VOLUME 24 JULY, 1936 NUMBER 7

PROCEEDINGS
of
The Institute of Radio Engineers

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Institute of Radio Engineers Forthcoming Meetings

CLEVELAND SECTION
September 24, 1936

DETROIT SECTION
September 18, 1936

LOS ANGELES SECTION
September 15, 1936

NEW YORK MEETING
October 7, 1936

PHILADELPHIA SECTION
September 3, 1936

WASHINGTON SECTION
September 14, 1936

PROCEEDINGS OF

The Institute of Radio Engineers

VOLUME 24 July, 1936 NUMBER 7

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The Institute of Radio Engineers

GENERAL INFORMATION

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

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

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

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

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Transferred to the Member Grade

New Jersey	Teaneck, 471 Claremont Ave.	Shelby, R. E.
New York	New York, Bell Tel. Labs., Inc., 463 West St.	Bair, R. S.
	New York, Hazeltine Service Corp., 333 W. 52nd St.	Dean, C. E.
	New York, Bell Tel. Labs., Inc., 463 West St.	Green, E. I.
	New York, Bell Tel. Labs., Inc., 180 Varick St.	Morrison, J. F.
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	New York, Bell Tel. Labs., Inc., 180 Varick St.	Tinus, W. C.
Pennsylvania	Emporium, 115 E. 5th St.	Steen, J. R.
Texas	Galveston, 204 Prudential Bldg.	McCabe, L. L.
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Elected to the Associate Grade

California	Menlo Park, 125 Princeton Rd.	Knar, J. M.
	San Francisco, 716 Masonic Ave.	Williams, P. A.
District of Columbia	Washington, Navy Department	Hull, D. R.
	Washington, [illegible] St., N.E.	Soukaras, K. M.
Georgia	Atlanta, 705 Piedmont Ave., N.E.	Parkins, F. A.
Idaho	Pocatello, 434 W. Bonneville	Wilson, T. J.
	Wallace, Fire Department	Heuser, E. R.
Illinois	Chicago, c/o RCA Institutes, 1154 Merchandise Mart	Hamm, J. L.
	Chicago, c/o RCA Institutes, 1154 Merchandise Mart	McMahon, R. J.
	Chicago, 5012 N. Meade Ave.	Thorson, W.
	Chicago, [illegible] E. 48th St.	Wilson, D. M.
Indiana	Crown Point, Box 212	Murphy, E. J.
Kansas	Manhattan, Elec. Eng. Dept., Kansas State College	Schumann, F.
Maine	Bangor, 221 Center St.	Lawson, N. H.
Massachusetts	Boston, Northeastern Univ., 316 Huntington Ave.	Oberg, R. O.
	[illegible] St.	Ouimet, V. E.
Minnesota	Minneapolis, Elec. Eng. Dept., Univ. of Minnesota	Hartig, H. E.
New Jersey	Bloomfield, 84 Davis Ave.	Fuller, F. L.
	Camden, 715 Cooper St.	Johnson, L. R.
	Camden, 820 N. 2nd St.	Joyner, A. A.
	Camden, 10[illegible] N. 6th St.	Martin, K. H.
	Pleasantville, 711 W. Washington Ave.	Strockbine, J.
New York	New York, 510 W. [illegible] St.	English, W. K.
	New York, Bell Tel. Labs., Inc., 463 West St.	Ingram, S. B.
	New York, Bell Tel. Labs., Inc., 180 Varick St.	Kuzela, E. V.
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	Rego Park, L.I., 61-30 Saunders St.	Wilkens, W. B.
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	Rochester, Delco Appliance Corp.	Wallis, C. T.
	Schenectady, 1014 Helderberg Ave.	Williams, W. R.
	Scotia, [illegible] Glen Ave.	Beggs, J. E.
	Snyder, 75 Chateau Ter.	Osborne, F. H.
	Tonawanda, 48 Fremont St.	Bauer, H. G.
	Tonawanda, [illegible] Keller St.	Best, T. A., Jr.
Ohio	Cleveland, USS Tahoma, C.G.	Jones, M. S.
	Cleveland, 7002 Home Ct.	Knife, W. J.
	Fairfield, 61 N. Main St.	Magee, E. E.
	Oxford, Oxford Radio Service	Damm, R. M.
Pennsylvania	Bethlehem, Elec. Eng. Dept., Lehigh Univ.	Formhals, W. H.
	Central City, U.S. Airways Station, R.F.D. 1	Sponeybarger, J. O.
Texas	San Antonio, 130 Belden Ave.	Dupree, J. E.
Wisconsin	Rhinelander, U.S. Forest Service	Music, J.
Australia	Ashfield, N.S.W., c/o Amalgamated Wireless (A'sia) Ltd., 551 Parramatta Rd.	Huey, R. M.
	Ashfield, N.S.W., A.W.A. Radio Electric Works	O'Donnell, A.
	Burwood, N.S.W., 51 Park Rd.	Freeman, A. C.

Canada	Toronto 4, Ont., 805 Davenport Rd	Cumming, A. A.
England	Cricklewood, London N.W. 2, 112 Gladstone Park Gardens	Ellen, R. G.
	Dudley, Worcs., The Grove, Kingswinford	Herdman, T. L.
	Islington, London N. 1, Culford Works, Kingsbury Rd	Andersen, R. C.
	Streatham, London, 63 Baldry Gardens	Harrison, K. W.
Germany	Berlin S.W. 11, Hallesches Ufer, 12	Moser, W.
Holland	Eindhoven, Bredalaan 77	Severs, J.
Japan	Kyotofu, 11 Nankai, Kamiueno, Mukomachi, Otokunigun	Uchida, H.
	Yokkaiti, Mie-ken, Nihon Musen Yokkaiti Syuttyozyo	Miyakosi, K.
North Wales	Old Colwyn, "Headlands," Bryn Ave	Oliver, C. J.
Norway	Oslo, Bjerregaardsgt. 29 E	Flood, S. W.
South Africa	Durban, Natal, African Broadcasting Co. Ltd	Heyes, O.
	Creey P. O., N. Transvaal, c/o Captain P. C. Sanderson, "Belmont"	Houghton, H. W.

Elected to the Junior Grade

California	Glendale, 1529 Ridgeway Dr	Tate, J. W.
Illinois	Chicago, 1401 E. 55th St	Kealy, D.
	Chicago, 1142 W. Grand Ave	Scaruffi, J.
	Evanston, 1225 Grant St	Chalberg, H. W. A.

Elected to the Student Grade

Connecticut	New Haven, 25 Irving St	Clements, S. E.
Massachusetts	Cambridge, M.I.T. Dormitories	Farmer, D. E.
	Cambridge, 273 Harvard St	Hedberg, C.
	Cambridge, 66 Wendell St	Mao, Y. Y.
Michigan	Ann Arbor, 1208 South University	Shannon, C. E.
Ohio	Columbus, 763 Mohawk St	Fritschel, P. G.
Texas	Austin, Dept. of Physics, Univ. of Texas	Jones, H. W.
Canada	Edmonton, Alta., 10731-97th St	Jordan, E. C.
England	London W. 12, 24 Shepherds Bush Green	Beebe, R. D.

APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Admissions Committee. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before July 31, 1936. Final action will be taken on these applications August 5, 1936.

For Election to the Associate Grade

California	Los Angeles, 7460 Beverly Blvd.	Hawkins, J. N. A.
	Los Angeles, 2114 Holly Dr.	Thias, E. P.
	Oakland, 733 Wesley Ave.	Braun, G. B.
	San Diego, Naval Air Station	Jackson, R.
District of Columbia	Washington, c/o Bureau of Navigation, Navy Dept.	Bennett, R.
Georgia	Atlanta, Box 162, Georgia Inst. of Tech.	Caulfield, T. B., Jr.
Illinois	Chicago, 5936 S. Cicero Ave.	Brunner, R. R.
	Chicago, c/o WCFL, 666 Lake Shore Dr.	Marquardt, M.
	Chicago, 951 E. 58th St.	Offner, F. F.
	Wheaton, 224 N. Blanchard St.	Fettweis, C. P.
Louisiana	New Orleans, 828 Gravier St.	Pendergrass, F. L.
Massachusetts	Brookline, 12 Boylston St.	Harvey, C. A.
	Cambridge, General Radio Co., 30 State St.	Cady, C. A.
Michigan	Ann Arbor, 147 East Physics Bldg.	Firestone, F. A.
Minnesota	Minneapolis, Radio Station WLB	Holtan, C. H.
	St. Paul, Radio Station KSTP, St. Paul Hotel	Carr, L. H.
Mississippi	Hattiesburg, 204 N. 25th Ave.	Davidson, C. E.
Nebraska	Kearney, c/o Tinytone Radio Co.	Beshore, P. S
New Jersey	Lakewood, R.F.D. No. 3	Bartolf, H. P.
New York	New York, Bell Tel. Labs., Inc., 463 West St.	Miller, R. A.
	New York, Bell Tel. Labs., Inc., 180 Varick St.	Wright, S. B.
	Scotia, R.F.D. No. 2, Spring Rd.	Hahn, W. C.
Ohio	Lakewood, 1632 Lauderdale Ave.	Grostick, G. E.
	Norwood, 4228 Franklin Ave.	Silver, J. F.
Oregon	Corvallis, Physics Dept., Oregon State College	Yunker, E. A.
Pennsylvania	Emporium, 326 W. 6th St.	Stringfellow, G. O.
	Germantown, Phila., Cloverly Apts., 437 W. School Lane	Wannamaker, W. H. Jr
Texas	Baytown, Box 331	Cornelius, L. W.
Wisconsin	Horicon, 108 Main St.	Pahl, H. H.
	West Allis, 1424 S. 72nd St.	Keller, O.
Canada	Winnipeg, Manit., 35 Essex Ave.	Sheffield, A. G.
England	Birkby, Huddersfield, 20 Bromley Rd.	Dale, J.
	Clacton-on-Sea, Essex, 125 Old Rd.	Budden, G. A.
	Clapham Park, London S.W.4, c/o British Broadcasting Corp., 87 King's Ave.	Bulow, V. A. M.
	East Finchley, London N.2, 66 Lincoln Rd.	Burnett, E. F.
	Kendal, Westmorland, 1 Kent View	Dougherty, J. B.
	Wanstead, Essex, 71 Cranbourne Ave.	Fielding, T. J.
Japan	Kannon Shinchiku, Formosa, c/o Kannon Receiving Station, Kokusai Denwa	Iida, M.
	Kawasaki-shi, c/o Nipponophone Co. Ltd., Kunesaki	Suko, N.
	Komuro Kitaadachi-gun, Saitama, c/o Komuro Receiving Station, Kokusai Denwa	Sakai, T.
	Nazaki Yuki-gun, Ibaraki, c/o Nazaki Transmitting Station, Kokusai Denwa	Kudo, M.
	Tokyo, c/o Kokusai Denwa Kaisha Ltd., Osaka Bldg.	Hayakawa, D.
Manchoukuo	Daidohiroba, Hsingking, Wireless Plant Office, M.T.T. Co.	Nagatake, N.
New Zealand	Wellington, 43 Roy St.	Firth, E. T.
South Africa	Johannesburg, 9th Fl., Main House, Main St.	Duff, W. A.
Spain	Madrid, Int. Tel. and Tel. Corp., Pi Y Margall 2	O'Neill, H. N.
Switzerland	Ste. Croix, Les Genets	Gautschi, W.
	Zurich-Zollikon, Hohestr.	Strohschneider, W.
Venezuela	Maracaibo, Lago Petroleum Corp.	Franks, W. E.

For Election to the Junior Grade

California	San Francisco, 1062 Divisadero St.	Croninger, B.
New York	Brooklyn, 1285 Dean St.	Hart, G., 3rd.
	New York, 16 E. 7th St.	Padrusch, J.
British West Indies	Basseterre, St. Kitts, Government House	Stewart, C. J.
China	Shanghai, c/o T. S. Wang, 346 Route de Zikawei	Lor, P. L.
Cuba	Havana, San Pedro B., Cerro	Iglesias Cobas, R.
New Zealand	Wanganui East, 153 Anzac Parade	Laird, J. N.

Applications for Membership

For Election to the Student Grade

California	Berkeley, 1615 Virginia St.	Okada, K.
	Long Beach, 3818 E. 4th St.	Smith, M. H., Jr.
	Los Angeles, 5913 Miramonte Blvd.	Bucknell, W. H.
	Oakland, 536-21st St.	Child, V. E.
	San Francisco, 1265-18th Ave.	Ames, E. K.
Indiana	West Lafayette, 471 N. Grant St.	Cheng, C. C.
	West Lafayette, 471 N. Grant St.	Liu, I. L.
Maine	Portland, 102 Dartmouth St.	Kenneson, H. C.
Massachusetts	Cambridge, 1619 Massachusetts Ave.	Pitzer, R. L.
Michigan	Ann Arbor, 611 Church St.	Miles, W. F.
	Ann Arbor, 727 S. State St.	Parker, C. V.
Minnesota	New Prague, 108-3rd St. N.W.	Williams, E. C.
	North Hibbing, 201 W. Washington St.	Jenkins, A.
Missouri	Normandy, 4510 Marlboro Ct.	Landwehr, I. H.
New Jersey	Plainfield, 915 Madison Ave.	Thompson, E. H.
New York	Troy, Waite Dormitory, Sage Ave.	Banker, J. R.
Ohio	Cleveland, 7908 Halle Ave.	Pacanovsky, G. J.
	Columbus, 3100 Olentangy River Rd.	Drayer, R. W.
Texas	Dallas, 5242 Richard Ave.	Norgaard, D. E.
West Virginia	Fairmont, 426 Morgantown Ave.	Mason, J. W., III
Wisconsin	Madison, 228 N. Mills St.	Peterson, D. W.
Canada	Montreal, P. Q., 2325 St. Luke St.	McLean, M. D.

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ELEVENTH ANNUAL CONVENTION COMMITTEE

Seated, left to right—R. A. Fox, Mrs. E. L. Gove, E. L. Gove, K. J. Banfer, Convention Committee Chairman; S. E. Leonard, and C. E. Smith. *Standing, left to right*—J. S. Hill, Bruce David, R. M. Pierce, Chairman of the Cleveland Section; and J. R. Martin.

INSTITUTE NEWS AND RADIO NOTES

Eleventh Annual Convention

The Institute's Eleventh Annual Convention was held in Cleveland, Ohio on May 11, 12, and 13, with headquarters at the Hotel Statler. The attendance totaled 589 of whom sixty-one were ladies.

There were nineteen papers presented during the five technical sessions which occupied the major portion of the convention time. Summaries of these papers appeared in the May issue of the PROCEEDINGS.

An exhibition which occupied forty booths was open for inspection during the entire duration of the convention.

The Institute Medal of Honor was presented to G. A. Campbell and the Morris Liebmann Memorial Prize to B. J. Thompson during the banquet which was held on the evening of May 12 and attended by 360.

Committee Work

ADMISSIONS COMMITTEE

A meeting of the Admissions Committee was held in the Institute office on Wednesday, April 1, and attended by C. M. Jansky, Jr., chairman; H. H. Beverage, R. A. Heising, L. C. F. Horle, F. A. Kolster, E. R. Shute, H. M. Turner, and H. P. Westman, secretary.

Three applications for transfer to Fellow were considered; one was approved and two were denied. Of twelve applications for transfer to Member, ten were approved and two were tabled pending further investigation. Three of eight applications for admission to Member grade were approved while two were denied and the remaining three tabled.

The Admissions Committee met again on Wednesday, May 6, in the Institute office and those present were C. M. Jansky, Jr., chairman; H. H. Beverage, R. A. Heising, C. W. Horn, F. A. Kolster, and H. P. Westman, secretary.

Two applications for transfer to Fellow grade and six for transfer to Member grade were acted on and approved. Two applications for admission to Member grade were approved and one was tabled.

CONSTITUTION AND LAWS COMMITTEE

A meeting of the Constitution and Laws Committee was held in the Institute office on Monday, May 25, and attended by H. M. Turner, chairman; Arthur Batcheller, Austin Bailey, and H. P. Westman, secretary.

Another meeting was held on Wednesday, June 3, in the Institute office and those present were H. M. Turner, chairman; Austin Bailey, and H. P. Westman, secretary.

The committee is making a review of the Institute Constitution and expects to propose a number of modifications of that document to the Board so they may be submitted to the membership for vote.

Convention Papers Committee

A meeting of the Convention Papers Committee which was attended by William Wilson, chairman; Haraden Pratt, and H. P. Westman, secretary, was held on Monday, April 6, in the Institute office and devoted to the preparation of the final program for the Eleventh Annual Convention technical sessions.

Nominations Committee

The Nominations Committee held a luncheon-meeting on Wednesday, May 6, in the McGraw-Hill dining room and those present were Stuart Ballantine, chairman; Melville Eastham, C. M. Jansky, Jr., William Wilson, and H. P. Westman, secretary. The committee prepared a slate of candidates for office which was submitted to the Board of Directors at its meeting that afternoon.

Sections Committee

The annual meeting of the Sections Committee was held in the Hotel Statler in Cleveland, Ohio, on the afternoon of Monday, May 11. Those present were H. P. Westman, acting chairman and secretary; D. S. Bond, A. B. Buchanan, H. L. Byerlay, F. W. Cunningham, E. C. Denstaedt, John Evans, N. B. Fowler, H. S. Gould, G. C. Gross, L. E. Hayslett, M. T. Kiebert, Jr., Ben Kievit, G. F. Platts, R. M. Pierce, G. T. Royden, H. J. Schrader, E. E. Schwarzenbach, Dayton Ulrey and H. C. Vance. The following sections were represented: Altanta, Buffalo-Niagara, Chicago, Cincinnati, Cleveland, Connecticut Valley, Detroit, Emporium, Philadelphia, Pittsburgh, San Francisco, Seattle, and Washington.

Reports on the fiscal affairs of sections, their meeting activities and membership status were presented and discussed in detail. Considerable attention was given to the seminar meetings held by the San Francisco Section at which papers which have appeared in the PROCEEDINGS are discussed. In addition to ten regular meetings each year, ten seminar meetings are held and are attended by almost as many as are present at the regular meetings.

In a number of instances, the attendance at section meetings is regularly so large as to indicate that many nonmembers are present.

Some sections encourage this by adding names of nonmembers to the list of those who receive meeting notices. It was felt that attempts should be made to encourage these nonmembers to join the Institute and thus lend their support to the organization. Where nonmembers request notices of meetings, it was considered reasonable to ask them to pay a nominal charge to defray the cost of mailing.

The Philadelphia Section recently circulated a questionnaire to those who attended one of its meetings in an effort to ascertain what types of papers were most interesting to its members. The possibility of making similar surveys at other section meetings was discussed.

As a method of encouraging more personal interest among section members and visitors, it is proposed that small cards be inscribed with the name of the wearer and worn conspicuously during each meeting. These cards may be collected as those in attendance leave and should assist the Membership Committee in solicitation work.

A discussion was held on methods of avoiding conflicting meeting dates when the subject matter is of interest to more than one group of engineers.

Methods of stimulating transfers to higher grades on the part of those who are qualified for advancement were considered.

STANDARDIZATION

Technical Committee on Electroacoustics

Meetings of the Technical Committee on Electroacoustics were held in the Institute office on April 17 and May 22. The earlier meeting was attended by H. F. Olson, chairman; Sydney Bloomenthal, C. H. G. Gray, Hans Roder, Julius Weinberger (visitor), V. E. Whitman, Harold Zahl, and H. P. Westman, secretary. Those present at the later meeting were: H. F. Olson, chairman; Sydney Bloomenthal, J. T. L. Brown (representing C. H. G. Gray), Knox McIlwain, Hans Roder, Julius Weinberger (visitor), Harold Zahl, and H. P. Westman, secretary.

These two meetings were devoted to the preparation and revision of a report on loud speaker testing employing both outdoor and indoor techniques.

Technical Committee on Electronics

The Technical Committee on Electronics met in the Institute office on March 20 and those in attendance were B. J. Thompson, chairman; R. S. Burnap, F. R. Lack, George Lewis, J. W. Milnor, G. D. O'Neill, O. W. Pike, P. T. Weeks, and H. P. Westman, secretary.

This was an organization meeting of the committee and after discussing the scope of its activities, it established seven subcommittees, designated chairmen, and proposed some names of members to serve on these subcommittees. The report of the 1935 technical committee which has not as yet been submitted to the Standards Committee was distributed with requests for notations of errors. No major changes were to be permitted.

Another meeting of the Technical Committee on Electronics was held in the Hotel Statler in Cleveland, Ohio, during the Eleventh Annual Convention on Wednesday, May 13. It was attended by B. J. Thompson, chairman; C. H. Bachman (visitor), R. M. Bowie, D. W. Epstein, F. A. Holborn, Ben Kievit, F. R. Lack, I. G. Maloff, G. F. Metcalf, G. D. O'Neill, H. W. Parker, Dayton Ulrey, P. T. Weeks, and H. P. Westman, secretary.

At this meeting, the chairmen of the various subcommittees reported on the status of their committees in regard to personnel and meetings held. Consideration was then given to a number of corrections which were proposed to be made in the report of the 1935 committee. That document is now suitable for submission to the Standards Committee for action.

Subcommittee on Cathode-Ray and Television Tubes
Subcommittee on Photoelectric Devices

The above two subcommittees which operate under the Technical Committee on Electronics met jointly in the Hotel Statler in Cleveland on Monday, May 11, during the Eleventh Annual Convention. The personnel of the two committees and the scope of their activities were considered. An operating schedule was also planned.

Institute Meetings

Buffalo-Niagara Section

The Buffalo-Niagara Section met at the University of Buffalo on May 20. L. E. Hayslett, chairman, presided and there were twenty-five present.

A paper on "The Application of Phenolic Materials to Radio Construction" was given by E. A. Russell of the Spaulding Fibre Company, North Tonawanda, N.Y. The manufacture of a vulcanized or parchmentized fibre from the raw materials to the finished product was described in detail. It was pointed out that the first portion of the process is essentially the manufacturing of a cotton cellulose paper. Laminated bakelite is produced by utilizing the interaction of phenol and formaldehyde in the presence of heat.

The meeting was concluded with the projection of several reels of motion pictures giving views of equipment and manufacturing operations.

Chicago Section

H. C. Vance, chairman, presided at the May 22 meeting of the Chicago Section which was held in the RCA Institutes Auditorium. There were 181 members and guests present and twenty-eight at the informal dinner which preceded the meeting.

The first paper presented at the meeting was entitled "An Analytical and Experimental Study of a Resistance Coupled Neutralizer Circuit" by C. S. Roys, assistant professor of electrical engineering at Purdue University. Dr. Roys first discussed the needs for a resistance neutralizer, some of the problems involved in the design of a device capable of supplying resistance neutralization, and a brief history of some of the attempts which have been made to develop units usable for this purpose. The design of a one-tube resistance neutralizer unit was described and the defects shown. Advantages of employing two tubes giving inherent balance were outlined and analyzed mathematically. Conditions of load and output were developed and the limitations of the device shown. It is not of extreme value where very small resistance is involved. The instrument is intended to provide a source of negative resistance for laboratory purposes. The paper was discussed by Messrs. Robinson, Sears, and White.

The second paper was on "Beam Power Tubes" and was presented by Otto Schade of the RCA Manufacturing Company, Harrison, N.J. This paper was presented in New York on April 1 and is summarized in the June issue of the Proceedings. It was discussed by Messrs. Kohler, Sandretto, and White.

Connecticut Valley Section

On February 19 a meeting of the Connecticut Valley Section was held in Springfield, Mass. M. E. Bond, chairman, presided and there were thirty-five present.

A paper on "The 'Q' Meter" was presented by C. J. Franks, engineer of the Ferris Instrument Corporation of Boonton, N.J. This paper was previously presented at the last Rochester Fall meeting.

Detroit Section

The May 29 meeting of the Detroit Section was held jointly with the Detroit Physics Club and comprised a tour of the University of Michigan at Ann Arbor. E. C. Densteadt, chairman, presided and the attendance was 160. There were fifty-eight at the informal dinner which preceded the meeting.

Prior to the inspection of the laboratories, brief talks were given by several members of the Physics Department describing the theory and operation of various instruments which were on exhibition. Small groups were formed and conducted through the laboratories and many pieces of research equipment were operated and questions answered by those in charge. Great interest was shown in the cyclotron or "atom smasher" now under construction. The magnet for it contains fifteen tons of copper and eighty tons of steel. Power for the high-frequency field for the machine will be supplied by a pair of 100-kilowatt water-cooled tubes operating at twelve megacycles. A number of other interesting exhibits were inspected and included voice modulated five-centimeter-wave equipment, a high potential source of ions for production of artificial radio activity, spectroscopic equipment for quantitative analysis, and equipment for analyzing sound by using calibrated microphones, a cathode-ray oscillograph, and selective electrical filters.

Emporium Section

On May 15 a meeting of the Emporium Section was held in the American Legion Club Rooms. R. R. Hoffman, chairman, presided and there were fifty present.

A paper on "Properties and Manufacture of Technical Glass" was the subject of a paper presented by A. E. Marshall, consulting engineer for the Corning Glass Works, Corning, N.Y. The subject was introduced with a historical sketch of the industry of glass manufacture covering a period from about 4000 years ago. It was pointed out that the fields in which glass is used have increased greatly in modern times and the paper was closed with a discussion of technical glass as applied to radio, astronomy, signaling, filters, heat transfer problems, household utilities, fabrics, cordages, threading, and many other items. A motion picture was then shown of the factory of the Corning Glass Works where many of the manufacturing processes were displayed. A demonstration was given of the filtration properties of technical glass.

Los Angeles Section

The Los Angeles Section met on May 19 at the Los Angeles Junior College. There were thirty present and the meeting was presided over by C. R. Daily, chairman. There were ten at the dinner which preceded the meeting.

The first paper presented was on "Radio Interference" by Frank Grimes, Manager of the Los Angeles Radio Interference Bureau. This bureau consists of four men and is supported by eight utilities, eight broadcast stations, radio manufacturers, dealers, clubs and interested parties. The bureau traces interference and takes what steps are necessary to eliminate it. During the winter approximately five hundred calls are received each month. About seventy per cent of the reports are found to be based on disturbances which are not within the scope of the Bureau but are due to either atmospherics or interference from other transmitting stations. The remaining thirty per cent was classified as within the scope of the bureau and involves audio- or radio-frequency interference due to public utility services.

A second paper on "Development Work on Vacuum Tubes for Use in High-Frequency Circuits" was presented by C. Y. Meng, a graduate student of California Institute of Technology. It covered work which Dr. Meng is doing in the production of one-centimeter waves. Tubes developed to produce this high-frequency were described. He outlined the theory and experimental work of Barkhausen and Kurz and Gill and Morrill. In these forms of oscillation, the generated frequency is determined by the electrode potentials and dimensions. A tube having a grid diameter of 0.2 millimeter, a plate diameter of 0.539 millimeter, and a filament of one millimeter diameter was made. The length of all elements was one millimeter. The power generated was insufficient to permit the frequency of oscillation to be measured and it was calculated.

New York Meeting

The June New York meeting of the Institute was held jointly with the Radio Club of America on the 3rd and was attended by 300. The first part of the meeting was held in the auditorium of the American Museum of Natural History and H. T. Stetson of the Institute of Geographical Exploration of Harvard University presented a paper on "Cosmic Cycles and Radio Transmission." The effects of the sun and moon in the ionization of the upper atmosphere upon which radio transmission depends were outlined and the results of recent experiments, including those made during solar and lunar eclipses, were employed in the interpretation of cosmic effects. It was indicated that the magnetic characteristics of the earth effect the velocity of transmission of radio waves travelling over long paths.

At the conclusion of the paper, the meeting was adjourned to the new Hayden Planetarium where W. H. Barton, Jr., associate curator, gave a special lecture and demonstration.

Philadelphia Section

The annual meeting of the Philadelphia Section was held on May 7 at the Engineers Club. Knox McIlwain, chairman, presided and there were 300 members and guests present. Twenty-four attended the dinner which preceded the meeting.

A paper on "Frequency Modulation" was presented by E. H

Armstrong, professor of electrical engineering, Columbia University. New York City. This paper was published in the May PROCEEDINGS.

In the election of officers, Irving Wolff of the RCA Manufacturing Company was designated chairman; A. F. Murray, Philco Corporation, vice chairman; and R. L. Snyder continues as secretary-treasurer.

PITTSBURGH SECTION

Lee Sutherlin, chairman, presided at the March 17 meeting of the Pittsburgh Section which was attended by sixty-four. The meeting was held in the auditorium of the C. C. Mellor Company.

A paper on "The Hammond Organ" was presented by C. Williamson of the physics department of the Carnegie Institute of Technology. Professor Williamson first outlined the fundamentals of acoustics and the construction of the musical scale. He then described the Hammond organ in which ninety wheels mounted on a shaft are driven by a synchronous motor. Each wheel is cut to produce a desired frequency when rotated in a magnetic field. The alternating currents generated are controlled by a keyboard and pass through an audio-frequency amplifier to the loud-speaker system. The amplifier and speakers are mounted together in a cabinet and connected by means of a cable to the keyboard and generator. Means are provided to control the harmonic content of each wave to simulate tones produced by various musical instruments. A number of these combinations are preset at the factory while others may be adjusted by the organist. In a demonstration various musical instruments were played and then imitated on the organ.

ROCHESTER SECTION

The annual meeting of the Rochester Section was held on May 7 at the Sagamore Hotel and presided over by E. C. Karker, chairman. There were sixty-seven members and guests present and nine attended the dinner.

A paper by H. C. Steadman, engineering consultant of the Spaulding Fibre Company, was presented on "The Rôle of Rags in Industry." The paper was chiefly in the form of motion picture films and the speaker pointed out that the title was chosen because the best material for hard fibre and laminated phenolic products is cotton rags which are collected all over the world. The films showed the manufacturing process whereby rags are converted into paper and then into fibre by chemical and mechanical agencies. Many industrial applications were then illustrated and it was pointed out that over half the output of the Spaulding Company is used in the radio industry.

In the election of officers, L. A. DuBridge, head of the physics department of the University of Rochester was named chairman; A. L. Schoen of the research department of the Eastman Kodak Company was designated vice chairman; and H. A. Brown of the Rochester Gas and Electric Corporation was re-elected secretary-treasurer.

San Francisco Section

The San Francisco Section met on May 2 at the Alameda Airport. There were eight members and guests present and R. D. Kirkland, chairman, presided.

The *Hawaiian Clipper* was first viewed as it took off for a regular trip across the Pacific. After this an inspection was made of the radio receiving station, the hangar where one of the large ships was being serviced, the radio repair shop, transmitters used for communication with the plane, the maintenance shops, and the meterological office. This was followed by a talk by G. W. Angus of Pan American Airways.

Another meeting of the San Francisco Section was held on May 20 at the Bellevue Hotel. R. D. Kirkland, chairman, presided and there were fifty-five in attendance. Eighteen were at the dinner which preceded the meeting.

A paper on "Engineering Methods for the Design and Calculation of Class C Amplifiers and Harmonic Generators" was given by F. E. Terman, professor of electrical engineering at Stanford University. The paper followed along the lines of papers recently published by the author but included some additional material. Mimeographed material giving design curves and examples of their use for both harmonic generators and class C amplifiers was distributed to all present.

Seattle Section

A meeting of the Seattle Section was held on May 29 at the University of Washington. J. W. Wallace, vice chairman, presided and there were thirty-eight members and guests present.

A paper on "Application of Revolving Vectors to the Analysis of Amplitude and Frequency Modulated Waves" was presented by J. R. Tolmie, engineer for the Pacific Telephone and Telegraph Company. He first briefly reviewed revolving vectors as used in the more conventional single frequency circuits of electrical engineering, pointing out that all such vectors revolve at the same speed and can therefore be considered to stand still. He then showed that in modulation, the problem is more complicated because the vectors revolve at different velocities.

In the case of amplitude modulation, the carrier vector, of length corresponding to the unmodulated amplitude revolves at carrier frequency, while two auxiliary vectors each of length equal to one half the modulation amplitude and rotating in opposite directions with respect to the carrier vector, add up vectorially with the carrier vector to give the instantaneous value of the modulated wave.

In the case of frequency modulation, Mr. Tolmie showed how the vector of constant length and varying angular velocity can be resolved into three or more component vectors of constant but different angular velocities. These vectors represent the carrier and the two or more side frequencies present in the modulated wave. By referring these vectors to a system of coordinates rotating at carrier frequency, great simplification is obtained.

His method further shows that for combined frequency and amplitude modulation, if synchronized, each of the above-mentioned component vectors acts as a carrier for amplitude modulation.

The paper was discussed by Messrs. Fisher, Kiebert, Libby, Wallace, Wilson, and Woodyard.

Errata

The following corrections have been received to the paper "Effective Resistance of Closed Antennas" by V. I. Bashenoff and N. A Mjasoedoff, which appeared in the May, 1936, issue of the PROCEEDINGS, pages 778–801.

On page 800, footnote 7 should read:

[7] See also R. H. Barfield, *Jour. I.E.E.* (London), vol 76, p. 48; N. A. Mjasoedoff, Patent's Application in U.S.S.R., (1929), No. 51,651.

On page 801, footnote 8 should read:

[8] V. I. Bashenoff and N. A. Mjasoedoff, U.S.S.R. Patent No. 28,551.

Personal Mention

F. T. Alker, formerly engineer-in-chief of the Hungarian Broadcasting Station at Lakihegy, is now an engineer at the international toll-cable repeater station at Budapest.

F. G. Allen, Lieutenant, U.S.A., has been transferred from Fort Monmouth N.J., to Chanute Field, Rantoul, Ill.

J. M. Allen has left Stewart-Warner Corporation to become superintendent of purchased material inspection for RCA Victor, Camden, N.J.

P. G. Andres, formerly with the Indiana State Police Radio System, has become chief engineer of the Electric Switch Coporation of Columbus, Ind.

Proceedings of the Institute of Radio Engineers
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TECHNICAL PAPERS

RAIN STATIC*

By
Howard K. Morgan
(Transcontinental and Western Air, Inc., Kansas City, Missouri)

Summary—*The problem of rain static, particularly as it affects aircraft, is treated.*

Rain static is due to particles of rain, snow, or dust striking aircraft antennas. It has been found that the disturbance is electrostatic in origin and that an electrostatically shielded loop antenna reduces it materially.

Ever since aircraft have been equipped with radio receiving apparatus, it has been known that under certain conditions, usually in rain or snow, reception was seriously limited by a heavy static interference. The interference became an increasingly serious problem as radio receivers became more sensitive and radio shielding was improved on the aircraft. The interference, whether from rain, snow, or sand, has been called rain static.

Few, except those intimately connected with aircraft radio work, have ever heard this type of static interference. It is particularly elusive and may occur but a few times in any particular locality in a year. In transcontinental operation of airlines, it may affect reception for extended periods and cover distances of several hundred miles of the route. It is particularly bad in the months from January to May and will be reported in most snow or rain conditions in those months.

The effect in the worst conditions is to prevent the reception of a 400-watt station on low (200 to 400 kilocycles) or high frequencies (3000 to 5000 kilocycles) at distances as close as a few miles. Under usual conditions, it will prevent reception at distances ten miles or more from these stations.

Rain static sounds quite different from ordinary electromagnetic thunderstorm static. It is often steady and gives anything from a low rumble to a high pitched squeal in the headphones. Sometimes it is quite intermittent, and this usually occurs in rain squalls or in conditions of varying rainfall.

That rain static should be noticed in aircraft and not usually at ground stations suggests that the speed of flying through the rain is

* Decimal classification: R114. Original manuscript received by the Institute, March 23, 1936.

largely responsible for this interference. This has been tested by slowing down a plane from 180 miles an hour to about 120 miles an hour, under conditions where the rain static was just barely blanketing out a station. The effect was to reduce the interference materially.

Rain static has been reported on land in the Great Lakes region at direction finding stations during gales accompanied by snow. The substitution of electrostatically shielded loops at these stations reduced the rain static to a negligible quantity. The velocity of the air over the antennas was probably not much in excess of sixty miles an hour during the interference.

Sand static was disturbing at the ground station of Transcontinental and Western Air, Inc., at Amarillo, Texas, with antennas which were erected about twelve feet above the ground and in conditions of wind which blew sand from the ground over the antenna structure. When these antennas were relocated on the top of the hangar, the interference was far less pronounced.

Although there is no general agreement on the mechanics of rain static production, it is believed that it is due to the fact that the particles of rain, snow, or sand (dust) are charged in a nonuniform manner. This may take the form of a difference in polarity of charge or merely a difference of potential with the same sign of charge. It is known that small droplets of water falling through free air will assume a different charge from large droplets falling in the same region. The velocity with which the particles fall is too low to cause serious or rapid change in potential of fixed antenna systems in still air. With a plane passing through the variously charged droplets, the change of charge may be very rapid and serious in effect.

The question immediately arises as to the necessity for the droplets to be variously charged in order to give interference. Would not the speed of a plane be sufficient to cause interference due to the number of impacts with charged droplets and resultant leakage of charge causing a current to flow in the antenna structure? Experimental proof shows that, under some conditions of flight, the plane will build up sufficient charge to show a plainly visible corona discharge (St. Elmo's fire) and yet the radio reception may be affected very slightly. When the corona discharge from parts of the plane becomes very pronounced, there will be interference, but this may actually be electromagnetic in origin from the heavy corona current flow.

It would be expected, on the basis of the explanation given, that the disturbance on an antenna located near the tail of the plane would be more serious than on one near the nose because not only is the velocity of air higher in the slip stream, but the propellers tend to mix the air and allow some particles to touch the metal fuselage structure and change charge. This is very definitely noticeable, and conditions have been encountered where the antenna near the nose is relatively free from rain static while the antenna near the tail will be disturbed with rain static. The ratios in these cases are not tremendous, and the conditions where the nose antenna is superior are difficult to find and represent a condition of light rain static.

Static leak arrangements have not been found to improve reception materially, and the problem has resolved itself into finding some method which would prevent the charged particles from striking the antenna structure—an electrostatic shield. From the experience noted at the Great Lakes and from the apparent theoretical explanation for the origin of rain static, it was decided to construct an electrostatically shielded loop and test its efficacy.

In May, 1935, a loop was constructed and mounted on the upper surface of the wing of a Ford transport on the center line of the plane. A condition developed over Kansas City, Missouri, which was believed to be conducive to rain static. A flight was made and it was found that hail, rain, and snow were present at various levels and in varying amounts. Some conditions of rain static interference were found which blanketed out the Kansas City radio range completely at distances as close as three or four miles. A switching arrangement was used to transfer the receiver quickly from the antenna to the loop. The improvement noted was from a condition of not being able to hear anything but rain static with the vertical antenna to one of being able to hear the radio range clearly, using the loop, with occasional small traces of the rain static. In each condition of severe rain static, the switch was used to transfer the receiver many times between the loop and antenna. Thus, it was proved beyond a shadow of doubt that the electrostatically shielded loop, at least, would materially aid in the reception of signals in rain static conditions. At estimated distances as great as thirty miles from the station, the same relative conditions existed. Tests were not carried on at greater distances than about thirty miles.

During the summer of 1935, another loop was mounted on a Douglas transport in airline operation on the top and near the tail. Unfortunately, this loop was within a few inches of the vertical fin antenna. It was believed that electrostatic charges built up sufficient current in the vertical antenna to induce rain static disturbance in the loop. No conditions of severe rain static were encountered by that plane during the summer months, and the results of the test were largely negative.

In February, 1936, a loop was installed on another Douglas transport under the nose, and the usual antenna used in that location was moved to the upper surface of the fuselage. Numerous rain static conditions were encountered shortly, and the tests again definitely proved that, in most cases, the loop gave good reception when the vertical antennas were useless because of rain static. Thus, the results of the tests of a year before were definitely re-established.

Loop construction for modern high speed transports becomes a problem. There have been constructed a number of loops for aircraft use which are housed in a streamlined "egg." These reduce air resistance to a negligible quantity. The loop may be more easily rotated in the fixed shell. Further, the inside of the compartment which surrounds the loop may be lined with an electrostatic shield of parallel and insulated conductors which are grounded at one end only.

The question arises as to whether this type of construction will reduce rain static, as well as the open metallically shielded loop. It may be that the electrostatic charges will form currents on the surface of the shell which will, in turn, set up an electromagnetic field which will produce serious interference in the loop, regardless of the intervening electrostatic shield. The metallically shielded and exposed loop does not have this disadvantage. It can have currents flowing on its surface, or parts of its surface, without there being an appreciable current set up in the loop itself because of the low resistance of the loop shield.

Some few tests on antennas located within fabric wings have shown little or no betterment in rain static conditions, and yet the charged particles certainly cannot strike the antenna as located. Currents are most probably formed in rain static on the wet fabric wing, inducing the disturbance or capacitively coupling the disturbance to the wing antenna.

The last problem concerns de-icing of the loop antenna. It was found that with vertical antennas, particularly short ones, no signal was heard on the receiver when some types of ice covered the antenna and its insulators. An antenna so iced will be so thoroughly shielded that signals of a 400-watt station at ten miles or sometimes less will be unheard. With the loop antenna, which is composed of a metal tubing with a short gap in the circumference, there is always the possibility of this gap icing over and shielding the loop from electromagnetic waves. With modern air-pressure-operated de-icer equipment the problem is rather easy, and a small rubber section which is periodically expanded should de-ice the gap. It would also be necessary to de-ice the remainder of the loop structure to prevent its loading up with ice and forming a hazard in the mechanical sense. It is known, by Transcontinental and Western Air flight tests, that an eighteen-inch diameter loop may ice to the extent that there is but a two- or three-inch diameter hole at the center, the remainder being filled in with built-up ice. The loop now being used is but ten inches in diameter, but a comparable problem exists.

Further reduction in rain static is to be expected when the thickness of the metal loop shielding is increased above the 0.040-inch wall aluminum tubing now used. The gap in the circumference is now about a quarter inch, and this can be shortened and a parallel wire shield placed over it with the wires grounded to the loop structure on one side of the gap.

Further research in the underlying principles of rain static and methods of reducing it are being undertaken by the Radio Technical Committee for Aeronautics.

Acknowledgment

Acknowledgment is made to Mr. D. S. Little, of the RCA Manufacturing Company, who suggested the use of a shielded loop, based on his experience with Great Lakes loop antennas.

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MODES OF OSCILLATION IN BARKHAUSEN-KURZ TUBES*

By
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Summary—Experiments are described in which "parasitic" oscillations are elicited, and suppressed, in a generator of the Barkhausen-Kurz type by means of either voltage or circuit adjustments. The performance of tubes of new design employing a straight filament, helical grid, and cylindrical plate cut transversely into three sections is described, and the bearing of the experimental results on the problem of the modes of oscillation of the resonating grid oscillator is indicated. A method of stabilizing frequency by the use of such tubes in conjunction with a tuned transmission line is given.

Introduction

The problem of eliminating parasitic oscillations in a Barkhausen-Kurz oscillator has been treated recently by Pfetscher and Müller[1] who designed a series of tubes to generate pure frequencies. The question is one which is intimately connected with that of the modes of oscillation possible in such tubes and it is of considerable importance, especially when it is desired to use the oscillations for communication. Parasitic oscillations are at times encountered in work at the lower frequencies, below 300 megacycles, with conventional circuits because in a system consisting of a tube generator and its associated circuits which has several degrees of freedom the conditions for oscillation in several modes may be met under a single set of operating conditions. As a result several frequencies may be generated at once with a reduction in power at the frequency desired. In an electronic oscillator the situation is aggravated by the added factor that the space charge between the electrodes contributes additional degrees of freedom to the system. However, in general only a few of the added modes of oscillation are consistent with the conditions imposed by the geometrical form of the electrodes, and by the potential differences and impedances between electrodes. The purpose of this paper is to present the results of studies which illustrate the complicated character of these oscillations and to point out the conditions favorable for the production, or suppression, of parasitic oscillations in certain cases. Tubes of new design using a helical grid and a plate cut transversely into three sections were constructed. The performance of such tubes is

* Decimal classification: R133. Original manuscript received by the Institute, February 27, 1936. Published with permission of the War Department.
[1] Pfetscher and Müller, *Hochfrequenz. und Electroakustik*, vol. 45, p. 1, (1935).

described and some results with them which throw light on the modes of oscillation of a resonating grid tube are presented. In addition, the advantages afforded by this type of construction are pointed out with special reference to the problem of stabilizing frequency.

Measurement of Wave Length

The determination of wave length is of primary importance in this study. Two methods were used concurrently. One method employed a parallel wire transmission line or Lecher wire system four meters long. The fixed end of the line was terminated physically by a short-circuiting bar and a crystal detector was coupled to this end of the line. The movable bridge across the line was a brass plate eight inches square soldered to long brass bushings fitting closely on and sliding on the wires. This arrangement insured that the plate was always held perpendicular to the wires. The plate was subjected to remote control by a simple cord-and-pulley system and its position could be read to the nearest millimeter. The output of the crystal detector was led either to a direct-current microammeter or to an audio-frequency amplifier when tone modulation was used with the generator. The coupling between the generator and the receiver was made as loose as possible to decrease the possibility of interaction between the two systems.

The second method of determining wave length employed the RCA-955 tube as an oscillating detector in a conventional Hartley circuit. One such circuit operated in the range from 2.0 to 3.0 meters and a second from 3.5 to 5.0 meters. Harmonics in these circuits were used to determine the wave length of the Barkhausen-Kurz oscillations most of which were less than one meter in length. These circuits were calibrated using the Lecher wires and afforded a rapid means for searching for parasitic frequencies. In case of doubt as to the order of a harmonic recourse was had to the Lecher wire system. The 955 tube was followed by one stage of amplification.

Parasitic Frequencies Dependent on Voltage Adjustments

A tube of given structure may often be made to oscillate in any one of several distinct frequency ranges by adjustment of the supply voltages and the filament current. This type of operation has been noted previously[2] and the oscillations variously referred to as "harmonic" oscillations, or as "normal" and "dwarf" oscillations. The performance of an experimental tube of General Electric manufacture illustrates

[2] The review of work prior to 1933 given by Megaw, *Jour. I.E.E.* (London), vol. 72, p. 313, (1933), is so comprehensive that no detailed list of general references is given in this paper.

certain features of this behavior for tubes of the Clavier[3] resonating grid type. The tube employs a straight filament, a helical grid whose two ends are connected to a transmission line external to the tube, and a cylindrical plate. For the present tube, line length has an influence of secondary importance in determining wave length compared to the effect of the applied voltages.

Table I shows some of the wave lengths which may be obtained from this tube together with the operating conditions. The last column gives values of $\lambda^2 E_g \times 10^{-4}$ which according to the Barkhausen law is constant for a given mode of oscillation.

TABLE I

λ (cm)	E_g (volts)	I_g (ma)	E_p (volts)	$\lambda^2 E_g \times 10^{-4}$
22.6	470	70	−75	24.0
22.8	450	70	−75	23.4
32.1	320	100	−15	32.9
34.5	280	100	−20	33.3
36.0	220	65	−17	28.5
88.1	155	60	−10	120.5
94.7	135	50	− 2.5	121.0
97.0	125	50	− 2.5	117.5

In short, this tube is capable of oscillating in any one of three well-defined regions with each region characterized by its own Barkhausen constant. Oscillations in the thirty-four-centimeter range are more powerful than those in the other ranges and these are attributed to resonance in the helical grid considered as an inductance tuned by its own distributed capacity. In the ninety-centimeter mode the grid must act largely as an equipotential electrode, while the twenty-two-centimeter oscillations may arise from resonance due to other features of tube structure such as lead length inside of the glass envelope.

When the applied voltages are selected to cause the tube to operate near the middle of any of these regions of oscillation no evidence of parasitic oscillations was found. On the other hand, if one works near the edge of an oscillation region, or selects operating voltages intermediate between those for the thirty-four-centimeter range and the ninety-centimeter range, or the thirty-four-centimeter range and the twenty-two-centimeter range, the tube generates not one wave length but several. These two transition regions display similar characteristics. Such a region is found when $E_g = 200$ volts, $I_g = 70$ milliamperes, and $E_p = -10$ volts. Fig. 1 shows the output of the crystal detector as a function of the length of the Lecher wire wavemeter under these conditions. It is clear that the peaks are divided into two separate families. The families are distinguished both by the magnitude of the peaks and their spacing. Table II gives the positions of the peaks on the centi-

[3] Clavier, *Elec. Communication*, vol. 12, p. 3, (1933).

meter scale used. Column 1 gives the ordinal number of a peak, column 2 the positions of peaks belonging to one family, column 3 the positions of peaks belonging to the second family, while columns 4 and 5 give the intervals between the peaks of columns 2 and 3, respectively.

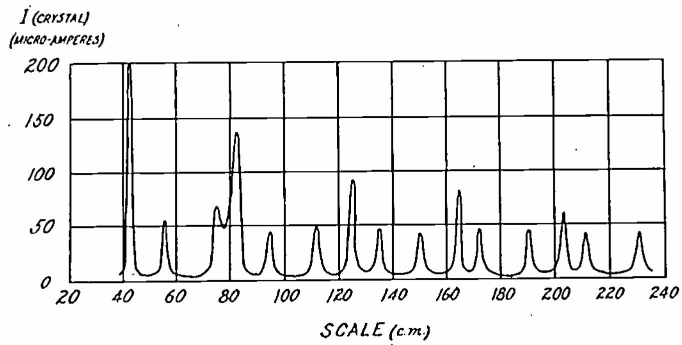

Fig. 1—Detector current when Lecher wires are used to receive radiation from a tube generating 81.0- and 38.6-centimeter waves simultaneously.

Thus the tube is oscillating in two modes simultaneously, one corresponding to a wave length of 81.0 centimeters and the second to a wave length of 38.6 centimeters. The value of $\lambda^2 E_g \times 10^{-4}$ for the former is 131.0 and for the latter 29.8 which agree reasonably well with those found for the Barkhausen constant in Table I.

TABLE II

1	2	3	4	5
1	42.0			
2		56.2		
3		75.6	40.2	19.4
4	82.2			19.0
5		94.6		
6		113.4	41.3	18.8
7	123.5			19.9
8		133.3		
9		151.6	40.0	18.3
10	163.5			20.4
11		172.0		
12		191.0	40.5	19.0
13	204.0			19.6
14		210.6		
15		230.0		19.4
	$\lambda_1 = 81.0$ cm.			
	$\lambda_2 = 38.6$ cm.			

Parasitic Frequencies Dependent on Circuit Tuning

The parasitic frequencies just noted arise when the operating voltages for a tube of given structure are improperly chosen. In other cases it is feasible to elicit, or suppress, parasitic frequencies by holding voltages constant and making appropriate circuit adjustments. The push-pull circuit of Kozanowski[4] lends itself well to this type of experi-

[4] Kozanowski, Proc. I.R.E., vol. 20, pp. 957–968; June, (1932), and Hershberger et al., *Physics*, vol. 4, p. 291, (1933).

ment. The circuit uses two tubes with the plates and filaments, or the plates and grids, connected to a four-wire transmission line. The tubes used were de Forest 552's. The operating conditions were as follows: $E_g = 500$ volts, $I_g = 400$ milliamperes, and $E_p = -90$ volts. The curves of Fig. 2 were obtained by varying the length of the plate Lecher wires. Abscissas give the distance from the axis of the tubes to the shorting bridge, while ordinates give the main wave lengths generated. The features of interest shown by the curves are the existence of a range of line lengths for which a single frequency is generated and of another

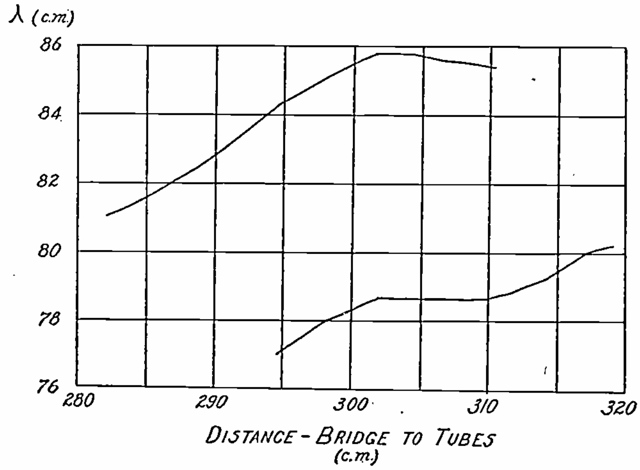

Fig. 2—Oscillations of two different wave lengths generated simultaneously as a result of improper circuit adjustments.

range for which two frequencies are generated. The range of lengths for which two frequencies are generated is one in which the line is approximately eight half-wave lengths long for one frequency and seven at the other. These frequencies exist side by side until the length of the line is adjusted to suppress one mode or the other.

It is often possible to choose such a combination of operating voltages and circuit adjustments as to elicit three, four, or more-modes from a circuit at once. Thus the Kozanowski circuit could be adjusted to generate 77.0-, 90.4-, and 104.8-centimeter oscillations simultaneously. Such an adjustment is held to be improper, first, because of the presence of the parasitic frequencies themselves, and, second, because under the conditions favorable for the production of parasitic oscillations, power output is considerable below the maximum attainable. It is possible that the multiplicity of frequencies noted by Pfetscher and Müller[1] for certain tubes were due to improper tube operation or circuit adjustments, as well as to faults inherent in tube design.

Helical Grid Tube With Trisection Plate

Tubes of new design were constructed which differ from those of Clavier in that the plate is cut transversely into three sections independently connected to the external circuit so they may be held at desired differences of potential with respect to each other. The two outer sections are usually held at the same potential. A photograph of one such tube is shown in Fig. 3. In a tube of this design the equipotential surfaces immediately inside of and concentric with the plates may be made either concave or convex inwards at will. This arrangement

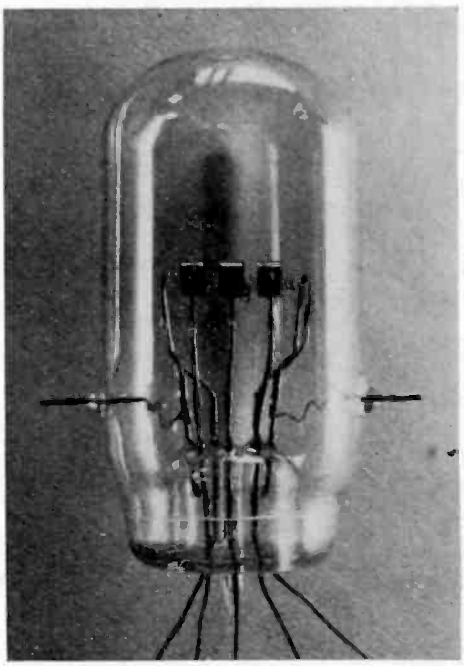

Fig. 3—Helical grid tube with trisection plate.

ment affords a flexible control on the boundary conditions to which the space charge in the grid-plate space is made subject.

In one tube of this type the leads inside of the glass envelope were given such a form as to isolate in effect the tube elements. No change in frequency could be measured on varying the length of a transmission line coupled to any pair of electrodes or to the ends of the helical grid. Two distinct sets of operating conditions were found. The curves of Fig. 4 show one region of oscillation for the tube. To excite this mode, we set $E_g = 250$ volts, $I_g = 150$ milliamperes, and the inner plate section was held α times as negative with respect to filament as the two outer sections. Abscissas in Fig. 4 give the negative bias on the outer sections while ordinates give the wave lengths generated. Curves are shown for

$\alpha=3$, 5, 10, 15, and 20. Oscillations were feeble for $\alpha=3$ or 20 but were much more powerful for the other values of α. Under the operating conditions outlined the central plate section obviously drew no current but its bias affected the current distribution along the length of the grid. The middle section of the grid becomes appreciably cooler when the middle plate section is given a large negative bias. The oscillations in the wave-length range from thirty-eight to forty centimeters are attributed to resonance in the grid helix coupled to the end plate sections. The natural period of the grid if isolated from the other elements is calculated to give rise to a wave length of the order of twenty-five

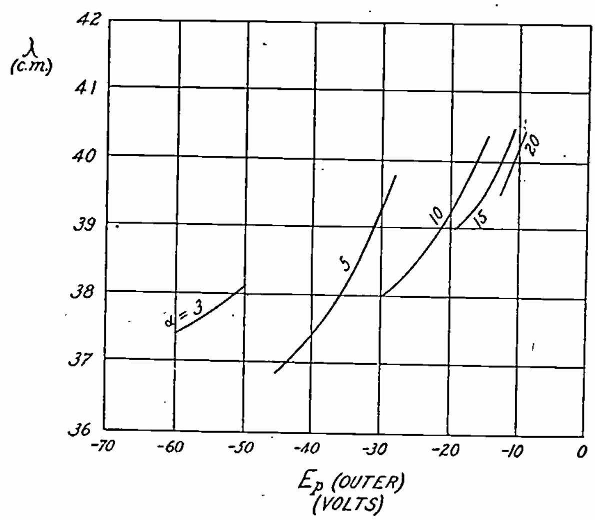

Fig. 4—Oscillation region for trisection plate tube. Inner plate section is held α times as negative with respect to filament as outer plate sections.

centimeters. The presence of the end plate sections with leads is sufficient to account for the discrepancy between this wave length and those actually observed.

The second mode of oscillation was elicited when $E_g=170$ volts, $I_g=130$ milliamperes, and the outer plate sections were held γ times as negative with respect to filament as the inner plate section. Under these conditions the end portions of the grid were appreciably cooler than the central portion. γ was given values 2, 3, 4, 6.5, and 10. The curves of Fig. 5 show the results obtained. Abscissas give the negative bias on the inner plate section, which now draws current, while ordinates give the wave lengths generated. The region from forty-eight to fifty-five centimeters included by the dotted lines marks the operating conditions for the greatest amount of power. Oscillations at wave lengths shorter than forty-eight centimeters were feeble while in the neighborhood of sixty centimeters a variety of parasitic oscillations were elicited. In this second mode of oscillation the helical grid is driven at its central portion and probably acts largely as an equipotential surface.

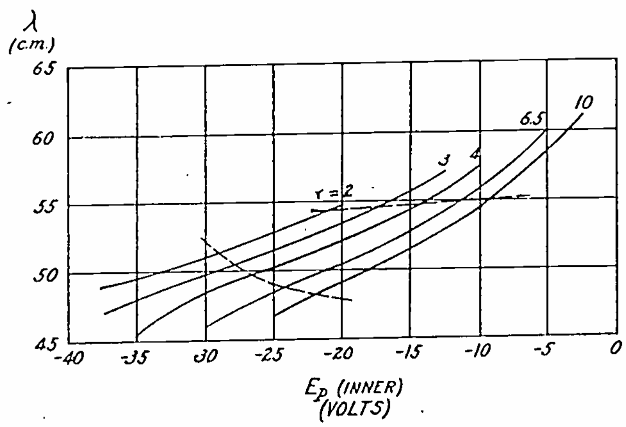

Fig. 5—Oscillation region for trisection plate tube. Outer plate sections are held γ times as negative with respect to filament as inner plate section.

In later tubes employing a three-section plate the leads from the grid and from the three plate sections were made as short and straight as possible to permit the use of external transmission lines. One tuned

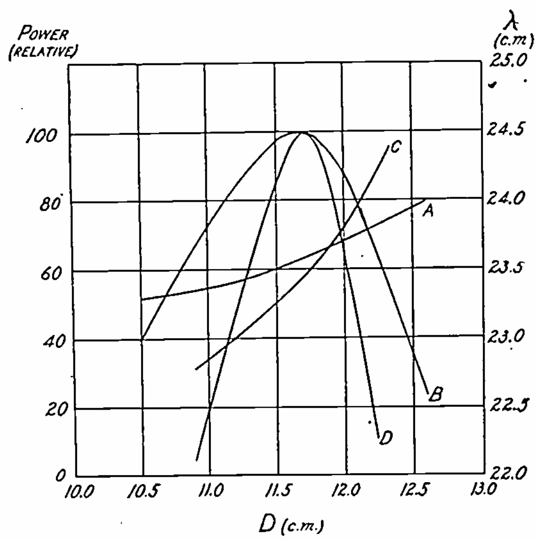

Fig. 6—Curves illustrating effect of grid tuning on wave length, curve A, and on relative power, curve B; also effect of plate tuning on wave length, curve C, and on relative power, curve D.

line was connected to the end sections of the plate, a second line to the ends of the grid, and the two lines coupled together as in the Kozanowski oscillator. Optimum operating conditions for one tube were as follows: $E_g = 200$ volts, $I_g = 80$ milliamperes, $E_p(\text{outside}) = -22$ volts, and $E_p(\text{inside}) = -125$ volts. That is, for this tube, the inner plate section was best held from five to six times as negative with respect to filament as the outer plate sections.

Fig. 6 shows the effect of the length of the lines on the wave length generated. Abscissas represent the distance from the shorting bars terminating the transmission lines to the point where the leads join the plate sections proper. This figure shows the effects obtained by varying the lengths of plate and grid transmission lines separately, with D_{pl} and D_{gr} indicating these lengths, respectively. Curve A represents the wave lengths obtained when $D_{pl} = 11.8$ centimeters and D_{gr} was varied while curve B represents relative power. Curve C was obtained on holding D_{gr} constant at 11.8 centimeters and varying D_{pl} while D is the relative power curve. It is to be noted that under these conditions plate tuning is approximately three times as effective in determining both wave length and power as is grid tuning.

The effect of the adjustments of the four-wire transmission line on the wave length generated shows that under appropriate working conditions a tube of the helical grid type operates in a true push-pull fashion. In fact, the entire system bears a striking resemblance to the two-tube circuit of Kozanowski. The elementary oscillators making up the Clavier generator employ one glass envelope and a common filament and plate, but owing to its helical construction, the grid offers a high impedance at a narrow frequency range and is thus able to support the necessary radio-frequency difference of potential along the direction of its length. One important distinction between the resonating grid and Kozanowski oscillators lies in the type of coupling between the elementary oscillators constituting them. In particular there is present in the former, but not in the latter, "electron" coupling, that is, the motion of electrons in one end of the Clavier tube plays a direct part in determining the forces on electrons in the other end. The "electron" coupling is subject to a degree of control in the tube with the three-section plate by adjustments of the direct-current potentials applied to the various sections.

The tube with the three-section plate makes possible the use of a plate transmission line and thus the best position for a shorting bar on the line may be determined. The tube may be operated under conditions approximating those for the Clavier tube if the bar terminating the plate line is placed at a distance from the plates exactly equal to an integral number of half-wave lengths and if the three plate sections are held at the same direct-current potential. The effects noted on departing from these conditions serve to determine whether or not the Clavier tube is operating under optimum conditions.

The merits of plate tuning were put to test by holding the length of the receiving Lecher wires used as a wavemeter constant and noting the power received as the lengths of the plate and grid transmission lines were adjusted simultaneously to generate the fixed wave length desired. Fig. 7 shows the relationship between D_{gr} and D_{pl} to generate 23.40-centimeter waves and the relative power curve. The power curve displays a maximum when $D_{pl} = 11.7$ centimeters and $D_{gr} = 11.25$ centimeters. It is very nearly symmetrical about $D_{pl} = 11.7$ centimeters. Thus the maximum amount of power is generated when the plate line is exactly one-half wave length long within the limits of the error of the measurements, but with the grid line a few millimeters shorter than a

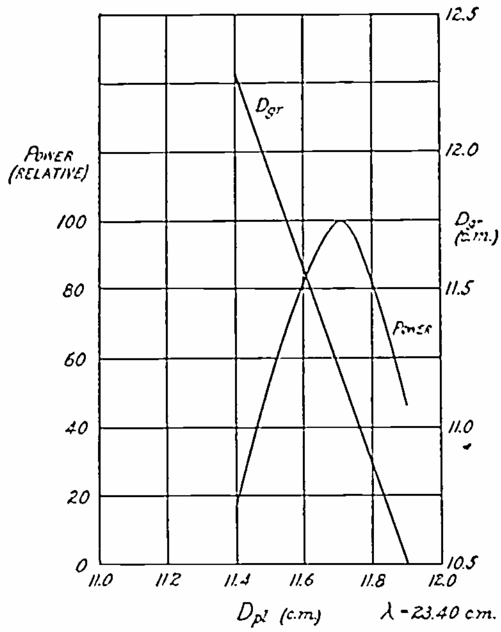

Fig. 7—Relationship between length of grid transmission line and that of plate transmission line to generate 23.40-centimeter waves, and the corresponding curve for relative power.

half wave length. This conclusion is borne out by experiments carried on throughout the range of wave lengths generated by the tube. In the curve shown, the received power drops to one half the maximum value when the difference between one-half wave length and the length of the plate line is plus or minus two millimeters. In the present instance, it is also noted that the length of the plate line has 3.5 times as great an effect in determining wave length as has the length of the grid line.

It thus becomes evident that for generating the maximum amount of power the shorting bar terminating the plate line could be placed to connect directly the end plate sections and this line eliminated. On the other hand, one advantage afforded by plate tuning is that, due to its marked influence on wave length, in combination with grid tuning a single tube may be used over a greater frequency range than if grid tuning alone were used. In addition, the use of the tuned transmission line instead of a shorting bar connected directly between the end plate sections permits one to take advantage of the selectivity afforded by such lines[5] and thus attain a high degree of wave length stability. In either case, we meet equally well the condition for generating the maximum amount of power but the shorted plate sections meet this condition for any wave length in the region of interest while the resonant line meets it for but an extremely narrow range of wave lengths.

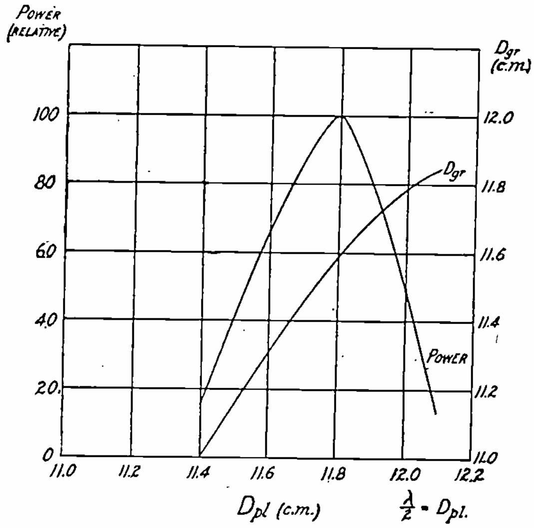

Fig. 8—Relative power curve and curve showing length of grid line under condition that plate line is given its optimum length, namely, one-half wave length.

Fig. 8 shows the effect on power of simultaneous adjustments of plate and grid transmission lines with the plate line always given its optimum length, that is, one-half wave length. An examination of the curve shows that the maximum amount of power for this tube is generated at a wave length of 23.6 centimeters and that the power drops to approximately one half the maximum value at 23.1 and 23.9 centimeters. This curve gives a measure of the sharpness of resonance of the grid helix itself. The second curve in Fig. 8 shows the length of grid line for maximum power. The fact that it is several millimeters shorter than the plate line is attributed to an end effect at the grid.

In Fig. 9 is shown the dependence of wave length and power on the applied grid voltage. Filament current and the plate voltages were held constant and the transmission lines were adjusted for maximum power

[5] Terman, *Elec. Eng.*, vol. 53, p. 1046, (1934).

at a wave length of 23.32 centimeters. The wave-length change was measured by noting the progressive change in position of the shorting plate on the Lecher wires used as a wavemeter when this receiving line was twenty half-wave lengths long. Under these circumstances the reflector position could be determined to the nearest millimeter. It will be observed that wave length changes from 23.34 centimeters at 180 volts on the grid to 23.31 centimeters at 210 volts. The high degree of wave-length stability with changes in grid voltage is attributed primarily to the resonant transmission line connected to the end plate sections. The effectiveness of this control is brought out forcibly by

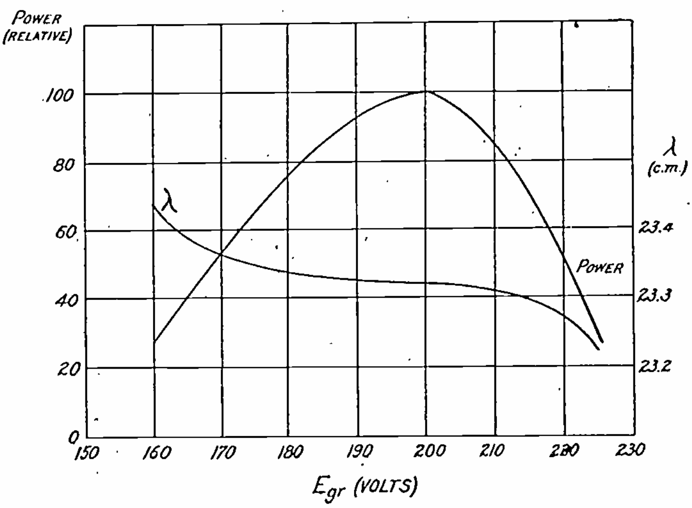

Fig. 9—Curve illustrating the frequency stability of trisection plate tube in conjunction with tuned transmission line. Grid voltage was varied and effect on power and wave length noted.

a comparison with the corresponding effect in the tube whose performance is outlined in Table I. In this case a change in grid voltage from 320 to 280 volts gave rise to a wave-length change from 32.1 to 34.5 centimeters.

The effect of varying the potential difference between the inner plate section and the outer plate sections and filament was determined under conditions of constant grid voltage and current. Observations were taken on wave length and relative power. To obtain the maximum power output for all adjustments the outer plate sections were held from eighteen to twenty-two volts negative with respect to filament in the course of a run in which the potential difference between the inner section and the filament was varied from minus ten to minus 300 volts. Wave length changed from 23.71 to 23.61 centimeters. The power output increased by a factor of three on increasing the negative bias on the inner plate from minus twenty-two volts, where it equalled that on the outer plate sections, to minus 125 volts. The power curve is shown in Fig. 10 and displays a broad minimum when all plate sections are given the same bias. The increased efficiency of the tube for large values of the negative bias given the central plate section is attributed to the fact that the end portions of the electrodes rather than the central portion play the predominant part in the generation of oscillations. The current drawn by the central portion of the grid probably adds little or nothing to the maintenance of oscillations.

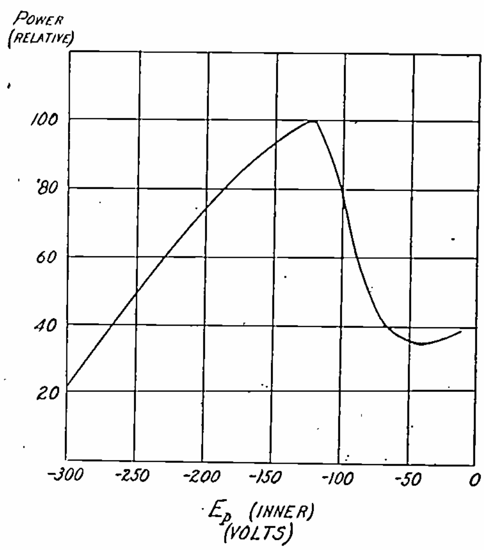

Fig. 10—Power generated by trisection plate tube as a function of the negative bias given the inner plate section.

The conclusions reached in this paper concerning the principal mode of oscillation of a microwave generator of the helical grid type are in general accord with the picture presented by Müller[6] as a result of his studies with this type of tube. The evidence afforded by the present design serves in particular to emphasize the importance of the end portions of both grid and plate when the tube is oscillating in the mode for which the grid resonates. In this mode the tube may be viewed as a frequency "doubler" whose operation is in some respects analogous to that of a conventional full-wave rectifier. The fundamental driving frequency, which may be detected as a relatively feeble oscillation, is given by the Barkhausen law. The strong second harmonic oscillation arises from distortion in the system accentuated by the resonant grid. Under these circumstances the second harmonic is the predominant one.

[6] Müller, *Ann. der Physik*, vol. 21, p. 611, (1935).

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HIGH VOLTAGE MERCURY-POOL TUBE RECTIFIERS*

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Summary—*Mercury-pool tubes of the ignitron type offer attractive features for power rectifier service. This paper discusses the application of this type to controlled high voltage, high power rectifiers for radio transmitters, and some problems peculiar to their use as such. A developmental rectifier of this type, which has been in trial operation, is described.*

MERCURY-POOL tubes of the ignitron[1] type have been used very successfully for some time in resistance-welder control. These tubes also offer attractive features for power rectifier service, particularly for certain power ranges, and several applications in this wide field have already been made. This paper will discuss the application of this type of tube for high voltage, high power rectifiers for radio transmitters, and some problems peculiar to their use as such. It will also describe such a developmental high voltage pool tube rectifier which has been in trial operation.

The ignitron type of mercury-pool tube has been adequately described in previous papers.[1] Briefly, it consists of an evacuated envelope with an anode, a mercury-pool cathode, and a third, or starter, electrode of high resistance material, dipping into the mercury pool. When a brief pulse of current is passed through the starter, an arc spot is established on the surface of the mercury pool and power current flows from the anode to the cathode. This action is repeated each time it is desired to start conduction. The tube is thus a controlled half-wave rectifier, which will not start conduction unless there is (1) positive voltage on the anode, and (2) an impulse of starter current passed through the starter. Such a device has very attractive possibilities as a controlled high voltage rectifier. These possibilities made it worth while to set up a developmental rectifier to work out the circuit and application problems involved. Before describing this developmental rectifier,

* Decimal classification: R356.3×621.313.7. Original manuscript received by the Institute, February 24, 1936. Presented before Philadelphia Section, February 6, 1936.

[1] J. Slepian and L. R. Ludwig, "A new method of starting an arc," *Elec. Eng.*, vol. 52, p. 605; September, (1933).
Ludwig, Maxfield, and Toepfer, "An experimental ignitron rectifier," *Elec. Eng.*, vol. 53, p. 75; June, (1934).
J. M. Cage, "Theory of the immersion mercury-arc ignitor," *Gen. Elec. Rev.*, vol. 38, pp. 464–465; October, (1935).
W. G. Dow and W. H. Powers, "Firing time of an ignitor type of tube," *Elec. Eng.*, vol. 54, p. 942; September, (1935).

let us consider what might be gained by using this type of tube in a high voltage rectifier; and to this end, let us make a comparison of such a rectifier with (I) a grid-controlled hot-cathode rectifier, and (II) a grid-controlled mercury-arc tank rectifier. It should be understood that this comparison is based on operation of a relatively small developmental unit. Tubes for higher voltages and currents are not at present available. There appears to be no fundamental reason, however, why such tubes of increased ratings cannot be made available when needed.

I. Comparison With Hot-Cathode Rectifier

One advantage which the pool tube has over the hot-cathode tube is that of emission. When more current is called for, the arc-spot area simply increases, and the electron demand is supplied. Currents of 100 or even 1000 times the average rating of the tube can be supplied easily. This eliminates any emission limitation from the cathode; and in case of an arcback, those tubes which carry high current in a forward direction are not affected adversely. In a hot-cathode tube, during an arcback started by one tube this high forward current may sputter the emitting surface of another tube to a harmful degree, and perhaps throw some of it to the anode. This causes the condition every radio station operator is familiar with, where one tube arcs back, thereby causing several others to arc back. Another advantage of this high-peak-current-carrying capacity is that an initial condenser filter can be used without bad effect on the tubes. Such a filter reduces the anode bombardment resulting from working into an initial inductance filter; this reduction makes arcbacks less probable for the same value of peak inverse voltage.

Another advantage over the hot-cathode tube is also connected with greater freedom from anode emission. The collection on the anode of particles of barium-emitting surface, particularly after cathode sputtering due to passage of high current, may result in reverse currents. This action of course does not exist in pool tubes.

Another point in favor of pool tubes is the absence of heating time, which permits quick replacement of a tube in case of a failure or other emergency, and immediate resumption of service.

Another advantage in the use of the ignitron type of tube is the fact that instead of an upper and lower condensed-mercury temperature limit (as must be set in both hot-cathode and tank rectifiers) there is only an upper limit, approximately the same as in hot-cathode tubes. We have operated a mercury-pool tube of this type in liquid air, at a temperature of at least −200 degrees centigrade with the mercury pool frozen solid and far below its melting point; under these conditions the tube would apparently start and operate as well as at room temperature. This single temperature limit considerably simplifies the cooling arrangements and control.

Last but not least is the point of operating life. So far we have not enough data to give definite figures; but what we have is encouraging, and we expect to get life at least as good as that of the best hot-cathode tubes.

II. Comparison With Mercury-Arc Tank Rectifier

An advantage the ignitor-starter pool tube has in comparison with the mercury-arc tank for radio transmitter service is the absence of ionization in the pool tube on the inverse part of the cycle. In the mercury-arc tank this ionization is always present, because of the continuous presence of the holding arc on the mercury pool, and intense positive-ion bombardment of the anode will, therefore, take place unless the anode is very thoroughly baffled and shielded.

Another important advantage is the fact that these tubes are single half-wave units; thus they can be used in circuits other than those requiring a common cathode. Tank rectifier circuits require a common cathode. In high voltage rectifier service this means that with tubes the three-phase full-wave circuit can be used, which (in comparison with the six-phase half-wave circuit used by tanks) results in only half the peak inverse anode voltage for a given direct-current output, and in better transformer utilization, which means a smaller transformer for a given output.

In addition, these tubes are sealed-off units, complete in themselves, and require no pumps or other auxiliaries. No time is required to pump them back down after an extended power failure; they have no heating or aging time requirement, and in general they are air-cooled rather than water-cooled; in most cases no temperature control is required on the cooling medium.

Finally, there is the question of reserve capacity. Reserve capacity in a tank means one additional tank, complete; otherwise a failure of one part or a fault in the tank proper will take the whole rectifier out of service until it can be repaired. A tube rectifier, on the other hand, requires only several tubes for spares.

There is also the further advantage of the ignitron tube, over both the grid-controlled hot-cathode tube and grid-controlled mercury-arc rectifier, in the matter of control. Control of output voltage can be obtained in ignitron tubes in exactly the same manner as in grid-controlled devices; i.e., by starting conduction at the desired point in the cycle. The difference is that in grid-controlled devices the interior is usually in a conducting condition before it starts; the cathode or the arc spot is emitting electrons, which are restrained from reaching the anode by the charge of the grid; in other words, the tube is more or less forcibly "held off." Loss of control power usually means that the device turns fully on. On the other hand, the ignitron tube is completely dead until "fired"; there is no source of electrons; and in case of control failure, the device fails "off," or in the safe condition. Grids may easily become contaminated and lose their "hold-off" power. Such is not the case in the ignitron tube, and a more positive control results.

There are several characteristics peculiar to ignitron tubes which must be considered in designing radio transmitter rectifier equipment to use them. The first, naturally, is the starter itself. Starting these tubes requires that a pulse of current be passed through the starter each time the tube is to start. This current is not critical, and needs last for only a very short time; reliability of firing, however, requires that a good excess of current be used for at least a certain part of a cycle. As an example, we might cite the starting circuit in use in the South Schenectady installation, which represents the maximum starting power we believe should be used. This circuit puts a half-sine wave of a higher frequency current, having a peak value of eighteen amperes and a duration of six electrical degrees on a sixty-cycle basis, through the starter. Using two starting pulses per cycle, this represents an average current through the starter of 0.38 ampere. The total control circuit power input, including all filaments and losses, is less than 200 watts per main power tube in this circuit. We hope that with improved starter design we can reduce this considerably. In this connection, it should be pointed out that the starter power and circuit are the same regardless of the size of the tube; all that is necessary, in any size of tube, is to establish the arc. The same power is required for a higher rated pool tube as for the smallest.

The second characteristic which must be considered is that the tube is a *controlled* rectifier, and that conduction cannot occur unless there is exciting current passed through the starter as well as forward voltage on the anode. If two tubes are to start conducting at the same part of the cycle, as they do in the three-phase full-wave circuit with capacitive load, they must be fired together.

The third characteristic is the fact that a certain minimum current is required to maintain the arc spot, once started, and hence to continue conduction. This minimum current is determined by the design of the tube, and a representative value is three or four amperes. When the current drops below this value, the arc spot goes out, and conduction in the tube ceases in a time of the order of a very few microseconds. If the current is flowing through an external circuit containing considerable inductance (such as the high voltage winding of a transformer), the stored energy in this inductance (1/2 LI^2) must be dissipated or transferred; usually this is accomplished by the voltage across the inductance rising until the energy is transferred by raising the charge on the distributed capacity of the windings. But if the distributed capacity is low, the voltage may rise to a very high value before this balance takes place, high enough to cause an insulation puncture or flashover. Fortunately, the problem is easily met by shunting each of the power windings by a small condenser, with somewhat more resistance in series with it than is necessary to damp critically the *LCR* circuit thus formed. This circuit has a high impedance to power frequency but a very low impedance to the extremely steep-fronted transient set up by the arc snap-out.

This type of transient voltage protection, incidentally, is also desirable on all rectifier installations, and as a matter of fact is an excellent thing to have on any inductance subject to surges.

III. Developmental Installation at South Schenectady

As a means of obtaining design and operation experience on an ignitor-starter high voltage mercury-pool rectifier, a developmental rectifier was set up at the South Schenectady transmitter station, which houses the transmitters of WGY, W2XAF, W2XAD, and WPGC. The power circuit (Fig. 1) consists of a three-phase transformer with a delta connected primary and a Y connected secondary with an ungrounded neutral. Each of the legs of the Y is shunted by a 0.02-microfarad condenser in series with a 1500-ohm resistor, which

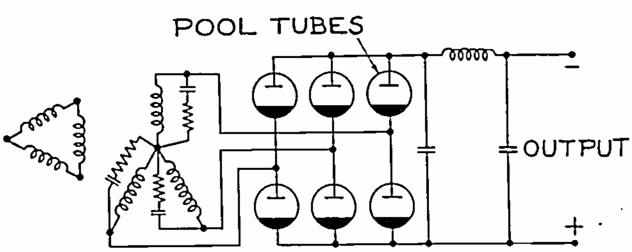

Fig. 1—Main power circuit.

acts as a transient voltage suppressor. This transformer feeds six ignitron tubes connected as a three-phase full-wave rectifier. Physically they are essentially a glass cylinder four inches in diameter and ten inches long, with an anode in the top and a mercury-pool cathode and immersion ignitor starter in the bottom. (See Fig. 1.) It should be understood at this point that these tubes were not designed for high voltage rectifiers, and that future high voltage tubes will contain design changes. It was desired, however, to get a high voltage rectifier installation into operation to work out the initial control and other design features of the rectifier, before undertaking development work on high voltage pool tubes.

The power transformer is capable of about seven amperes direct-current output, at a voltage determined by taps on the transformer and having a maximum of twenty kilovolts. The rectifier output is fed into a filter, which has an initial condenser of 3.7 microfarads, a filter choke of ten henrys, and a final condenser of three microfarads. The negative point of this output is grounded, and the positive point can be

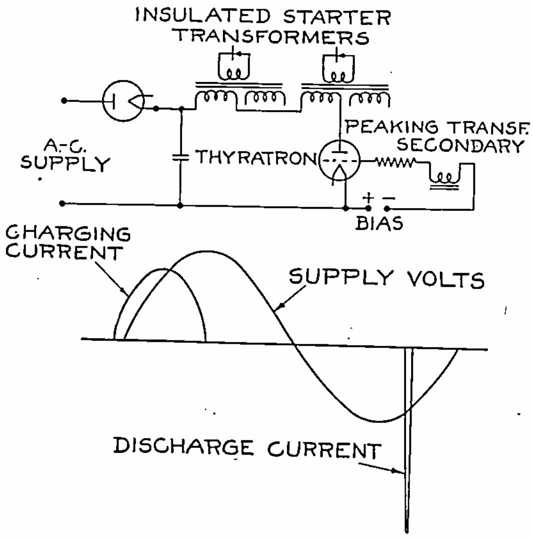

Fig. 2—Elementary firing control circuit.

switched to the transmitter of either W2XAD or W2XAF, or to a large adjustable resistance load.

The main tubes are fired through small transformers (about the physical size of a 150-watt transformer), one for each tube, which have their secondary windings insulated for cathode voltage to ground and connected across the starter of the tube they are to fire. The core of this coil is at ground potential, and the primaries at the low voltage of the control circuit.

The control circuit consists essentially of six type 866 rectifier tubes (rated 0.5 ampere average, two amperes peak) charging six two-microfarad Pyranol condensers, which are in turn discharged through the firing transformer primaries by six FG-57 thyratrons (rated 2.5 amperes average, fifteen amperes peak), thereby sending a pulse of current lasting about six electrical degrees through the secondaries and starters. An elementary diagram of one firing circuit is shown in Fig. 2. The condenser charges to the peak of the applied alternating voltage; and at any desired point during the half of the cycle when charging current to the condenser is held off by the rectifier, the condenser may be discharged through the primaries of the firing transformers by the thyratron. The exact point is determined by a six-coil peaking transformer which fires the thyratrons, which is in turn supplied by a small three-phase Selsyn (four inches diameter by six inches long). Turning this Selsyn determines at what point in the cycle the main tubes are fired, and by simply turning its dial, smooth control of the output voltage from zero to maximum is obtained.

It will be noticed that each thyratron fires two main tubes at the same time, one carrying current to the positive side of the load and the other carrying the return current from the negative side of the load,

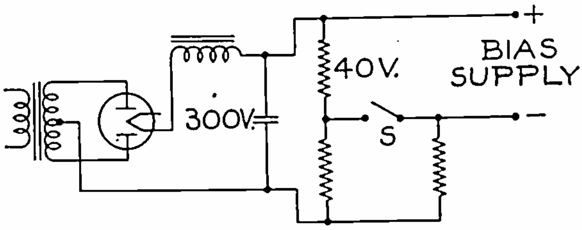

Fig: 3—Bias control circuit.

and that each tube is, therefore, fired twice during each cycle, once initially and again sixty electrical degrees later. This type of firing is necessary when working into an initial condenser load, as can easily be seen from an analysis of the power circuit.

High speed automatic relaying and fault-clearing may be secured by control of the hold-off bias on the thyratrons, either by electronic or mechanical means. In this installation we obtained high speed clearing simply by connecting the contacts of the existing instantaneous overload relays into the bias rectifier circuit, as shown in Fig. 3. When the relay trips open, it opens *S* and instantly raises the bias on the thyratrons to a point where the peaking transformer excitation cannot fire them; no more excitation is then applied to the starters. In event of an arcback, this easily clears the power before the inverse time-overload relays in the transformer primary circuit could operate.

The results we obtained with this developmental rectifier are very encouraging. The rectifier would operate very satisfactorily up to the arcback limit of the tubes. W2XAF was operated on several test programs, drawing five amperes at sixteen kilovolts from the rectifier, and W2XAD was also operated on a number of programs, drawing five amperes at eight kilovolts The operation was completely satisfactory from the standpoint of hum and noise, even when operating at two-thirds maximum voltage with Selsyn control "cutback." It might be expected that the discontinuous power output of a controlled rectifier operating at reduced voltage would result in ripple at the filter output; but in the three-phase full-wave circuit, power is delivered over six intervals during each cycle, and with an initial condenser filter of normal design, no trouble is experienced.

Five amperes at sixteen kilovolts, however, represents about the best we could get out of these particular tubes with a fair degree of reliability. Application of this type of rectifier must, therefore, await the development of high voltage ignitrons especially designed for this class of service. The possibilities are so attractive, however, that this development will undoubtedly be undertaken; and we hope to have available ignitrons to meet any high voltage, high power rectifier requirements.

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RECENT DEVELOPMENTS OF THE CLASS B AUDIO- AND RADIO-FREQUENCY AMPLIFIERS*

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Summary—*Class B audio-frequency and radio-frequency amplifiers have many applications and the distortion can be kept to a very low value if the necessary precautions are taken to prevent nonlinearity of such amplifiers. Undoubtedly, the most important factor in the design of a class B amplifier for low distortion is the characteristic of the driver system. Tube characteristics and the use of a proper load are also important but are more definite and more generally understood.*

The purpose of this paper is to present the results of recent developments of the class B audio and radio amplifiers to reduce distortion. The results of the investigations indicate that a heavily loaded driver system in general is undesirable because of the power consumed and because such loading results in greater distortion than obtainable by other means.

The general procedure adopted to reduce distortion was to prevent distortion in each unit of the amplifier system. Distortion balancing schemes are not only critical to adjust but are likely to introduce higher order harmonics and sum and difference tones which may be more objectionable than a higher measured value of lower order harmonics. Actual performance data are presented for medium and relatively high powered audio and radio systems. The necessary input requirements to permit the performance obtained are discussed.

Sufficient theory is given to make the paper complete and to show that the actual performance of such amplifiers can be quite accurately predicted if the necessary tube characteristics are known.

Introduction

THE general tendency toward high fidelity performance for radiophone transmitters as well as high modulation capability has necessitated appreciable development of the class B audio- and radio-frequency amplifiers to reduce distortion. The class B radio-frequency amplifier has, for several years, been used almost exclusively for the output system of radio transmitters. However, the application of the class B audio amplifier for high level modulation is gaining much recognition in the past few years, although such an amplifier was successfully used for modulating a 1000-watt broadcast transmitter as early as 1928.[1] The application of the class B audio amplifier for

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[1] Loy E. Barton, "A plate modulation transformer for broadcast stations," Univ. of Arkansas, Eng. Exp. Station Bulletin No. 8.

high level modulation has also been applied to a 500-kilowatt station[2] and to a number of medium powered stations.

This general trend permits considerable saving in power costs, reduces the number of tubes operating in parallel for a given output, and is inherently a more simple system as regards design, adjustments, and general maintenance.

For power outputs such that no convenient or economical tube complement is available, the saving of power in high level class B modulated systems may be offset by the cost of tubes and the modulation transformer and reactor. In such cases, the class B radio-frequency amplifier may be used to an advantage. The advisability of using a low level modulated system in such cases may be questionable when an economical class B audio amplifier for modulation is developed for the particular power classes now using low level modulation. Because of the similarity of the class B radio- and audio-frequency amplifiers, it is convenient to discuss some of the recent developments of both amplifiers for low distortion output.

Definitions

Since there is some question as to the true definition of the various types of amplifiers, it is desirable to define and discuss briefly the principle types of amplifiers as a basis for the discussions in this paper and to clarify some of the misconceptions regarding the various types of amplifiers.

A class A amplifier is an amplifier in which the plate current is essentially proportional to the grid voltage for the entire signal cycle or 360 electrical degrees.

A class B amplifier is an amplifier in which the plate current or resultant plate current is essentially proportional to the grid voltage for 180 electrical degrees of the signal and the power output for each tube is appreciably lower on one half cycle than the following half cycle.

A class AB amplifier is an amplifier that fulfills the conditions for a class A amplifier for low grid swings and fulfills the conditions for a class B amplifier during the upper part of a cycle for high grid swings.

A class C amplifier is an amplifier in which the grid excitation and bias are such that the direct-current value of the plate current and output current is proportional to, or tends to be proportional to, the direct plate voltage for intervals of time large as compared to the period of time for a radio-frequency cycle.

[2] J. A. Chambers, L. F. Jones, G. W. Fyler, R. H. Williams, E. A. Leach, and J. A. Hutcheson, "The WLW 500-kilowatt broadcast transmitter," Proc. I.R.E., vol. 22, pp. 1151–1180: October, (1934).

It should be noted that in the above definitions the value of bias is not given nor is the grid swing stated. These values for the amplifier tubes are not given because of the various types of tubes that may be used for any, or all, three types of amplifiers. For instance a zero-bias class B tube may be used as a class A amplifier if a positive bias is used such that a signal voltage applied to the grid will result in a plate-current change essentially propotional to the applied grid voltage for the entire 360 electrical degrees. Therefore, it is evident that a class A amplifier cannot be defined as a function of grid bias but it can be defined in terms of the relation of the grid voltage and plate current over certain ranges of grid voltage. The limit of grid swing and grid bias depends entirely upon the grid conditions that permit the plate current to meet the requirements of a class A amplifier.

The proper grid conditions for a class B amplifier are also the conditions that will permit the plate current to behave according to the definition for the class B amplifier given above. It is well known that no tubes are yet available and probably will not be available that have a linear plate current with respect to grid voltage from zero plate current to the limit of plate swing. Therefore, for low distortion in a class B amplifier the resultant plate current obtained from the slope of the plate-current vs. grid-voltage curve must be essentially linear from the operating bias point. This resultant plate current will be discussed in detail later.

From the definition of a class AB amplifier (sometimes called class A prime amplifier) the resultant slope of plate-current vs. grid-voltage is different at low and at high values of grid swing. If the plate resistance of the tube used is very high compared to the load resistance the resultant slope of plate current of two tubes in push-pull over the class A part of the cycle will approach two times the slope at points where only one tube is working. For very low plate resistance tubes, the two slopes may approach the same value if the load resistance is several times the value of plate resistance. It may be seen that in any case the class AB or A prime amplifier will inherently introduce distortion to a degree depending upon the plate resistance of the tubes used. A class B amplifier may have the proper bias for low amplification factor tubes and appreciable ouput may be obtained with no positive grid swing. This type of operation is a class B amplifier and not a class A prime amplifier as sometimes called.

The class C amplifier will not be discussed further except to state that for plate modulated service the grid bias and excitation necessary to cause the direct plate current to be proportional to plate voltage may vary with instantaneous value of plate voltage. In general, however, the class C amplifier conditions are usually met if the grid is operated at so-called saturated condition as regards bias and alternating-current excitation for all values of plate voltage to be applied to the amplifier.

Class B Audio Amplifier

The essential circuit and tube requirements for successful operation of a class B audio amplifier are much more severe than the requirements for a good class B radio amplifier because the output of the audio amplifier is a function of the instantaneous value of the plate current, whereas, in the case of radio-frequency amplifiers a distorted plate current is smoothed out by the tuned plate circuit.

The two most important factors in a good class B audio amplifier are the driver system supplying the audio voltage to the grids of the class B stage and the relation of the plate current to grid voltage for the particular tube. These two factors are discussed at some length because of their importance in the design of a low distortion class B audio amplifier.

Operation of the Driver Tubes

The grid current characteristic of a tube operating as a class B audio amplifier for a given plate load resistance must be known in order that a good driver system may be devised. The plate characteristics are also necessary for output calculations so that by using the circuit as shown in Fig. 1, the grid and plate characteristics may be obtained simultaneously. Such characteristics are shown in Fig. 1 for two 849 tubes and are used to illustrate more fully the driver requirements. It should be noted that the characteristics of the plate and grid currents plotted for various load resistances are for two individual tubes and that the grid current characteristics may vary widely for individual tubes. However, for a given load resistance and grid voltage, the plate current will be quite uniform for various tubes of the same type.

Referring to the grid current curves in Fig. 1, tube *A* is somewhat more erratic than tube *B*. The positive grid resistance represented by the slope of the grid current curve of each tube at a point soon after grid current starts is approximately 600 ohms. As the grid swings further positive, the resistance increases to infinity and then decreases to a negative resistance of approximately 600 ohms for tube *A* but the minimum negative resistance of tube *B* is not nearly as low. At peak grid swings of about +110 volts the instantaneous grid resistance of each tube is about 600 ohms positive resistance.

It will be noted that the plate current curves are essentially linear for grid voltages between −75 or −80 volts and +100 to +120 volts when the grid voltage swings in a positive direction. Therefore, if no voltage distortion is applied to the grids there should be no distortion in the plate circuit from approximately the −75-volt bias point to the plate current limit. However, the grid current wave to the grids of the 849 will be badly distorted as can be readily observed. If the grid current is applied from a source having appreciable impedance, the

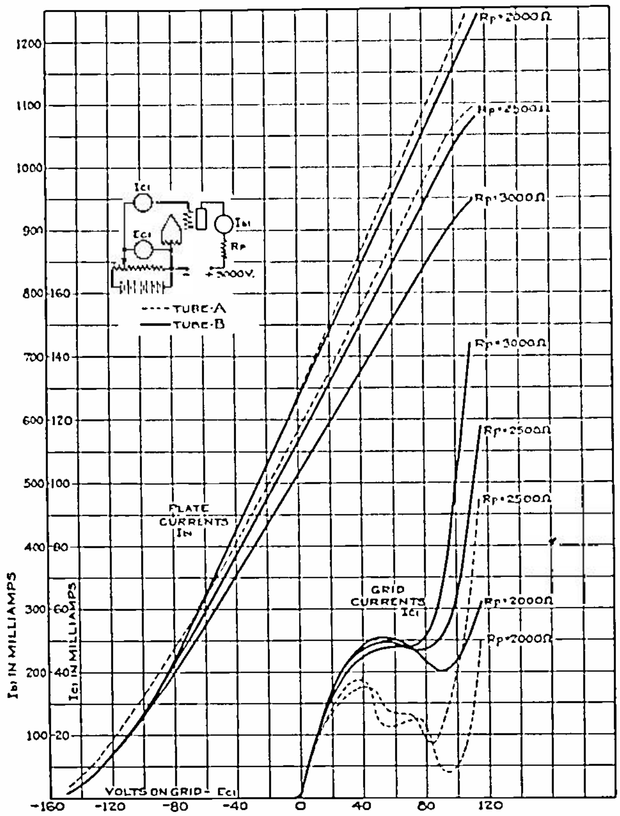

Fig. 1—Dynamic transfer characteristics of RCA 849.

voltage wave applied to the grid will be distorted, which will cause distortion in the plate circuit. Therefore, the internal resistance or impedance of the source supplying audio voltage to the grids of the class B tubes must be as low as practical in order to reduce voltage distortion to the grids.

Since the driver system is essentially aperiodic, it is obvious that any change in impedance in the grids of the class B tubes is reflected directly through the coupling means to the plate circuit of the driver tubes provided the coupling transformer approaches the performance of an ideal transformer. If an ideal coupling transformer is assumed the lowest equivalent resistance in series with the grids of the class B tubes is obtained when there is no loading resistance across the grids of the class B amplifier, thus permitting a maximum step-down ratio of the driver transformer, provided low impedance driver tubes are used. There has been considerable disagreement on the question of loading the secondary of the driver transformer to improve regulation to the grids of the class B audio amplifier tubes so that a short proof of the above statement is given. A proof of this statement is also given by McLean.[3] As will be shown later, this general statement also applies to class B radio amplifiers in so far as the audio cycle is concerned.

The equivalent circuit for a driver system in a class B audio amplifier is shown in Fig. 2 in which two tubes operating as a push-pull class

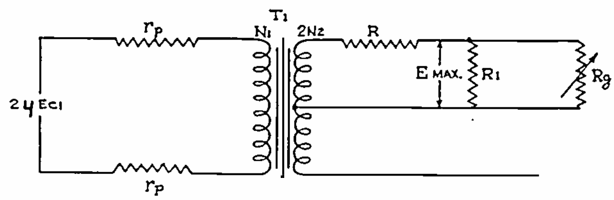

Fig. 2—Equivalent circuit of driver for class B audio amplifier.

A amplifier are assumed to be connected to the primary of the driver transformer, T_1. The generated peak voltage in the plate circuit of the two tubes in push-pull is $2\mu E_{c1}$ and the plate resistance in series with each lead to the transformer primary is r_p. The turn ratio of the driver transformer T_1 from total primary to total secondary is N_1 to $2N_2$ and the ratio to one side of the secondary is N_1 to N_2. The resistance, R, is the equivalent resistance transferred from the primary of T_1, in series with each grid; R_1 is the equivalent loading resistance across one side of the secondary, and R_g is the resistance of the grid of the tube which in general varies from a very high value to some low value. The variation of R_g depends on the particular class B tube and conditions under which it operates. From the characteristics of a tube such as shown in Fig. 1, the peak grid voltage $E_{\max}$ for a given power output may be read. From the characteristics of the driver tubes the expression $2\mu E_{c1}$ can be found. The general quantities in Fig. 2 may be used in the following expression to determine the equivalent series resistance R for a required peak grid voltage $E_{\max}$ for an ideal transformer:

$$2r_pN^2 = R \tag{1}$$

[3] True McLean, "An analysis of distortion in class B audio amplifiers," Proc. I.R.E., vol. 24, pp. 487–509; March, (1936).

in which $N = N_2/N_1$

$$2N\mu E_g \frac{(R_L)}{N^2 2r_p + R_L} = E_{\max} \tag{2}$$

in which R_L = parallel resistance of R_1 and R_g.

When R_1 is omitted the variation of R_L is from a very high value to the minimum grid resistance of the class B tube. If a loading resistance R_1 is used, the maximum resistance of R_L is then R_1. It is obvious

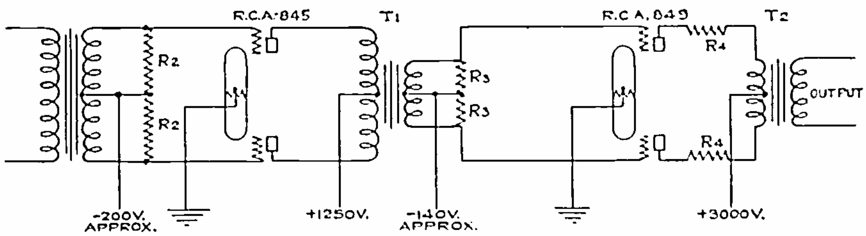

Fig. 3—Circuit for class B audio amplifier and driver.

from (1) that the regulation of the driver system improves as the plate resistance r_p is decreased so that tubes with minimum plate resistance should be used. It is not so obvious from (2) that R_1 should be omitted

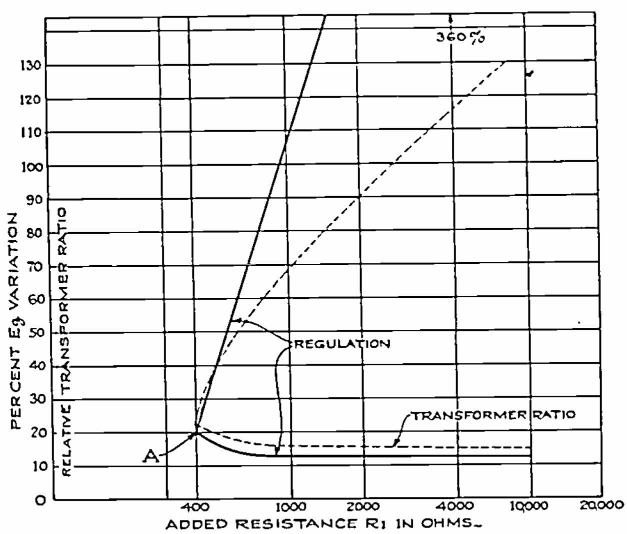

Fig. 4—Relative transformer ratio and voltage regulation to grids of RCA 49 class B audio amplifier.

for best regulation of the voltage $E_{\max}$ when R_g varies from a very high value to a comparatively low value for any practical driver system, therefore the essential circuit constants are assumed for a system such as shown in Fig. 3 in order that relative values may be substituted in (2) to obtain data for the curves in Fig. 4.

It will be noted that there are two solutions for N in (2) that will satisfy the value assigned for E_{max}. At point A on the regulation curves, the equation has only one solution and for lower shunting resistances the solution of the equation is imaginary for the output voltage assumed for E_{max}. Therefore, the output voltage assumed cannot be obtained for lower values of shunt resistance. It is obvious that if the higher transformer ratio solution is used the regulation is improved very much by making the shunt resistance as low as possible for a given voltage output. On the other hand, if the lower transformer ratio solution is used the regulation is improved by omitting the shunting resistance, provided the transformer ratio is reduced accordingly.

The above curves and solutions are obtained on the basis of ideal operation of the driver tubes as a class A push-pull amplifier and that the driver transformer is an ideal impedance coupling device. A study of the plate characteristics of the 845 tube indicates that the tubes operate more nearly as ideal class A amplifiers when the load resistance is very high. If the load resistance is comparatively low, the plate resistance of the tube increases at the extreme negative grid swing which in turn raises the resistance in series with the primary of transformer T_1. Therefore, in practice the regulation of the driver system becomes worse than is indicated by the curves in Fig. 4 when shunting resistance is added. The variations of plate resistance upward at extreme negative grid swings is perhaps the greatest disadvantage in using shunting resistors across the secondary of the driver transformer to improve regulation. This is the result of grid swings causing very low values of plate current for an appreciable part of a cycle. Other tubes, such as the 2A3 at maximum plate voltage may require loading to obtain class A operation of the tubes.

It is also possible to operate low plate resistance driver tubes as class B amplifiers without loading. In this case, for grid swings equal to the class A condition, it is obvious from Fig. 2 and (1) that the effective resistance R in series with the secondary of the driver transformer is double the value for class A operation of the tubes. By driving the tubes as class B amplifiers into the positive region, some improvement is obtained but is not equal to class A operation of low plate resistance tubes unless a greater step-down driver transformer can be used because of greater plate voltage swing.

The effective impedance in series with the grids of the class B amplifier tubes is the combined resistance transferred to the secondary of the driver transformer from the plate circuit of the driver tubes plus any loss and effective leakage reactance. The transformer ratio is determined by methods discussed above but the losses in the transformer and effective leakage reactance in series with each half of the secondary winding is a function of the transformer design. The losses in the transformer can be kept relatively low but the leakage reactance of a driver transformer is not as easily kept to a low value.

A convenient way in which to obtain the approximate effective impedance in series with each side of the driver transformer secondary with the normal tube plate resistance connected to the primary terminals is indicated by the circuit in Fig. 5. The curves, also shown in Fig.

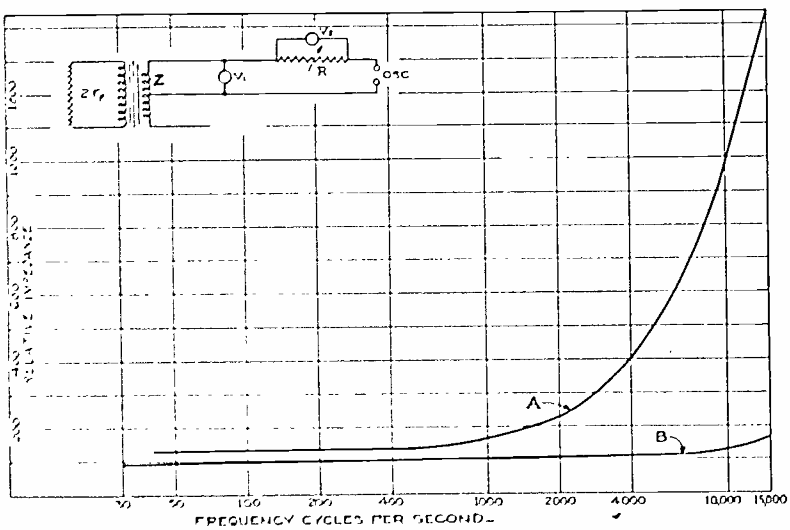

Fig. 5—Equivalent impedance in series with each grid of a class B audio amplifier showing effect of leakage reactance for two transformer designs.

5, are the results of tests on two driver transformers designed to drive 849 tubes as class B audio amplifiers. It will be seen that the effective impedance in series with each side of the secondary of the transformer represented by curve *A* increases quite rapidly as the frequency increases because of the leakage reactance to each secondary winding. The curve of impedance in series with the grids for the transformer represented by curve *B* is essentially flat over the audio-frequency band.

Referring to the curves for the grid current in Fig. 1, it will be noted that the grid current for the 849's will be composed of higher order harmonics so that these currents correspond to frequencies in the upper audio-frequency range for comparatively low fundamental frequencies. If such grid current must be supplied through a relatively large effective impedance as the result of high transformer leakage, it is obvious that a grid voltage similar to the grid current wave will be superimposed on the signal, resulting in appreciable distortion. Another objection to the use of a transformer with high leakage reactance is that its impedance at high frequencies may be greater than the negative resistance of the grids of the class B tubes over certain portions of a cycle, in which case a condition for dynatron oscillation is present. The use of a transformer in which the leakage is very low, reduces the distortion to a negligible value and prevents the generation of high parasitic voltages across the output transformer.

Output Circuit Requirements

It is obvious that for low distortion from class B audio amplifiers the bias must be approximately correct so that the resultant output from the two tubes in push-pull at points near the zero signal axis will have the same slope as the output at higher values of signal. The plate current curves in Fig. 1 indicate that the proper bias is about-135 volts for approximately twenty-five milliamperes no-signal plate current. However, due to the fact that the cutoff characteristic of the 849 is somewhat extended under operating conditions because of the increased instantaneous plate voltage, as a result of its push-pull relation with the other tube, the no-signal plate current for minimum distortion is about forty to sixty milliamperes with a corresponding reduction in bias. The extended plate-current cutoff effect is present because of the relatively low plate resistance of the 849 as compared to the plate load resistance. Therefore, it will be seen that the resultant plate current as stated in the definition of class B amplifiers must be linear from the bias point, and that the no signal plate current is not necessarily essentially zero but may be appreciably higher than usually given.

Another factor affecting the distortion in a class B audio amplifier is the leakage reactance in the output transformer shown as T_2 in Fig. 3. It will be noted that the total power output is supplied from one tube through one half of the output transformer primary during most of alternate half cycles. Since the power is transferred through one half of the primary winding, due consideration must be given to the leakage reactance of this transformer. Undue leakage in the output transformer not only reduces the response at high audio frequencies but also increases the percentage distortion of the output signal.

Experimental Results

The above discussion on the class B audio amplifier indicated the procedure in recent developments of this class of amplifier. The characteristics of the 849 tube were used to illustrate the various items of importance to develop a low distortion class B audio system. Although an amplifier of this type may be used for any purpose for which approximately 1000 watts of audio power is desired, the most common use for the amplifier is for high level modulation of a 1000-watt radiophone transmitter. Distortion measurements were made on a complete high level class B modulated 1000-watt transmitter using 849 class

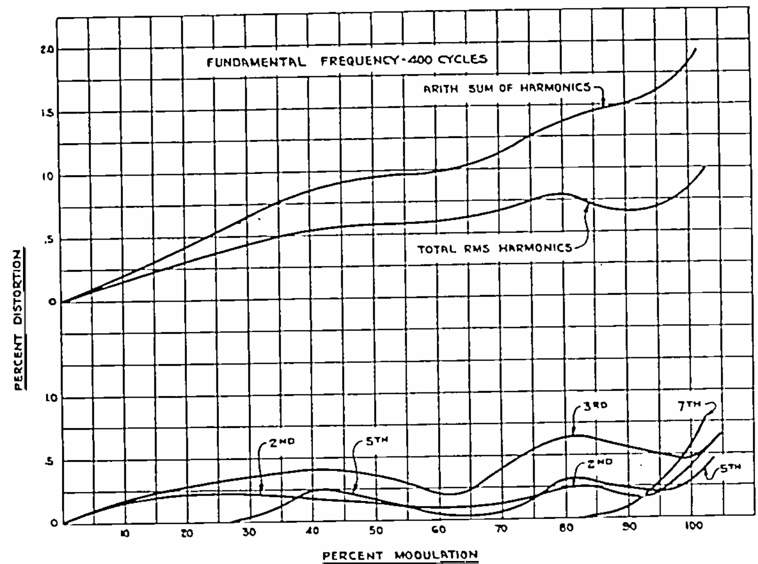

Fig. 6—Performance of a 1000-watt high level class B modulated transmitter.

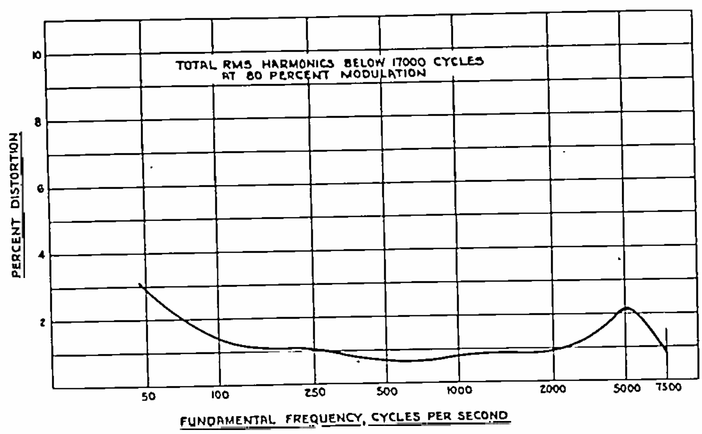

Fig. 7—Distortion at various frequencies for transmitter as adjusted for performance in Fig. 6.

B modulators. The input signal level was approximately 1.8 volts across 500 ohms and a rectified portion of the radio-frequency output was applied to the analyzer.

The data obtained from the above distortion measurements are plotted in Figs. 6 and 7. It is believed that a distortion approximating the values shown in Figs. 6 and 7 is sufficiently low to meet any present high fidelity requirements. This is especially true when no visible disturbance is indicated on the carrier envelope by a cathode-ray oscillograph.

Zero-Bias Class B Amplifiers

The plate-current grid-voltage characteristic for a vacuum tube follows approximately a three-halves power curve. For tubes having a low plate resistance as compared to the load resistance in series with the plate, the resultant curve becomes nearly linear so that little distortion results from the use of such tubes as class A amplifiers, or as class B amplifiers if the proper bias is used. On the other hand if the tube is to be used at zero bias the plate resistance of the tube must be high compared to the load resistance resulting in little change in the normal three-halves power curve for the plate-current grid-voltage dynamic transfer curve or the curve in which the plate load resistance is included in the plate circuit. Therefore, it is improbable that a zero-bias tube can be made to have low distortion output with essentially zero no-signal plate current. A possible solution is to allow appreciable zero-signal plate current so that the resultant plate current as related to the grid voltage will be linear from the operating point for one-half cycle as indicated in the definition of a class B audio amplifier. The maximum zero-signal plate current is limited by the plate dissipation of the tube so that if maximum output is to be had, the average plate current will rise appreciably at peak outputs.

By using a relatively high zero-signal plate current for zero-bias tubes the normal three-halves power curve of plate current vs. grid voltage can be altered appreciably by the use of a special grid structure so that the resultant plate current slope of two tubes in push-pull at points near the zero-signal axis will be equal to the slope of one tube during peak positive grid swings at which point the plate current will be zero for the other tube.

The new 838 tube has approximately the characteristics desired for a zero-bias tube as outlined above. Because of the difficulty and complications in design and manufacture, the 838 was so designed that it could be used for zero-bias operation as a class B audio amplifier for applications in which some distortion could be tolerated, and for applications requiring very low distortion a small bias could be used.

The dynamic transfer curve for an individual 838 is shown in Fig. 8 and represents approximately the characteristic of this type of tube. Two resultant plate currents are drawn which indicate the slopes of the resultant plate current for zero-bias operation and for −15-volt bias operation. The circuit with which the curve data were taken is the same as in Fig. 1.

The resultant curve A, drawn for a bias of -15 volts, is much more linear than the resultant curve B. It should be noted that the grid current for this tube represents approximately a constant resistance as compared to the erratic grid resistance of the 849 tube shown in Fig. 1. As a result of the more or less constant value of grid resistance

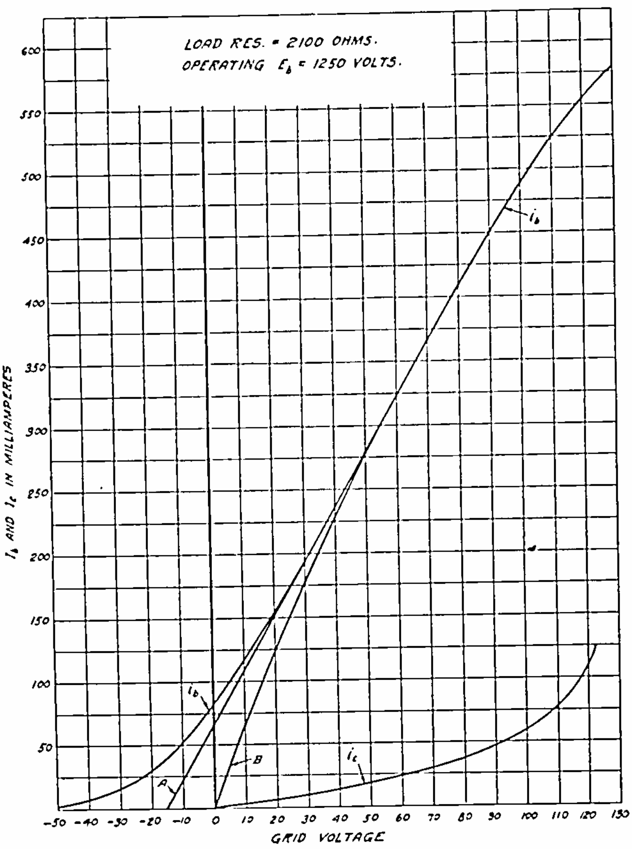

Fig. 8—Dynamic transfer characteristics of an 838 tube.

the 838 is much easier to drive than the 849; however, the output of a pair of 838's is approximately 250 watts as compared to 1000 watts output for two 849's.

Distortion measurements were made for operation of the 838 at zero bias and are shown in Fig. 9. Distortion curves are shown for the -15-volt bias operation in Fig. 10. These curves agree quite well with distortion that may be predicted from the dynamic transfer curves in Fig. 8. The driver systems for the above curves were practical class A driver systems. The impedance of the driver systems for Fig. 10 was somewhat higher than the system driving the grids for the curves in Fig. 9 which accounts to some extent for the somewhat higher output power per pair of 838's in Fig. 9.

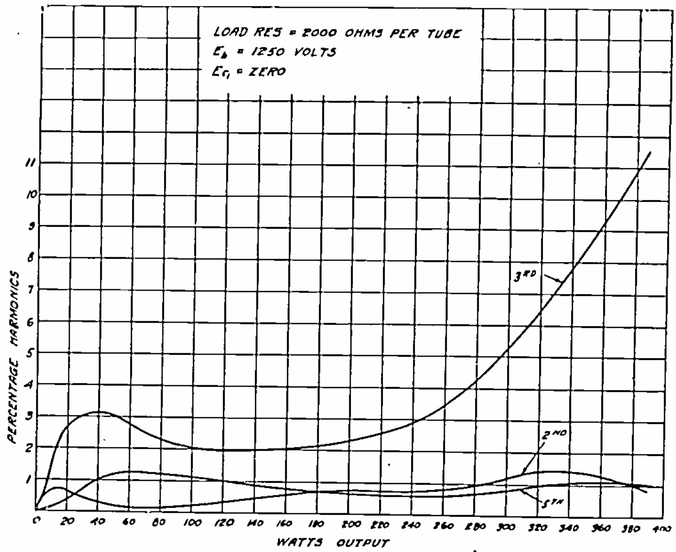

Fig. 9—Performance of two 838 tubes in a class B audio amplifier with zero bias.

The 838 tube may be used for class B radio and class C amplifiers. When the tube is used for class C radio service the bias can be obtained

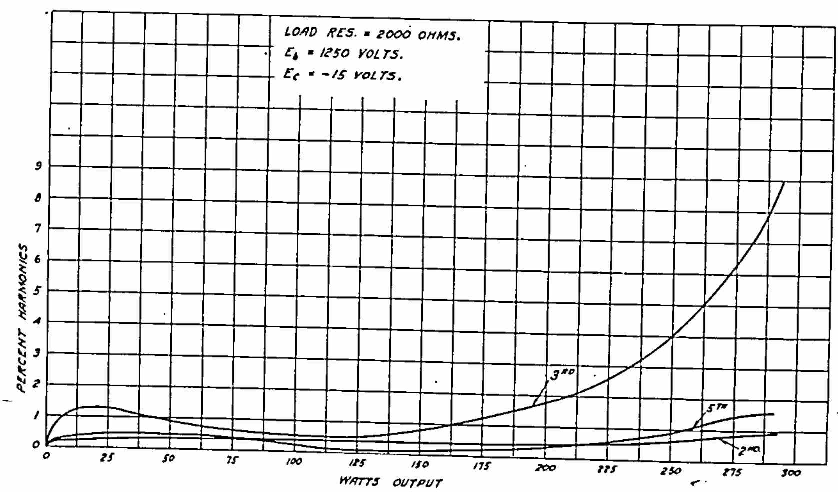

Fig. 10—Performance of two 838 tubes in a class B audio amplifier with fifteen volts bias.

from a grid leak, thus eliminating the bias problems in case excitation is lost, because the plate dissipation at zero bias is normal operation for the tube. Therefore, this tube should be welcomed in radio-frequency systems because the problem of bias source is simplified and the power required to drive the tube is about equal to the power required to drive other tubes of similar size.

High Bias Class B Audio Amplifiers

The high bias type of class B audio amplifier is quite interesting and useful in applications where the power requirements permit the use of an available tube. For such amplifiers, the driver problem is not nearly as severe because the grids need not be driven into the positive region. In general, the plate resistance of the high bias, low amplification factor tubes increases appreciably during that part of the signal cycle at which one tube approaches plate current cutoff. As a result of the increasing instantaneous plate resistance with the usual decrease in grid control at high negative grid voltages, the tailing-off effect of the dynamic transfer curve of the low-mu, low plate resistance tubes is extended beyond the value normally expected. This is especially true for the small 2A3 and the water-cooled 848 types of tubes. The extended dynamic plate current for two of these tubes operating in push-pull results in a comparatively high value of no-signal plate current in order that the resultant plate currents will meet the condition for class B audio amplifier operation as defined above.

In the case of the small 2A3 tube, the no-signal plate current for 300 volts on the plate is very nearly as great for class B operation of the tubes as the normal plate current for approximately class A operation at 250 volts on the plate. At higher plate voltages, the tube is not very satisfactory for class B audio operation, because the plate dissipation limits the high no-signal plate current needed for low distortion. The maximum class B audio output of two 2A3's is about fifteen watts.

The 845 type of tube does not have such an extreme tailing-off characteristic as the other two tubes mentioned, but this characteristic is present to some extent. The maximum output of two 845's as class B amplifiers is approximately 100 watts.

The 848 type of tube has a plate current tailing-off effect similar to the 2A3 and in some respects these two tubes have plate current characteristics strived for in the zero-bias 838 type of tube except that the 838 tube operates at essentially zero bias. The maximum output of two 848 tubes as class B audio amplifiers at approximately 13,000 volts on the plates is about ten killowatts. However, the plate voltage is limited to a value appreciably below the above value because of the relatively high no-signal plate current unless a high velocity of water is used around the plate for cooling purposes. The high velocity of water is necessary because the tube has a comparatively coarse meshed grid for a low amplification factor, resulting in local hissing or boiling of the water at hot spots. The hot-spotting of the plate is the result of electrons focusing at points on the plates for high values of grid bias and plate voltage.

Distortion measurements were made on two 848's as class B audio amplifiers and the results are plotted in Fig. 11. It will be noted that the distortion is very low for the comparatively high no-signal current operation, which is also true for class B operation of the other two types of low amplification factor tubes mentioned above. However

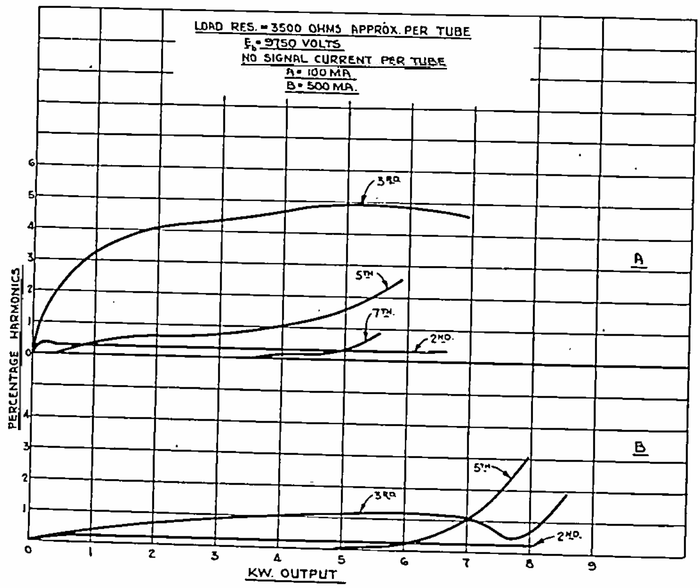

Fig. 11—Performance of two UV 848 tubes in a class B audio amplifier.

the curves shown in Fig. 11 for low no-signal current operation indicated appreciable distortion because of the long "tail-off" characteristic of the 848 tube. The low distortion from the high bias type of class B audio amplifier when operated at the proper no-signal plate current is made possible because of the remote cutoff of the plate current, the relatively low plate resistance as compared to load resistance, and the relatively simple problem of driving the tubes when the grids are not driven into the positive grid region. However, it must be remembered that if maximum output is needed from a particular size tube, the grid must be driven into the positive region necessitating good regulation of the driver system. Because of the high grid swings necessary for high bias class B amplifiers, it is not practicable to drive the high bias tubes into the positive grid region and still have low distortion. This point is evident from (1) and (2) given above. Approximately 250 watts can be obtained from the zero-bias 838 tube, which is physically about the same as the 845, and approximately twenty-five kilowatts can be obtained from the 863 tube, which has the same structure as the 848 except for grid mesh. The bias for the 863 is zero to about -100 volts, depending upon the plate voltage while the bias for the 848 is -1200 to -1600 volts, which also depends on the plate voltage.

Class B Radio Amplifier

As indicated at the beginning of this paper, there are certain power ranges of transmitters in which there is no economical tube comple-

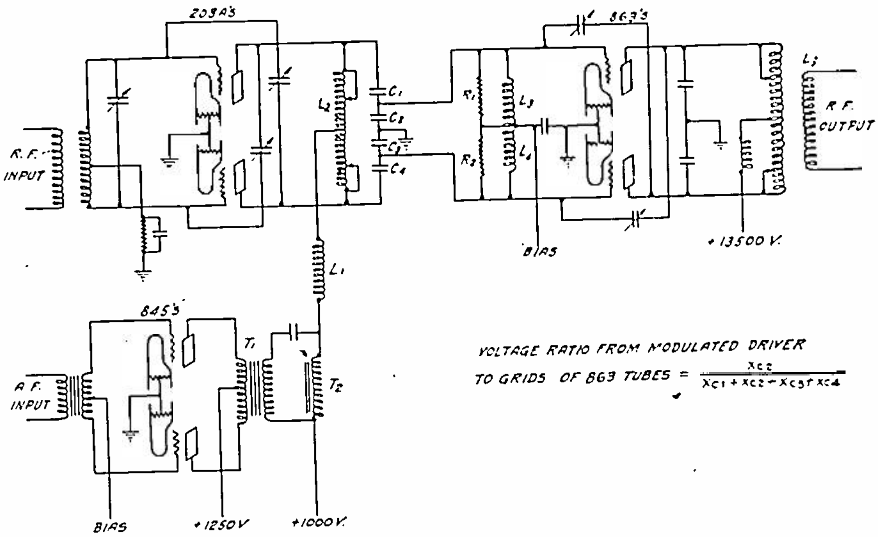

Fig. 12—Essential circuit for driver and output systems for a recently developed five-kilowatt transmitter.

ment available for high level modulated systems. In such cases, fewer tubes are needed for a class B radio output system and, of course, the high power modulation transformer and reactor are not needed. The possible saving in number of total tubes used, plus the saving of the cost of modulation transformer and reactor, may more than offset the disadvantages of the class B radio output system. The class B radio amplifier need not be unnecessarily critical if the circuit constants are correctly chosen for a typical circuit as shown in Fig. 12.

According to the definition of a class B amplifier, two tubes in push-pull will function in the same manner whether the frequency is audio or radio. At radio frequency, a single tube may be used because the plate circuit can be tuned to the desired frequency and the tube can store energy in the tank circuit during one half cycle to be dissipated over the complete cycle. Referring to Fig. 12, 863 tubes are shown in the output system for a transmitter developed for five killowatts output. It will be noted that the tubes are connected in a usual push-pull fashion except for the fact that the grids are coupled directly to the condenser part of the modulated class C amplifier tank circuit. The general discussion of the driver and output systems for the class B audio amplifier above applies equally well to the class B radio amplifier. Therefore, it is very desirable to reduce the leakage of the driver system as much as possible. A low impedance system is obtained by using as large a ratio of C_2 and C_3 to C_1 and C_4 as possible without appreciable loading by R_1 and R_2. The resistors R_1 and R_2 are used to suppress parasitic oscillations and have a relatively high value of resistance. The radio-frequency choke coils L_2 and L_3 are used for grid returns to a bias supply.

Since the grids of the class B radio tubes are excited from low impedance condensers and low power is supplied to the tank circuit, $L_2C_1C_2C_3C_4$, by the 203A class C amplifier, no regulation problem is present for any constant radio-frequency voltage applied to the 863's. However, during modulation of the class C amplifier, the power taken by the class B amplifier grids varies over the audio cycle, so that the problem of low impedance audio driver system becomes very important.

The direct-current value of grid current to the 863's at normal carrier conditions is usually quite low, because of negative grid current over a portion of the radio-frequency cycle. The approximate relative direct grid current curve for the 863's operating under the conditions as shown in Fig. 12 is shown in Fig. 13 for various instantaneous values of upward and downward modulation. The change of slope of this curve at various instances over an audio cycle indicates the variable resistance the driver system must work into. The curve shown is only approximate because the actual curve depends upon individual tubes and the load under which the tubes operate. The curve can be used for the purpose of discussion and in any case, it is obvious that a high impedance driver system will result in distortion depending upon the particular shape of the grid current curve such as is shown in Fig. 13.

The impedance of the driver system over an audio cycle is the resulting impedance of the modulator through the class C amplifier, which in turn is coupled to the class B radio amplifier. The impedance of the modulated class C amplifier is comparatively low corresponding to 1000 ohms or less for each of the class C tubes. That is, within reasonable limits, a change of load resistance on the class C amplifier causes the amplifier to draw more plate current with little change in output voltage. Therefore, since the regulation of the grid voltage applied to the 863's is a function of the regulation of the audio voltage applied to the plate, the regulation of the driver system is largely a function of the regulation of the output of the modulator. Because of this fact, low impedance 845 tubes were used for modulators and as explained above, according to (1) and (2), the lowest impedance driver system can be obtained when the loading resistance is a maximum, thus permitting a maximum step-down ratio of the modulation trans-

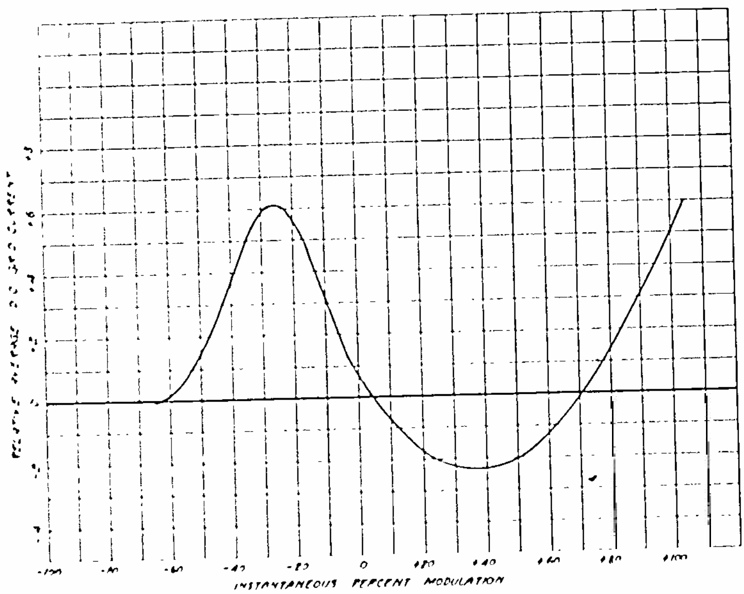

Fig. 13—Approximate grid current characteristic of the UV 863 tube as a 2.5-kilowatt class B radio amplifier.

former T_1 for a maximum operating voltage applied to the plate of the class C amplifier. A maximum direct voltage on the class C amplifier permits a minimum output impedance to the grids of the 863's because a maximum step-down ratio can be used to apply the necessary voltage to the grids of the class B radio amplifier. The 845 tubes were operated as class B audio amplifiers and, of course, according to (1) and (2) four modulator tubes will result in a driver system with lower impedance than two tubes. If sufficient driver tubes are used to permit class A operation, a still lower impedance driver system could be had. However, two modulator tubes operating as a class B audio amplifier quite successfully drove the class B radio amplifier through the class C amplifier with a power input of only 80 to 100 watts to the class C tubes. Distortion curves are given for such operation in Fig. 14 for five kilowatts output from the 863 output system.

As a matter of interest, four tubes were used in the modulated class C amplifier and the resistors R_1 and R_2 were decreased to such a value that two 203A's as modulators were driven to approximately full power output. The output of the class C amplifier was about 250 watts. The measured distortion in the output of the 863's under these conditions was appreciably higher than the distortion shown in Fig. 14. This result was due principally to the fact that the decreased value of the loading resistance R_1 and R_2 was not as effective in reducing driver impedance as the use of the low impedance modulator tubes.

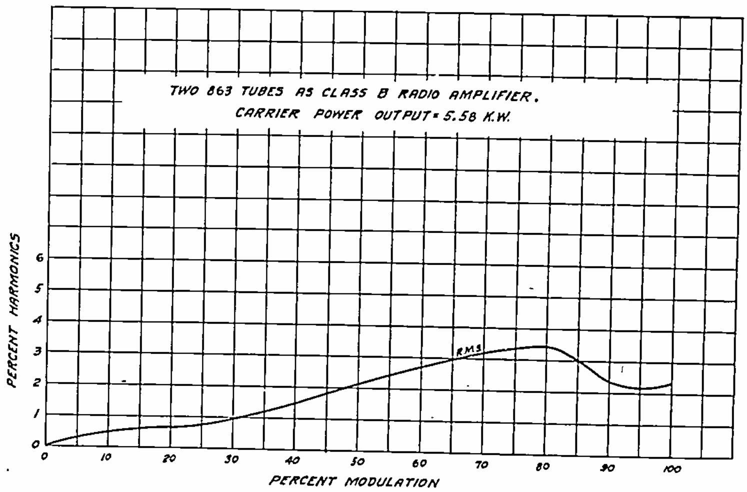

Fig. 14—Distortion introduced by a recently developed five-kilowatt broadcast transmitter.

The driver system for a fifty-kilowatt class B radio output system used the same low impedance radio-frequency coupling means as used for the five-kilowatt system shown in Fig. 12. Because of greater driver power requirements for the 898 water-cooled tubes for fifty kilowatts output, it was found most convenient to replace the high plate resistance 863 tubes in Fig. 12 with low impedance 848 tubes and so biased that the tubes operated as nearly as a class A radio amplifier as possible. The output impedance of the 848 driver stage as a class A radio amplifier can be determined approximately by (1) and (2) and it was found that the highest value of loading resistance across the grids of the 898, connected in the same manner as R_1 and R_2 in Fig. 12, resulted in the minimum distortion from the 898 class B radio output system. As shown in Fig. 4, if the loading resistance is increased, a decrease in driver impedance is obtained only if the step-down ratio of the coupling condensers to the grids of the 898's is increased. It is interesting to note that under normal operation with a fifty-kilowatt carrier, the direct grid current to the 898 tubes was negative, and that the power dissipated by the parasitic resistors across the grids of the 898's was approximately 750 watts. The general shape of the direct grid current at instantaneous values of upward and downward modulation for the 898 tubes was the same as shown in Fig. 13, except that the negative values over certain grid regions were much greater than shown in this figure.

The grids of the 848 tubes were not driven into the positive grid region so that the output impedance of the plate modulated system

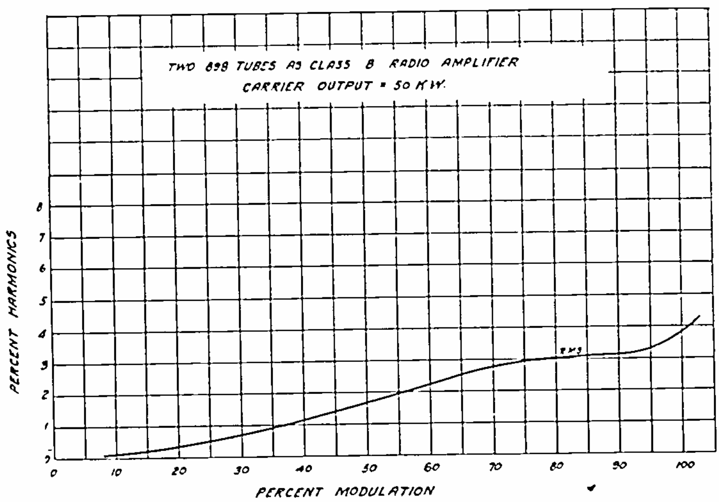

Fig. 15—Distortion introduced by a recently developed fifty-kilowatt broadcast transmitter.

as shown in Fig. 12 did not have to be as low as was necessary when 863's were being driven. The success with which the 898 tubes were driven with the low impedance 848 driver system is indicated by the curve shown in Fig. 15, data for which was taken at normal fifty-kilowatt carrier output and was the distortion introduced by the entire transmitter from an audio level of approximately 1.5 volts signal across 500 ohms.

Conclusions

The theoretical considerations for the driver systems for class B audio- or radio-frequency amplifiers and the experimental results which substantiate the accuracy of the theory indicate that a low impedance driver system is very important for low distortion class B amplifier systems.

Another important consideration for class B amplifiers, especially for audio systems, is the plate current characteristic of the tube. This characteristic should be such that the disturbance in the output wave near the zero axis of plate current may be as low as possible. Practically this result can best be obtained by purposely causing the plate current to tail-off appreciably, which results in appreciable no-signal plate current. The remote plate current cutoff is particularly helpful in reducing distortion when high plate resistance tubes are used as class B audio amplifiers.

The class B amplifier plate load is somewhat critical for maximum power output and minimum distortion. However, this phase of the subject has been discussed in other papers and space does not permit a review of the proper plate load for the class B audio and radio amplifiers.

If the above factors affecting class B amplifier design are satisfactorily solved, class B audio- and radio-frequency amplifiers can be built for very low distortion output systems with existing tube designs.

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GENERAL THEORY AND APPLICATION OF DYNAMIC COUPLING IN POWER TUBE DESIGN*

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Summary—*This paper presents a simplified method of driving a power tube without the need of coupling devices and grid-biasing means. The power section is one whose useful plate-current versus grid-voltage characteristic is realized only with positive values of grid voltage. Its low input impedance is in series with the cathode-ground circuit of the driver tube. This impedance, also, automatically provides a negative bias for the grid of the driver, thereby eliminating external biasing. Since the electronic coupling of the two tubes varies with signal excursions, this method of amplification is termed "dynamic coupling."*

Practical considerations show immediately that the driver must operate into an impedance which is considerably lower than its own plate impedance. It is shown that the distortion which is produced when working with such ratios is minimized partly by making the driver circuit degenerative in order to nullify the varying effect of the driver's mu. This is treated first with a pure resistive load. When the grid impedance is the load, a further reduction in distortion is shown, for then the ratio of plate-to-load impedance remains more nearly constant throughout a signal excursion.

The remaining part of the paper deals with a commercial application of these principles. Certain design considerations for a tube embodying both driver and power sections are discussed. A current surge phenomenon caused by secondary emission is discussed and a practical means for its elimination is given. The delayed point at which the driver's grid begins drawing current is shown to be particularly advantageous. The tube's electrical and economical advantages are compared with those of contemporary audio systems.

Introduction

In the endeavor to obtain economically relatively high audio output for radio receivers, methods of driving power tubes have become popular. By driving is meant the supplying of signal power to the grid in order to permit signal excursions extending into the region where the grid is positive. The driver tube is usually coupled through a step-down transformer to the power tubes. The sudden changes in the grid impedance of the power tubes, caused by intermittent flow of grid current, reflect such load conditions on the transformer as to make it nearly impossible to supply a voltage on the grids, the wave form of which is an exact replica of that applied to the driver grid. This is especially true if the ratio of the transformer and size of the driver tube are consistent with the economical reasons for using

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such a system. The harmonic distortion resulting from the discontinuities in the impressed wave include the third and higher harmonics. The audible effect of these higher order harmonics, even when their percentages are small, in so-called high fidelity receivers is most unpleasant; the notes are not clean cut and there is a background of buzzing noise. Systems operating as class AB, in which the grid current does not flow until a fairly high output level is reached, are not free from this trouble. Since harmonic measurements can be taken readily, it would be most convenient to have a "yardstick" in terms of percentages of individual harmonics for practical classification of amplifiers. To this end, investigations were conducted to correlate the permissible distortion which could be tolerated without being objectionable to the ear, with the quantitative values of the individual harmonics. Details of the technique used are too lengthy to be related here, but suffice it to say that wave outputs originating from the simultaneous application of two pure notes of different frequencies not harmonically related were used extensively. This method is very effective because the resulting spurious frequencies (which appear as sum and difference terms) caused by the higher order harmonics are so readily distinguishable. As anticipated, the distortion was discernible even when the harmonic percentages were incredibly small, providing the order of the harmonic was high and the predominating modulation products occurred at frequencies within the sensitive region of the audible spectrum of the ear. It was then decided to listen to a system having higher order harmonics the generation of which was not caused by intermittent operation into grid current. A system was devised in which the power tubes were operating in their grid-current region throughout the signal excursions. The higher order harmonics were caused in this case by slight wavers in the grid-current characteristic of the power tubes. Operating conditions were selected so that the harmonics were present at nearly all power levels. One essential difference between the systems, then, is the manner in which the higher order harmonics make their appearance; in one case they appear abruptly at some specific power level; while in the other, they are somewhat of a residual quantity, being present at nearly all power levels. It was at once evident that some factor had been omitted in determining the "yardstick" because the harmonic percentages could be several times larger with the new system than that which could be tolerated when operating intermittently into grid current. Just what is the missing factor is not yet known. It might be speculated that since we are dealing with human reactions, this work cannot be correlated simply on an electrophysical basis. If the related human equation could be stated, the variance in psychological reactions might explain the audible difference. On the other hand, the measuring tools or the technique used might be faulty in disclosing the true picture. The investigations will continue and it is hoped that an adequate theory can be found which will explain the phenomenon. However, the practical advantages of the new system are of sufficient interest to warrant the presentation of the design theory and the commercial applications at this time.

This paper is now devoted to describing a method of dynamic coupling of the driver to the power tube, thereby eliminating the usual coupling transformer. The working range of the power section's plate-current versus grid-voltage characteristic is attained only with positive values of grid voltage. This eliminates abrupt changes in the input impedance of the section. This impedance also establishes an automatic negative bias for the driver grid, making external biasing unnecessary. Unlike other driven systems, push-pull or push-push operation of the power tubes is not necessary. A single power tube and its associated driver may operate in combination as a true class A system.

I—Fundamental Theory

Complete understanding of the operation of the system can best be obtained by a study of the characteristics of the two individual tubes which comprise the system. First, we shall consider the power tube. Since grid current is to be present throughout the operating cycle, the grid potential is limited to the positive region of its plate-current versus grid-voltage characteristic. At zero potential the plate current should be a minimum, as remaining current will be unusable, and therefore, wasteful. If the amplification factor of a triode is sufficiently increased, this desired characteristic can be obtained. The tube would have plate characteristics similar to those in Fig. 1. It should be noted that the curves are given only for positive values of grid voltage. A few grid-current curves are also shown in Fig. 1. Since we desire to operate the triode as class A, the operating axis 0 and the corresponding optimum plate load, represented by line MN, are found by the conventional method used in determining the same factors for negatively biased tubes. The operating point 0 will be at some mean value of plate current and grid voltage such that if the grid voltage is increased and decreased in equal amounts about point 0, the rise in plate current will approximately equal the corresponding fall in current. From the intersections of the load line with the characteristics the I_b vs. E_c curve may be plotted, as shown in Fig. 2. The I_c vs. E_c curve is also shown here and it can be determined by constructing an auxiliary load line on the grid-current curves so as to reflect the load in the plate circuit. By inspection of Fig. 2, it is appar-

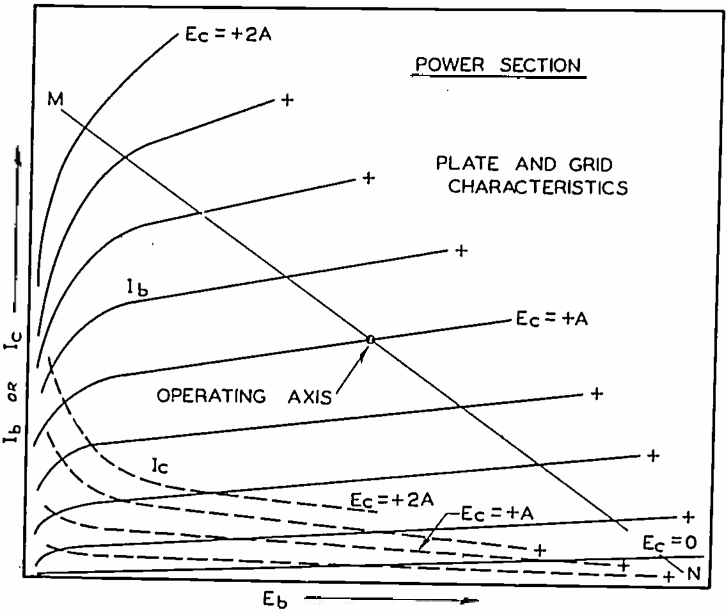


Fig. 1—Typical plate and grid characteristics of the power section.

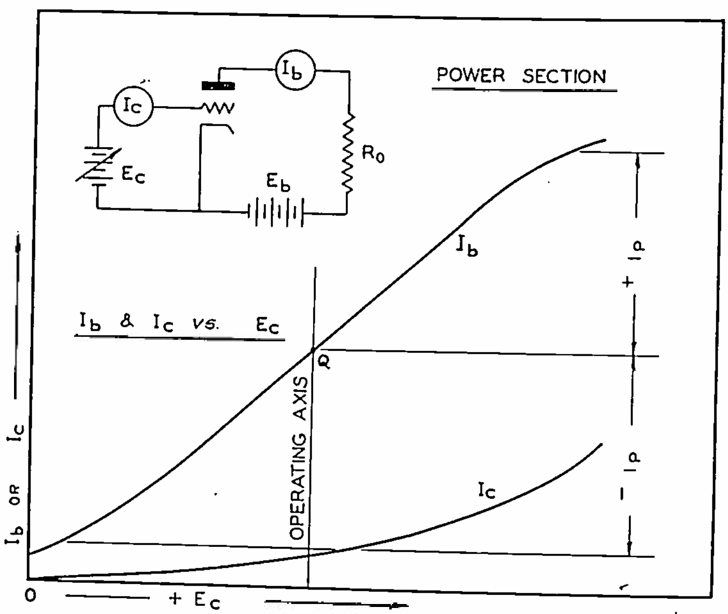


Fig. 2—Typical power section; I_b and I_c versus E_c curves.

ent that over a considerable portion of the I_b curve the current responds linearly with grid voltage. The departure from a straight line is at the ends, where the curvatures approach equality in variation and are opposite in direction. This results in the generation of odd harmonics with the even harmonics tending to cancel. Calculations of the distortion could readily be made, but, since it would add little to this treatment, it is omitted. Turning now to the I_c curve, it will be noted that its shape is quite different from that of the I_b curve. Although a certain portion is fairly straight, the slope increases continuously in one direction. This means that, looking into the tube, the input impedance varies with signal, being relatively high at low grid voltage, and conversely, low at high grid voltage. In practice, an eight-to-one ratio at the extremities of the curve is not unusual. In order that no distortion shall be introduced because of this variable input load, special means must be provided whereby the changing impedance will have little or no effect upon the impressed wave. A coupling transformer and driver tube could hardly satisfy this requirement without, at least, introducing considerable amplitude distortion, even though the proportions of the apparatus were ample in size. It should be mentioned here that the ratio of average plate-to-grid impedance, obtained from practical design data, is roughly twelve to one, and since it is desirable to keep down the plate impedance, the grid impedance is necessarily relatively low. The solution of the problem is direct dynamic coupling to the driver such that the varying input impedance causes no appreciable distortion of the signal supplied by the driver. In fact, it will be shown that this changing load causes less distortion than that produced by a constant load of equivalent magnitude.

Our first consideration of the driver, however, is not the changing load but rather the ratio of the driver's plate impedance to this load. A driver whose plate impedance is sufficiently low to match the average grid load will not be considered, since such a design is impracticable because it would necessitate both excessive physical proportions and current consumption. Consequently, a step-down ratio must be used. A triode operating into a step-down load under ordinary circuit operation is quite hopeless from a distortion consideration, but if the circuit is made degenerative by placing the load between cathode and ground rather than between plate and B+, and the signal applied between grid and ground, the distortion is greatly reduced. Let this load be represented by a resistor; then, during no input signals, the direct-current drop across it produces a negative bias similar to usual self-biasing. When an input signal is applied, the change in voltage appearing across the resistor is in phase with the signal with respect to ground; hence, this developed voltage appears in the input circuit as a subtractive quantity. In other words, the voltage appearing directly across the grid to cathode at any instant is always equal to the arithmetical sum of the input voltage and the voltage across the resistor. The reduction in distortion due to this degenerative action can be readily appreciated by comparing the curves relating the voltage appearing across the resistor with the input voltage. The curve $A-B$ in Fig. 3 is for the degenerative circuit and it is much more linear than the curve $C-D$ which is for the nondegenerative operation. The resistance load used was approximately one fifth the plate impedance of the triode. For simplicity, the nondegenerative operation is for the case in which the input signal is applied directly across grid and cathode.

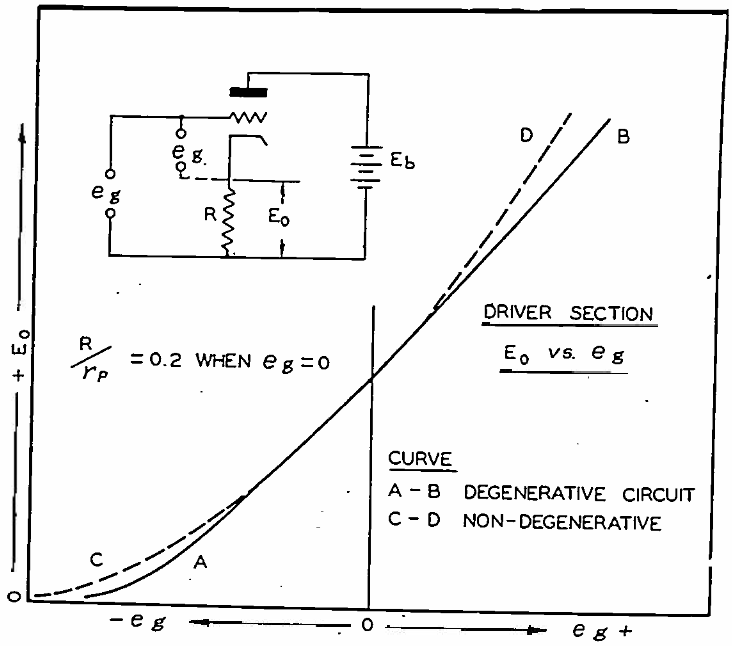


Fig. 3—Driver section; E_0 versus e_g curves with resistive load.

Of course, the result is exactly the same as if the load had been between plate and B+. Also, the bias voltage necessary to obtain the same plate current when e_g equals zero as was present for curve $A-B$, has been purposely omitted from the circuit drawing for further simplification. Incidentally, since the nondegenerative circuit is the more sensitive, and, in order to compare the curvature more readily, the slopes of the two curves were made nearly to coincide by enlarging the input voltage scale for curve $C-D$. To express mathematically the degenerative performance in complete form is tedious; however, a few simplified expressions will aid in appreciating the action. The current in the nondegenerative circuit (neglecting second and higher order terms) is

$$i_p = \frac{\mu e_g}{r_p + R} \tag{1}$$

whence e_0, the voltage across R, is

$$e_0 = i_p R = \frac{\mu e_g R}{r_p + R}; \tag{2}$$

whereas, in the degenerative circuit, the voltage appearing across R is a subtractive quantity, looking into the input circuit. The plate current becomes

$$i_p = \frac{\mu(e_g - i_p R)}{r_p + R} \tag{3}$$

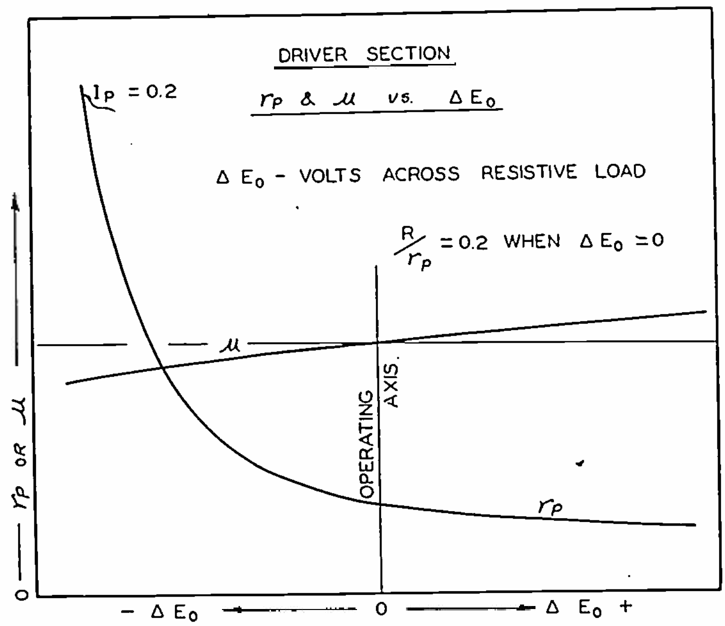


Fig. 4—Calculated variation in the plate impedance and amplification factor of the driver plotted against the change in voltage across a resistive load.

and the voltage e_0 becomes

$$e_0 = i_p R = \frac{\mu e_g R}{r_p + R(1 + \mu)}. \tag{4}$$

By comparing (2) and (4), we find the only difference between the two is the manner in which μ enters. Obviously, if μ remained constant throughout the signal excursion, the distortion would be the same regardless of the mode of circuit operation. However, the μ is not a constant; in fact, when R is small compared with r_p, the variations in both μ and r_p are large. This is graphically shown in Fig. 4, where μ and r_p are plotted against the change in voltage across the output load. The ΔE_0 range is equivalent to the E_0 range indicated in Fig. 3. These curves were taken using the same driver as was used for the previous curves. Obviously, the distortion in the degenerative circuit is less because the effect caused by shifting the μ term is greatly diminished due to the $(1+\mu)$ term in the denominator. Furthermore, when μ is made large compared to unity, the distortion produced by changes in μ asymptotically disappears.

We have seen that by introducing degeneration it is possible to operate into a step-down load with considerably reduced distortion. This fact suggests the possibility of using a triode for driving with its load between cathode and ground. Let us substitute the grid-cath-

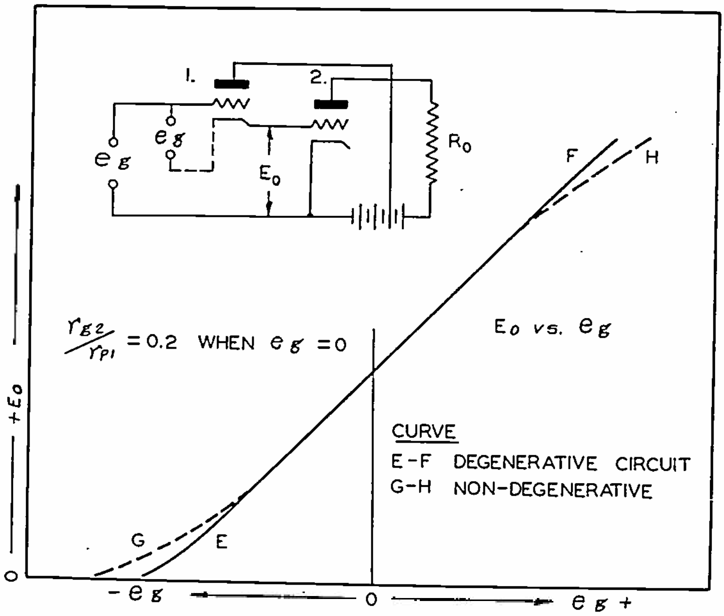


Fig. 5—Driver section; E_0 versus e_g curves with grid impedance load.

ode impedance of the output tube previously described for the resistor, as shown in Fig. 5. Then the voltage drop across this impedance becomes the negative bias for the grid of the input section as did the voltage across the resistive load in the previous circuit; and, also, it becomes the positive bias for the output grid. This positive bias establishes the operating axis for the I_b vs. E_c characteristic of the output section. In designing, the optimum positive bias and the corresponding grid current are determined as previously described. With this information, the driver is simply designed so that at a negative bias equal to the selected positive bias the plate current of the driver will equal the determined grid current. For the sake of direct comparison, the resistance R used for Fig. 3 was made equivalent to the value of the power section's grid-cathode impedance, measured at the operating axis. (See Fig. 2.)

The question now arises, how will the voltage appearing across the grid-cathode impedance vary with respect to the input voltage? Curve $E-F$, Fig. 5, shows that this characteristic is very nearly linear. The curve $G-H$ is included to show the relation without degeneration, and the greater curvature is plainly evident. It will be noted that both curves are considerably straighter than those taken with a resistor load, Fig. 3, the improvement being more noticeable in the nondegenerative case.

The reason for this is as follows. The grid impedance is not a simple fixed value as was the resistor. It is actually a quantity varying as a function of the voltage appearing across it. This is obvious by inspection of the I_c curve in Fig. 2. Moreover, since the grid current changes with plate voltage its slope is not independent of the output load. Since an expression of the grid impedance in terms of the tube's parameters is irrelevant in showing why the driver produces less distortion with this variable load than with a constant load, it will be omitted. For simplicity, the voltage developed by the driver should be examined in terms of the ratio of the driver's load to its plate impedance. Let this ratio $R/r_p = K$. Then R may be written as Kr_p and this term substituted for R. For the nondegenerative operation, (2) now becomes

$$e_0 = \frac{\mu e_g K}{1 + K} \tag{5}$$

and for the degenerative circuit, (4) becomes

$$e_0 = \frac{\mu e_g K}{1 + K(1 + \mu)}. \tag{6}$$

It is obvious that if K remains constant throughout the entire cycle no distortion will result in e_0 due to any variations in either the plate or load impedance. As previously shown by Fig. 4, r_p varies considerably; hence the factor K cannot remain constant if a pure resistive load is used. If the load varies in the correct relation with respect to r_p, a more constant ratio is possible, which results in a reduction in distortion. This is precisely what happens with the system in question. The circuit's phase relation is such that both the plate impedance of the driver and the grid impedance of the output section vary in the same manner. Both increase and decrease in phase. Consequently, a nearly constant ratio is maintained throughout the signal excursion. The writer has chosen the term "dynamic coupling" to describe this method of coupling.

Returning now to the plate-current versus grid-voltage characteristic of the output section, it will be remembered that the curve was practically linear over the operating range. From the foregoing discussion of dynamic coupling we have seen that a driver can impress voltage across the output section's grid impedance which will have a linear relation with its input voltage. Thus the composite relation of output plate current to input volts is also linear.

From the foregoing treatment one might conclude that the matching of characteristics is essential to obtain good performance. This is true with respect to the selection of the characteristics for the composite design. However, the normal variations encountered in modern tube production have no more detrimental influence than they do with any ordinary power tube. It should be remembered that the static voltage appearing on the grids of the tubes is similar to the voltage across a self-biasing resistor in that variations in resistance or plate currents tend to be restricted because of the well-known self-biasing action. Thus current variations from perfectly centered individual sections have less effect in the combined system than if the self-biasing action were not present. With regard to the harmonic generation caused by variation in matching to the driver, the curves in Figs. 3 and 5 for the degenerative operation readily demonstrate what a wide variation may be tolerated, for here in one case a pure resistive load was used and in the other a variable grid impedance. Although there is a noticeable difference in curvature, it should be realized that this represents an extreme change in load conditions. In practice, the normal variations in the grid impedance produce differences which are negligible compared with changes shown in Fig. 5.

Since the output plate current flows during the 360 electrical degrees of the cycle with equal negative and positive current peaks, the power supply need not be different from that required for any class A operation of equivalent drain. The unequal peak currents of the driver are of such low magnitude compared with the output current that their effect on the power supply regulation can usually be neglected. Furthermore, the plate efficiency of the system is high because the power tube has no negative grid field and therefore the same plate current swings can be obtained at much lower plate voltages.

Until now there has been no mention of grid current in the input circuit. It should be clearly understood that no power is required in the driver's grid circuit. By no power, it is meant here that no convection current is flowing between the driver's grid and its cathode; hence the input impedance is high, limited only by reactive admittance and electrical leakage. The system may be so designed that the power developed throughout the entire normal portion of its over-all I_b vs. E_c characteristic can be realized long before the input grid is swung sufficiently far to draw current. Then there is no advantage whatever in driving the grid positive. This unique grid overload characteristic functions as follows. When e_i (the signal input voltage) is zero, E_0 (the voltage across cathode and ground) is the static bias of the tube. At positive values of e_i, E_0 rises beyond the established bias at a rate dependent upon the gain of the driver circuit. By (6) it is seen that the gain must always be less than unity. So the increase in e_i is faster than the increase in E_0. Eventually, e_i will overtake E_0. At this point, the potential between grid and cathode reverses, with the grid becoming positive with respect to its cathode. Therefore, the point at which grid current begins flowing may be expressed as

$$+e_i \geqq -(E_0 + \Delta E_0) \pm \mathcal{E} \tag{7}$$

where $\mathcal{E}$ represents the contact potential between grid and cathode. Since the gain, G, of the circuit is equal to $\Delta E_0/e_i$, then $e_i G$ may be substituted for ΔE_0, and clearing, (7) becomes

$$+e_i \geqq -\left(\frac{E_0}{1-G}\right) \pm \mathcal{E}. \tag{8}$$

It is obvious that the higher the gain the more extended the grid overload characteristic. In practice, the limiting condition is usually a signal which is sufficiently great to make the instantaneous voltage on the power tube plate as low as the voltage across its grid. When this occurs its grid impedance decreases so rapidly that the driver's gain actually decreases.

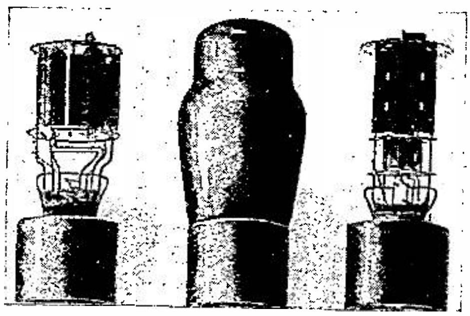
II—Commercial Application—6B5 Tube

The remaining part of this paper deals with a commercial application of dynamic coupling, the new type 6B5 power tube. This tube combines both the driver and power section to make a six-prong unit; the terminals being common heater, output cathode, driver grid, driver plate, and output plate. Such a tube provides a simplified power output system, one without the need of any biasing means. Justification of this construction is established by two major points. First, the cost of the combination tube is less than two single tubes; and second, since the performance is interdependent upon the functions of each section, it is logical that the system be treated as a whole rather than as two individual units. There are several comments regarding the design of such a tube which are worthy of considerable attention.

In consideration of gain, distortion, and input grid current, the grid impedance of the output section should be as high as possible and the impedance versus grid-voltage curve should descend smoothly and gradually. Two grids of different diameters but connected together should be used in the output section because, by proper distribution of the mean field, it is possible to obtain a higher mu with a closer approach to the desired characteristics than can be obtained if only a single grid is used. Any irregularities in the impedance curve appear in the wave shape of the driving voltage and hence their effects are transferred to the output current. Consequently, the smoothness of the impedance curve is important with respect to distortion. Slight wavers in the curve produced by a small amount of secondary grid emission will cause the generation of higher order harmonics. Consequently, preventive measures should be taken to suppress secondary emission. Since it is difficult to eliminate completely all traces of secondary emission, the tube should be designed so that the slight irregularities occur at or near the quiescent point on the I_b vs. E_c curve as represented by Q in Fig. 2. The resulting higher order harmonics will, therefore, be present at nearly all power levels rather than appear abruptly at some particular level.

Another important point is the plate-current cutoff characteristic of the driver section. By inspection of Fig. 2 it is seen that if the output plate current is to be moved down to a minimum value corresponding to zero grid, the grid current must likewise go down to its minimum value. Usually this latter current is relatively small. Now if the driver has a rounded I_b vs. E_c curve at the extreme low current end, it will not lower the output plate current at the proper rate. Hence the driver should have a sharp cutoff characteristic.

In the design calculations, a current surge was overlooked. The surge would start during the cathode heating-up period of the combination models, and once started, the output current would rise rapidly to a very large value. Often the current would become sufficiently great to destroy the tube. If the cathodes were allowed to reach their normal operating temperature before the plate voltage was turned on, the surge did not occur. The surge, also, did not occur when the heater and plate voltage were applied simultaneously, providing the output cathode reached its operating temperature considerably faster than the input cathode. Once the surge started, the input plate current would nearly disappear. To obtain such high output current, the output grid must be subjected to an abnormally high positive voltage. Obviously, secondary emission from the output grid was the cause of the surge, because the only manner in which the input plate current can go nearly to cutoff is by having high resistance in the input section's cathode-ground circuit, and this resistance could only be obtained by the subtractive effect of electrons emitted from the grid. This explanation is satisfactory but certainly not obvious from an inspection of the grid-current versus grid-voltage curves of the output section, for they show no evidence of sufficient secondary emission to reduce the grid current to zero. However, these curves were taken after the cathode had reached its operating temperature and are not necessarily indicative of the grid resistance during the cathode heating period. The rate of emitted secondary electrons does not always increase proportionately with the number of electrons leaving the cathode; and there is a saturation limit of secondary emission for any fixed voltage condition, just as there is with primary emission. For example, when the cathode is warming up, the bombarding electrons are comparatively few, and so the secondary electrons may number as great as or be even greater than the bombarding quantity. By the time the cathode emits copiously, the secondary emission may have exceeded its saturation limit to the extent that the secondary current becomes negligible as compared with the primary current. However, should the positive grid voltage continuously increase during the cathode heating period and thereby increase the impact force of the bombarding electrons, the point at which the secondary emission saturates may never be reached; then the number of bombarding electrons cannot appreciably exceed the number of secondary electrons. Returning now to the combination tube, the voltage on the output grid obviously increases as the cathodes warm up and stability is attained with tubes which do not surge when the cathodes have reached their operating temperature. The increasing grid voltage of tubes which do surge, however, grows so rapidly due to the high grid resistance produced by the secondary emission phenomenon, that the tubes become unstable. The tubes, which were designed to have the output cathode arrive at its operating temperature considerably faster than the input cathode, do not surge upon normal heating-up simply because the output cathode passes its critical emitting value before its grid could obtain an appreciable positive potential. Unfortunately, the solution to the difficulty was not simply a heating time differential because the differential could not be made to maintain its proper relation unless the cathodes were actually cool each time the tube was turned on. In practice this is not always the case. For example, the tube might be turned off and then turned on again before the cathodes had become completely cooled. Naturally the percentage of time at which the surge would be encountered under these particular conditions would be indeed very small. Nevertheless, the possibility of such failures existed. One positive remedy was to introduce between the output grid and cathode a protective resistance low enough to prevent a voltage developing on the grid that might be sufficient to cause instability. The value of a satisfactory resistance was still sufficiently high so that its shunting effect on the dynamic operation of the tube could be neglected. Since the resistance is truly a protective device, it was incorporated within the tube. This required the design of a special miniature resistor. Had it been left to the user to connect the resistor externally, it is probable that in many applications where economy is rigidly practiced it would have been omitted. No doubt, in some applications there would be no detrimental result, but in others, the surge would ultimately ruin the tube and probably cause considerable damage to the equipment. As an additional safety factor for the final design, a heating time differential of two to one was employed without disturbing the identical operating temperatures of the two cathodes.



1 2 3

Fig. 6—1. Type 6B5, side-by-side cathode construction, manufactured by Hygrade Sylvania Corporation. 2. Showing over-all size of the 6B5. 3. Type 6B5, tandem cathode construction, manufactured by Triad Manufacturing Company and the Ken-Rad Corporation.

The 6B5 tube is now being produced commercially by several manufacturers. One group[1] utilizes a tandem arrangement of the elements. Here the cathodes are of different diameters, with the end of the smaller telescoped into the larger and separated by a small insulator bushing. A common heater supplies the cathode power. The physical proportions of the input section with respect to the output can be seen from Fig. 6. The diamond-shaped plate was chosen because a particularly concentrated field pattern at the minor axis could be obtained without appreciable sacrifice in the dissipating area of the plate. The power grids are also diamond-shaped, the major axis being made so great that the electronic power absorbed by the grid posts is negligible. Another manufacturer[2] uses a construction in which the two sections

[1] Triad Manufacturing Company and the Ken-Rad Corporation.
[2] Hygrade Sylvania Corporation.

are mounted side by side, requiring two separate heaters. Again the special plate and grid design are used. This construction is also shown in Fig. 6.

The average static plate characteristics for a fixed input plate voltage are shown in Fig. 7. Naturally, if a different input plate voltage

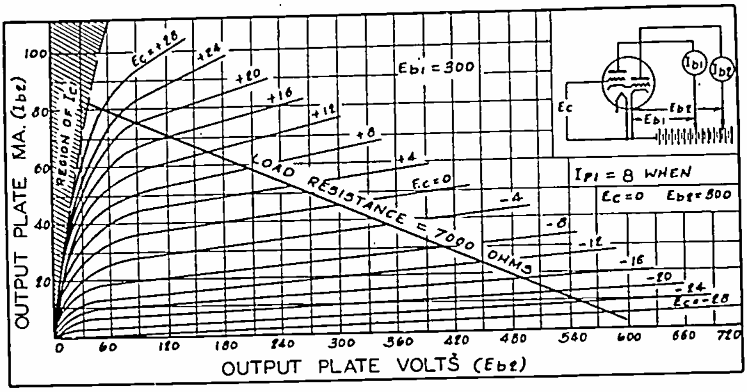


Fig. 7—Type 6B5—Composite output plate characteristics.

is used, a new set is necessary. The region where input grid current is present is clearly indicated. These curves are useful for calculation purposes. However, not too much reliance should be placed on the accuracy of the curves with high positive grid values. These were taken by the usual static method, rather than by some instantaneous method,[3] which would more accurately represent the true dynamic condition.

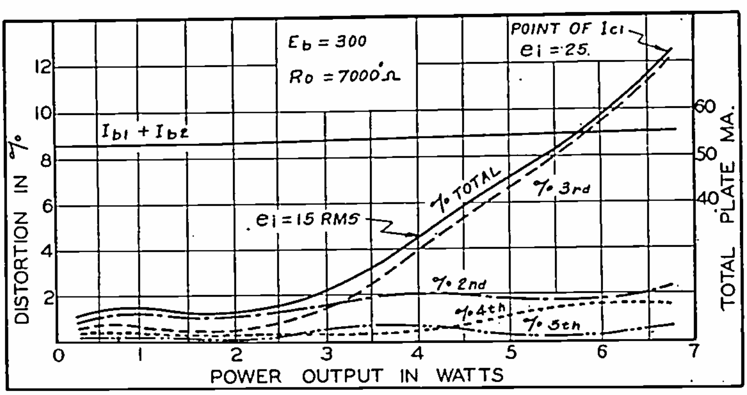


Fig. 8—Single 6B5—Harmonic distortion curves and total plate current as functions of power output.

Unfortunately, the required apparatus was not readily available. The practical performance is more readily appreciated by inspecting the dynamic operating results.

The harmonic distortion versus power output for single tube operation is shown in Fig. 8. The rated power of four watts at slightly under

[3] H. N. Kozanowski and I. E. Mouromtseff, Proc. I.R.E., vol. 21, pp. 1082–1096; August, (1933).

five per cent total distortion is obtained with a fifteen-volt signal. The power sensitivity is only a little less than the conventional American pentode. It should be noted that grid current does not begin flowing until the signal reaches twenty-five volts. The fourth and fifth harmonics are present at all levels. Higher harmonics than the fifth were far too small to be shown with these curves. As previously explained in the introduction, the effect of these residual higher order harmonics upon the actual performance of the tube is much less than it would be if it were produced by intermittent operation into grid current. Repeated listening tests under high fidelity conditions have shown no perceptible difference between 6B5 operation and undriven triode

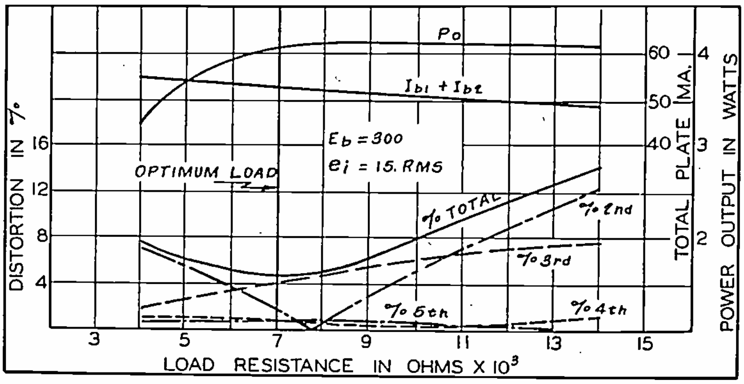


Fig. 9—Single 6B5—Harmonic distortion curves, total plate current, and power output as functions of load impedance with constant input signal.

class A, provided the actual sound-pressure frequency response has been treated so that both systems are identical in this respect. If no attempt is made to equalize the frequency response, the 6B5 system will usually be more brilliant; that is, the "highs" are not attenuated to the same degree as they are in triode performance. This can be anticipated from the relatively flat power curve in Fig. 9. The input signal for rated output was held constant and the output load varied over a wide range. The ratio of the plate impedance to the optimum 7000-ohm load is roughly three to one, compared with a modern triode ratio of three tenths to one, or the typical pentode ratio of fourteen to one. Therefore, the power does not fall off as rapidly as it does with triode operation; nor does it rise, as it does with a pentode. This compromise is advantageous, for when the reflected speaker impedance is varying with the signal frequency, the passages are not as rapidly lost nor are they objectionably accentuated. Until quite recently, a step-down ratio of plate-to-load impedance was considered important for minimizing transient response of the speaker. Recent investigations have shown that high-frequency transients caused by center moving modes of the cone are sufficiently damped even when the core flux is low and the plate impedance of the power tube is relatively high. High-frequency transients caused by other modes of vibration are not affected by the electrical load on the voice coil. Low-frequency transients are apparently not objectionable if the speaker's natural resonant period is one octave below the speech vowel fundamental. This is usual practice in the design of large speakers. In midget speakers the natural period is at a higher frequency, but they are almost invariably employed in small cabinets where the acoustical low-frequency attenuation is normally great. In such applications an apparent sense of low response is ob-

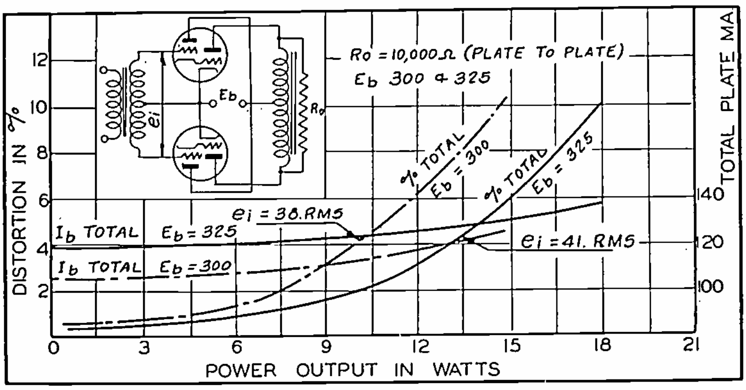


Fig. 10—Push-pull 6B5—Harmonic distortion and total plate current curves as functions of power output.

tained by allowing the natural resonance to accentuate the low frequencies. This condition is further supported if the power generated in the voice coil is relatively independent of the normal variations in the voice coil's impedance. From the foregoing it will be appreciated that the flat 6B5 power characteristic offers more latitude in speaker design. Fig. 9 also shows the harmonic distortion with the load as the variable parameter. The only harmonic that increases at any appreciable rate with increasing load impedance is the second. The rise in distortion is, of course, much slower when input signals of smaller magnitude are used.

The tubes can be used in conventional push-pull circuits. Fig.10 shows the performance at supply voltages of 300 and 325. Since the total distortion values are made up almost entirely of third harmonic, the individual harmonics have been left out purposely. The output of ten watts at two per cent distortion for 325-volt operation exemplifies the excellent performance. The total plate-current curves show the slow rate at which the drain increases with power output. Fig. 11 shows the push-pull performance with respect to output load for two different input signals. The optimum load has been selected at 10,000 ohms plate to plate rather than twice the rated value for single tube operation. This is feasible due to the particular order of harmonic distribution and the flat power curve.

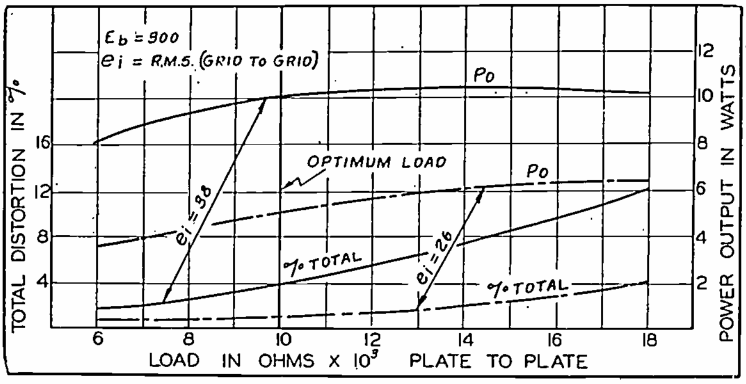


Fig. 11—Push-pull 6B5—Harmonic distortion and power output curves as functions of load impedance with a constant input signal.

From the foregoing it is obvious that the 6B5 was expressly designed for general purpose work. Should higher plate efficiency be essential for certain applications, a similar tube could be designed having its normal operating axis below the geometric center of its I_b vs. E_c characteristic. The maximum current wave from such a tube would

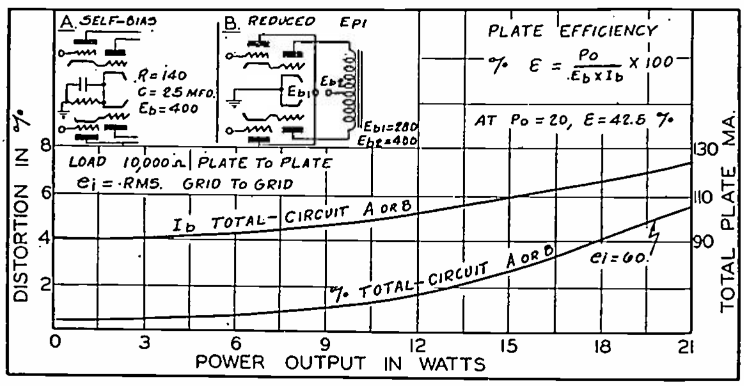


Fig. 12—Demonstration of push-pull high plate efficiency operation.

have a pronounced flattening on the negative peak which would limit the tube's usefulness to push-pull operation where this flattening disappears in the composite wave. The total plate current would rise with power output, but not at an objectionable rate. No input power would be required. To indicate this possibility the operating axis of the 6B5 in push-pull was lowered by introducing an ordinary bias resistor with suitable by-pass condenser in series with the common cathode lead. The plate supply was increased to 400 volts but the bias limited the plate current so that the static dissipation per plate was no greater than it was when operating at 325 volts without bias. Fig. 12 shows this performance. Here a power of twenty watts is obtained at five per cent distortion and at a high plate efficiency of forty-three per cent. It will be noted that the total plate current increases about thirty per cent from no signal to full signal. This is not a great change and the power supplies need not have as excellent regulation as are needed for class B operation. The same result can be obtained by limiting the plate current by lowering the input plate voltage rather than by the biasing means. This method is the preferable one because the driver plate dissipation is somewhat less. In either condition, the input grid does not draw current until the twenty-watt power level is considerably exceeded. Incidentally, it should be remembered that the slope of the output grid impedance curve rather than the plate voltage of the driver is the important factor with regard to the flow of input grid current. The 400-volt curves are used to demonstrate high efficiency operation. However, it should not be assumed that the 325 maximum voltage rating is to be exceeded. Of course, when it is necessary to economize on current drain, as with automobile receivers, advantage may be taken of the high efficiency operation at low voltage.

A quality possessed by the 6B5 and lacking in triodes and pentodes is the ability to stand considerable grid overload. When triodes or pentodes are fed from high impedance sources, a slight increase in signal above the value for rated output produces grid current which loads the preceding stage. If the coupling circuit has high resistance, grid rectification will also shift the power tube's operating axis. This overload causes severe distortion of the "crack-up" type which, for most receivers, is the practical limit for maximum output. The 6B5 operation is such that grid current does not flow until the signal actually increases to more than sixty per cent above the rating. This does not mean that the power output will increase proportionately, but rather that peak signals can exceed the rating to a considerable extent before causing "crack-up." In practice this is a decided advantage as the 6B5 system may be operated at a higher average level than an equally rated pentode or triode system. The effect of the 6B5 peak signal overload may be likened to automatic audio muting. Furthermore, this overload characteristic permits the practical employment of resistor coupled phase inverter circuits in 6B5 push-pull applications. The desirability of using phase inverter circuits is apparent when cost and frequency response are considered.

While the 6B5 tube combines desired qualities of many of the common types of power tubes, dynamic coupling is by no means limited to the medium audio power output field. The same principles are applicable to much larger systems as well as to transmitting tubes for both audio and radio frequency. Preliminary investigation of oscillator and frequency doubler applications show promise of certain advantages. No doubt, as more technicians explore the possibilities, many new uses will be found.

Acknowledgment

In conclusion, the writer wishes to acknowledge his appreciation for the assistance of Mr. Donald L. Alfred, who collaborated in the solution of many of the problems, and to the organization of the Triad Manufacturing Company.

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IONOSPHERE STUDIES DURING PARTIAL SOLAR ECLIPSE OF FEBRUARY 3, 1935*

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Summary—*Virtual height and critical frequency measurements of the several regions of the ionosphere were made during the day of the solar eclipse of February 3, 1935, and during several days before and after the eclipse day. The eclipse was found to produce a decrease of the critical frequency of each region. The decrease of critical frequency was approximately in time phase with the eclipse, thus indicating an ionizing agency (probably ultraviolet light), originating in the sun and propagated at approximately the velocity of light. The decrease of equivalent electron density in each region during the eclipse was compared with the decrease of the exposed area of sun's disk, and found to indicate that the ionization of the normal E region was diminished when the ionizing agency was decreased, by recombination of plus and minus charges, while the ionization of the F_2 region and a high stratum of the E was diminished by a process of attachment of electrons to neutral particles.*

Introduction

PREVIOUS to the solar eclipse of August 31, 1932, no quantitative determinations of the ionization densities of the several regions of the ionosphere had been made at the time of an eclipse. Measurements made by the National Bureau of Standards[1] and other North American observers[2,3] during this eclipse indicated that the main source of ionization of the E and F_1 regions was ultraviolet light. Similar effects for the F_2 region were not definitely found. Rose[4] reported a thirty per cent decrease in ionization density but was not sure that he had followed the F_2 critical frequency all of the time. The National Bureau of Standards found a small decrease of ionization density during this eclipse but the effect found was of approximately the same magnitude as variations found on some of the control days. The observations of Kenrick and Pickard[5] and Mimno and Wang[6] seem to apply to the F_1 region instead of the F_2.

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1 S. S. Kirby, L. V. Berkner, T. R. Gilliland, and K. A. Norton, *Bur. Stand. Jour. Res.*, vol. 11, pp. 829–845; December, (1933); Proc, I.R.E., vol. 22, pp. 247–264; February, (1934).

2 J. P. Schafer and W. M. Goodall, *Science*, vol. 76, pp. 444–446; November 11, (1932).

3 John T. Henderson, *Canadian Jour. Res.*, vol. 8, pp. 1–14; January, (1933).

4 D. C. Rose, *Canadian Jour. Res.*, vol. 8, pp. 15–28; January, (1933).

5 G. W. Kenrick and G. W. Pickard, Proc. I.R.E., vol. 21, pp. 546–566; April, (1933).

6 H. R. Mimno and P. H. Wang, Proc. I.R.E., vol. 21, pp. 529–545; April, (1933).

The 1932 eclipse was total through the Province of Ontario, Canada and Northeastern United States, where it could easily be observed. At Washington about ninety per cent of the area of the sun's disk was obscured. The eclipse occurred during early and midafternoon. The season, hour of day, and percentage of totality seemed to be favorable for definite results. Such results were obtained for the E and F_1 regions but not for the F_2.

The partial solar eclipse of February 3, 1935, offered an opportunity for further study of the effects of eclipses on the ionosphere. The hour of day (late forenoon) seemed favorable but the season and percentage of totality (thirty-five per cent of area of sun obscured) did not seem favorable for good results.

The effect of a winter eclipse on the F_2 region was of especial interest. It had been observed for several years by the National Bureau of Standards[7,8] at Washington and the Bell Telephone Laboratories[9] at Deal, New Jersey, that the ionization density of the F_2 region was not a simple function of the intensity of solar radiation. The ionization densities reached maxima at about sunset in summer and shortly after noon in winter, the winter day values being higher than the summer day values.[10] The ionization density followed nearly in phase with the intensity of received solar radiation during the winter day. Therefore there was a possibility that the effect of an eclipse in the F_2 region would be more clearly indicated in the winter than in the summer.

The technique and equipment available for the observations in 1935 were considerably better than those available in 1932. The multifrequency automatic recorder[11] was available to measure the virtual heights and critical frequencies for the band 2500 to 4400 kilocycles with a continuous frequency variation. This frequency band was swept through ten times per hour on the eclipse day and also the two days before and the two days after. Each sweep through this band required 4.75 minutes. The records obtained gave the critical frequencies for the E and F_1 regions and during the eclipse three points for the F_2 region. The virtual heights and critical frequencies for the frequency band from 4500 kilocycles up to the F_2 critical frequencies were determined by manually operated equipment with which the

[7] S. S. Kirby, L. V. Berkner, and D. M. Stuart, Proc. I.R.E., vol. 21, pp. 757–758; June, (1933).

[8] S. S. Kirby, L. V. Berkner, and D. M. Stuart, *Bur. Stand. Jour. Res.*, vol. 12, pp. 15–51; January, (1934); Proc. I.R.E., vol. 22, pp. 481–521; April, (1934).

[9] J. P. Schafer and W. M. Goodall, *Nature*, vol. 131, p. 804; June 3, (1933).

[10] E. O. Hulburt has given a reasonable theory for these phenomena. *See* E. O. Hulburt, *Phys. Rev.*, vol. 46, pp. 822–823; November 1, (1934); *Terr. Mag.*, vol. 40, pp. 193–200; June, (1935).

[11] T. R. Gilliland, *Bur. Stand. Jour. Res.*, vol. 11, pp. 561–566; October, 1933; Proc. I.R.E., vol. 22, pp. 236–246; February, (1934).

frequency was varied in steps usually 200 kilocycles. Some manual observations were also made at frequencies below 2500 kilocycles. The general method of observation was to increase the frequency of the pulse transmitter from low values to high and measure the virtual heights until the critical frequency was exceeded. This general procedure has been followed at the National Bureau of Standards since August 1930,[12] the technique improving with experience and improvements of equipment. The ionization densities were computed from critical frequencies by equation (1) given later.

Experimental Data

The data include critical frequency and virtual height measurements of the several ionized regions of the ionosphere, made by the methods just described. Fig. 1 shows the photographic records obtained on the frequency band 2500 to 4400 kilocycles. Scales for virtual heights and frequencies are indicated in the lower left-hand corner of Fig. 1. Critical frequencies are indicated by a more or less sharp rise of virtual height or a point of inflection. The critical frequency[13] of a given ionized region is the lowest frequency which will penetrate that region. The f_E and $f_{F_1}{}^0$ are indicated on these records. In addition to the manual measurements three points of $f_{F_2}{}^0$ were obtained with the automatic recorder near the maximum of the eclipse.

The E region was complex throughout this series of measurements. It appeared to be stratified. Two or three critical frequencies appeared much of the time. The values of these critical frequencies varied independently, and frequently one critical frequency would disappear or a new one appear between sweeps of six minutes separation. The two criticals appearing on February 3, between 1020 and 1112 E.S.T., Fig. 4, are examples of this complexity. It was often impossible to follow a particular E critical frequency through a series of records taken during successive six-minute periods. Some experimenters[9] have distinguished the several stratifications below the F_1 region by separate names. With the conditions existing during the period of these experiments it did not appear that this could be done consistently. For the reasons given above the multiple E critical frequencies have been averaged for the four control days.

[12] T. R. Gilliland, G. W. Kenrick, and K. A. Norton, *Bur. Stand. Jour. Res.*, vol. 7, pp. 1083–1104; December, (1931); Proc. I.R.E., vol. 20, pp. 286–309; February, (1932).

[13] f_E = critical frequency for the E region.
$f_{F_1}{}^0$ = critical frequency for the F_1 region, ordinary ray.
$f_{F_1}{}^x$ = critical frequency for the F_1 region, extraordinary ray.
$f_{F_2}{}^0$ = critical frequency for the F_2 region, ordinary ray.
$f_{F_2}{}^x$ = critical frequency for the F_2 region, extraordinary ray.

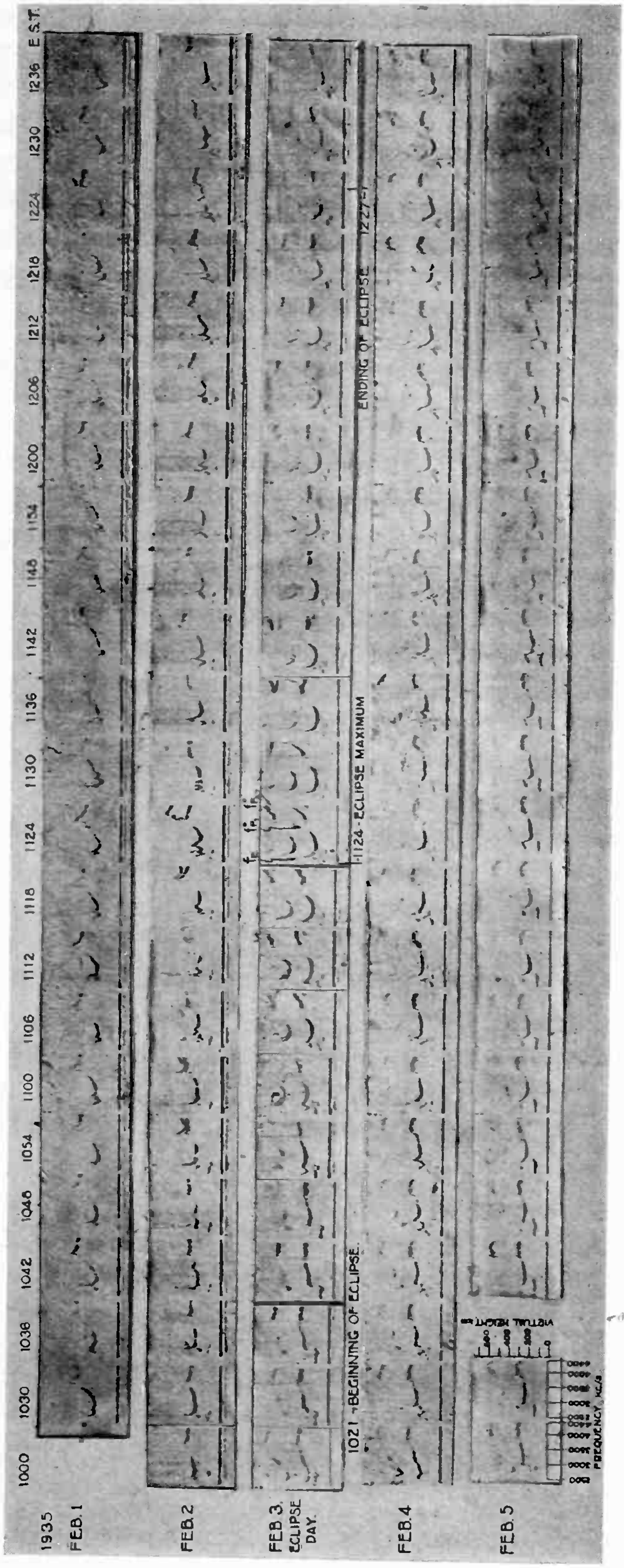


Fig. 1—Automatic multifrequency records of ionosphere heights and critical frequencies for day of eclipse and two days before and two days after eclipse.

The critical frequencies of the F_1 region were single but not sharply defined. The poor definition is characteristic of winter f_{F_1}. No double refraction was observed.

Three of the photographic records taken during the eclipse indicate $f_{F_2}{}^0$. On the first of these the record beginning at 1124 E.S.T., $f_{F_2}{}^0$ is indicated at the end of the record 4400 kilocycles. On the record beginning at 1130 E.S.T., $f_{F_2}{}^0$ is indicated clearly at 4330 kilocycles. The time at which this critical frequency was recorded was approximately 1134.5 E.S.T. On the record beginning at 1136 E.S.T. $f_{F_2}{}^0$ was again

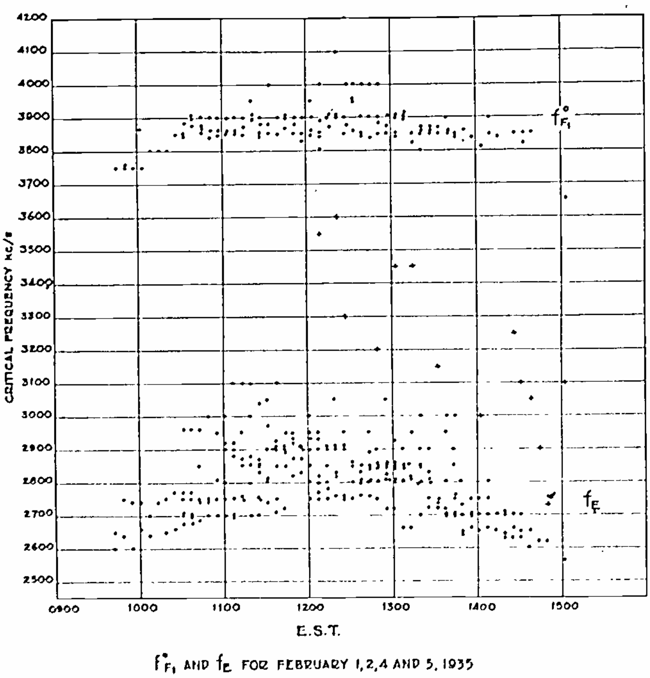


Fig. 2—Critical frequency data for control days for E and F_1 regions.

indicated at 4400 kilocycles. These three records indicate the time of minimum $f_{F_2}{}^0$ positively to within six minutes and by interpolation probably to the nearest minute.

In Fig. 2 the data are shown for f_E and $f_{F_1}{}^0$ for February 1, 2, 4, and 5 obtained from the records of Fig. 1. No attempt was made in this figure to distinguish critical frequencies for any individual day.

In Fig. 3 are shown the data for $f_{F_2}{}^0$ for the eclipse day and several days before and after the eclipse. The data for each day are shown by a separate graph in this figure. It will be noted that there was considerable variation of critical frequency from day to day. This makes it somewhat difficult to compare the results of any given day on which a special phenomenon such as an eclipse occurs with the average of several normal days because it is not known for certain what the results would have been on the given day had not the special phenomenon occurred. The Cheltenham Magnetic Observatory of the United States Coast and Geodetic Survey reported moderate magnetic disturbances for February 1 and 2. The magnetograms indicated that the disturbance of February 2 was much less severe than that on February 1. The F_2 critical frequencies were much lower on these two days than on the other control days. Some other data obtained at the National Bureau of Standards indicate that this correspondence between magnetic disturbances and low f_{F_2} frequently exists.

Figs. 1, 2, and 3 present the experimental data obtained in this investigation.

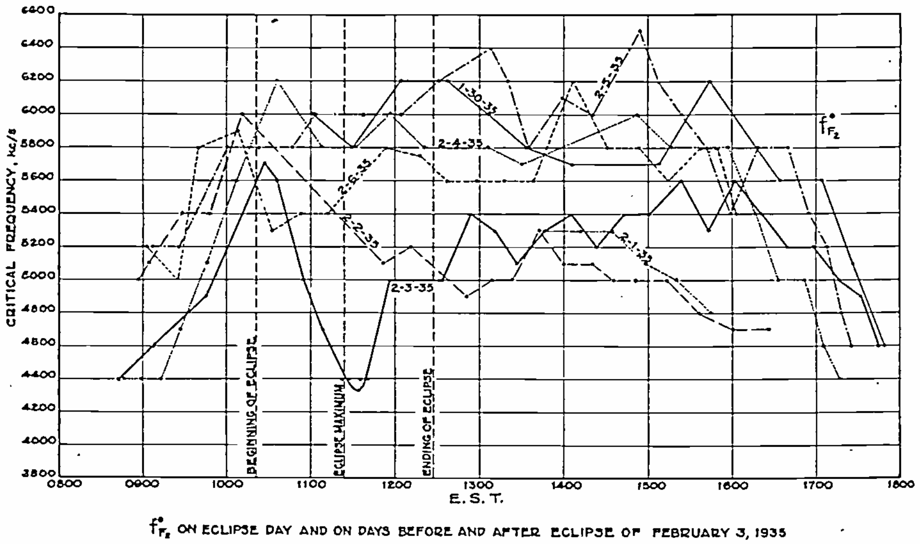
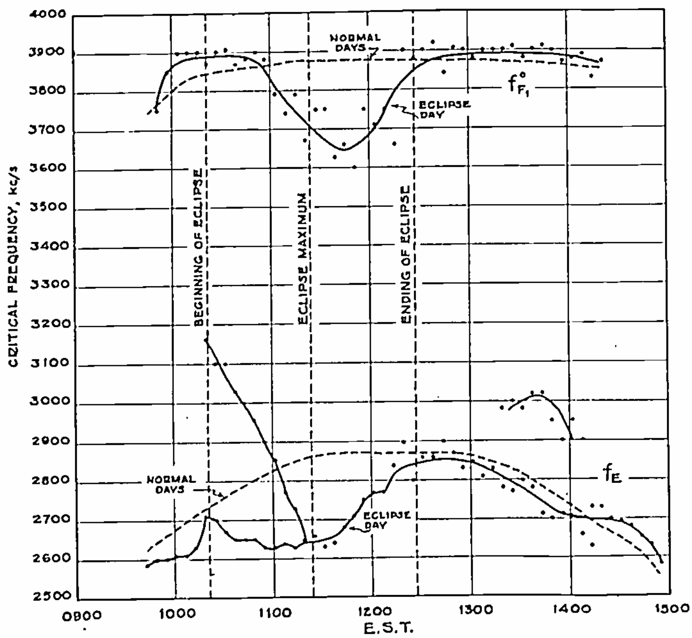


Fig. 3—Critical frequency data for eclipse day and control days for F_2 region.

Discussion of Data

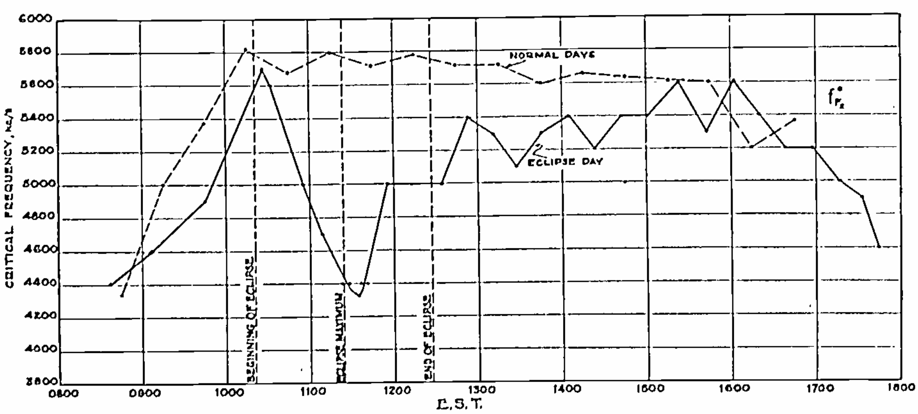
Since there was a complexity of critical frequencies in the E region and the critical values for a given region could not be traced through with certainty, an average was made of all the points obtained except a few widely scattered points indicated by + in Fig. 2. These points were not used in the average simply because they were so widely scattered that they did not seem to belong to the E region. This average, and the individual critical frequencies obtained on the eclipse day, are shown in the lower part of Fig. 4. The branch beginning with a critical frequency of 3160 kilocycles at about 1020 E.S.T. and ending about 1120 E.S.T. by combining with the more normal appearing f_E graph is of special interest although its significance is not clear. The critical frequencies forming this branch can be seen clearly in Fig. 1. Except for this branch the f_E was less complex on the eclipse day than on the control days. This figure indicates a decrease of f_E during the eclipse.



f_E AND $f^{\circ}_{F_1}$, FOR ECLIPSE DAY AND FOR DAYS BEFORE AND AFTER ECLIPSE

Fig. 4—Averaged critical frequency data for control days for E and F_1 regions compared with corresponding data for eclipse day.

The upper part of Fig. 4 shows $f_{F_1}{}^0$ for the eclipse day and the average for the control days. This figure indicates a decrease of $f_{F_1}{}^0$ during the eclipse.



$f^{\circ}_{F_2}$ FOR ECLIPSE DAY AND AVERAGE FOR DAYS BEFORE AND AFTER ECLIPSE

Fig. 5—Averaged critical frequency data for control days for F_2 region compared with corresponding data for eclipse day.

The $f_{F_2}{}^0$ for the eclipse day and the average for the control days are shown in Fig. 5. All of the data for the control days shown in Fig. 3

were used for the average. This figure shows a pronounced decrease of $f_{F_2}{}^0$ during the eclipse but with a slight phase lag behind the eclipse. At a height of 250 kilometers the eclipse maximum occurred at 1125 E.S.T. The lowest value of $f_{F_2}{}^0$ was observed at 1134.5 E.S.T. (on the record of Fig. 1 beginning at 1130 E.S.T.). Since this critical frequency was symmetrically spaced in time with respect to critical frequencies of 4400 kilocycles obtained six minutes before and after, it seems reasonable to consider that the minimum critical frequency was very nearly 4330 kilocycles and occurred very nearly at 1134.5 E.S.T. or 9.5 minutes after the maximum of the eclipse.

Immediately before the eclipse, $f_{F_2}{}^0$ was nearly normal while after the eclipse maximum it rose rapidly at first but later ceased to rise rapidly but reached normal between 1500 to 1600 E.S.T. This effect might conceivably have been caused by the eclipse but it is more likely that the normal value of $f_{F_2}{}^0$ was low during the eclipse day as on February 1 and 2. During the night following the eclipse $f_{F_2}{}^x$ remained below 2500 kilocycles from the hour 2030 E.S.T. to sunrise. This represents an exceptionally long sustained low value of ionization density of the F_2 region but it has been recorded on other nights. It is not likely that this effect was caused by the eclipse. The ionization density during the night preceding the eclipse was also very low.

In the following paragraphs, let N_1 equal the ion density on the normal or control days and N_2 the ion density on the eclipse day, and $f_1{}^0$ and $f_2{}^0$ the critical frequencies of the ordinary ray on the control days and eclipse day, respectively. The ratios N_2/N_1 and $(N_2/N_1)^2$ will be compared with the percentage of the sun's disk which was exposed on the eclipse day.

From familiar theory we may write the following equation expressing the relation between critical frequency and equivalent electron density

$$N_\epsilon = 1.24(f^0)^2 10^{-2} \tag{1}$$

where N_ϵ = equivalent electrons per cm³ for a given region

and f^0 = the critical frequency in kilocycles for the ordinary ray for the same region.

For heavy ions (1) would have the same form but with a greater numerical constant.

Equation (1) is well known and is obtained directly from the fundamental theory of Eccles[14] and Larmor.[15]

[14] W. H., Eccles, *Proc. Roy Soc.*, vol. 87, pp. 79–99; August 13, (1912).
[15] Joseph Larmor, *Phil. Mag.*, vol. 48, pp. 1025–1036; December 3, (1924).

From (1) we obtain

$$\frac{N_2}{N_1} = \left(\frac{f_2{}^0}{f_1{}^0}\right)^2. \qquad (2)$$

The values of N_2/N_1 and $(N_2/N_1)^2$ for the E region are shown in Fig. 6. The values of $f_1{}^0$ and $f_2{}^0$ were taken for each six-minute period from Figs. 4 and 5 so that allowance has been made for the varying zenith angle of the sun. The ratios of the exposed area of the sun's disk on the eclipse day to the total area of the sun's disk are shown in the

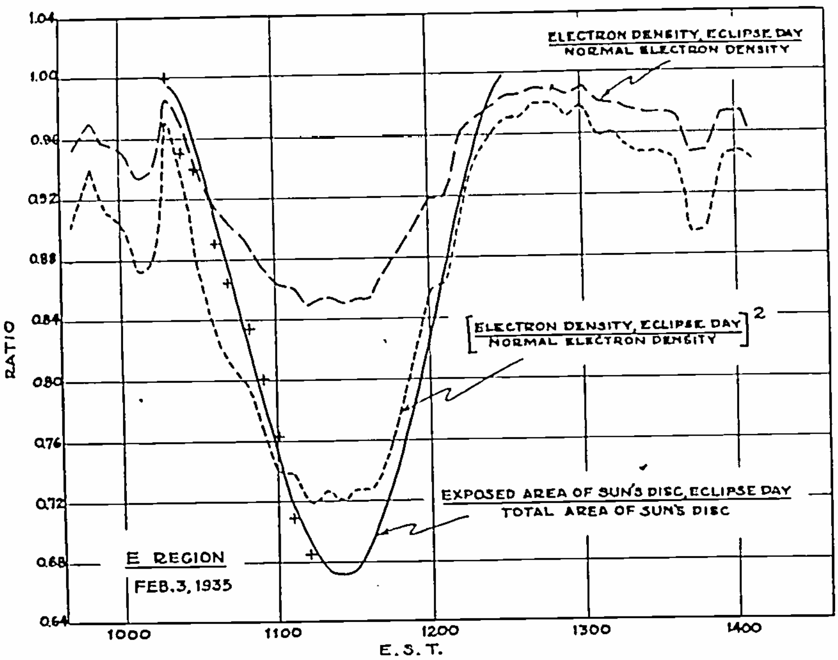


Fig. 6—Comparison of normal electron density of E region during 1935 eclipse, with exposed area of sun's disk. The plus marks = electron density eclipse day divided by normal electron density for upper branch of E region shown in Fig. 4.

same figure. It may be seen that the graph for $(N_2/N_1)^2$ for the main E region corresponds closely to the graph showing the percentage of the exposed area of the sun's disk. The plus marks in this figure represent values of N_2/N_1 instead of $(N_2/N_1)^2$ for the upper branch of the E region shown in Fig. 4. These values follow closely the graph showing the percentage of exposed area of sun's disk. The value measured just before the eclipse began was assumed to be normal and was used for N_1. From the results shown in this figure it would appear that there were two different laws of recombination for these two strata of the E region. The ionization density for the lower E region varied as the square root of the ionizing energy, while the ionization density of the upper branch produced during the first half of the eclipse varied as the first power of the ionizing energy. The ionizing energy of the sun is assumed to be proportional to the percentage of the exposed area of the sun's disk.

Fig. 7 is a reproduction of Fig. 3 of the Kirby, Berkner, Gilliland, and Norton[1] paper as published in PROCEEDINGS with the addition of a graph of $(N_2/N_1)^2$ against hour of day obtained from data taken during the 1932 eclipse. These results indicate that the ionization

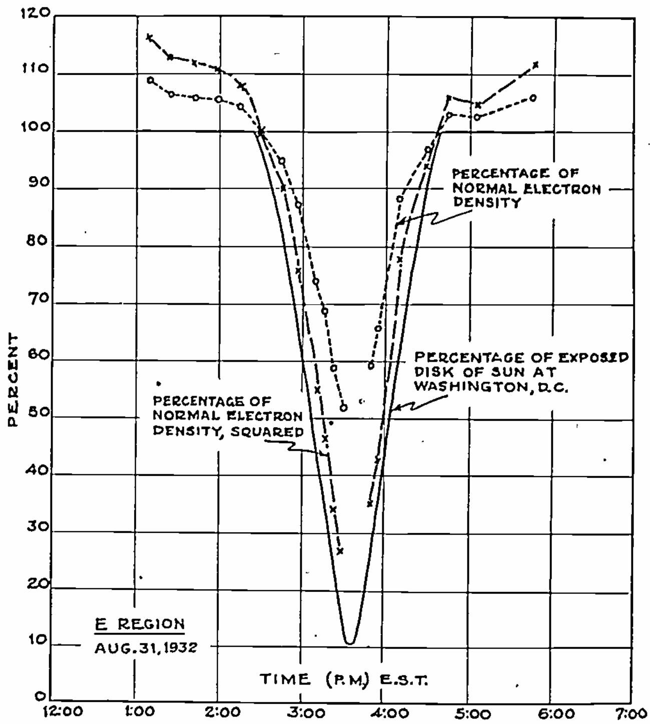


Fig. 7—Comparison of normal electron density of E_1 region during 1932 eclipse, with exposed area of sun's disk.

density of the E region was proportional to the square root of the ionizing energy during the eclipse of August 31, 1932.

The values of N_2/N_1 and $(N_2/N_1)^2$ for the F_1 region are shown in Fig. 8 along with the percentage of the exposed area of the sun's disk on February 3. Neither the graph N_2/N_1 nor $(N_2/N_1)^2$ correspond very closely with the graph showing the percentage of the exposed area of the sun's disk on eclipse day. However, the values of $(N_2/N_1)^2$ agree much more closely than those for N_2/N_1. These results indicate that it was more probable that the ion density of the F_1 region was proportional to the square root rather than the first power of the ionizing energy.

The values of N_2/N_1 and $(N_2/N_1)^2$ for the F_2 region are shown in

Fig. 9 along with the percentage of the exposed area of the sun's disk on February 3. In this case the graph N_2/N_1 corresponds closely with the graph showing the percentage of the exposed area of the sun's disk. This indicates that in the F_2 region at the time of this eclipse the ionization density was proportional to the first power rather than the square root of the ionizing energy of the sun.

Pedersen[16] shows that if there are $N-$ and $N+$ ions per cm^3 that the rate of recombination is proportional to N^2. Furthermore, if I pairs

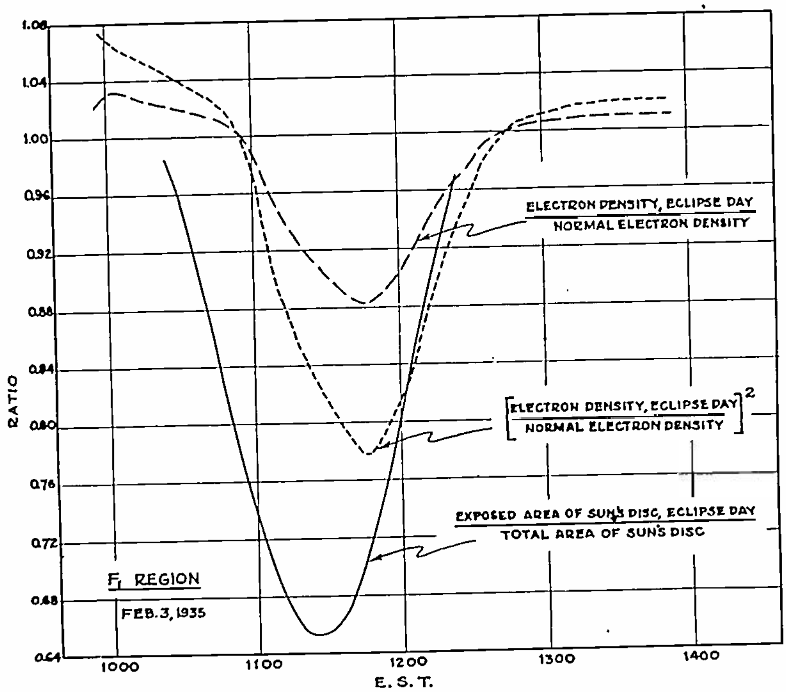


Fig. 8—Comparison of normal electron density of F_1 region during 1935 eclipse, with exposed area of sun's disk.

of such ions are produced per cm^3 each second by some agency such as ultraviolet light, and I is proportional to the ionizing energy of this agency, then

$$\frac{dN}{dt} = I - \alpha N^2 \tag{3}$$

where α is the coefficient of recombination.

However, if the ionization is effectively of one sign such as negative electrons surrounded chiefly by neutral gas particles, the effect of the electrons can be effectively destroyed by attachment to these neutral gas particles. The density and temperature of the gas will be assumed

[16] P. O. Pedersen, "The Propagation of Radio Waves," Chap. V.

constant. Then the rate at which electrons are lost by attachment is proportional to N rather than N^2. An equation for this condition corresponding to (3) may be written as follows:

$$\frac{dN}{dt} = I - \beta N \tag{4}$$

where β is the coefficient of attachment.

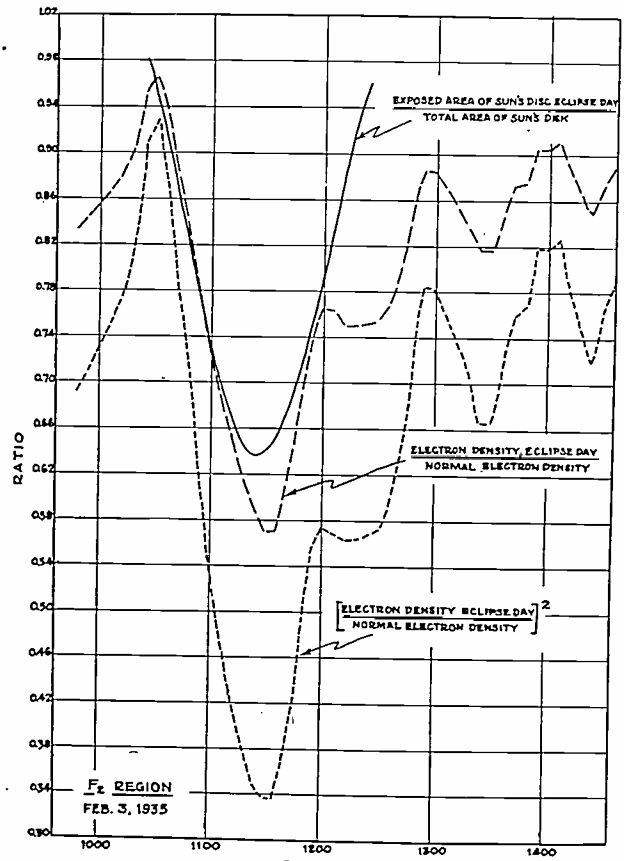


Fig. 9—Comparison of normal electron density of F_2 region during 1935 eclipse with exposed area of sun's disk.

For both (3) and (4) $dN/dt=0$ at the lowest points of the graphs of N vs. time. In Figs. 6 and 9 there were very small time differences between the minimum of exposed area of the disk of the sun and the minimum ionization densities. Also the graphs of $(N_2/N_1)^2$ vs. time for the normal E region in Figs. 6 and 7 coincide very closely throughout the eclipse with the graph of the ratio:

$$\frac{\text{exposed area of sun's disk on eclipse day}}{\text{total area of sun's disk}} \text{ vs. time.}$$

The graphs of N_2/N_1 vs. time for the upper branch of the E region in Fig. 6 and for the F_2 region in Fig. 9 coincide very closely throughout the eclipse with the graph of the exposed area of the sun's disk vs. time. Therefore we may safely assume that dN/dt was very small in comparison with either term in the right-hand members of (3) and (4) or for practical purposes $dN/dt=0$ at any point on the graphs mentioned in this paragraph for the conditions specified. This is equivalent to saying that equilibrium and recombination is reached quickly. Similar conclusions regarding the F_1 region at the time of this eclipse can hardly be drawn.

It now appears that Figs. 6 and 7 indicate that the ionization densities of the E region obeyed (3) and that N for these regions was proportional to the square root of the ionizing energy. These results indicate the probability that the ionization of the normal E region at the times of these experiments was made up of plus and minus charges which recombined with one another when the ionizing energy was decreased. This condition could be satisfied with heavy plus and minus ions or with heavy plus ions and minus electrons.

Also Figs. 6 and 9 indicate that the ionization densities of the region producing the upper branch of E region shown in Fig. 2, and the F_2 region, were proportional to the ionizing energy. These results indicate the probability that the ionization of these two regions at the times of these experiments was made up of electrons. Magnetoionic splitting is positive evidence that the F_2 region is electronic. The eclipse evidence checked this and in addition indicated that the effect of these electrons was destroyed by attachment to neutral gas particles rather than by recombination with plus ions.

Conclusions

The ionization density of the E region was reduced to a minimum of 0.85 normal by the eclipse of February 3, 1935, during the maximum of which 0.33 of the area of the sun's disk was covered. The ionization density of the F_1 region was reduced to 0.88 normal by the eclipse at whose maximum 0.35 of the area of the sun's disk was covered. The ionization density of the F_2 region was reduced to 0.57 normal by the eclipse at whose maximum 0.36 of the area of the sun's disk was covered.

There was no measurable time lag of ionization changes after the eclipse for the E region.

There appeared to be a time lag of ionization changes behind the eclipse for the F_1 region. This lag amounted to about twenty minutes. This estimate is not very certain because of the absence of sharp critical frequencies in the F_1 region during the winter. There was a definite lag of ionization changes after the eclipse for the F_2 region. The minimum ionization density occurred 9.5 minutes after the maximum of the eclipse.

In the E region the ratios of ionization densities squared on the eclipse day to the ionization densities squared on normal days was approximately proportional to the ratio of the exposed area of sun's disk on eclipse day to the total area. This result taken with the rapid recombination indicates that the ionization of the normal E region was probably made up of heavy ions with plus and minus charges or heavy plus ions and minus electrons and that these were destroyed by a process of recombination.

On the eclipse day the changes in a higher stratum of the E region indicated that the ionization there was electronic.

The nature of the ionization of the F_1 region was not very clearly indicated probably because of poor definition of the F_1 critical frequencies during the winter.

In the F_2 region the ratios of the first powers of ionization densities on the eclipse day to those on normal days were approximately proportional to the ratios of the exposed area of the sun's disk on the eclipse day to the total area. This result is in agreement with (4) and indicates that the ionization of this region was probably electronic and that the ionization was diminished chiefly by a process of attachment to neutral gas particles.

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A NEW ELECTRON TUBE HAVING NEGATIVE RESISTANCE*

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Summary—*The dynatronic and magnedynatronic systems commonly used to obtain a negative resistance by means of triodes or tetrodes are characterized by the fact that the potential of the active electrode (the anode) is not the highest potential in the system; an electrode (the grid or the screen grid) at a higher potential is necessary in order to take off the secondary emission current from the anode in such a way that a falling characteristic is produced in the anode circuit.*

The present paper deals with a three-electrode tube of such design that the negative resistance is obtained in the circuit of the electrode possessing the highest potential in the system.

THE NEW TUBE AND THE PRINCIPLE OF ITS ACTION

THE tube here described is a triode with a grid of special form (that of a venetian blind) as shown in Fig. 1. The electrode system is cylindrical: a straight wire filament, a cylindrical anode, and a grid consisting of a number of ribs inclined at an angle α (with regard to the radius).[1]

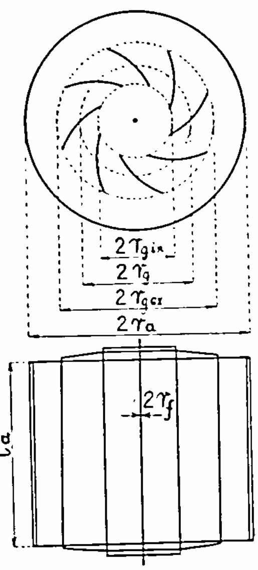

Fig. 1

* Decimal classification: R330. Original manuscript received by the Institute, December 5, 1935. Résumé from the paper in Polish, Wiadomości i Prace Państwowego Instytutu Telekomunikacyjnego, vol. 6, (1935).

[1] A grid of similar shape was used by F. B. Haynes (in a triode in which the electrostatic control was replaced by an electromagnetic one) *Physics*, vol. 1 p. 192, (1931), and by K. Okabe, *J.I.E.E.* (Japan), February, (1927).

The tube (connected with direct voltage sources as shown in Fig. 2) is placed in a magnetic field which is parallel to the axis of the electrode system. With the anode potential e_a, grid potential e_g and total emission current i_s constant, the dependence of the anode current i_a and

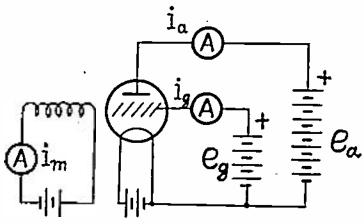

Fig. 2

the grid current i_g upon the magnetic field current i_m is given by the curves in Fig. 3. As seen, the curves i_a and i_g are not symmetrical with respect to the axis $i_m=0$. This is caused by the special form of the grid, which confers upon the tube different characteristics for the two direc-

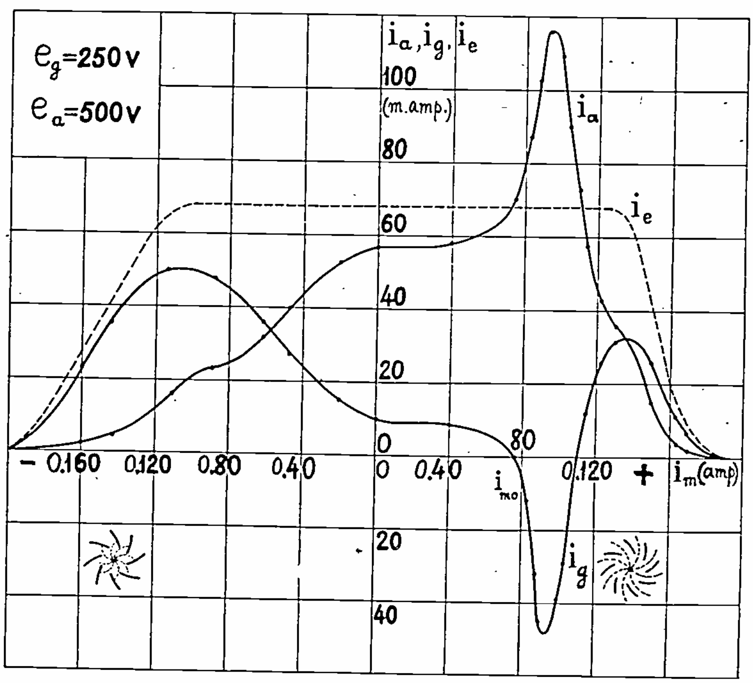

Fig. 3

tions of the magnetic field. Under the action of the magnetic field the electron paths are turned ribwise (sign +) or antiribwise (sign −), according to the magnetic field direction. For both directions there are critical values of i_m for which the electrons do not reach the grid: beyond these values the currents i_a and i_g become zero.

It is possible to establish the following equations: In the saturation range $e_a \gg 0$, $e_g > 0$, the total emission current i_s must be the sum of the primary grid and anode currents i_g' and i_a';

$$i_s = i_g' + i_a'. \tag{1}$$

The primary current distribution between the anode and the grid is given by the function F;

$$\frac{i_a'}{i_g'} = F\left(\frac{e_a}{e_g}\right). \tag{2}$$

The secondary emission efficiency of the grid can be defined as the ratio of the secondary grid current to the primary grid current;

$$\gamma = \frac{i_g''}{i_g'}. \tag{3}$$

The (resultant) anode current consists of the primary anode current i_a' and of the secondary grid current i_g'', according to the equation

$$i_a = i_a' + i_g''. \tag{4}$$

From (1), (2), (3), and (4) we obtain

$$i_a = i_s \frac{\gamma + F}{1 + F}. \tag{5}$$

If γ is greater than 1, we can write

$$\gamma = 1 + \beta \qquad (\beta > 0) \tag{6}$$

and (5) will be

$$i_a = i_s\left[1 + \frac{\beta}{1 + F}\right]. \tag{7}$$

The exact expression for the function F is not known; it can be represented in the form

$$F\left(\frac{e_a}{e_g}\right) \cong k\left(\frac{e_a}{e_g}\right)^m \tag{8}$$

where k is a constant depending on the geometrical shape of the electrodes ($k \cong 1$), and m lies between ½ and ⅓. Thus, we can write (7):

$$i_a = i_s\left[1 + \frac{\beta}{1 + k\left(\frac{e_a}{e_g}\right)^m}\right]. \tag{9}$$

Equation (9) is an equation of the falling characteristic

$$i_a = f(e_a). \tag{10}$$

If we suppose i_s, e_g, k, and β to be constant, the differentiation of (9) with respect to e_a gives the negative resistance of the anode circuit of the tube

$$\rho = \frac{de_a}{di_a} = -\frac{1}{km\beta i_s}\left(\frac{e_g}{e_a}\right)^m\left[1 + k\left(\frac{e_a}{e_g}\right)^m\right]^2 e_a. \tag{11}$$

Fig. 4 shows the anode-current—anode-voltage ($i_a - e_a$) characteristic at e_g, i_s, and i_m constant. In the same figure the grid current—anode voltage ($i_g - e_a$) characteristic is given. We see that the current i_g di-

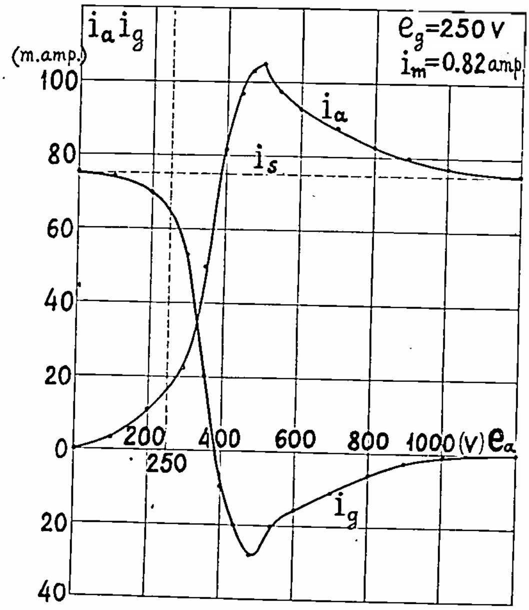

Fig. 4

minishes with increase of e_a, passes through zero, and changes sign (because of the increase of secondary electrons from the grid).

With regard to the magnetic field, its rôle is to make critical the influence of the anode voltage e_a on the current i_g''. Suppose the electron path in the proximity of the grid rib is nearly tangent to the rib, that is, the electron strikes the exterior rib surface at an angle near zero (Fig. 5). Then a small change of the anode voltage can bend the electron path in such a way that the electrons do not touch it any more; the current i_g' diminishes and i_a' increases. In consequence of the diminishing of the current i_g', the current i_g'' diminishes. If the secondary emission from the grid is intense (γ greater than one), the anode current i_a diminishes also. Thus we can obtain a falling characteristic in the anode circuit.

The Equations of Optimum Operation

We can establish the equations which express the relation between the dimensions of the electrodes, their potentials, and the magnetic field strength, in the range of negative resistance. As a basis we assume that; first, the greatest secondary emission efficiency of the grid occurs in the case when the electrons strike the grid at an angle near zero; second, a change of anode voltage causes a deviation of the electron paths near the grid ribs; third, the secondary emission from the grid is totally led away to the anode.

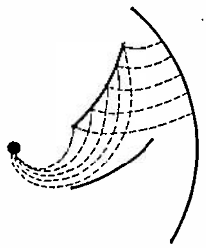

Fig. 5

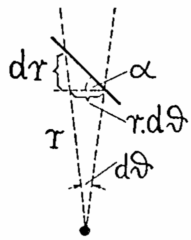

Fig 6.

The discussion relates to the cylindrical electrode system, disregarding space charge.

1. *The electrons reach the grid tangentially to its ribs.*

In order to obtain the relation between the operating conditions (e_a, e_g, H) and the constants of the tube (the radius of the grid cylinder r_g, the angle α, and the radius of the filament r_f, Fig. 1), we write a differential equation of the tangential forces acting on the electron which moves from the cathode to the anode. In polar co-ordinates (r, θ) we have

$$mr\frac{d^2\theta}{dt^2} + 2m\frac{d\theta}{dt}\frac{dr}{dt} = -\epsilon H\frac{dr}{dt} \tag{12}$$

where m and ϵ are the mass and the charge of the electron. After the integration

$$r^2\frac{d\theta}{dt} = -\frac{1}{2}\frac{\epsilon}{m}Hr^2 + C. \tag{13}$$

For $\theta=0$, $r=0$, hence $C=0$. From (13) we have, therefore,

$$\frac{d\theta}{dt} = -\frac{1}{2}\frac{\epsilon}{m}H. \tag{13a}$$

The law of energy conservation for an electron gives us

$$\frac{m}{2}\left[\left(\frac{dr}{dt}\right)^2 + r^2\left(\frac{d\theta}{dt}\right)^2\right] = \epsilon e. \tag{14}$$

If the electron has to reach the grid at the angle α (Fig. 6), there must be fulfilled the condition

$$\tan \alpha = \frac{1}{r}\,\frac{dr}{d\theta}. \tag{15}$$

Taking into consideration that $e = e_g$ for $r = r_g$, we obtain from (14) and (15)

$$\frac{m}{2}\left[r_g^2(\tan^2 \alpha + 1)\left(\frac{d\theta}{dt}\right)^2\right] = \epsilon e_g. \tag{16}$$

Putting (13a) into (16) we find the expression we want,

$$H = \frac{1}{r_g}\sqrt{\frac{8m}{\epsilon}\,\frac{e_g}{1 + \tan^2 \alpha}}\ \mathrm{lg} \tag{17}$$

or,

$$H = \frac{6.72\ e_g^{1/2}}{r_g\sqrt{1 + \tan^2 \alpha}}\ \text{e.m.u.} \tag{18}$$

2. *The electrons enter the anode field tangentially to the grid ribs.*

In order that the path of the electrons which pass from the grid field into the anode field does not change, this field should be the same on both sides of the grid as in a tube without grid; thus the potential equation must be given by

$$e = e_a \frac{\log_\epsilon \dfrac{r}{r_f}}{\log_\epsilon \dfrac{r_a}{r_f}}. \tag{19}$$

Since $e = e_g$ for $r = r_g$, we have, therefore,

$$e_g = e_a \frac{\log_\epsilon \dfrac{r_g}{r_f}}{\log_\epsilon \dfrac{r_a}{r_f}} \tag{20}$$

($\log_\epsilon$ = nat. log.)

3. *The secondary electrons from the grid have to reach the anode.*

On the other hand, the potential difference between the anode and the grid should be sufficient for the secondary electrons (leaving the grid with initial velocity very small) to reach the anode.

The equation of the motion of electrons in the grid-anode space we get from (13), putting there

$$C = \frac{1}{2}\frac{\epsilon}{m} H r_g^2 \tag{21}$$

since for

$$\theta = 0, \qquad r = r_g.$$

Hence,

$$r^2 \frac{d\theta}{dt} = -\frac{1}{2}\frac{\epsilon}{m} H(r^2 - r_g^2) \tag{22}$$

or,

$$\frac{d\theta}{dt} = -\frac{1}{2}\frac{\epsilon}{m} H \frac{r^2 - r_g^2}{r^2}. \tag{23}$$

The equation of energy conservation is here

$$\frac{m}{2}\left[\left(\frac{dr}{dt}\right)^2 + r^2\left(\frac{d\theta}{dt}\right)^2\right] = \epsilon(e_a - e_g). \tag{24}$$

In extreme cases, when $r = r_a$, $\alpha = 0$, and therefore

$$\frac{dr}{dt} = 0. \tag{25}$$

Putting (25) into (24) we have

$$\frac{m}{2} r_a^2 \left[\frac{1}{2}\frac{\epsilon}{m} H \frac{r_a^2 - r_g^2}{r_a^2}\right]^2 = \epsilon(e_a - e_g) \tag{26}$$

hence,

$$e_a - e_g = \frac{1}{8}\frac{\epsilon}{m} H^2 \left(\frac{r_a^2 - r_g^2}{r_a}\right)^2 \tag{27}$$

or,

$$e_a - e_g = \frac{1}{45} H^2 \left(\frac{r_a^2 - r_g^2}{r_a}\right)^2. \tag{28}$$

(e in volts, H in *c.g.s. e.m.u.*, r in *cm*)

4. *The optimum dimensions of the tube.*

By eliminating e_a, e_g, and H from (18), (20), and (28) we obtain the expression which gives the most suitable proportion of geometrical dimensions of the tube:

$$\frac{\log_\epsilon \dfrac{r_a}{r_f}}{\log_\epsilon \dfrac{r_g}{r_f}} - 1 = \frac{(r_a^2 - r_g^2)^2}{r_a^2 r_g^2} \cdot \frac{1}{1 + \tan^2 \alpha}. \tag{29}$$

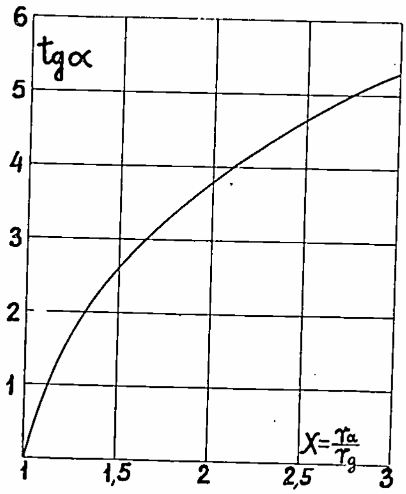

Fig. 7

The expression (29), after transformation, can be written as follows:

$$\tan \alpha = \sqrt{\left(\frac{r_a}{r_g} - \frac{r_g}{r_a}\right)^2 \frac{\log_{10} \dfrac{r_g}{r_f}}{\log_{10} \dfrac{r_a}{r_g}} - 1}. \tag{30}$$

($\log_{10}$ = dec. log.)
For values of r_g/r_f between 50 and 200, $\log_{10} r_g/r_f$ varies from 1.7 to 2.3. Assuming $\log_{10} r_g/r_f \cong 2$, and denoting $r_a/r_g = x$, we write (30) in the form

$$\tan \alpha \cong \sqrt{\frac{2\left(x - \dfrac{1}{x}\right)^2}{\log_{10} x} - 1}. \tag{31}$$

The curve $\tan \alpha = f(x)$ is shown in Fig. 7.

6. *Experimental results.*

One of the experimental tubes had the following data:

$r_a = 2.50$ cm	$c_a = 2.20$ cm	$e_f = 6$ volts
$r_{g\,\mathrm{ext}} = 2.00$	$c_g = 3.00$	$i_f = 2$ amps.
$r_{g\,\mathrm{int}} = 1.00$	$c_f = 2.50$	$i_s \cong 70$ ma.
$r_{g\,\mathrm{av}} = 1.50$	$\tan\alpha = 0.8$	
$r_f = 0.006$	no. of ribs $= 8$	

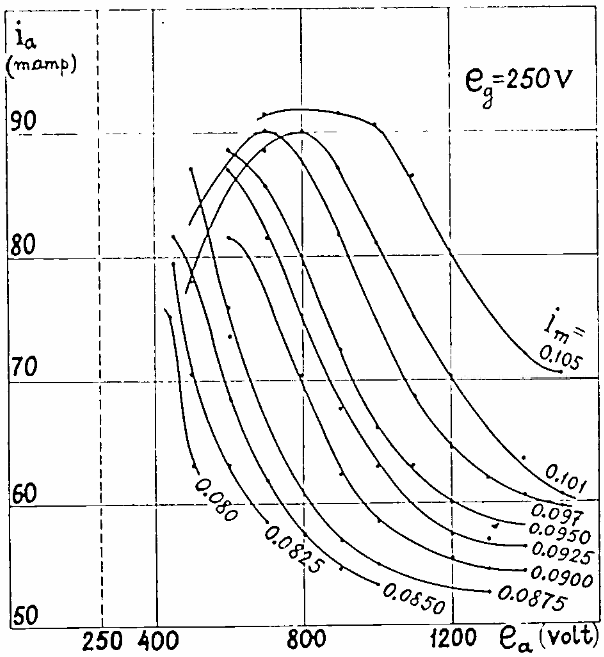

Fig. 8

The anode and the grid were made of nickel, the cathode of tungsten. The tube characteristics are shown in Figs. 3, 4, and 8. In the last figure we have the falling parts of the anode current characteristics for various magnetic field currents. As seen, it is possible to obtain, by suitable adjustment of the magnetic field, a minimum negative resistance of the order of 5000 ohms.

In the oscillatory conditions when operating with large anode voltage swing on the *LCR* circuit, a power of several watts could be obtained with this tube.

Acknowledgment

In conclusion, I wish to express my thanks to my assistants, Mr. S. Ryżko and Mr. Z. Jelonek, for their help in measurements, and to Mr. W. Gorka for the construction of the tubes.

BOOKLETS, CATALOGS, AND PAMPHLETS RECEIVED

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

Westinghouse electronic tubes are described in a catalog issued by the Westinghouse Lamp Company, Special Products Sales Dept., Bloomfield, N. J.

A general 1936 catalog has been issued by Shure Brothers of 215 W. Huron St., Chicago, Ill. Data sheet No. 101 covers three models of carbon microphones.

A data booklet on electrical steel sheets has been issued by the American Steel and Tin Plate Company, Pittsburgh, Pa.

Bulletin 8851 of the Ward Leonard Electric Company, Mt. Vernon, N. Y., covers a controlled rectifier to provide control of a direct-current motor from an alternating-current supply.

Heintz and Kaufman, South San Francisco, Calif., has issued data sheets on the following tubes: type 354, 150-watt, general purpose triode; 1554, 750-watt, general purpose triode; 3054, 1500-watt, general purpose triode; 55, 75-watt, gammatron; 155, 150-watt, general purpose triode; 255, 500-watt, gammatron.

Weston Electrical Instrument Corporation at 614 Frelinghuysen Ave., Newark, N. J., has issued a leaflet on tube base data connections and charts. Another leaflet covers its checkmaster broadcast receiver servicing equipment.

Engineering Bulletin CEB 36–7 on pentagrid converter oscillator considerations has been issued by the Ken-Rad Corporation of Owensboro, Ky.

The Insuline Corporation of America, 23 Park Pl., New York City, has issued a leaflet on its sectional standard construction rack.

Catalog No. 36 on radio coils and allied products has been issued by J. W. Miller Company of 517 S. Main St., Los Angeles, Calif.

Hygrade Sylvania Engineering News Letter No. 24 is on the subject of "Degeneration in Audio Amplifiers." Technical data sheets have been issued on the following tubes: 1F6, double diode pentode; 1E7G, double pentode power amplifier; 1F7G, double diode pentode; 5V4G, full-wave rectifier; 5X4G, full-wave rectifier; 5Y4G, full-wave rectifier; 6B4G, power amplifier; 6J5G, general purpose amplifier; 6K5G, high-mu triode; 6K6G, intermediate power pentode 6L6, power amplifier; 6N7, class B power amplifier; 6N7G, class B power amplifier; and 6R7G, double diode medium-mu triode. The issuing organization may be addressed at Emporium, Pa.

The Ken-Rad Corporation of Owensboro, Ky. has issued an Engineering Bulletin on "The 6L6 beam power amplifier."

National Union Laboratories, 365 Ogden Street, Newark, N. J., has issued three engineering reports on capacitance, converter operation, and current regulators.

Galvonometers and dynamometers manufactured by the Leeds and Northrup Company are covered in their Catalog ED1936 which may be obtained by addressing them at 4902 Stenton Ave., Philadelphia, Pa.

Earl Webber Company of 1217 Washington Blvd., Chicago, Ill., has issued a leaflet on its official radio service laboratory.

CONTRIBUTORS TO THIS ISSUE

Barton, Loy E. Born November 7, 1897, at Fayetteville, Arkansas. Received B.S. degree in E.E., University of Arkansas, 1921. Instructor, University of Arkansas, 1921–1925. Received E.E. degree, University of Arkansas, 1925. Student engineer, radio department, General Electric Company, 1925–1927. Associate professor, electrical engineering, University of Arkansas, 1927–1929. Tube problems and tube testing equipment, Radio Victor Company and RCA Victor Company 1929–1935; RCA Manufacturing Company, RCA Victor Division, 1935–1936; Philco Radio and Television Corporation, 1936. Member, Institute of Radio Engineers, 1929.

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* Paper published in May, 1936, issue of the PROCEEDINGS.

1923; assistant to Dr. L. W. Austin, radio transmission research laboratory, Bureau of Standards, 1923–1932; wave phenomena group, National Bureau of Standards, 1932 to February, 1936; Bureau of Yards and Docks, Navy Department, February, 1936 to date. Associate member, Institute of Radio Engineers, 1926.

Kirby, Samuel S.: See PROCEEDINGS for June, 1936.

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***Mjasoedoff, Nicolai A.:** Born 1902 at Moscow, U.S.S.R. Received E.E. degree, Moscow Technical High School, 1928. Engineer in charge, radio department, State Electrical Research Institute, Moscow, 1927–1933. Lecturer in Industrial Institute, Gorky, 1933 to date; engineer in charge of radio engineering, Gorky. Nonmember, Institute of Radio Engineers.

Morgan, Howard K.: Born February 22, 1906, at New York City. Received B.S. degree, University of California, 1929. Radio test course, General Electric Company, 1929–1930; special receiver engineering, RCA Victor Company, 1930–1931; test apparatus development engineering, International Resistor Company, 1931; receiver development and production engineering, Pilot Manufacturing Company, 1931; receiver development and aircraft experimental radio engineering, Transcontinental and Western Air, Inc., 1932 to date. Associate member, Institute of Radio Engineers, 1927.

Stromeyer, Charles F.: Born May 10, 1905, at Mount Vernon, New York. Engineering department, Northern Manufacturing Company, 1926–1927. Vacuum tube production engineer and later chief engineer, Cable Radio Tube Corporation, 1927–1932; chief engineer, Revelation Patents Holding Company, 1932 to date. Associate member, Institute of Radio Engineers, 1929.

* Paper published in May, 1936, issue of the PROCEEDINGS.

The Institute of Radio Engineers

Incorporated

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

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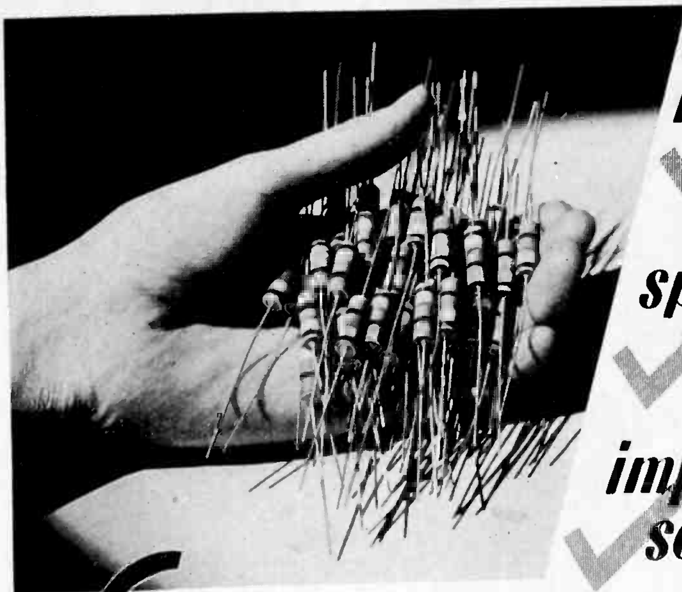

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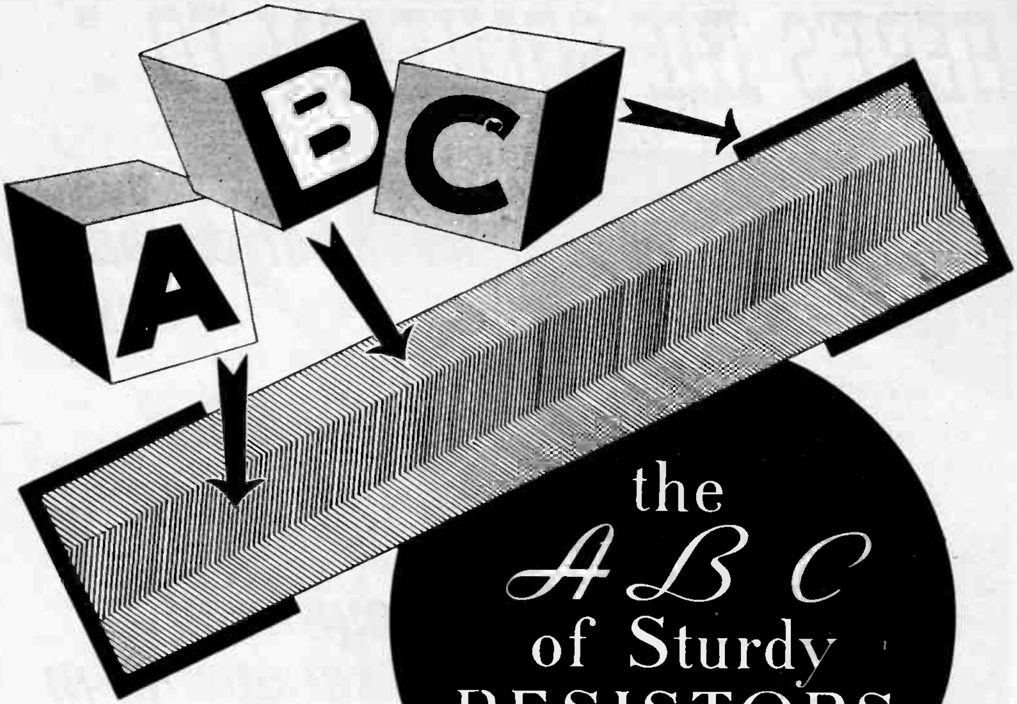

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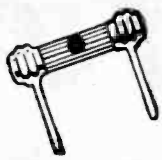

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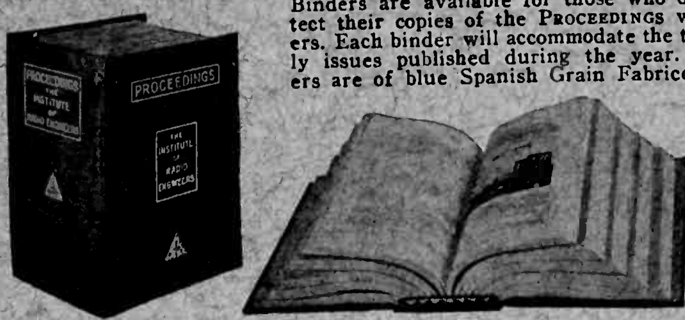

BACK ISSUES OF THE PROCEEDINGS

Complete volumes of back issues of the Proceedings are available in unbound form for the years 1918, 1921, 1922, 1923, and 1926 at $6.75 per year (six issues) to members. Also complete unbound for 1931, 1932, 1933, 1934, and 1935 at $7.50 per year to members. A list of the available miscellaneous copies published from 1913 on may be obtained from the Secretary. Single copies published prior to 1927 are sold to members at $1.13 each, and those issued during and after 1927 are $0.75 each. Foreign postage on single copies is $0.10 per copy, making $0.60 on unbound volumes before 1927.

Issues bound in blue buckram for 1922 (six numbers) are available at $8.75 to members. For 1930, 1931, and 1935 they are priced at $9.50 each to members. Foreign postage is $1.00 additional per volume. Volumes for 1922 and 1926 bound in Morocco leather are available at $10.25 each to members.

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An Index of the Proceedings for the years 1909-1930, inclusive, is available at $1.00 per copy. It is extensively cross indexed. The earlier references are to the Proceedings of the Wireless Institute; the Proceedings of the Institute of Radio Engineers has been published since 1913.

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