0.7 - 0.8)$ are lacking. for under these conditions the curve β' is much lower than $1/\omega C$. This, however, does not prevent the circuit from oscillating since it has been seen that if the anodic circuit has a sufficiently high equivalent inductance, the required value of angle $\psi + \psi'$ corresponds to a value of β' which is less than $1/\omega C$.

parallel brass plates with machined surfaces and separated at a distance of 4.65 mm in such a manner as to provide a space of 0.15 mm between the upper plate and the quartz. The plate of quartz was of the usual type ($33 \times 33 = 4.5$ mm) and oscillated on its highest frequency. The oscillation frequency was measured by another piezo-oscillator by means of a direct-reading frequency meter.[7]

It was ascertained with both insertion schemes, that when lightly tapping the quartz holder, there were produced, when the oscillating circuit was at a distance from resonance, significant frequency variations with jumps between extreme values, ranging up 1 part in 35,000 and in exceptional cases when the decrement was particularly high, 1 in 25,000. If these results are compared with the data in Table I the size of the variations of the decrement can be appreciated. The conditions

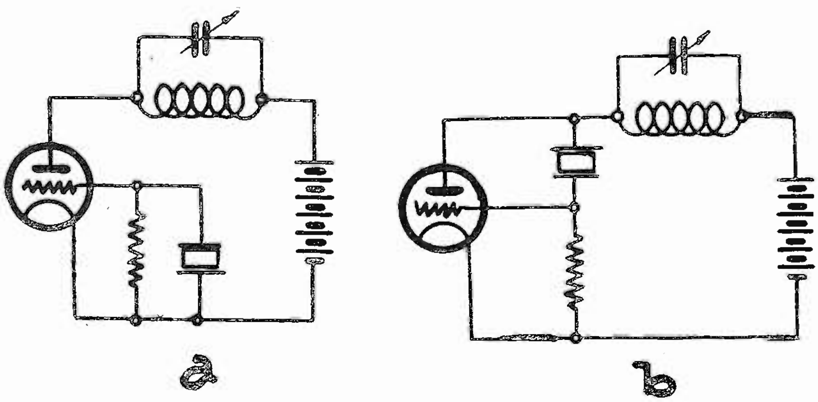

Fig. 16.

governing the adjustment of the plate and corresponding to the important low values of the decrement are not thoroughly constant and if the quartz is left undisturbed it will tend to establish itself in an intermediate condition.

By means of scheme (a) (Fig. 16) on the assumption that the capacity of the oscillating circuit increases, it would appear that the effect of the variations of the decrement on the frequency decreases; nevertheless, too much importance should not be attached to the result, owing to reasons already advanced. Vice versa, the experimental confirmation obtained of the inversion of the effect, when the oscillating circuit is almost in resonance, is of greatest interest and doubtless. With scheme (b) instead there was no inversion noticed.

[7] F. Vecchiacchi and F. Guarnaschelli, "A heterodyne frequency meter and a direct reading frequency meter for the acoustic scale. "*L'Electrotecnica,*" **17**, 224; April 5, 1930; and *I. E. C. Publications*, No. 48.

(16) The decrement exerts a lesser influence on the two values of the frequency at which the admittance of the quartz becomes highest and lowest.

The admittance $|A|$ is expressed as a function of the electric constant equivalents of the quartz, by

$$|A|^2 = \omega^2 C_2'^2 \left(\frac{C_1'}{C_0'}\right)^2 \frac{\phi^2 + \left(q + \dfrac{C'}{C_1'}\right)^2}{\phi^2 + \left(q + \dfrac{C'}{C_0'}\right)^2} \tag{16-1}$$

The maximum of $|A|^2$ is obtained by that value of q (q being a function of the frequency) which is produced by

$$o = \frac{d}{dq}|A|^2 \tag{16-2}$$

and, therefore,

$$o = q^2 + q\left(\frac{C'}{C_0'} + \frac{C'}{C_1'}\right) + \left(\frac{C'^2}{C_0'C_1'} - \phi^2\right) \tag{16-3}$$

which, solved for q, gives:

$$q = -\frac{1}{2}\left[\left(\frac{C'}{C_0'} + \frac{C'}{C_1'}\right) \pm \sqrt{\left(\frac{C'}{C_0'} - \frac{C'}{C_1'}\right)^2 + 4\phi^2}\,\right] \tag{16-4}$$

and, finally, in virtue of (13–4),

$$\frac{df}{fs} = \frac{1}{4}\left[\left(\frac{C'}{C_1'} - \frac{C'}{C_0'}\right) \pm \sqrt{\left(\frac{C'}{C_1'} - \frac{C'}{C_0'}\right)^2 + \phi^2}\,\right]. \tag{16-5}$$

It will be remembered that $1/2(C'/C_1' - C'/C_0') = \Delta f/f_s$, in which Δf is the difference between the parallel resonance frequency and that of the series resonance, as results from (1–1) and (1–2); an examination of (16–5) shows that the two frequencies of the highest and lowest admittance are respectively smaller and greater than the two series and parallel frequencies; and these differences will increase more as the decrement is greater and, in the same manner for the same value of the latter.

The following frequency differences for the various values of ϕ were calculated for the usual quartz plate:

TABLE III

$\phi = 10^{-1}\times$	0.3	0.4	0.5	0.6	0.7	0.8
$\frac{df}{fs} = \pm 10^{-6}\times$..	0.42	0.75	1.17	1.68	2.28	2.98

and the minus sign will be used for the maximum admittance and the plus sign for the minimum. Comparing the above table with Table I, it will be noticed that, with a scheme permitting the quartz to oscillate on the frequency of the highest or lowest admittance the influence of the decrement on the frequency can be reduced to almost 1/10 of that obtained with Pierce's schemes. It must be remembered that with these schemes it is advantageous, for other reasons, to hold the oscillating circuit at a distance from resonance and when this is done with

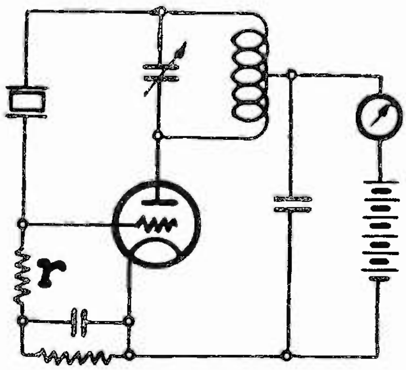

Fig. 17.

scheme (a), Fig. 16, which is the most stable and the most used arrangement, it is in the most unfavorable condition in regard to the influence of the decrement.

(17) The quartz can be made to oscillate at the frequency at which the admittance is lowest with the circuit disposed as in Fig. 17, *the peculiarity of which consists not in the arrangement itself, but in the value of the resistance r, which should be about the size of the minimum impedance of the quartz.*

If we examine the equivalent scheme of this circuit we will discover that when the oscillating circuit is perfectly syntonized, I_a will be in phase and V_a in opposition to E and, therefore, V_a' considered as the e.m.f. agent in the left mesh of the circuit is in opposition to V_a and, therefore, in phase with E. The oscillation frequency of the system will then be that by which the quartz Z_c equals a pure ohmic resistance, and V_c will be in phase with E. Moreover, on account of the condition of stability the quartz must become capacitative by a decrease of frequency, inductive by an increase. These requirements can only be met by the value of the frequency at which the admittance is at its highest as is shown by the comparison between (16–3) and the first part of (13–3) and Figs. 4 and 5.

It is necessary that the resistance r be noninductive and noncapacitative and that the grid-filament capacity which is in parallel with r be reduced as low as possible. It would also be well for r to have a very low value to reduce the effect of the parasitic capacities. With the I.E.C. type of quartz plate producing a minimum impedance of 685 ohms, a resistance r of approximately 800 ohms (Loewe 0.5 watts) is used. When this is decreased to less than 600 ohms the quartz will no longer oscillate, as there will not be sufficient coupling between plate and grid.

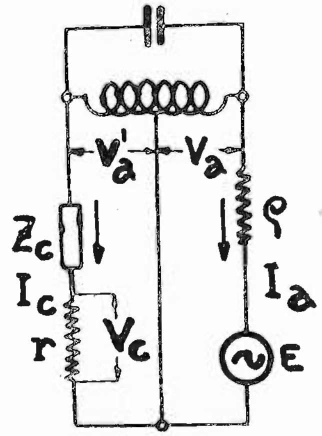

Fig. 18.

In order that the voltages V_a and V_a' between the two extremities of the oscillating circuit and the center tap be exactly in opposition the oscillating circuit should have a rather small L/C ratio and a low decrement. It is also advisable that the triode have a low internal resistance; a Philips B405 may be used.

The tuning of the anode circuit should be found by the minimum of the anode current (if the polarization of the grid is obtained by shunted resistance as shown in Fig. 17.) By varying the capacity of the oscillating circuit around the resonance value, the oscillation frequency varies as shown in Fig. 19. If a simple, nonbalanced milliameter is used to measure the anodic current, the lowest point will be determined with a frequency error smaller than $2/10^6$, if the polarization resistance has a sufficiently high value (100,000 ohms); if, on the other hand, a balanced microameter of 200 μa is used, a precision of $0.4/10^6$ is reached. The frequency corresponding to the minimum of the anodic current is not exactly that at which the admittance of the quartz is maximum, although it varies only by about $-2/10^6$. This is probably due to the parasitic capacity which is in parallel with resistance r.

Placing an external capacity across the resistance r, the oscillating

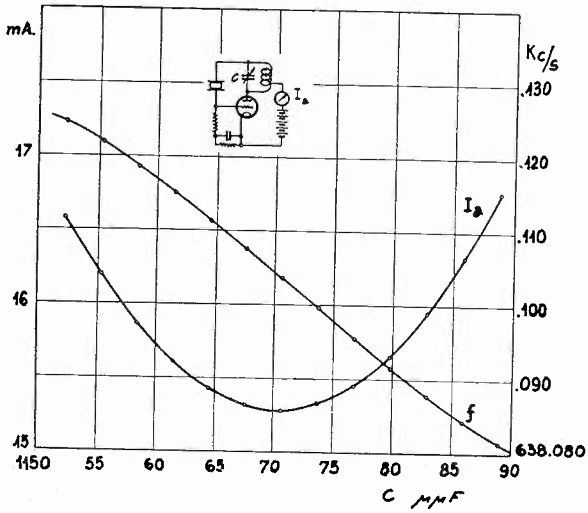

Fig. 19.

frequency decreases (Fig. 20). The effect of a capacity in parallel with the quartz is negligible, for, even at 50 $\mu\mu$f there exists a variation of the frequency smaller than $1/10^6$.

(18) In order to arrive at definite experimental results concerning the influence of decrement on frequency with the scheme described

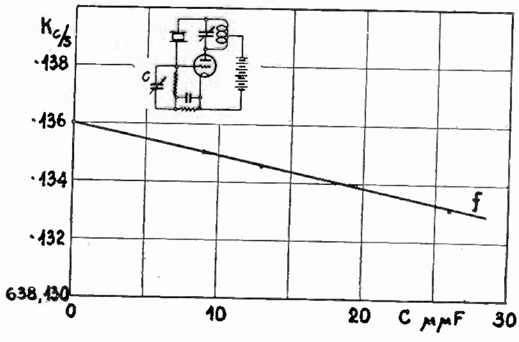

Fig. 20.

above as well as with that of Pierce a circuit arrangement was used by means of which it was possible to switch from one arrangement to another in rapid succession. In this manner it was possible to verify that an equal variation of decrement produces with the Pierce scheme, with the quartz inserted between grid and filament, a frequency variation of about ten times that with the new scheme. By lightly tapping the quartz holder it was observed that the frequency variations were much less in the latter case being of the order of 1/350,000, and in the former, about 1/25,000. It, therefore, appears that the experiment confirms the conclusions obtained from the study of the equivalent circuits.

(19) The piezo-oscillator scheme described above, in addition to affording a noteworthy interest or fact because of its application to frequency standards, will be of some use in the field of quartz measurements. Moreover, by varying the capacity of the oscillating circuit, the

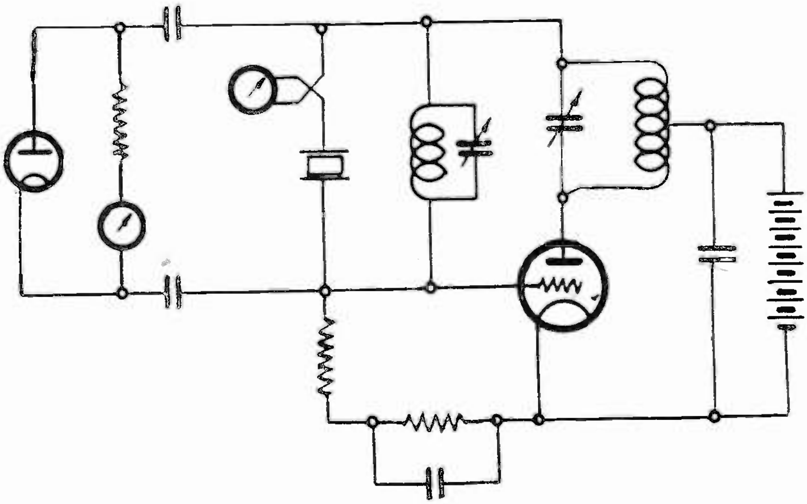

Fig. 21.

oscillation frequency can be varied somewhat around the point of the highest admittance; and by measuring the quartz current and the voltage between the electrodes it is possible to obtain a part of the resonance curve, namely, sufficient to determine the decrement of the quartz plate and the equivalent electric constants. In Fig. 21 is shown the arrangement of the circuit in the following manner: the quartz current is measured by a thermocouple and the voltage by a tube voltmeter for peak voltages; the oscillating circuit in parallel with the quartz, syntonized on the oscillating frequency, avoids the errors due to harmonics. This method of measuring has the advantage over the usual system, with separate oscillator, that the frequency stability does not have to be worried about.

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THE OPERATING FREQUENCY OF REGENERATIVE OSCILLATORY SYSTEMS*

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Summary—*The operating frequency of regenerative oscillatory systems is quantitatively derived in terms of the natural frequency, the damping constant, and the phase of the driving force. As an example the results are used to calculate the change in rate of a pendulum clock due to a given variation in the phase of the driving impulses. Applications to other types of systems are briefly indicated.*

HORTON[1] investigated qualitatively the behavior of regenerative oscillatory systems by means of a graphical method. He found that the operating frequency of the system differs from the natural frequency of the vibrating element by an amount which depends upon the damping of the vibrating element and upon the phase of the regenerative driving force with respect to the velocity of the vibrating element. Although he dealt with an electrical system the results are quite general and apply equally well to other types of regenerative systems such as clocks, tube driven tuning forks, and quartz oscillators. In the following treatment the quantitative relations are derived and a few applications are briefly indicated.

Following Horton it will be convenient to consider the regenerative system as being made up of a frequency element and an amplifier element. The frequency element controls the input to the amplifier element. The output of the amplifier element is coupled to the frequency element and supplies sufficient power to maintain the system in an oscillating steady state condition. However, in practice the functions of the two units are not so clearly separated. The amplifier element frequently exhibits mass, resistance, and restoring force properties, which are added to those of the frequency element by virtue of the coupling that exists between the two. In applications of the theory it must therefore be considered that the results refer to the effective elements rather than to the actual ones. The manner of determining the effective elements will be indicated later on.

In the steady state condition a regenerative system may be treated as a case of forced vibration with the frequency element as the vibrator and the amplifier element as the driving force. The fact that the

* Decimal classification: R133. Original manuscript received by the Institute, December 23, 1930.

[1] *Bell Sys. Tech. Jour.*, **3**, 508; July, 1924.

amplifier input is excited by the frequency element itself does not affect the argument, since in the steady state the frequency is constant and consequently the source of excitation immaterial so long as the given phase relations are maintained. Accordingly the equation of motion for the steady state is

$$\frac{d^2x}{dt^2} + 2\epsilon\frac{dx}{dt} + \omega_0{}^2 x = A \sin \omega t \tag{1}$$

where x is the displacement, ϵ the damping coefficient, ω_0 the undamped frequency of the vibrating element, ω the operating frequency of the system, and A a constant. The equivalent form of (1) in the case of electric oscillators is

$$\frac{di}{dt} + \frac{R}{L} i + \frac{1}{LC}\int idt = \frac{E_0}{L} \sin \omega t. \tag{1a}$$

The solution of (1) for velocity is

$$\frac{dx}{dt} = \frac{A\omega \sin\left[\omega t - \left(\delta - \frac{\pi}{2}\right)\right]}{\left[(\omega_0{}^2 - \omega^2)^2 + 4\epsilon^2\omega^2\right]^{1/2}} \tag{2}$$

and,

$$\tan \delta = \frac{2\epsilon\omega}{\omega_0{}^2 - \omega^2} \tag{3}$$

$\delta-\pi/2$ is the phase displacement of the velocity of the vibrating element with respect to the driving force. Therefore $\pi/2-\delta$ is the phase displacement of the driving force with respect to the velocity of the vibrating element. Letting $\phi=\pi/2-\delta$, equation (3) gives

$$\tan \phi = \frac{\omega_0{}^2 - \omega^2}{2\epsilon\omega} \tag{4}$$

which is the desired general relation. In systems which are designed for frequency stability ϕ is always small, so that in these cases $\tan \phi=\phi$ and (4) becomes

$$2\epsilon\omega\phi = \omega_0{}^2 - \omega^2. \tag{5}$$

Neglecting terms of higher order in ϕ the solution of (5) is

$$\omega = \omega_0 - \epsilon\phi. \tag{6}$$

Substituting frequencies for angular velocities

$$\nu = \nu_0 - \frac{\epsilon\phi}{2\pi}. \tag{7}$$

Equation (7) is the relation which is the most generally useful. It is in accordance with the results of Horton.

It will be noticed by reference to (1) that the driving force has been assumed to be of a sinusoidal form. As a matter of fact this condition does not exist in practice. In electrical oscillators of known forms a strictly sinusoidal driving force indicates a condition of unstable equilibrium or else one in which the amplitude is varying with the time. In general the steady state equilibrium condition is reached only when the output is no longer proportional to the velocity of the vibrating element. In electrical oscillators this condition is brought about by overloading of the amplifier. In mechanical systems of the clock type the output from the amplifier, which in this case is an escapement mechanism, is always independent of the amplitude. Hence the departure from sinusoidal form is greater in these than in any other form. In general, the driving force is of complex form and may be represented by a Fourier series. Equation (1) then takes the form

$$\frac{d^2x}{dt^2} + 2\epsilon\frac{dx}{dt} + \omega_0{}^2x = \sum_1^n A_n \sin(n\omega t + \delta_n). \tag{1b}$$

Since (1b) is linear the solution is the sum of the solutions for the separate terms of the series. The motion of the vibrating element due to one term of the series is therefore independent of that due to other terms of the series. Consequently, in computing the fundamental frequency of the system it is necessary to consider only this frequency and its phase. The validity of (4) and (7) is therefore independent of the wave form.

As an example of the use to which the theory may be put it will be instructive to compute the change in rate of a clock due to a given change in operating conditions. A good pendulum clock is a type of mechanical regenerative system in which the actual elements are very nearly equivalent to the effective elements. The pendulum is the vibrating or frequency element and the escapement mechanism is the amplifier element. It is evident from (7) that variations in the rate of the clock will depend upon the damping of the pendulum and upon the phase of the driving impulses communicated by the escapement. It will be assumed that the pendulum has an undamped natural frequency $\nu_0 = 0.5$ cycles per second and a damping factor[2] $\epsilon = 1.5 \times 10^{-4}$. It will be

[2] A simple way to determine ϵ is to measure the time t required for the amplitude of the oscillations to die down to $1/n$ of their value when the driving force is released. ϵ is then given by

$$= \frac{\log_e n}{t}.$$

assumed further that on account of dust, wear, or imperfections of the escapment the time of application of the driving impulses varies 0.01 second from that which is maintained when the clock runs at its normal rate. The resulting variation in phase is $\Delta\phi = \pi \times 10^{-2}$ radians. From (7) the corresponding increment in frequency is $\Delta\nu = \epsilon\Delta\phi/2\pi = 7.5 \times 10^{-7}$ cycles per second. The change in daily rate is $\Delta\nu/\nu \times 86{,}400 = 0.130$ second per day.

In the foregoing example the effective elements are easily derived from the actual elements. This will be found true for most systems which are designed for frequency stability. In other systems such as the various tube oscillators, the effective elements may depart considerably from the actual elements due to additions from the amplifier and coupling circuits. In order to determine the effective elements in these cases it is necessary to set up the equation of the system in the form of (1). The desired elements are then the coefficients of dx/dt and x. Furthermore, in calculating the effects of circuit variations in these systems it must be remembered that usually the elements are not independent of each other. Thus, for example, a variation of filament emission or of plate voltage will not only change ϵ and ϕ but ν_0 as well. With quartz oscillators the mechanical constants of the crystal element may be taken for the effective elements to a first approximation, although for higher precision it is necessary to take into account the effects of the electric circuit elements.

Proceedings of the Institute of Radio Engineers
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SUPPLEMENTARY NOTE TO THE "STUDY OF THE HIGH-FREQUENCY RESISTANCE OF SINGLE LAYER COILS"*

By

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SINCE the publication of the above paper in the December, 1930, number of the Proceedings of the Institute of Radio Engineers, the attention of the authors has been called to two articles on the alternating-current resistance of solenoidal coils, by S. Butterworth, which we regret to state were overlooked in the preparation of our paper.

In the first of these articles[1] Butterworth has explained the disagreement of results, calculated by a formula,[2] derived by him in 1921, and the experimental results of Hickman[3] as being due to a misinterpretation by the latter of the scope of this formula. The measurements of Hickman were carried out with long coils, while the constants in the formula cited apply only to coils whose length is less than their diameter. Butterworth shows that, properly extended, his formula correctly represents Hickman's experimental values of the resistance ratio.

In the second of these articles[4] Butterworth gives the derivation of this extended expression, formula (30), together with that of two other formulas, (43) and (53). It will be convenient to refer to these formulas by their numbers.

Formula (30) applies to a single layer coil of a large number of turns. Both the tangential and the transverse components of the magnetic field of the coil are taken into account in its derivation.

The second formula (43) applies to short coils of a small number of turns. Limitations of its use are that the ratio of the wire diameter d to the pitch of the winding D does not exceed 0.6, and that the frequency is not very high. The case of short coils with closer winding is covered by Butterworth's formula (53) which holds up to $d/D=0.8$, or even 0.9 if the frequency is not too high.

* Decimal classification: R144. Original manuscript received by the Institute, April 11, 1931.

[1] *Phys. Rev.*, **11**, 752–755; June, 1924.
[2] *Phil. Trans.* (London), **A222**, 57, 1921.
[3] *Bureau of Standards Scientific Paper* No. 472.
[4] *Proc. Roy. Soc.*, **A**, **107**, 693, 1925.

We have made calculations of the resistance ratio using these formulas and Butterworth's tables of constants for a number of the coils included in our measurements. The following table includes the principal data of the coils, the measured resistance ratio, and the values calculated by Butterworth's formulas. Since it is not evident in every case which of the formulas is best adapted, calculations are made in all cases by all three formulas. Coils *A* and *B* are those specifically mentioned on pages 2056 and 2057 of our article, and the measured resistance ratios cited are the values found for these coils at the Bureau of Standards and used in checking our measurements. The remaining measured values are taken from our curves.

TABLE

Coil	No. of Turns	Wire Diam. in.	Pitch in.	Freq. kc	d/D	Length Diam.	Z	Resistance Ratio			
									Calculator		
								Meas.	by (30)	by (43)	by (53)
A	45	0.051	0.0638	685	0.8	0.696	11.49	15.65	16.57	12.20	14.31
A	"	"	"	1375	"	"	16.28	37.7	25.3	17.9	22.3
B	50	0.032	0.0445	685	0.73	0.73	7.20	7.61	7.54	6.42	7.10
B	50	"	"	1375	"	"	10.20	15.8	11.22	9.30	10.54
C	20	0.032	0.0445	1000	0.704	0.238	8.71	7.08	9.18	6.70	6.98
C	20	"	"	1500	"	"	10.67	10.88	11.60	8.28	8.67
D	20	0.020	0.0298	1500	0.67	0.156	6.66	6.30	6.14	4.77	4.88
E	30	0.020	0.0298	1000	0.67	0.333	5.45	4.4	4.55	4.10	4.26

From an inspection of the table it is apparent that for the coils measured by us, Butterworth's formula (30) gives the best agreement with the measured values in all except a few cases. Coils *C*, *D*, and *E*, which have small ratio of length to diameter, would, because of this reason, be expected to lie within the scope of formula (43), but, as the table shows, they have a ratio of wire diameter to pitch which lies above the limit 0.6 prescribed for that formula. Formula (53) for which the limiting value of d/D lies higher gives a better agreement, but not as good as does formula (30), except in the case of coil *C* at 1000 kilocycles.

The parameter z, which is the same as Russell's x, and is equal to $2\sqrt{\lambda}$ in our paper, gives a rough measure of largeness of the resistance ratio to be expected in any given case. Those cases where z is greater than 10 are to be regarded as high-frequency cases, and considering that, in addition, the ratio d/D is also large for all these coils, such cases are to be regarded as lying close to the limits of applicability of the formulas. Thus these measurements furnish a severe test of the formulas.

The discrepancy between the measured and computed values of the resistance ratio for coils *A* and *B* at the higher frequency is probably to be explained as principally due to irregularities in the winding pitch at different parts of the coils. Butterworth's Table VI in the second article gives the limiting value of the coiling effect possible in coils of different lengths and ratio d/D. The limiting value is seen to increase very rapidly with increasing d/D, as this ratio approaches unity.

In our work the winding pitch is measured as an average value for the whole coil. Thus regions where the winding is closer than the average would have an effect on the resistance ratio which would not be compensated by the effect of those regions where the winding is slightly more open. Thus the measured value of the resistance ratio corresponds to an effective value of d/D which is greater than that found by measurements of the average value of the pitch, and consequently computations made by the formulas using the average value of the pitch would necessarily give a value smaller than the measured.

This explanation is further substantiated by the consideration that in the higher frequency values of the resistance ratios of Coils A and B, the observed coiling effect is, in both cases, greater than the maximum possible for the assumed pitch, as found from Butterworth's Table VI. This indicates that the ratio d/D for these coils is effectively greater than the average values quoted in the table above.

In conclusion, the general agreement of our measured values and the values of resistance ratio computed by Butterworth's formulas may be regarded as satisfactory in all cases where the assumptions of the formulas are fulfilled. For a more rigid comparison, measurements are necessary with coils of more open winding and with greater attention paid to obtaining uniformity of spacing over the length of the coil.

(Errata in the previous paper:

p. 2055, equation (15), read $\dfrac{0.707}{x}$ instead of $0.707x$

p. 2056, equation (A), last term read π^4 instead of x^4.)

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DISCUSSION OF OSCILLATION IN TUNED RADIO-FREQUENCY AMPLIFIERS*

By

B. J. Thompson

J. R. Nelson:[1] B. J. Thompson in the above article makes the following statement—"Hull[3] and more lately Nelson[4] have analyzed the conditions for stability in terms of the maximum amplification per stage, ignoring the all-important effect of feed-back on amplification." He evidently did not understand the significance of the symbol Av in either of the above articles. The effect of feed-back has not been ignored in determining the conditions for stability. The voltage amplification Av is not the amplifaction realized with a tuned input but is the amplification obtained with an untuned input or the true repeater gain of the stage, which quantity may be easily measured. The desirability of expressing the results in terms of Av will be discussed later.

The equations derived by Thompson for one stage and for two stages are exactly the same as derived by the writer. Because the writer defined part of the equation as Av does not mean that the effect of feed-back was ignored. The condition for sustained oscillation for a single tuned circuit as given by equation 24 in Thompson's article is

$$C_o = \frac{2g_1g_2}{\omega g_m} \tag{24}$$

assuming $g_1g_2 = g$ as is done in the article

$$C_o = \frac{2g^2}{\omega g_m}.$$

The above expression may be written as

$$\frac{C_o}{g_m} = \frac{2g^2}{\omega g_m{}^2}$$

or,

$$\frac{g_m{}^2}{g^2} = \frac{2g_m}{\omega C_o}$$

g_m/g is the nonregenerative amplification per stage defined as Av[3,4]. Thus the expression may be written as

$$Av = \sqrt{\frac{2g_m}{\omega C_o}}$$

which expression is exactly the same as given by the writer for $n=1$ in his expression

$$Av = \sqrt{\frac{2g_m}{n\omega C_o}}.$$

* Proc. I.R.E., 19, 421; March, 1931.
[1] Engineer, Raytheon Production Corporation, Newton, Mass.
[3] A. W. Hull, "Measurements of high frequency amplification with shielded grid pliotrons," *Phys. Rev.*, 27, 439; April, 1926.
[4] J. R. Nelson, "Circuit analysis applied to the screen grid tube" Proc. I.R.E., 17, 320; February, 1929.
[3,4] *Loc. cit.*

Similarly, his equation for two stages may be shown to be exactly the same as given by the writer.[3]

Some time ago H. A. Wheeler in his comments as a member of the papers committee, on another paper by the writer[5] showed that the expression for stable amplification should be written as

$$Av < \sqrt{\frac{2g_m}{\omega C_o} f(n)}$$

instead of

$$A < \sqrt{\frac{2g_m}{\omega C_o} \frac{1}{n}} \text{ as given by the writer.}$$

The writer, because $f(n)$ is $1/n$ for one stage and for two stages, assumed that this law held for n stages without checking further. This did not affect the validity of the results obtained in the article[5] as the results were given for a single stage. The writer checked the case since that time and found that his general expression is not quite true for more than two stages as $2f(n)$ came out to be 0.765 instead of 0.667 for three stages. The above value checks closely with that given by Thompson, 0.764.

It is hoped by the writer that Wheeler will publish his results as the value of $2f(n)$ may be found for any number of stages by his method. In other words his results would be general, so that the value of $2f(n)$ could be found for n stages.

In conclusion it might be said that Thompson's article is not a rigorous general solution of the subject which would close it. He, however, has found the value of two more stages than given by the writer for $2f(n)$ in Wheeler's general formula

$$Av < \sqrt{\frac{2g_m f(n)}{\omega C_o}}$$

as the value of his symbol A is really $2f(n)$ for the number of stages considered.

The writer believes that the results expressed in terms of Av are more useful to the set designer than when they are expressed in terms of C_o. For example, the set designer has to design a receiver to cover a certain frequency range, which determines ω. He has to do this with tubes having certain values of C_o and g_m and he does not have much control over the minimum value of C_o or the maximum value of g_m for a certain type of tube. He does, however, have considerable control up to a certain maximum value over Av. The value of Av may be easily measured for any given conditions in a single stage set-up. Thus the set designer may easily design his transformers to give stability for any combination of C_o, g_m, and ω.

As a matter of interest the constant current generator concept of a vacuum tube given in Appendix A of Thompson's proper was also given in a paper by H. A. Wheeler and W. A. MacDonald.[6]

[5] J. R. Nelson, "Note on the stability of a balanced high-frequency amplifier," Proc. I.R.E., 18, 88; January, 1930.

[6] H. A. Wheeler and W. A. MacDonald, "Theory and operation of tuned radio-frequency coupling systems," presented at 1930 Convention, Toronto; published Proc. I.R.E. 19, 738; May, 1931.

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BOOK REVIEW

Radio Theory and Operating, by Mary Texanna Loomis. Fifth Edition. 1006 pages. 803 figures. Price $4.25. Loomis Publishing Company, Washington, D. C.

This is an elementary book treating a great variety of radio subjects, as well as other subjects closely related to radio. The strong points of this book are the many circuit diagrams, charts and tables, descriptions of modern applications of radio, especially those of popular interest, amateur and commercial operating, and receiving set trouble shooting. In the discussion of elementary electricity and magnetism and principles of radio, which consume many pages, the treatment is superficial and unsatisfactory. Errors and incorrect uses of scientific terms are prevalent.

S. S. Kirby

* Bureau of Standards, Washington, D. C.

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

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

A comprehensive supplement to their general catalog and known as Catalog F, Part 2, has been recently issued by the General Radio Company of Cambridge, Mass. Part 2 describes all of the new G-R instruments and announces many important specification changes in others. Among the important new items are listed the type 513-B a-c operated beat frequency oscillator having a range of from 10 cycles per second to 10 kilocycles, the type 575-AR rack mounted piezo-electric oscillator with temperature control, the type 457-A modulation meter for determining the percentage modulation of modulated transmitters, and the type 408 oscillograph camera for use with the string oscillograph.

A sixteen-page pamphlet, Series C, Experiment C-13, issued by the Central Scientific Company of Chicago, Ill., lists laboratory equipment for the measurements of potential difference and gives a brief outline of the theory of the potentiometer. A thirty-two-page pamphlet known as Series C, Experiment C-21, describes laboratory equipment for resistance measurements and contains an outline of nine experiments on the measurement of resistance.

"Midget Condensers" is the title of a four-page folder describing small tuning and neutralizing condensers manufactured by the Hammarlund Manufacturing Company, 424 West 33rd St., New York, N. Y.

A twelve-page bulletin, known as Bulletin No. 56, published by the Operadio Manufacturing Company of St. Charles, Ill., lists a complete line of microphone amplifiers, control boxes, phonograph units, and other equipment for public address systems.

REFERENCES TO CURRENT RADIO LITERATURE

THIS IS a monthly list of references prepared by the Bureau of Standards, and is intended to cover the more important papers of interest to the professional radio engineer which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the "Classification of radio subjects: An extension of the Dewey Decimal System," Bureau of Standards Circular No. 385, which appeared in full on pp. 1433-56 of the August, 1930, issue of the PROCEEDINGS of the Institute of Radio Engineers. The classification numbers are in some instances different from those used in the earlier version of this sytem used in the issues of the PROCEEDINGS of the Institute of Radio Engineers before the October, 1930, issue.

The articles listed are not obtainable from the Government or the Institute of Radio Engineers, except when publications thereof. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. RADIO

R007 Felix, E. H. Engineering testimony before official bodies. PROC. I.R.E., 19, 851-855; May, 1931.

This paper discusses the requirements for the effective presentation of testimony on engineering subjects, and the most important qualifications of a witness on technical matters.

R100. RADIO PRINCIPLES

R111 van der Pol, B. Über die Ausbreitung elektromagnetischer Wellen (Electromagnetic wave propagation). *Zeit. für Hochfrequenz.*, **37**, 152-156; April, 1931.

An equation is derived which gives the value of the electric or magnetic field at a point in terms of the distance between the point and the antenna, and the power radiated from the latter.

R111 Sommerfeld, A. Das Reziprozitäts-Theorem der drahtlosen Telegraphie (The reciprocity theorem applied to wireless telegraphy). *Zeit. für Hochfrequenz.*, **37**, 167-169; April, 1931.

An earlier paper with the same title contained an error in calculation which is corrected here. The statement of the reciprocity theorem is, however, in no way affected

R111.2 Kiebitz, F. Über die Masseinheiten der Strahlung (On units of measure for radiation). *Zeit. für Hochfrequenz.*, **37**, 136-139; April, 1931.

It is pointed out that units for antenna radiation and wave propagation are in part based on assumptions and must be used with care in practice.

R113.1 Merritt, E; McLean, T; Bostwick, W. E. Visual studies of radio fading. *Jour. Franklin Institute*, **211**, 539-566; May, 1931.

An oscillographic method of determining the amplitudes and relative phase of two components of the resultant magnetic field at the receiving station is described, and results of observations made with signals from broadcasting stations are reported.

R113.5 Yokoyama, E. and Nakai, T. Effects of sun spots and terrestrial magnetism on long-distance reception of low-frequency waves. PROC. I.R.E., **19**, 882-890; May, 1931.

The results of a series of field intensity measurements which were conducted in the Pacific area during the period of more than a year are correlated with sun spots, magnetic storms, etc.

R113.6 Elias, G. J. Reflection of electromagnetic waves at ionized media with variable conductivity and dielectric constant. Proc. I.R.E., **19**, 891–907; May, 1931.

Using the electrical properties of the higher atmosphere which are determined by the ionization caused by the ultra-violet radiation of the sun and the corpuscular rays sent out from it, conclusions are drawn as to the reflected amplitude, and the height where electromagnetic waves are reflected. In this way results can be obtained in good agreement with the observations. The influence of hydrogen in the upper atmosphere is discussed. Finally the reflection time for a signal is calculated.

R113.6 Hartree, D. R. Optical and equivalent paths in a stratified medium, treated from a wave standpoint. *Proc. Royal Soc.*, **131**, 428–450; May, 1931.

The limitations of a ray treatment of reflection of electromagnetic waves by a stratified medium are discussed, and it is shown that a wave treatment is essential for the interpretation of some of the phenomena of reflection from the Heaviside layer.

R113.62 Joos, G. Zur Frage nach der Natur der Langzeitechos (On the nature of delayed echoes). *Zeit. für Hochfrequenz.*, **37**, 136; April, 1931.

The author suggests an experimental test to determine which of the two delayed echo theories, that due to Störmer or that due to van der Pol, is correct.

R114 Barkhausen, H. Pfeiftöne aus der Erde (Whistling tones from the earth). *Zeit. für Hochfrequenz.*, **37**, 123–125; April, 1931.

Two possible explanations are offered as to the origin of the peculiar whistling tones that are often heard through an a-f amplifier connected directly to the antenna.

R125 Dieckmann, M. Die Wirksame Höhe kurzer Linearantennen. (The effective height of short linear antennas). *Zeit. für Hochfrequenz.*, **37**, 126–127; April, 1931.

A formula is given for the effective height of a linear antenna which has a geometric length that is small in comparison with the wave length.

R125
×R270 Bäumler, M; Krüger, K; Plendl, H; Pfitzer, W. Radiation measurements of a short-wave directive antenna at the Nauen high power radio station. Proc. I.R.E., **19**, 812–838; May, 1931. *Zeit. für Hochfrequenz.*, **36**, 1–13; July, 1930.

This paper is a report on the experimental investigation of the radiation of a short-wave directive antenna at Nauen. Measurements were made on the ground and also in the air, using an airplane. The results, plotted in the form of radiation characteristic curves, were compared with calculated radiation characteristics. In the case of the horizontal radiation characteristics, the agreement between measurement and calculation is comparatively good.

R125.1 Krüger, K. and Plendl, H. Horizontale Strahlungskennlinie einer Kurzwellen-Richtantenne mit gespeistem Reflektor. (Horizontal radiation characteristic of a short-wave directive antenna with direct excited reflector). *Zeit. für Hochfrequenz.*, **37**, 142–145; April, 1931.

The experimentally determined radiation characteristic of a directive antenna with direct excited reflector is compared with that of a similar antenna with a reflector excited by induction.

R130 Pungs, L. and Schulze, K. Beobachtungen über den Einfluss der Belichtung auf die Arbeit von Elektronenröhren (Observations on the influence of light on the operation of a vacuum tube). *Zeit. für Hochfrequenz.*, **37**, 157–159; April, 1931.

A photo-electric effect was observed when screen-grid, oxide-filament vacuum tubes were subjected to intense light.

R132 Kniepkamp, H. Die Anweichungen der Verstärkerröhrenkennlinien vom $e^{3/2}$=Gesetz (The variation of the amplifying vacuum-tube characteristic from the $e^{3/2}$-law). *Telegraphen und Fernsprech Technik*, 20, 71–76; March, 1931.

The variation of the static characteristic from the $e^{3/2}$-law and the causes of such variation in different types of vacuum tubes are discussed.

R132 Oatley, C. W. The distortionless amplification of electrical transients. *Exp. Wireless and the Wireless Eng.* (London), 8, 244–249; May, 1931.

The extent of distortion in the amplification of transients by thermionic amplifiers of various types is investigated.

R133 Möller, H. G. and Hinsch, W. Reissdiagramme bei Barkhausenschwingungen und ihre Theorie (The theory of Barkhausen oscillations). *Zeit. für Hochfrequenz.*, 37, 145–149; April, 1931.

The results of a theoretical and experimental study of Barkhausen oscillations are given.

R133 Vecchiacci, F. Oscillations in the circuit of a strongly damped triode. Proc. I.R.E., 19, 856–872; May, 1931.

A study of the particular action produced by a triode oscillator when the relation of the inductance to the capacity, L/C is greater than the square of the internal resistance ρ^2, and when the reactive coupling between plate and grid is greater than the limit required for starting oscillation. The shape of the oscillation curve is clearly other than sinusoidal, the frequency is much lower than that usual in the LC circuit, and is determined essentially from the constants of the triode and from the current.

R134 Robinson, G. D. Test procedure for detectors with resistance coupled output. Proc. I.R.E., 19, 806–811; May, 1931.

A simple circuit arrangement is presented for use primarily in determining the response to modulation of a detector with resistance-coupled output.

R134 Chaffee, E. L. Small-signal detection. *Electronics*, 2, 641–643; May, 1931.

The theory of detection in diodes and triodes is reduced to its simplest form and presented with graphical aids to a visualization of the phenomena.

R139 Feldtkeller, R. and Kautter, W. Zur Theorie der Rückkopplung bei Hochfrequenzempfängern (The theory of regeneration in high-frequency radio receivers). *Elek. Nach. technik*, 8, 93–103; March, 1931.

A mathematical study of regeneration at high frequencies showing how the regenerative circuit and tube may be replaced with negative circuit elements to obtain an equivalent circuit arrangement that facilitates analysis.

R142 Wheeler, H. A. and MacDonald, W. A. Theory and operation of tuned radio-frequency coupling systems. Proc. I.R.E., 19, 738–805; May, 1931.

An improved r-f coupling system for broadcast receiving sets allows the gain of the amplifier to be held uniform, or made to vary with frequency in any desired manner consistent with the amplifying ability of the tube and the tuned secondary circuit, and without appreciable loss of selectivity.

R143 Zobel, O. J. Extensions to the theory and design of electric wave filters. *Bell Sys. Tech. Jour.*, 10, 284–341 April, 1931.

An exhaustive treatise in which the problem of terminal wave-filter impedance characteristics is considered.

R143 Band-pass filters in radio receivers. *Exp. Wireless and the Wireless Eng.* (London), 8, 233–237; May, 1931.

A brief analysis of preselector, band-pass filters is given.

R145.3 Dwight, H. B. The magnetic field of a circular cylindrical coil. *Phil. Mag.*, 7 series, No. 72, 948–957; 1931.

The collection of formulas given in this paper covers nearly all the field of a cylindrical coil.

R148 Eckersley, T. L. Radio telephony distortion. *Marconi Rev.*, 12–18, March–April, 1931.

A particular type of distortion in short-wave telephony and telegraphy reception, involving a partial disappearance of modulation with retention of carrier signal has been noticed on many short-wave stations. An attempt to explain this phenomenon is given.

R162 Bayly, B. de F. Selectivity, a simplified mathematical treatment. Proc. I.R.E., **19**, 873–881; May, 1931.

This paper gives a simple formula for finding the voltage gain of a resonant circuit at different frequencies in terms of that at resonance. Tables and curves are given in decibels below the resonant value so that calculations are quickly made.

R163 Koepping, E. D. Notes on superheterodyne design. *Radio Eng.*, **11**, 25–27; May, 1931.

It is shown that a high order performance attained at the loud speaker is made possible by point-to-point coördination in superheterodyne receiving set design.

R163 Langley, R. H. Undesired responses in superheterodynes. *Electronics*, **2**, 618–620; May, 1931.

Design problems encountered in superheterodyne receiving sets are discussed.

R163 Hermanspann, P. Berechnung der Verstärkung des Zwischenfrequenz-Gleichrichters (Calculating the amplification of the intermediate-frequency detector). *Zeit. für Hochfrequenz.*, **37**, 134–135; April, 1931.

The expansion of the detector characteristic into a series, leads to a relation between the rectifying ability and the amplifying ability of an amplifier tube. An example of this method of analysis shows that the optimum amplification of an intermediate detector is one-half the optimum high-frequency amplification of the same vacuum tube.

R165 McLachlan, N. W. The lower register in moving coil loud speakers. *Wireless World and Radio Rev.*, **28**, 479–481; May, 1931; 514–516; May, 1931.

An explanation of the peculiar properties of the loud speaker diaphragm at the lower speech frequencies is given.

R190 Winter-Günther, H. Selbsterregung von Systemen mit periodisch veränderlichen Induktivitäten (Self-excitation of systems having periodically changing inductances). *Zeit. für Hochfrequenz.*, **37**, 172–174; April, 1931.

A theoretical and experimental investigation of self-excitation phenomena that occur in oscillating circuits having periodically changing inductance.

R200. Radio Measurements and Standardization

R223 Schlesinger, K. Hochfrequenz-Messeinrichtung zur Bestimmung der Dielektrizitätskonstanten (High-frequency measuring apparatus for determining the dielectric constant). *Elek. Zeit.*, **52**, 533–534; April, 1931.

A bridge arrangement is described by means of which the dielectric constant of liquids, etc., may be measured with less than 2 per cent error.

R240 Colebrook, F. M. and Wilmotte, R. M. A new method of measurement of resistance and reactance at radio frequencies. *Jour.* I.E.E., (London), **69**, 497–506; April, 1931.

A method of measuring the radio-frequency resistance is described. It depends on the change in reactance of a valve-maintained oscillating circuit when it is coupled to the circuit under investigation.

R241 Dannatt, C. and Goodall, S. E. The permittivity and power factor
×R281 of micas. *Jour.* I.E.E., (London), **69**, 490–496; April, 1931.

Means are described by which electrode contact effect with mica may be entirely eliminated and test results on power factor and permittivity of ruby, green, and amber mica are given, together with variations produced by change of stress and temperature.

R241 Stuhlman, O. Jr. A thermionic megger with linear scale. *Jour. Franklin Institute,* **211**, 617–625; May, 1931.

A voltmeter, recording the potential across filament and plate of a vacuum tube having a negative characteristic, is calibrated to read resistances directly.

R242.1 Fortescue, C. L. and Moxon, L. A. A method of comparing ammeters at very high frequencies. *Jour. Sci. Instruments* (London), **8**, 94–97; March, 1931.

A reliable and accurate method of overcoming difficulties experienced in comparing ammeters at frequencies above one million cycles per second is described.

R243.1 de Cola, R. A dynatron vacuum-tube voltmeter. *Electronics,* **2**, 623–624; May, 1931.

A pliodynatron vacuum-tube voltmeter is described which has a greater sensitivity than the ordinary single-tube voltmeter.

R265.2 Strutt, M. J. O. On the amplitude of driven loud speaker cones. Proc. I.R.E., **19**, 839–850; May, 1931; *Exp. Wireless and the Wireless Eng.* (London), **8**, 238–243; May, 1931.

Bragg's method for measuring small amplitudes of vibration was developed technically for the measurement of amplitudes of driven loud speaker cones and several different loud speaker systems were tested.

R281 Williams, J. W. Recent dielectric constant theory and its relation to problems of electrical insulation. *Jour. Franklin Institute,* **211**, 581–606; May, 1931.

The importance of a knowledge of the chemical and capillary structures of insulating materials, if electrical energy losses in them are to be explained, is indicated, and available methods of determining molecular structure are demonstrated.

R300. Radio Apparatus and Equipment

R320.8 ×R320 Gerten, N. Radio towers and antennas. *Radio Eng.* **11**, 21–24; May, 1931.

Some notes on the electrical properties, optimum height, and base insulation of antenna towers are given.

R330 Dunsheath, J. New and better tubes. *Radio Eng.*, **11**, 28–30; May, 1931.

New types of vacuum tubes for broadcast receivers have been recently introduced. They include the variable-mu type, the pentode type, and special tubes for automobile receiving sets. The characteristics and uses of these tubes are briefly discussed.

R332 ×R133 Craig, P. H. Oscillations produced by gaseous diodes. *Electronics,* **2**, 639–640; May, 1931.

An apparatus is described, which consists of a diode and a special transformer and is capable of generating either audio or radio-frequency oscillations.

R335 Hull, R. A. Putting the pentode to work. *QST*, **15**, 16–20; June, 1931.

The construction of a small, three-tube high-frequency radio receiver of high sensitivity is described.

R335 Briggs, R. S. Using the pentode tubes in the low powered transmitter. *QST*, **15**, 27–29; June, 1931.

The functioning of the pentode as a radio-frequency power amplifier and frequency doubler is described.

R355.6 ×R550 Gerth, F. and Hahnemann, W. Moderne quartzgesteuerte Gleichwellensender (Modern piezo-controlled common frequency transmitters). *Elek. Nach. tech.*, **8**, 131–134; March, 1931.

A piezo control for common frequency broadcast stations is described. The frequency is held constant to better than 1 part in a million.

R355.7 ×R390 Everitt, W. L. Output networks for radio-frequency power amplifiers. PROC. I.R.E., **19**, 725–737; May, 1931.

The design of reactance networks to connect a resistive load efficiently to a source of power, and at the same time attenuate undesired harmonics, is accomplished by the theory of image impedances.

R357 Cocking, W. T. Frequency changers. *Wireless World and Radio Rev.* **28**, 473–476; May, 1931.

A study of different methods of frequency changing in the superheterodyne and their comparative advantages.

R357 Sammer, F. Zur Frage des Wirkungsgrades und der Nutzleistung von Frequenztransformatoren (The efficiency of frequency transformers). *Zeit. für Hochfrequenz.*, **37**, 159–161; April, 1931.

A frequency-multiplying circuit arrangement is described, in whieh several harmonics of a magnetic frequency converter are utilized to increase the output at the frequency desired.

R361 Gardiner, E. L. The performance of the Stenode. *Wireless World and Radio Rev.*, **28**, 449–451; April, 1931.

An American manufacturer has made some tests on a commercial model of the Stenode radiostat. The resulting over-all curves of selectivity and fidelity are given.

R361 Palmgren, A. Experiments with a quartz crystal receiver. *Exp. Wireless and the Wireless Eng.* (London), **8**, 250–252; May, 1931.

A detailed experimental study of the Stenode radiostat type of receiving set was made and the results of these tests are given by the author.

R361.5 Gall, D. C. New instrument galvanometer for the direct reception of very short waves. *Jour. Sci. Instruments* (London), **8**, 109–110; March, 1931.

A d-c galvanometer, crystal, and a loop of wire are used for ultra-high-frequency reception.

R365.2 Neumann, H. and Trendelenburg, F. Über Hochleistungsblatthaller (Powerful plate loud speakers). *Zeit. für Hochfrequenz.* **37**, 149–151; April, 1931.

A recent technical development in loud speakers is described.

R365.3 Smith-Rose, R. L. An automatic recorder of signals from a rotating beacon transmitter. *Jour. Sci. Instruments* (London), **8**, 81–88; March, 1931.

A description is given of an automatic apparatus which has been developed for the recording of wireless bearings from a rotating beacon transmitter of the type now in operation at Orfordness, Suffolk, England.

R380 Hormann, E. Über den Frequenzgang von Wechselstrom Messinstrumenten mit Trockengleichrichtern (On the frequency variation of a.c. instruments of the rectifier type). *Zeit. für tech. Physik*, **12**, 222–224, No. 4; 1931.

Methods of compensating the frequency error of rectifier type a-c instruments are described.

R381 Moullin, E. B. A variable-capacitance cylindrical condenser for precision measurements, and a wavemeter for short wave lengths. *Jour.* I.E.E., (London), **69**, 507–518; April, 1931.

A form of cylindrical condenser for precision measurements at very high frequencies is described.

R383 Wien, M. Über Hochfrequenzwiderstände (On high frequency resistance elements). *Zeit. für Hochfrequenz.*, **37**, 169–172; April, 1931.

It is shown that thin wire, carbon, or suitable wet cells may be used as pure resistances up to frequencies of the order of 100 megacycles per second.

R385.5 Grützmacher, M. and Just, P. Über Kohlemikrophone (Carbon microphones). *Eleck. Nach. technik*, **8**, 104–114; March, 1931.

Methods of measuring frequency and intensity response of carbon microphones are given.

R388 Osbon, W. O. A new cathode-ray oscilloscope. *Elec. Jour.*, **28**, 322–324; May, 1931.

An improved type portable oscillograph is described, by means of which recurring phenomena can be visually inspected on a linear time axis in an ordinary lighted room.

R400. Radio Communication Systems

R423.21 ×R590 Centralized radio and public address system in a large hotel. *Radio Eng.*, **11**, 31; May, 1931.

A typical public address system for hotels is described.

R440 Felstead, Chas. Remote control systems for radio transmitters. *Radio Eng.*, **11**, 34–35; May, 1931.

The operation of transmitting equipment from operating room, required connection facilities which are discussed by the author.

R500. Applications of Radio

R512 Wireless direction finding systems for marine navigation. *Marconi Rev.*, 1–11; March–April, 1931.

The relative merits of two systems of direction finding, the Marconi-Bellini-Tos system and the rotating loop antenna method, are discussed.

R512.1 Smith-Rose, R. L. Some observations on the Orfordness rotating beacon. *Jour. I.E.E.*, (London), **69**, 523–532; April, 1931.

An account is given of observations, which have been made over a nine-month period on the Orfordness beacon with the object of ascertaining the utility of this type of radio aid to navigation.

R525 Diamond, H. and Davies, G. L. Characteristics of airplane antennas for radio range-beacon reception. *Bureau of Standards Jour. Research*, **6**, 901–916; May, 1931. B. S. Research Paper No. 313.

This paper gives the results of an investigation on the characteristics of airplane receiving antennas to determine whether an antenna arrangement could be devised which would have all the desirable electrical properties of the vertical pole antenna and yet be free from the mechanical difficulties encountered in the use of the pole antenna. The symmetrical transverse T antenna and the symmetrical longitudinal T antenna, with vertical lead-in portions, are both found to fulfill the desired requirements.

R583 Ives, H. E. Some optical features in two-way television. *Bell Syst. Tech. Jour.*, **10**, 265–272; April, 1931.

A description of some recent improvements that have been incorporated in the two-way television system now being demonstrated between the American Telephone and Telegraph Company's building and the Bell Telephone Laboratory in New York.

R590 Clarke, W. C. Signal Corps equips army transports with radio. *Radio Eng.*, **11**, 45–46; May, 1931.

The modern centralized radio and public address systems with which transport ships are equipped are described.

R600. Radio Stations

R612.1 Semm, A. Der Grossrundfunksender Heisberg (The high-power broadcast transmitter at Heisberg). *Telegraphen und Fernsprech Technik*, **20**, 103–109; April, 1931.

A technical description of the installation is given.

R800. Nonradio Subjects

535.3 Ramadanoff, D. Photo-electric properties of composite surfaces at various temperatures and potentials. *Phys. Rev.*, **37**, 884–896; April, 1931.

Experiments designed to determine the variation of photo-electric current with temperature and plate potential for barium photo-electric cells are described.

535.38 Schröter, F. Neuere Formen lichtelektrischer Zellen (New types of photo-electric cells). *Zeits. für tech. Physik*, **12**, 193–200, No. 4; 1931.

A review of recent developments in the field of photo-electric phenomena.

535.38 Baker, W. R. G.; Fitzgerald, A. S.; Whitney, C. F. Power control by means of phototubes. *Electronics*, **2**, 632–633; May, 1931.

Light operated relays and their applications are described.

536.83 Fischer, W. Die Frage der Frequenz bei der induktiven Erwärmung (The question of frequency in connection with inductive heating). *Zeits. für Hochfrequenz*, **37**, 127–133; April, 1931.

Formulas are derived from which dimensions may be determined, on the one hand for obtaining maximum efficiency in an induction furnace, and on the other hand for preventing heating losses in structural parts.

538 Hoppe, W. H. Permanent magnets for electronic instruments. *Electronics*, **2**, 636–638; May, 1931.

Permanent magnets represent an inconspicuous but important part of many electronic devices; such as measuring instruments, sound pick-ups, etc. Principles of design of such magnets are treated.

621.375.1 Sobin, C. J. Tubes and cells at work. *Radio Eng.*, **11**, 42; May, 1931.

Various types of vacuum tubes that are being employed in commercial and industrial operations are described.

621.385.96 Schlenker, V. A. Acoustic treatment for sound-picture theatres. *Electronics*, **2**, 625–627; May, 1931.

A new method of diagnosis and treatment of acoustics in sound-picture theatres and other auditoriums is described. The method employs electronic acoustimeters as an improvement over former methods.

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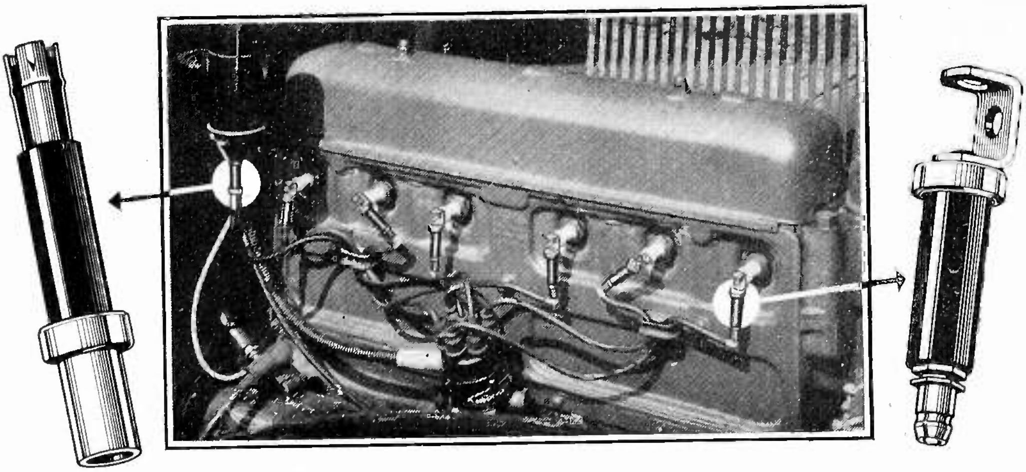

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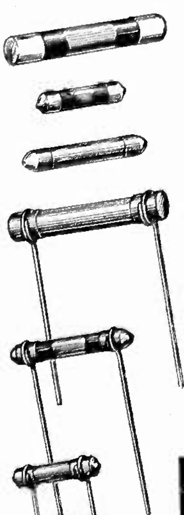

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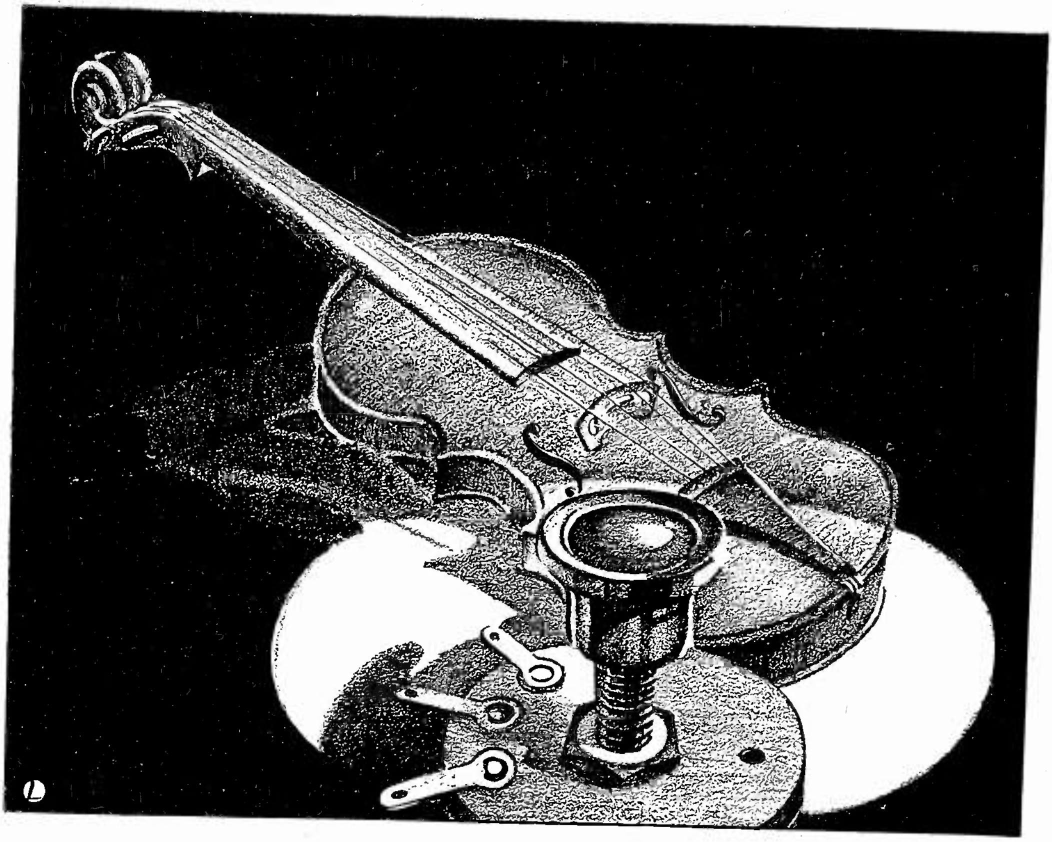

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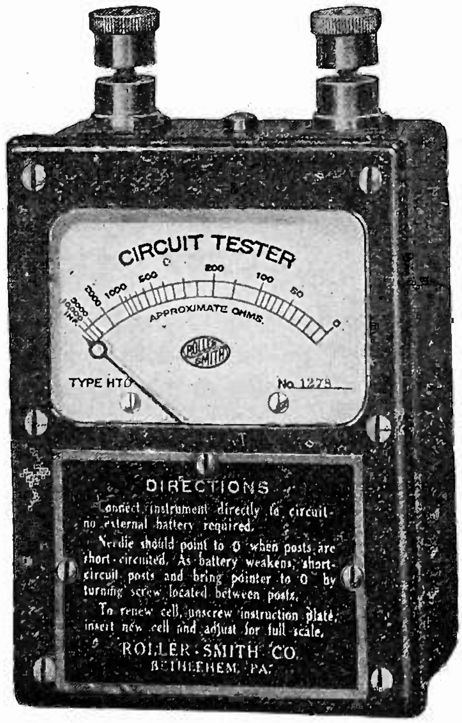

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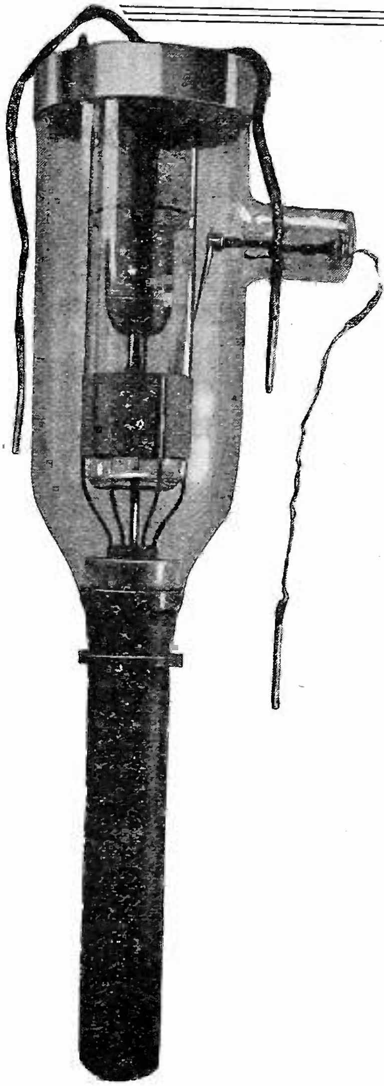

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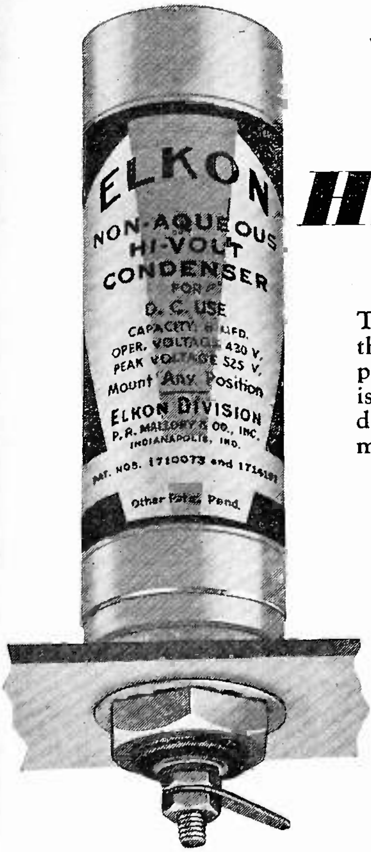

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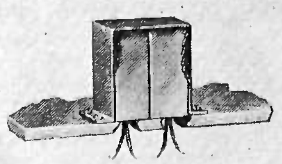


If your set is already designed for condensers on top of the chassis, the Elkon condensers may be placed under an inexpensive drawn metal cover.

otherwise—

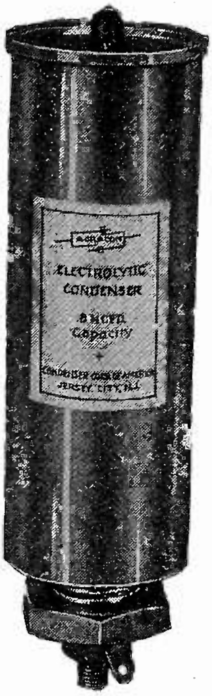
—the condensers may be mounted directly to the under side of the chassis.

ELKON DIVISION

P R. Mallory & Company, Incorporated

Indianapolis, Indiana

When writing to advertisers mention of the PROCEEDINGS *will be mutually helpful.*



Specify the Acracon Electrolytic Condenser and Know You're Getting the Best

These 19 characteristics will convince any engineer of Acracon's superiority in electrical efficiency and mechanical construction:

1. Metal cover for protection and appearance.
2. Live, rubber nipple.
3. Nipple spun into aluminum shell. Absolutely leak-proof.
4. Anode spiral cold welded into anode, giving rigid construction.
5. One piece extruded aluminum container.
6. Retaining flange for rubber gasket.
7. Tapered anode stem for snug fit.
8. Large cadmium plated steel mounting nut, concave to insure tight connection.
9. ¾″-16 thread neck for mounting.
10. Metal washer.
11. Anode nut.
12. Anode soldering tab.
13. Large size insulating washer.
14. Tapered hole to take tapered anode.
15. Special live, rubber insulating gasket free from impurities.
16. Heavy, rigid, anode stem of high purity aluminum.
17. High purity anode, spiral, so wound as to eliminate the necessity of insulating liner between anode and container.
18. Special, high, critical voltage electrolyte, well over anode to insure long life.
19. Leak-proof rolled seam as used in canning industry.

The Acracon unit is now available in capacities up to 16 microfarads at either 440 or 475 volt peak in the single anode type. Write today for further information!

ACRACON

Electrolytic Condensers

Condenser Corporation of America

259 Cornelison Ave., Jersey City, N.J.

Factory Representatives In:

Chicago Cincinnati St. Louis San Francisco Los Angeles Toronto
And Other Principal Cities

When writing to advertisers mention of the PROCEEDINGS *will be mutually helpful.*

"Here's a New SPRAGUE INVERTED CONDENSER that does the work"

Metal enclosed vent

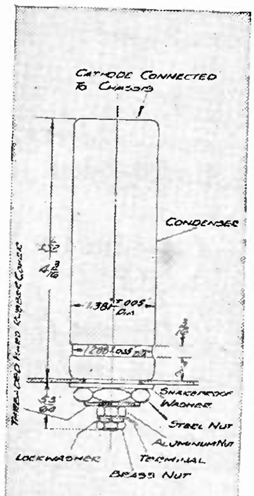
One piece anode

One hole mounting

Can—dull nickel finish

Small space—high capacity

High voltage—self-healing



RADIO engineers of 1931 receiving sets will receive with unqualified approval this announcement of a Sprague Electrolytic Condenser that helps materially in their problems to make the new radio receivers as simple as possible, more efficient than ever before—and at the same time conserve time, space and expense.

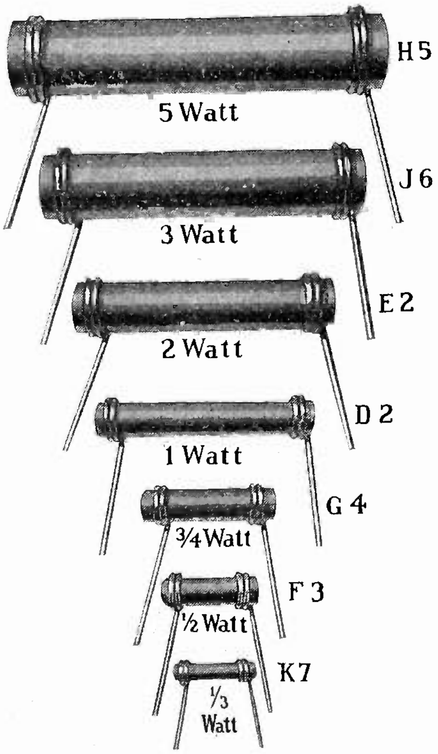
Only an organization like SPRAGUE, with superb modern manufacturing facilities, highly developed research laboratories, an engineering department devoted to the problems of the set manufacturer, and a highly trained personnel could produce the tremendous improvement represented by this new SPRAGUE INVERTED ELECTROYLTIC CONDENSER in tremendous quantities with unqualified UNIFORMITY.

Send for illustrated booklet, diagrams, etc. Address Dept. 13K

SPRAGUE SPECIALTIES COMPANY
North Adams, Mass.

SPRAGUE
Electrolytic **CONDENSERS**

Continental Resistors



RESISTOR CHARACTERISTICS:

NOISELESS
RESISTANCE CONSTANT
SMALL TEMPERATURE COEFFICIENT
RUGGED
FIRMLY SOLDERED TERMINALS
DEPENDABLE
COLOR CODED

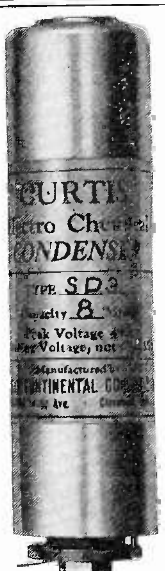
UNITS SHOWN HALF SIZE

CONTINENTAL SUPPRESSORS

For Radio-Equipped Cars

Stop ignition noise in the automobile radio set when used with suitable by-pass condensers. The distributed capacity is extremely small. The resistors are enclosed in a tough ceramic tube of high crushing and tensile strength, of low coefficient of expansion and of high dielectric strength. They are hermetically sealed in the tube and this renders the suppressors moisture proof.

Suppressors shown full size



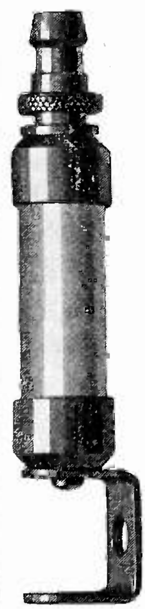
Curtis
Electro
Chemical
Condenser

Essential Characteristics

- Full capacity at all voltages
- Uniform capacity at all frequencies
- Low freezing point
- Low internal resistance
- Low leakage
- No nipples, new breather principle
- One hole mounting terminal at bottom

For Further Particulars Write to

Curtis Continental Corp.
13900 Lorain Ave.,
Cleveland, O.



SPARK PLUG SUPPRESSOR

SCREW IN CABLE TYPE SUPPRESSOR

DISTRIBUTOR SUPPRESSOR

Write for Information and Prices

CONTINENTAL CARBON INC.
13900 Lorain Ave.
WEST PARK, CLEVELAND, O.

The Institute of Radio Engineers

Incorporated

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

APPLICATION FOR ASSOCIATE MEMBERSHIP

(Application forms for other grades of membership are obtainable from the Institute)

To the Board of Direction
Gentlemen:

I hereby make application for Associate membership in the Institute.

I certify that the statements made in the record of my training and professional experience are correct, and agree if elected, that I will be governed by the constitution of the Institute as long as I continue a member. I furthermore agree to promote the objects of the Institute so far as shall be in my power, and if my membership shall be discontinued will return my membership badge.

Yours respectfully,

....
(Sign with pen)

....
(Address for mail)

....
(Date) (City and State)

References:
(Signature of references not required here)

Mr. .... Mr. ....

Address .... Address ....

Mr. .... Mr. ....

Address .... Address ....

Mr. ....

Address ....

The following extracts from the Constitution govern applications for admission to the Institute in the Associate grade:

ARTICLE II—MEMBERSHIP

Sec. 1: The membership of the Institute shall consist of: * * * (d) Associates, who shall be entitled to all the rights and privileges of the Institute except the right to hold the office of President, Vice-president and Editor. * * *

Sec. 5: An Associate shall be not less than twenty-one years of age and shall be: (a) A radio engineer by profession; (b) A teacher of radio subjects; (c) A person who is interested in and connected with the study or application of radio science or the radio arts.

ARTICLE III—ADMISSION

Sec. 2: * * * Applicants shall give references to members of the Institute as follows: * * * for the grade of Associate, to five Fellows, Members, or Associates; * * * Each application for admission * * * shall embody a concise statement, with dates, of the candidate's training and experience.

The requirements of the foregoing paragraph may be waived in whole or in part where the application is for Associate grade. An applicant who is so situated as not to be personally known to the required number of members may supply the names of non-members who are personally familiar with his radio interest.

(Typewriting preferred in filling in this form) No.................

RECORD OF TRAINING AND PROFESSIONAL EXPERIENCE

1 Name ....
(Give full name, last name first)

2 Present Occupation ....
(Title and name of concern)

3 Permanent Home Address ....

4 Business Address ....

5 Place of Birth........................ Date of Birth............Age......

6 Education ....

7 Degree ....
(college) (Date received)

8 Training and Professional experience to date...............................

NOTE: 1. Give location and dates. 2. In applying for admission to the grade of Associate, give briefly record of radio experience and present employment.

DATES HERE	

9 Specialty, if any..

Receipt Acknowledged............ Elected............ Deferred............
Grade............ Advised of Election........... This Record Filed..........

PRE-EMINENT

LEADERSHIP

Produces LOWER PRICES

DURING PAST TWO YEARS

101

RADIO MANUFACTURERS

in the United States used

Cornell

"CUB" CONDENSERS

and were pleased with their performance.

Such leadership and large production enable us to announce, a new and lower price list with the same high standard of quality.

Cornell Electric Mfg. Co., Inc.

LONG ISLAND CITY, NEW YORK

Filter Condensers—By-Pass Condensers—Radio Interference Filters—Power Factor Correction Banks Uniformly High Insulation Resistance—Paper Dielectric Condensers (All Types)—Resistors (All Types)

When writing to advertisers mention of the PROCEEDINGS *will be mutually helpful.*

CHANGE IN MAILING ADDRESS OR BUSINESS TITLE

Members of the Institute are asked to use this form for notifying the Institute office of a change in their mailing address or any change in the listing of their company affiliation or title for the Year Book membership list.

The Secretary,
THE INSTITUTE OF RADIO ENGINEERS,
33 West 39th Street,
New York, N.Y.

Dear Sir:

Effective .... please note change in my address
(date)

for mail as follows:

FROM

....
(Name)

....
(Street Address)

....
(City and State)

TO NEW ADDRESS

....
(Street Address)

....
(City and State)

Also for the membership list for next year's Year Book note change in my business address (or title) as follows, this is / is not my mailing address:

....
(Title)

....
(Company Name)

....
(Address: Street, City and State)

PLEASE FILL IN, DETACH, AND POST TO THE INSTITUTE PROMPTLY

Alphabetical Index to Advertisements

EMPLOYMENT PAGE

Advertisements on this page are available to members of the Institute of Radio Engineers and to manufacturing concerns who wish to secure trained men for positions. For rates and further information address the Advertising Department, The Institute of Radio Engineers, 33 West 39th Street, New York City, New York.

MANUFACTURERS and others seeking radio engineers are invited to address replies to these advertisements at the Box Number indicated, care the Institute of Radio Engineers. All replies will be forwarded direct to the advertiser.

CHIEF ENGINEER of eastern broadcast station, with twelve years experience on design and manufacture of receiving and transmitting equipment, desires responsible position in the western portion of the United States or in some foreign country. Previously in charge of relay broadcasting station with four high frequency assignments. Is a Member of the Institute. Age, 30. Box 66.

RADIO OPERATOR AND SERVICE MAN, with four years experience in charge of service department and two years experience as commercial operator desires position in radio field where past experience will be most useful. Is interested in laboratory work on broadcast receivers. Has completed radio courses in two radio schools as well as eight months college work. Age, 24. Box 67.

ENGINEER experienced in design and development of manual and automatic testing equipment for telephone, radio, and sound picture circuits, having knowledge of industrial applications of photo-electric cells and vacuum tubes, desires connection in testing equipment or communication field. Age, 31. Box 68.

RADIO OPERATOR holding commercial first-class unrestricted operator's license, having two years operating experience and three years experience broadcast receiver construction and repair, desires permanent position as operator with broadcast station or radio telegraph communication company. Age, 23. Box 69.

RADIO ENGINEER B.S. in M.E., 1924, with seven years experience in testing, manufacture, design, and installation of radio transmitting apparatus desires position as consulting or chief engineer of a medium power broadcasting station. Is well acquainted with methods of frequency control and modulation and could rebuild a station to conform to latest regulations of the Federal Radio Commission. Age, 28. Box 70.

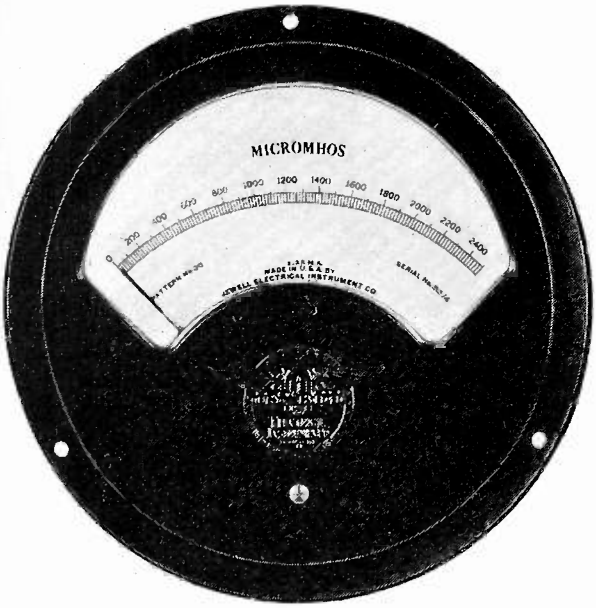
GRADUATE of Massachusetts Institute of Technology, B.S. in E.E., 1925, experienced in research, development and production problems desires suitable affiliation. Age, 31. Box 71.

GRADUATE ENGINEER with four years of radio experience, two years with large radio manufacturer, desires change of employment. Experience includes in part the design and construction of laboratory television receivers, a complete broadcasting station, and a large number of high frequency receivers. Position desired is one in which engineering knowledge, foresight and judgment are essential. Married. B.S. in E.E., 1929. Age, 25. Box 72.

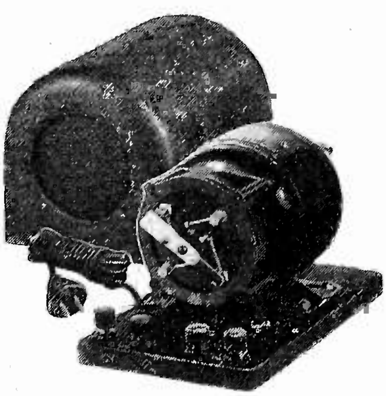
COMMUNICATIONS ENGINEER with ten years experience in carrier current telegraph and telephone systems desires change in affiliation. Age, 34. Box 73.

COLLEGE GRADUATE AVAILABLE, with experience in radio sales and service, design and construction of service equipment, and construction of public address and talking picture systems. B.A., 1928. Age 23. Box 63.

More Accurate Direct Measurement of Mutual Conductance



A 7-inch instrument calibrated from 0 to 2500 micromhos is standard. Additional scales or instruments of other sizes are optional.



The synchronous converter unit with dust case removed.

An instrument that makes more sensitive and more accurate direct measurement of vacuum tube mutual conductance has been developed by the Jewell laboratories.

This equipment is adaptable to wide application in either laboratory or production testing, the indicating instrument being a high torque, rugged direct current instrument of standard design. In conjunction with Jewell Instrument Relays, rejection or acceptance of tubes may be accomplished by automatic machinery.

Previous methods of direct measurement of mutual conductance usually gave results which were of necessity modified by the relatively heavy plate circuit load in the tube under test. The Jewell Synchronous Commutator method increases the plate impedance less than 1 per cent, giving a true reading of this most important tube characteristic under any and all conditions and regardless of the mu of the tube.

Normal operating potentials are applied to the tube under test, assuring the utmost of accuracy for the measurements. Variable-mu tubes may be tested at a series of control grid potentials over their entire working range.

An engineering data sheet completely describing the "Synchronous Commutator" method and Jewell equipment for applying it to your needs, has been prepared. Write for it today.

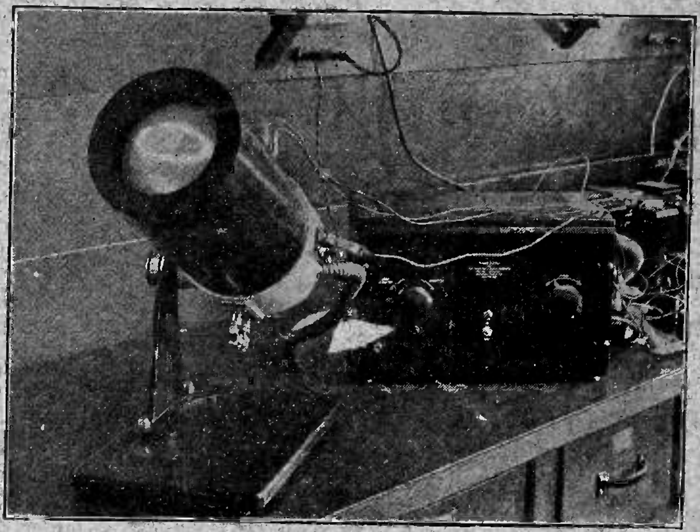
Jewell Electrical Instrument Co.

1642-D Walnut St., Chicago, Ill.

31 YEARS MAKING GOOD INSTRUMENTS

JEWELL

NEW DEVELOPMENT IN OSCILLOGRAPHY



The new General Radio Cathode-Ray Oscillograph, from a daylight photograph, unretouched except in the background

BRILLIANT patterns, long tube life, complete alternating-current operation, and a convenient universal mounting are features of the new General Radio cathode-ray oscillograph. All these you'll appreciate, if you have ever used a cathode-ray oscillograph for your own experimental work.

The price of the oscillograph is $280.00 complete Write for a copy of Bulletin F-303-R which describes the instrument in detail.

GENERAL RADIO COMPANY
OFFICES LABORATORIES FACTORY
CAMBRIDGE A, MASSACHUSETTS

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