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PROCEEDINGS
of
The Institute of Radio Engineers

1929 CONVENTION
Washington, D. C.
May 13-15

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General Information and Subscription Rates on Page 198

Institute of Radio Engineers

Forthcoming Meetings

FOURTH ANNUAL CONVENTION
Washington, D. C., May 13-15, 1929

ATLANTA SECTION
Atlanta, Georgia, February 8, 1929

CLEVELAND SECTION
Cleveland, Ohio, February 15, 1929

NEW YORK MEETING
New York, N. Y., February 6, 1929 and March 6, 1929

PHILADELPHIA SECTION
Philadelphia, Penna., February 19, 1929

PITTSBURGH SECTION
Pittsburgh, Penna., February 21, 1929

TORONTO SECTION
Toronto, Canada, February 20, 1929

WASHINGTON SECTION
Washington, D. C., February 14, 1929

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 17 February, 1929 Number 2

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PART I

The Institute of Radio Engineers

GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains papers and discussions thereon submitted for publication or for presentation before meetings of the Institute or its Sections. Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

Subscription rates to the PROCEEDINGS for the current year are received from nonmembers at the rate of $1.00 per copy or $10.00 per year. To foreign countries the rates are $1.10 per copy or $11.00 per year.

Back issues are available in unbound form for the years 1918, 1920, 1921, 1922, 1923, and 1926 at $9.00 per volume (six issues) or $1.50 per single issue. Single copies for the year 1928 are available at $1.00 per issue. For the years 1913, 1914, 1915, 1916, 1917, 1918, 1924, and 1925 miscellaneous copies (incomplete unbound volumes) can be purchased for $1.50 each; for 1927 at $1.00 each. The Secretary of the Institute should be addressed for a list of these.

Discount of twenty-five per cent on all unbound volumes or copies is allowed to members of the Institute, libraries, booksellers, and subscription agencies.

Bound volumes are available as follows: for the years 1918, 1920, 1921, 1922, 1923, 1925, and 1926 to members of the Institute, libraries, booksellers, and subscription agencies at $8.75 per volume in blue buckram binding and $10.25 in morocco leather binding; to all others the prices are $11.00 and $12.50, respectively. For the year 1928 the bound volume prices are: to members of the Institute, libraries, booksellers and subscription agencies, $9.50 in blue buckram binding and $11.00 in morocco leather binding; to all others, $12.00 and $13.50, respectively. Foreign postage on all bound volumes is one dollar, and on single copies is ten cents.

Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership, etc., are priced at seventy-five cents per copy per year.

Contributors to the PROCEEDINGS are referred to the following page for suggestions as to approved methods of preparing manuscripts for publication in the PROCEEDINGS.

Advertising rates to the PROCEEDINGS will be supplied by the Institute's Advertising Department, Room 802, 33 West 39th Street, New York, N. Y.

Changes of address to affect a particular issue must be received at the Institute office not ater than the 15th of the month preceding date of issue. That is, a change in mailing address to be effective with the October issue of the PROCEEDINGS must be received by not later than September 15th. Members of the Institute are requested to advise the Secretary of any change in their business connection or title irrespective of change in their mailing address, for the purpose of keeping the Year Book membership catalog up to date.

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.

All correspondence should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York, N. Y., U. S. A.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered, and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer de visions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^3/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportions.

Abbreviations—Write a.c. and d.c., kc, μf, μμf, emf, mh, μh, henries, abscissas, antennas Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right, in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

Publication of Paper

Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

Reprints—With the notification of acceptance of paper for publication reprint order form is sent to the author. Orders for reprints must be forwarded promptly as type is not held after publication.

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L. W. Austin

President of the Institute of Radio Engineers, 1914

L. W. Austin was born at Orwell, Vermont, October 30, 1867. He received the degree of Bachelor of Arts from Middlebury College in 1889, and the Ph.D. degree from the University of Strassburg in 1893. From 1893 to 1901 he was instructor and assistant professor at the University of Wisconsin, following which he did research work at the University of Berlin in 1901–1902. Since 1904 Dr. Austin has been with the Bureau of Standards, Washington, D. C.; head of U. S. Naval Radio Research Laboratory, 1908–1923; chief of Radio Physics Laboratory, 1923 to date.

Dr. Austin was President of the Institute in 1914, served on the Board of Direction from 1915 to 1917, and was awarded the Institute Medal of Honor in 1927. His contributions to the Proceedings have been frequent. Dr. Austin was elected to Associate membership in the Institute in 1913, transferred to Member grade in 1913 and to the Fellow grade in 1915.

INSTITUTE NEWS AND RADIO NOTES

Fourth Annual Convention of the Institute

Washington, D. C., May 13th to 15th

The tentative program for the 4th Annual Institute Convention, to be held in Washington, D. C. on May 13 to 15th has been arranged. The Convention Committee Chairmen are as follows: C. B. Jolliffe, Convention Chairman; T. Mc. L. Davis, Registration and Arrangement; S. S. Kirby, Trips; F. P. Guthrie, Dinner and Entertainment; Mrs. L. W. Austin, Ladies; A. E. Kennelly, Fellowship; W. G. H. Finch, Publicity.

Convention Headquarters will be in the Mayflower Hotel where a number of rooms are being set aside for reservation by Institute members. Registration, Committee meetings, several functions for the ladies, the banquet, etc., will be held in the Mayflower, which is conveniently located with respect to the U. S. Chamber of Commerce Building in which the technical sessions are to be held, and can be reached readily either by auto, streetcar or bus.

The tentative program is as follows:

SUNDAY, MAY 12

2 P.M. to 6 P.M.—Registration at the Mayflower Hotel.

MONDAY, MAY 13

8:00 A.M.	Registration at the Mayflower Hotel.
10:00 A.M.	Opening Session; Speeches by President A. Hoyt Taylor, F. P. Guthrie and C. B. Jolliffe; Symposium on "Technical Problems of Radio Regulation."
11:00 A.M.	Meeting of representatives of Sections.
12:30 P.M.	Lunch.
1:15 P.M.	Inspection Trip leaving for Naval Research Laboratory, Bellevue, Anacostia, D.C., returning to the Mayflower Hotel at 5:15 P.M.
8:00 P.M.	Popular Lecture by Professor M. I. Pupin.

TUESDAY, MAY 14

9:00 A.M.	Symposium on Photo Radio.
10:30 .M.A	Inspection trip to Arlington Radio Station via Potomac Park, Arlington, and Arlington Cemetery.
12:30 P.M.	Lunch.

1:15 P.M.	Inspection trip to Bureau of Standards via 16th Street and Rock Creek Park.
4:15 P.M.	Short speech by Director of Bureau of Standards.
4:30 P.M.	Tea at the Bureau of Standards, returning to Headquarters at 5:30 P.M.
7:30 P.M.	Banquet at the Mayflower Hotel, including speeches by President A. Hoyt Taylor, several Institute members, presentation of Institute prizes, talking moving picture feature, and facilities for dancing.

Wednesday, May 15

Annual meeting of the American Section, International Scientific Radio Union, to which members of the Institute are invited. This meeting will consist of some twenty technical papers, averaging fifteen minutes each. The papers will cover developments in the fields of work of the Union's technical Committees. Besides the technical papers on particular developments, there will be a general report on the status of the field by the chairman of each technical committee, as follows: radio measurements, J. H. Dellinger; measurements of interference, E. F. W. Alexanderson; wave propagation, L. W. Austin; wave direction, G. Breit; phenomena above 3000 kilocycles, A. H. Taylor; atmospherics, H. T. Friis; cooperation, A. E. Kennelly; radio physics, E. L. Chaffee.

The program of specific papers in these fields will be published in a later issue. It is expected that preprints of the papers will be available in advance of the meeting.

Change in Proceedings Format

With the January issue the arrangement of PROCEEDINGS pages was changed with a view to making the various departments of each issue more readily available to members. With the April issue it is contemplated that the *type* size of each page will be increased to allow approximately fifteen per cent additional reading matter to be printed on each page.

Comments from the membership, both as to the typographical arrangement of the PROCEEDINGS and as to its contents, will be appreciated by the Board of Direction. Constructive criticism is always welcomed in the endeavor to make the PROCEEDINGS as attractive as possible to all grades of membership.

January Meeting of the Board of Direction

The following were present at the meeting of the Board of Direction of the Institute held on January 2nd in the Institute office: Alfred N. Goldsmith, President; L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Ralph Bown, Junior Past President; Arthur Batcheller, W. G. Cady, J. H. Dellinger, R. A. Heising, J. V. L. Hogan, R. H. Marriott, and John M. Clayton, Secretary.

The count of the ballots for 1929 Officers and new Board members brought the following results: President, A. Hoyt Taylor; Vice-President, Alexander Meissner; Managers elected for three-year terms, Arthur Batcheller and C. M. Jansky, Jr.

The Board appointed the three Managers with one-year terms as follows: Lewis M. Hull, R. H. Marriott, and L. E. Whittemore.

The Board approved the action of the Committee on Admissions through the election or transfer of the following: elected to the Fellow grade: Lt. Cmdr. T. A. M. Craven; transferred to the Member grade: Avery G. Richardson, Harvey Meisenheimer, and L. W. Branch; elected to the Member grade: Knox C. Black, W. B. Morehouse, and Albert Kofes.

Seventy-eight Associate members and eight Junior members were elected.

Paul A. Greene, of the Columbia Broadcasting System, was appointed to membership on the Committee on Broadcasting.

Notices of New York Meetings

As authorized by the Board of Direction, and effective immediately, notices of regular New York meetings of the Institute will, in the future, be sent only to members residing within the states of Connecticut, Delaware, Maryland, Massachusetts, New Jersey, New York, Pennsylvania, and District of Columbia, unless specific request be made once a year by individual members in other states.

1928 Bound Volumes

The 1928 volume of the PROCEEDINGS is available in a handsome blue buchram binding at $9.50 (one dollar additional for foreign postage). Only a limited number of volumes have been bound. Members desiring to obtain the 1928 volume in bound form are also cautioned that unbound copies for the entire year are *not* available.

The above prices are to members of the Institute.

Standard Frequency Transmissions by the Bureau of Standards

The Bureau of Standards announces a new schedule of radio signals of standard frequencies, for use by the public in calibrating frequency standards and transmitting and receiving apparatus. This schedule includes many of the border frequencies between services as set forth in the allocation of the International Radio Convention of Washington which went into effect January 1, 1929. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to 1,000 miles from the transmitting station.

The transmissions are by continuous-wave radiotelegraphy. The modulation which was previously on these signals has been eliminated. A complete frequency transmission includes a "general call" and " standard frequency" signal, and "announcements." The "general call" is given at the beginning of the 8-minute period and continues for about 2 minutes. This includes a statement of the frequency. The "standard frequency signal" is a series of very long dashes with the call letter (WWV) intervening. This signal continues for about 4 minutes. The "announcements" are on the same frequency as the "standard frequency signal" just transmitted and contain a statement of the frequency. An announcement of the next frequency to be transmitted is then given. There is then a 4-minute interval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequency points are received, persons can obtain as complete a frequency meter calibration as desired by the method of generator harmonics, information on which is given in the letter circular. The schedule of standard frequency signals is as follows:

Radio Transmissions of Standard Frequency; Schedule of Frequencies in Kilocycles

Eastern Standard Time	March 20	April 22	May 20	June 20	July 22
10:00–10:08 P.M.	1500	4000	125	550	1500
10:12–10:20	1700	4500	150	600	1700
10:24–10:32	2250	5000	200	700	2000
10:36–10:44	2750	5500	250	800	2300
10:48–10:56	2850	6000	300	1000	2700
11:00–11:08	3200	6500	375	1200	3100
11:12–11:20	3500	7000	450	1400	3500
11:24–11:32	4000	7300	550	1500	4000

Report of Federal Radio Commission

The annual report of the Federal Radio Commission to the Congress of the United States for the year ending June 30, 1928, together with a supplementary report for the period from July 1, 1928 to September 30, 1928, can be procured from the Superintendent of Documents, Government Printing Office, Washington, D. C., for twenty-five cents per copy.

The report is a work of some two hundred and sixty pages and is divided into three parts; part 1 contains a statement of the personnel and organization of the Commission; part 2 is devoted to the Commission's work with regard to the broadcast band; part 3 is devoted to low and high-frequency bands. The report also contains copies of all General Orders issued by the Commission between July 1, 1927 and October 26, 1928.

Institute Meetings

Atlanta Section

The Atlanta section held a meeting in the Chamber of Commerce Building, Atlanta, Ga., on January 9th. Major Van Nostrand, chairman of the section, presided.

A paper, "Modern Merchandising methods in the Radio Field," was presented by Pierre Boucheron, of the Radio Corporation of America.

Seventeen members of the section attended the meeting.

The February 8th meeting will be held in the Chamber of Commerce Building. At this meeting J. K. Clapp, of the General Radio Company, will present a paper "A Convenient Method for Referring Secondary Standards to a Standard Time Interval."

Boston Section

A meeting of the Boston section was held in the Cruft Laboratory, Harvard University, Cambridge, Mass., on December 12th. This was a joint meeting with the Boston Signal

Post. Dr. G. W. Pierce and D. S. Boyden presided. Major Richard H. Ranger, of the Radio Corporation of America, presented a paper, "Recent Developments in Photoradio." Messrs. Sewall Cabot, A. F. Murray, P. S. Bauer, H. W. Lamson, and R. W. Hall participated in the discussion of the paper.

One hundred and sixty-two persons were present at the meeting.

Buffalo-Niagara Section

A meeting of the Buffalo-Niagara section was held on November 14th in Edmund Hayes Hall, University of Buffalo. L. C. F. Horle, chairman of the section, presided.

John F. Morrison, of Radio Station WKBK, presented a paper, "The Modulation System of WKBK."

Edward Stanko, of Station WGR, presented a paper, "Broadcast Antenna Characteristics," and Francis D. Bowman presented a paper, "Present Day Broadcasting from the Advertiser's Viewpoint."

Forty members of the section attended the meeting.

At the December 12th meeting of the Buffalo-Niagara section D. E. Replogle presented a paper, "Television." Messrs. Horle, Hector, Johnson, Smith, Freck, and Henderson participated in the discussion which followed.

Sixty members of the section attended the meeting.

On January 17th L. Grant Hector, of the University of Buffalo, will present a paper, "Apparent Equality of Loudspeaker Output at Various Frequencies."

Chicago Section

A meeting of the Chicago section was held January 11th at the Electric Club, Chicago. John H. Miller, chairman of the section, presided.

Fred H. Schnell, of the Burgess Battery Company, presented a paper, "Short Wave Aircraft Radio." The paper explained the work of the Burgess Battery Company on dry cell operated transmitters and receivers for aircraft. A rather detailed analysis was given of the requirements for equipment for radio service on airplanes, and it was shown that this varied to a wide extent. Actual tests were given of light weight transmitters. A description of the apparatus installed on the Rockford Plane, which flew to Greenland, was given. Lantern slides covering this paper were shown.

Following the presentation of the paper Messrs. Oxner, Miller, Wilcox, Minnium, and Kennedy participated in the discussion.

The new officers of the Chicago section were elected as follows: H. E. Kranz, chairman; B. J. Minnium, vice-chairman; John H. Miller, secretary-treasurer.

Cleveland Section

A meeting of the Cleveland section was held in the Case School of Applied Science in Cleveland, Ohio, on December 14th. Professor John R. Martin, chairman of the section, presided and presented a paper on "The Use of the Phonodeik in the Study of Loud Speakers and the Electric Pick-up." The talk was illustrated with lantern slides and was accompanied by a demonstration of a special model of the phonodeik. Professor Martin emphasized the difficulty in making tests by ear, as the personal factor of likes and dislikes affects the observer. The advantage of a sound photographing device in comparing tones throughout the audible range lies in its ability to actually record the air motion or its electrical counterpart at any point in its transmission. The direct method eliminates the personal element in studying sound reproducing, amplifying, and pick-up devices.

Extensive discussion followed the presentation of the paper. The report of the Nominating Committee for 1929 officers of the Cleveland section was unanimously adopted. These officers are: B. W. David, chairman; Ralph Farnham, vice-chairman; D. Schregardus, secretary-treasurer; D. M. Ward, chairman of Committee on Membership; Prof. J. R. Martin, chairman of Program Committee.

Sixty-eight members of the section attended the meeting.

Connecticut Valley Section

A meeting of the Connecticut Valley section was held on January 18th. K. S. Van Dyke, vice-chairman of the section, presided. J. K. Clapp, of the General Radio Company, presented a paper on "Short Wave Short Distance Radio Transmission."

The paper dealt with the electronization of the atmosphere and ideal paths of transmission. Data of a series of tests run between Boston and South Dartmouth, Mass., working on various wavelengths around 40 meters which gives a very logical explanation of the so-called "skip distance," "flutter zone," and other short-wave phenomena were presented. Results of tests made with an antenna which could be rotated or placed at different angles with the earth were shown.

Messrs. Laport, Van Dyke, and others participated in the discussion which followed.

The officers for 1929 were elected and are as follows: Q. A. Brackett, chairman; E. A. Laport, vice-chairman; F. C. Beekley, secretary-treasurer.

Thirty-seven members of the section were present at this meeting.

Detroit Section

A meeting of the Detroit section was held on November 23rd in the Detroit News Building.

Earle D. Glatzel, chairman of the section, presided. Professor Roy S. Glasgow, of Washington University, St. Louis, Mo., presented a paper on "Some Recent Developments in Broadcast Receiving Set Design."

The paper discussed factors affecting selectivity, sensitivity, and fidelity of receiving sets, after which a new type of receiver employing a band pass filter tuning system followed by a fixed transformer coupled radio-frequency amplifier was described. Design of the filter to give a constant width of received band throughout the broadcast band range was discussed. The radio-frequency amplifier utilizes variation of the input capacitance of a vacuum tube with the plate circuit reactance automatically tuning the amplifier circuits to the frequency being received.

Forty-five members of the Institute and guests were present at this meeting.

The Detroit section held its December meeting on the 21st of the month in the Detroit News Building, 615 Lafayette Blvd. E. D. Glatzel, chairman of the section, presided.

C. G. Hall, of the Cardon Corporation of Jackson, Michigan, spoke on "Manufacturing Production of Vacuum Tubes." A detailed description of the processes involved in turning out vacuum tubes at the rate of 10,000 a day was given, together with an explanation of methods and apparatus used in manufacturing the glass parts of the tube, making and assembling the elements, degassing, sealing, and evacuating, and the final testing of the product.

Los Angeles Section

On November 19th a meeting of the Los Angeles section was held in the Elite Cafe, Los Angeles. Don C. Wallace, chairman of the section, presided. Lt. Commander Lowell Cooper, U.S.N., presented a talk on "Naval Radio Procedure and Practice." The address covered the general training, routine, and practice employed in naval radio control as well as the method of transmitting messages from different squadron groups to the flag ship. A superficial description of the types of radio equipment used was also given.

Lieutenant Dean Farrand, U. S. Army reserve, presented a talk on "Aircraft Beacons and Radio Equipment" in which were described the radio beacons in use along the coast and in Hawaii. Blackboard diagrams of their field distribution were given, and the speaker described some of the difficulties and problems of commercial aviation and the limitations imposed on radio equipment. The necessity for visual rather than oral indicators for the pilot was stressed.

W. S. Halstead, U. S. Forestry Service, gave a talk on "Radio as Applied to Fire Control and Forestry." By means of lantern slides a number of interesting views by a combined radio transmitter and receiver suitable for portable service in forestry and fire control work were shown. The talk included various technical data descriptive of the radio equipment and operating methods.

Forty-six members of the section attended the meeting.

The following officers were elected for the coming season: Thomas F. McDonough, chairman; Theodore C. Bowles, vice-chairman; W. W. Lindsay, secretary-treasurer.

The section's Board of Directors was elected as follows: Don C. Wallace, Theodore C. Bowles, Captain F. E. Pierce, Robert B. Parrish, Lee Yount, W. W. Lindsay, and Thomas F. McDonough.

New Orleans Section

A meeting of the New Orleans section was held on December 29th, presided over by Pendleton E. Lehde, chairman.

Anton A. Schiele presented a paper, "Control of Street Lights by Means of Tuned Circuits." Messrs. Jones, Gallo, and Mackie participated in the discussion which followed. Twelve members of the New Orleans section attended the meeting.

The New Orleans section held a meeting January 10th at which Chairman Pendleton E. Lehde presided.

Professor C. M. Jansky, Jr. read a paper entitled "Location of Salt Domes by Geo-Physical Explorations." The following members took part in the discussion that followed: Professor Ricker, J. N. DuTreil, and George Deiler.

Forty members and guests attended the meeting.

On February 11th a meeting of the New Orleans section will be held at which time J. K. Clapp, of the General Radio Company, will present a paper on "A Convenient Method for Referring Secondary Standards to a Standard Time Interval.

New York Meeting

A New York meeting of the Institute was held on January 2nd in the Engineering Societies Building, 33 West 39th Street, presided over by Alfred N. Goldsmith. V. Zworykin, of the Westinghouse Electric and Manufacturing Company, presented a paper on "Facsimile Picture Transmission." This paper will be published in an early forthcoming issue of the Proceedings.

Following its presentation the following participated in the discussion: Messrs. Goldsmith, McCullough, Gallagher, Bonn, Nyman, Ballantine, Hagglung, Dewhirst, Shannon, and Ranger.

Three hundred members of the Institute and guests attended this meeting.

Pittsburgh Section

The January 15th meeting of the Pittsburgh section was held in the Fort Pitt Hotel. L. A. Terven, vice-chairman of the section, presided. J. G. Allen and A. Mag presented "A Symposium on Radio Interference."

The paper showed how the source of noise is traced from the high voltage down to the original house current. It was stated that less than fifteen per cent of the radio interference cases reported were due to public utility companies. Electrical appliances are responsible to a great degree for the interference caused. It was also stated that appliances should be made "interference proof" at the factory. A demonstration of "noise makers" was given including universal motors, heating pads, tree grounds, etc. Messrs. Froelich, McKinley, Horn, and Terven participated in the discussion which followed.

Twenty members of the section attended the meeting.

On February 21st there will be a meeting of the Pittsburgh section in the Fort Pitt Hotel. J. K. Clapp, of the General Radio Company, will present a paper on "A Convenient Method for Referring Secondary Standards to a Standard Time Interval."

The Pittsburgh section held a joint meeting with the Pittsburgh section of the American Institute of Electrical Engineers in the Chamber of Commerce Building on December 11th.

H. A. Iams presented a paper on "Facsimile Picture Transmission."

Twenty-one members of the Institute were present.

Rochester Section

The Rochester section held a meeting on November 23rd in the Rochester Chamber of Commerce Building. A. B. Chamberlain, chairman of the section, presided.

Three minute speeches were made by J. P. Boylan on "Telephony"; F. W. Reynolds on "Broadcasting"; C. E. K. Mees on "Research"; A. T. Haugh on "Merchandising"; C. L. Cadle on "Power."

L. B. F. Raycroft, vice-president in charge of Radio Division of the National Electrical Manufacturers' Association, talked on "The Radio Industry."

One hundred and thirty-two members attended the meeting.

On December 7th a meeting of the Rochester section was held in the Sagamore Hotel, Rochester. A. B. Chamberlain, chairman of the section, presided.

The meeting was addressed by Conan A. Priest, of the General Electric Company, on the subject of "Short Waves." The speaker presented a summary of recent progress in the development and use of short waves explaining the various operating and economic advantages of their use. This was a joint meeting between the Rochester section of the American Institute of Electrical Engineers, the Rochester Engineering Society, and the Rochester section of the Institute of Radio Engineers.

Sixty-three members of these societies attended the meeting.

San Francisco Section

On November 21st a meeting of the San Francisco section was held in the Bellevue Hotel, 505 Geary Street, San Francisco. Leonard F. Fuller, chairman of the section, presided.

Major Henry L. Dunn, Signal Corps Reserve, presented a talk on "The Radio Intelligence Service in the World War."

Major Dunn spoke of many phases of the Intelligence Service of the U. S. Army during the late war. The means of obtaining information was given, the source of information being from spies, enemy prisoners, airplane observers, and observers using electrical methods. The use to which the information was put in keeping track of enemy forces was outlined. Methods of securing information directly from German telephone lines by actual contact, induction, and grounded lines with low-frequency amplifiers was described in detail. The development of radio compass intercepts stations in locating secret transmitters was mentioned. The talk was illustrated with lantern slides and was made interesting by an account of many humorous incidents.

Twenty-nine members and guests attended the meeting.

Seattle Section

A meeting of the Seattle section was held on November 30th in the Washington Engineer's Club, Arctic Building, Seattle. Walter A. Kleist, chairman of the section, presided.

Austin V. Eastman presented a paper on "Four-Electrode Tube as an Audio Amplifier." The speaker outlined some of the short-comings of three-electrode tubes as audio-frequency amplifiers and showed the need for the four-electrode type. He briefly outlined the use of the 222-type tube as a radio-frequency amplifier and next discussed this tube as an audio-frequency amplifier. Characteristic curves were drawn and a description of laboratory measurements of the overall voltage gain of a stage of impedance coupled amplification using the 222-tube as a space charge tube was given. Curves were presented showing the gain as the frequency was varied from less than a hundred cycles to six thousand cycles. The limit of input voltage with a consideration of distortion effect was explained.

Messrs. Tolmie, Kleist, and Willson discussed the paper.

Thirty-two members of the section attended this meeting.

Toronto Section

The November meeting of the Toronto section was held in the Electrical Building, University of Toronto, on the 23rd of the month, presided over by V. G. Smith.

R. H. Langley, of the Crosley Radio Corporation of Cincinnati, Ohio, presented a paper on "First National Broadcasting Station."

The paper described in some detail the new 50 kw station placed in operation in October of 1928 by the Crosley Radio Corporation. Description of the transmitter and the antenna lay-out was given. Following the presentation of the paper Messrs. Lowry, Hackbusch, Smith, Pipe, and others participated in its discussion.

On January 16th the Toronto section held a meeting in the Electrical Building, University of Toronto. A. M. Patience, chairman, presided.

J. M. Thompson, of the Ferranti Electric Company, presented a paper on "Transient Conditions in Output Transformers." Following the presentation of the paper Messrs. Cline Hackbush, Bayley, and others participated in its discussion.

Forty-five members of the section were present at this meeting.

On February 20th there will be a meeting of the Toronto section in the Electrical Building, University of Toronto, at which time Mr. Thompson, of the Marconi Company, will present a paper on "Marconi Beam System."

A meeting of the Toronto section was held in the Electrical Building, University of Toronto, on December 19th, presided over by A. M. Patience.

Lieut. W. L. Laurie, Royal Canadian Signals, presented a talk on radio activities of Hudson Straits Expedition which was sponsored by the Canadian Government in July of 1927 and returned in the fall of 1928. The members of the expedition established three bases along the coast, using radio to keep them in touch with civilization at Ottawa. Three aeroplanes were used in the expedition and were equipped with lightweight radio sets built by the Air Force Laboratories in Ottawa. The sets were operated by dry batteries for filament supply and wind driven generators for high voltages. Both telephony and telegraphy were used with a maximum radius varying between 300 and 900 miles. Messrs. Pollack, Choat, Lowry, and others participated in the discussion which followed.

Forty-five menbers of the section were present at the meeting.

Washington Section

On December 13th a meeting of the Washington section was held in the Continental Hotel, North Capitol Street, N.W., Washington, D. C. F. P. Guthrie, chairman of the section, presided.

E. O. Hulburt, of the Naval Research Laboratory, presented a paper, "Recent Developments in the Theory of High Atmospheres of the Earth," which included a brief summary of four papers which are to be published on the following subjects: "A Theory of Auroras and Magnetic Storms," "The Ion Ring Around the Earth at a Distance of about 45,000 Kilometers," and "Comets and Magnetic Storms."

Following the presentation of the paper A. H. Taylor, H. G. Dorsey, August Hund, W. H. Crew, G. D. Robinson, F. P. Guthrie, W. B. Burgess, and Charles Speaker participated in its discussion.

Preceding the meeting a dinner was held at which thirty-nine members and guests were present. The meeting was attended by seventy members of the section.

The Washington section held a meeting in the Continental Hotel, Washington, D. C. on January 10th. F. P. Guthrie chairman of the section, presided.

L. A. Hyland, of the Aircraft Section, Naval Research Laboratory, presented a paper "The Elimination of Ignition Interference to Aircraft Radio Reception."

The paper gave an outline of the procedure followed in the past few years to eliminate interference from aircraft radio reception. The various forms of interference were pointed out and their effects and remedies discussed. The several forms mentioned consisted of wind noises, propellor whine, exhaust noises and motor vibration, all of which produced microphonic disturbances from vibration of the plane or the surrounding air, and finally the ignition interference which was considered as the worst offender. The use of tuned audio systems were recommended for reducing the microphonic disturbances and shielded ignition systems for ignition interference elimination. Particular reference was made to recent shielded spark plug development, which has resulted in a design that not only makes shielding practical but embodies features that should increase the life of the plug with a possible increase in ignition efficiency.

Messrs. Dorsey, Price, Robinson, Guthrie, Burgess, Mirick, and Jackson participated in the discussion which followed.

Fifty-two members of the section attended this meeting.

The February 14th meeting will be held in the Continental Hotel and J. K. Clapp, of the General Radio Co., will present a paper "A Convenient Method for Referring Secondary Standards to a Standard Time Interval."

Committee Work

Committee on Standardization

During the first few months of 1928, the Committee on Standardization, L. E. Whittemore, chairman, continued as during the previous year, with the formulation of material by the subcommittees which had been organized. These subcommittees and their chairmen were as follows: Subcommittee on Vacuum Tubes,—L. A. Hazeltine, chairman, 1927, C. B. Jolliffe, chairman, 1928; Subcommittee on Receiving Sets,—J. H. Dellinger, chairman, 1927, E. T. Dickey, chairman, 1928; Subcommittee on Electro-Acoustic Devices,—R. H. Manson, chairman; Subcommittee on Circuit Elements,—H. M. Turner, chairman; Subcommittee on Power Supply,—W. E. Holland, chairman; Subcommittee on Use of the Transmission Unit,—J. V. L. Hogan, chairman; Subcommittee on Bibliography,—C. A. Wright, chairman.

The material prepared by these subcommittees up to May 25, 1928 was printed in a Preliminary Draft Report bearing that date. Notices of the copies of this Preliminary Report available at that time were printed in the PROCEEDINGS. Copies were also distributed at the conventions of the two associations of radio manufacturers held in Chicago, in June, and were sent to others who were likely to be able to give constructive comments.

Comments and criticisms of this report were compiled and sent to members of the committee for consideration prior to a series of meetings held in New York during the latter part of 1928. Meetings were held on October 2, October 16, November 8, and December 6, 1928.

At these meetings consideration was given to the material contained in the Preliminary Report and to the comments and additional material submitted for adoption. A committee on form and arrangement, consisting of Haraden Pratt, E. T. Dickey, and the chairman, was authorized to put the material adopted by the committee into final form for submission to the Board of Direction of the Institute. The scope of the report is indicated by the following items included in the table of contents of the preliminary draft:

Standard Graphical Symbols
Letter Symbols; Vacuum-Tube Notation
Standard Methods of Measuring the Important Characteristics of Vacuum Tubes
Performance Indices of Electro-Acoustic Devices
Standard Methods of Measuring Properties of Circuit Elements; Chart and Bibliography

The Institute had continued to have the cooperation, through membership on the committee, of representatives of the American Institute of Electrical Engineers, the radio division of the National Electrical Manufacturers' Association, and the Radio Manufacturers' Association. The policy formulated in 1927 under which the field of radio standardization has been divided between the Institute of Radio Engineers on the one hand and the manufacturers' associations on the other has been continued. Partly as a result of this policy, it is believed that there is a growing support of the work of the Institute by other branches of the radio industry.

Through interlocking committee personnel and by other means every effort has been made to take advantage of the results of related work done by other organizations and to avoid duplication of effort.

In accordance with the plan adopted by the Board of Direction it is anticipated that the final report will be published as a part of the 1929 Year Book to be issued early in 1929.

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED JANUARY 2, 1929

Elected to the Fellow grade

Dist. of Columbia	Washington, Naval Communications, Navy Dept	Craven, T. A. M.

Transferred to the Member grade

Michigan	Muskegon, 1390 Palmer Avenue	Richardson, Avery G.
New York	New York City, 195 Broadway Room 1607	Misenheimer, Harvey N.
Hawaii	Hilo, Box 923	Branch, L. W.

Elected to the Member grade

Massachusetts	Cambridge, 50 Kirkland Street	Black, K. Charlton
New Jersey	Bloomfield, 120 Berkeley Avenue	Morehouse, W. B.
Germany	Berlin-Charlottenburg, Cauerstrasse 19III	Kofes, Albert

Elected to the Associate grade

Arkansas	Camden, Box 661	Tagart, Sam W.
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California	Los Angeles, 947 Francisco Street	Heimberger, Albert E.
	San Francisco, 378 Golden Gate Ave., Apt. 235	Kelsey, Karl D.
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	Chicago, 2530 Kedzie Blvd.	Wilm, C. F.
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	Boston, 468 Massachusetts Ave.	Wass, Howard H.
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	Cambridge, 28 Gorham Street	Tsao, T. C.
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	Springfield, 38 Greenacre Square	Nystrom, Raymond A.
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Montana	Forsyth, Box 1064	Roberson, Carl
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	Buffalo, 605 Auburn Avenue	Perau, Frederic Henry
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	Gardenville, 447 Potter Road	Felmet, Albert
	New York City, c/o Norton Lilly, 26 Beaver Street	Berry, Harold C.
	New York City, 140 West 16th Street	Castaneda, Santiago
	New York City, 118 East 103rd Street	Doubles, David J.

	New York City, 41 West 86th Street	Feldstein, Martin A.
	New York City, c/o Norton Lilly, 26 Beaver	Francis, Philip
	Brooklyn, 326 Utica Avenue	Gottdenker, Martin
	New York City, R. C. A., 326 Broadway	O'Connor, John G.
	New York City, 195 Broadway	Reinken, Louis W.
	Rochester, 7 Edmonds Street	Steneri, Arthur John
	Schenectady, 848 Union Street	O'Neill, John P.
	Staten Island, USS *Acushnet*	Miller, Paul E.
North Carolina	Henderson	Woolard, E. W.
North Dakota	Fargo, 905-5th St., North, No. 19 Rust Apts	Cook, Tedd W.
Ohio	Cleveland, 7617 Myron Avenue	Crocus, John Stanley
	Cleveland, 3516 Storer Avenue	Irvine, Robert P.
	Marion, 177 East Center Street	Brown, David Ashton
	Zanesville, 606 Ridge Avenue	Bell, Lewis M.
Oklahoma	Oklahoma City, Oklahoma Gas & Electric Co	Bathe, C. E.
Oregon	Portland, 873 Belmont Street	McCargar, S. Harold
Pennsylvania	Allentown, 246 S. Madison Street	Bowman, Charles W.
	Allentown, 226 N. 6th Street	Haines, A. J. D.
	Allentown, 2014 Highland Avenue	Muthart, John A.
	Allentown, 949 Hamilton Street	Rauhofer, Frank
	Easton, 426 S. 21st Street	Clendaniel, John Edwin
	Easton, 18 N. 9th Street	Raesly, James B.
	Easton, 1815 Fairview Avenue	Weller, Everett Clare
	Philadelphia, 2209 South Chadwick Street	Johnson, Harmon
	Phialdelphia, 765 S. 10th Street	Mattia, Ralph F.
	Wilkinsburg, 215 Ross Avenue	Reynolds, C. C.
Rhode Island	Providence, 4 Pemberton Street	Brewster, O. H.
West Virgina	Charleston, 4 Maple Terrace	Moore, Thomas H.
Wisconsin	Milwaukee, 1667 Oakland Avenue	Hough, W. E.
	Milwaukee, 1405 Bremen Street	Strassman, Irving H.
British Guiana	Demerara, Georgetown, 61 Hadfield St	Tasker, Joseph Thomas
Chile	Valparaiso, Casilla 1653	Vierling, Gustav
England	Stockport, North Reddish, 446 Gorton Road	Howard, C. Alexander
India	Baroda, Babajipura	Dighe, K. S.
Ireland	60 Clifton Road, Bangor, Co. Down	Jamison, A.
Mexico	Puerto Mexico, Veracruz, Apartado, 86	Bourgeois, Allen B.

Elected to the Junior grade

Illinois	Champaign, 61 East Green Street	Pennington, D. J.
Indiana	Valparaiso, 712 Calumet Avenue	Clark, Edgar J.
	Valparaiso, 156 S. Franklin Street	Woodworth, Elwyn Crane
Massachusetts	Quincy, 189 Common Street	D'Alessandro, Genaro
Michigan	Royal Oak, 286 West Ten Mile Road	Alain, J. E.
New York	Brooklyn, 140 Vanderbilt Avenue	McGonigle, William J.
Oregon	Salem, 2405 Center Street	Poujade, Donald G.
Pennsylvania	Allentown, 38 S. Jefferson Street	Foley, W. R.

APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before March 1, 1929. These applicants will be considered by the Board of Direction at its March 6th meeting.

For Transfer to the Fellow grade

New York	New York City, 67 Broad Street	Pratt, Haraden

For Election to the Fellow grade

Italy	Rome, Via Tevere No. 20	Pession, Guiseppe

For Transfer to the Member grade

New Jersey	Boonton, Radio Frequency Laboratories	Loughlin, W. D.

For Election to the Member grade

Connecticut	Hartford, 221 Holcomb St.	Bourne, Roland B.
Oklahoma	Ponca City, 604 Marland Drive	Wyckoff, Ralph D.
England	Cornwall, Bodmin, Radio Beam Station	Struthers, G. A.

For Election to the Associate grade

Alabama	Mobile, c/o Adams Glass and Co., Inc.	Blakeney, George H., Jr.
Arkansas	Little Rock, 214 West 4th Street	Bilheimer, Joe Allen
California	Burbank, 320 No. Tujunga Avenue	Herrnfeld, Frank P.
	Glendale, 202 East Broadway	Mitchell, G. A.
	Monterey, 460 Alvarado Street	Newman, Clarence D.
	Oakland, 1609 Grand Avenue	Wild, Sidney J.
	Pasadena, 230 South Hudson Avenue	Taylor, Kenneth
	San Francisco, 328 Custom House	Kellogg, Frank L.
	San Francisco, 9 Duncan Street	Pinkney, J. L.
	San Pedro, USS California, Box 17, c/o Postmaster	Hart, Clyde Armand
	Stockton, 543 So. California Street	Green, Alfred H.
Connecticut	New Haven, 104 Lake Place	Gounarides, A. D.
	New Haven, 1523 Chapel Street	Rhoades, A. L.
	Waterbury, Y.M.C.A.	Knowlton, John J.
Dist. of Col.	Washington, Radio Laboratory, Bureau of Standards	Doherty, William H.
	Washington, U. S. Naval Research Laboratory	Wallace, James D.
Florida	Fort Meyers, 1138 Franklin Street	Johnson, Carl C.
	Jupiter, Box 158	Miller, Joseph Anthony
	Winter Park, Hamilton Hotel	Emerson, John
Georgia	Atlanta, 912 Candler Bldg.	Hobart, James C.
	Atlanta, Box 1593	McKinney, Joe H.
	Fort Benning, P.O. Box 615	Sosebee, Frank E.
Illinois	Brookfield, 126 So. Vernon Avenue	Courter, H. L.
	Chicago, 4849 No. Lawndale Avenue	Eppstein, Ralph M.
	Chicago, 4514 B. N. Ashland Avenue	Frederick, William J.
	Chicago, 1450 Argyle Street	Hitzenhammer, Anthony de V.
	Chicago, 4747 Kenmore	Riggs, John H.
	Wilmette, 1025 Sheridan Road	Freis, J. H.
Louisiana	St. Francisville, P.O. Box 232	Harris, David H.
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	Portland, 56 Lafayette Street	Giro, Daniel J.
Maryland	Baltimore, 1611 Lexington Bldg.	Mathers, Earl S.
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	Cambridge, 57 Gorham Street	Shen, S. Z.
	Northampton, 126 Franklin Street	Mettey, Bradford Joseph
	Quincy, 100 Bay View Avenue	Renton, Ralph James
	South Dartmouth, M. I. T., Round Hills	Houghton, Henry G., Jr.
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	Ann Arbor, 423 Fuller Street	Embury, Milburn S.
	Ann Arbor, 402 N. Main Street	MacCaffray, R. S., Jr.
	Ann Arbor, 403 S. Fifth Avenue	Martin, J. Faber
	Ann Arbor, 109 Packard Street	Pickett, Willis N.
	Ann Arbor, 1666 Broadway	Richmond, Clifford A.

Michigan (Cont'd)	Ann Arbor, 1108 Michigan Avenue	Wilber, Harold
	Birmingham, 33 Linden Road	Soeters, Raymond A.
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	Fordson, 5615 Horger Avenue	Knight, Donald M.
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	Keansburg, 389 Palmer Avenue	Herrmann, Edwin
	Orange, 459 Hepwood Avenue	Harris, Gwin C.
New York	Brooklyn, 400 Lincoln Place	Rojas, Fernando R.
	Jamaica, 12 Allen Street	Yenoli, Dominick J.
	New York City, City Island, Harlem Yacht Club	Appel, Henry W.
	New York City, 1160 Bryant Ave., Bronx	Atkin, Robert
	New York City, 2510 Davidson Ave., Bronx	Bond, Elmer Frederick
	New York City, 411 Fifth Avenue	Frank, James, Jr.
	New York City, R.C.A., 70 Van Cortlandt Park So.	Geoghegan, Eamonn
	New York City, 3985 Saxon Avenue, Bronx	Hall, John N.
	New York City, R.C.A. 70 Van Cortlandt Park So.	Hardin, L. L., Jr.
	New York City, 814 East 160th St., Bronx, c/o Rosenfeld	Hockner, Curtis
	New York City, 411 Fifth Avenue	Hulan, A. G.
	New York City, 2175 Walton Ave., Bronx	Knights, Alexander H.
	New York City, 117 West 89th Street	Lawrence, Walter L.
	New York City, 55 West 95th Street	Levy, A. Kingdon
	New York City, 26 Broadway, Room 560	Marshall, Floyd Wardell
	New York City, 50 Church Street, Room 1472	Patterson, John
	New York City, 40 Rector St., c/o British Marine Wireless Service	Verralls, J. Maynard
	Port Jefferson, L. I., 119 Thompson Street	Ernst, Murray C.
	Rensselaer Falls	Stiles, John L.
	Richmond Hill, L. I., 8640—110th Street	Brothers, James T.
	Riverhead, Box 1133	Stagg, Carl
	Rochester, 490 Ames Street	Melchiors, Frants Anagar
	Rochester, Columbus Apts., No. 605	Steneri, Arthur John
	Schenectady, General Electric Co., Radio Eng. Dept.	James, Wallace
	Schenectady, 337 Summit Ave.	Short, Donald William
	Syracuse, 117 Henry Street	Pierson, Theodore
	Whitestone Greens, 149-20—23rd Ave.	Lederhaus, Herman Wm.
	Woodhaven, L. I., 8637—77th St.	Stellwagen, Frank W.
Ohio	Athens, 291 East State Street	Green, Darrell B.
	Cincinnati, 3838 Columbia Ave.	Rice, Clarance
	Columbus, 90 West Northwood Ave.	Anderson, J. E.
	Toledo, 3255 Cottage Avenue	Sparaga, Leo. L.
Oklahoma	Lawton, 309 Dearborn St.	Newsom, Theo.
	Tulsa, 2044 East 12th Place	Murphy, Paul L.
	Tulsa, 504 Mary Brockman Apts.	Stinson, Lawrence W.
Pennsylvania	Allentwon, 1015 Allen Street	Heimbach, Charles W.
	Allentown, R. F. D. No. 2	Jones, Robert B.
	Allentown, 712 St. John St.	Pond, Edward W., Jr.
	Bangor, 28 Market St.	Bruschi, Lewis John
	Bath, Walnut Street	Graver, Frank S
	Bethlehem, 520 W. Broad St.	Hottel, C. W.
	Carnegie, No. 1 Rosslyn Road	Yoder, Leo E.
	Danville, 211 Church Street	Deaner, Haydn M.
	Darlington	Rohrmann, Edward R.
	Easton, 801 Cattell St.	Beans, Floyd L.
	Frackville, 320 Middle St.	Horwat, Peter
	Philadelphia, Phila. Storage Battery Co.	Hyatt, C. Brown
	Philadelphia, 64 Ashland Ave., W. Manayunk	Sweeney, Harold V.
	Pittsburgh, N. S., 316 North Ave., W.	Bower, Richard
	Pittsburgh, 5915 Alder St., E. E.	Rafferty, Charles H.
	Quakertown, 142 So. 3rd St.	Bartholomew, Robert G.
Texas	Sherman, 217 N. Travis St.	Morris, Truman S.
	Wichita Falls, 917 Scott St.	Ridling, Carrol W.
Washington	Everett, 2412 Summit Ave.	White, Keith
	Ferndale, Box 565	Legoe, Herbert S.
	Seattle, 801 Dexter-Horton Bldg.	Albert, J. H.
West Virginia	Wheeling, 502 So. Front St.	Thomas, W. Raymond
	Wheeling, 1229 Main St.	Weis, F. J.
British West Indies	Jamaica, Gregory Park P. O.	Morales, Seymour J.
Canada	Aurora, Ontario, C. F. R. B.	Sharp, Valentine
	Toronto, 4. 588 Bathurst St.	Brown, Alvin H.
	Toronto, 136 Shuter St.	Leonard, Sydney L.
	Toronto, 16 Chicora Ave.	Hope, Frederick C.
	Windsor, Ontario, 211 Bridge Ave.	Fisher, Albert E.
	Winnipeg, Manitoba, 421 Assiniboine Ave.	Jackson, Arthur

England	Burgess Hill, Sussex, Inholmes Park	Somerset, E. T.
	London, 67 Morley Road, West Ham	Beaven, Thomas G.
	London W1, Rothermel Radio Corp., 24 Maddox St.	Hannah, W. J.
	Portsmouth, 49, Wykeham Road, North End	Palmer, Reginald J
Japan	Osaki-machi, Teishinsho Denki Shikenjo	Nakai, Tomozo
	Osaki-machi, Teishinsho Denki Shikenjo	Namba, Shogo
Scotland	Gourock, Western House	Greenlees, Leslie M.
South Africa	Muizenberg, Cape, Main Road	Gilmour, Peter

For Election to the Junior grade

Florida	Lake Worth, P. O. Box 131	Dickins, Godfrey
	Lake Worth, P. O. Box 922	Lord, Wilfred F.
Michigan	Ann Arbor, 109 Packard St.	Harvey, Marvin
New York	Buffalo, 414 Northampton St.	Bergmann, Howard J.
Ohio	Cleveland, WHK, Engineers' National Bank Bldg.	Ostrander, Russell
Virginia	Roanoke, 206 Virginia Ave., Virginia Heights	Avery, Robert D.
Canada	Toronto, Ont., 86 Geoffrey St.	Knight, Frank
England	London N. 16, 48, Rectory Road	Goldstein, S.
New Zealand	Christchurch, Merivale, 43 Papanui Road	Wills, Syd, Jr.

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Part II

TECHNICAL PAPERS

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ON THE MECHANISM OF ELECTRON OSCILLATIONS IN A TRIODE*

By

H. E. Hollmann, Dr. Eng.

(Physical Institute of the Technical High School, Darmstadt, Germany)

Summary—*This paper deals with the different types of electron oscillations which occur in the Barkhausen and Kurz retarding field of a triode having a high positive potential applied to its grid, and zero or small negative potential applied to its anode. Four different frequency ranges were found as follows:*

(1) *In the retarding field between the electrodes there occur electron oscillations whose frequency in general increases with increasing field strength. The wavelengths can be calculated from the potentials and the tube dimensions under the assumption that the electrons vibrate about the grid: Barkhausen-Kurz oscillations.*

(2) *If there is an oscillation system connected to the tube electrodes, alternating fields are superposed on the steady fields produced by the d.c. voltages. As is theoretically shown in a simplified case, these tend to increase the frequency of the electrons and a building up process takes place, the final state of which is the natural frequency of the oscillation system: Gill and Morrell oscillations. Between both types of oscillations there is a constant transition region where the frequency follows neither the natural wave of the oscillation system nor the Barkhausen-Kurz laws.*

(3) *If in Barkhausen-Kurz oscillation the electrons vibrate about the grid* (1), *oscillations of higher frequency can appear with a grid of close mesh, which no longer occupy the whole anode-cathode space, but are confined to the anode-grid space. The frequency of a hypothetical oscillation in the grid-cathode space can be determined experimentally from the difference of the "longer" and "shorter" waves. This must remain constant for a constant grid-cathode space during a change of the grid-anode space, as was confirmed experimentally. Disturbances by Gill-Morrell oscillations* (2) *were suppressed by capacitive blocking of the electrode system. The "shorter" electron oscillations thus represent a special form of the Barkhausen-Kurz oscillations.*

(4) *From these "higher frequency electron oscillations," "higher frequency Gill-Morrell oscillations" can be produced in turn. The necessary high-frequency tuning of the electrode system is attained by half-wave excitation; thus with comparatively large electrodes 18 mm in external diameter, waves less than 20 cm long of considerable strength could be produced. A further reduction of the wavelength can only be obtained by reducing the dimensions of the electrodes.*

* Dewey decimal classification: R133. Original manuscript received by the Institute, August 16, 1928. Translation received, September 13, 1928. This paper summarizes the results of a recent publication of the author in *Ann. d. Phys.*, 86, 129, 1928.

Introduction

IN the generation of short electric waves by means of regenerative arrangements there is a lower limit occasioned by the fact that the time of transit of the electrons within the tube can no longer be ignored as compared with the period. Consequently the ordinary regenerative coupling conditions, as far as they concern the phase relations of the grid and anode alternating potentials, are no longer sufficient. For commercial tubes this lower limit is at about 1 meter, but the intensity of the oscillations is so slight as to be suitable only for measuring purposes.[1]

Shorter waves can be generated by the Barkhausen-Kurz method,[2] in which the frequency is mainly determined by the stay of the electrons in the inter-electrode space. In contrast with the ordinary method of operating the three-electrode tube, the grid potential is highly positive and the plate potential is negative. Because of the distribution of electric force inside the tube, there is a to and fro movement of the electrons about the grid, since those with high velocity that have passed through the meshes of the grid are reversed by the retarding field of the anode and are driven back to the grid; they then pass through it a second time and the process is repeated in the grid-cathode space.

A confirmation of the views of Barkhausen and Kurz is found in the fact that the wavelengths produced are determined only by the dimensions of the tube and the applied potentials, and are not influenced by the constants of the associated circuits. With the simplifying assumption of plane electrodes and equal electrode spacings and disregarding the space charge, Barkhausen and Kurz give the following formula for the wavelength:

$$\lambda = \frac{1000 d_a}{\sqrt{E_g}}$$

in which E_g is the grid potential in volts, d_a the diameter of the anode, while the potential of it, as well as that of the cathode, is zero. The dependence of the wavelength upon the determinative factors, electrode distances, and voltage is very well verified by experiment.

[1] Huxford, *Phys. Rev.*, 25, 686, 1925; Hollmann, *Radio-Umschau*, 177, 1927; Deubner, *Ann. d. Phys.* 84, 429, 1927; Bergmann, *Ann. d. Phys.*, 85, 961, 1928; Englund, Proc. I.R.E., 15, 914, November, 1927.

[2] Barkhausen and Kurz, *Zeits. f. Phys.*, 21, 1, 1920.

The theoretical considerations of Barkhausen and Kurz were extended to cylindrical electrodes by Scheibe[3] for the practical case under consideration. Under certain conditions he found two different frequencies and observed a good agreement of the "longer waves" with the theory. The assumption that the "shorter" wave represented a harmonic of the "longer" wave was contradicted by the experiment. Scheibe obtained a considerable increase in the strength of the oscillations by coupling to the tube an oscillation system comprising two parallel wires, shunted by a bridge, and connected to plates and grid, respectively (Fig. 1). According to Scheibe's results this external system has no effect on the wavelength, in agreement with the theory, but acts merely to increase the energy.

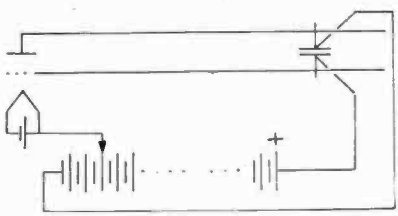

Fig. 1

The method of Barkhausen and Kurz has been used by various investigators for research, in the region of short waves, and in some of this work there have appeared small deviations from the theory, in that the wavelength is not always determined solely by the electrode potentials and spacing, but also depends upon the associated oscillation system. If the tuning system is directly at the tube electrodes as is shown in Fig. 1, its effect upon the electrode frequency is very great.[4,5,6,7,8] Two short wires, connected with the grid and anode as antennas, affect the wavelength as well as the determinative factors.[9,10,11] If such an antenna is coupled with a Lecher system for the measurement of the wavelength, an influence on the wavelength can still be

[3] Scheibe, *Ann. d. Phys.*, 73, 54, 1924.
[4] Gill and Morrell, *Phil. Mag.*, **44**, 161, 1922; **49**, 369, 1925.
[5] Grechowa, *Zeits. f. Phys.*, **35**, 50, 1926.
[6] Kapzov, *Zeits. f. Phys.*, **35**, 129, 1926.
[7] Sahanek, *Zeits. f. Phys.*, **26**, 368, 1925.
[8] Tank, *Helv. Phys. Acta*, 100, 1928.
[9] Bock, *Zeits. f. Phys.*, **31**, 543, 1925.
[10] Schaefer and Merzkirch, *Zeits. f. Phys.*, **23**, 166, 1923.
[11] Kapzov, l.c. 6.

detected since the resonance maxima are not fixed but "slide along" as the bridge is shifted.[12]

Gill and Morrell[4] in particular discovered oscillations whose frequency was given exclusively by the oscillation system connected with the tube electrodes and was not affected by the electrode potentials. Kapzov and Gwosdower[13] state that these oscillations and those of the Barkhausen-Kurz type can occur in one and the same arrangement.

I. Simultaneous Occurrence of Different Types of Electron Oscillations

1. Oscillations of the Types of Barkhausen-Kurz and of Gill-Morrell

In order to test the relationship observed by the different authors between the wavelength and the adjustment of the

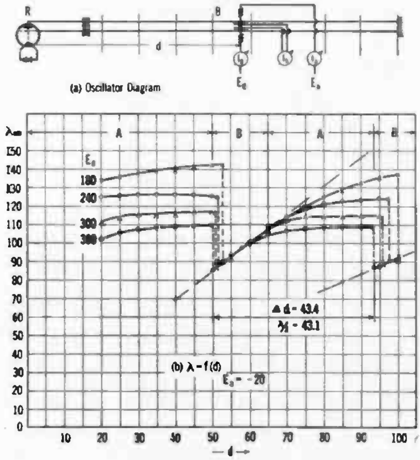

Fig. 2

oscillation system associated with the tube and also to check the discordant data of Barkhausen-Kurz and Scheibe, an experimental arrangement as described by Scheibe and by Gill and Morrell was used. The oscillator is shown diagrammatically in Fig. 2a. The high-frequency system consists of two wires, connected to grid and plate of tube R, along which the bridge B can be slid. In the bridge, capacitively insulated, is a thermocouple for measuring the oscillation current. It was calibrated

[12] Schriever, *Ann. d. Phys.*, 63, 645, 1920.
[13] Kapzov and Gwosdower, *Zeits f. Phys.*, 45, 114, 1927.

by comparison with a hot-wire amperemeter at a frequency of 10^6. Chokes are employed to minimize the disturbing effects of the leads.

The tube *R* is a Schott transmitter tube, and is especially suitable for the generation of Barkhausen-Kurz oscillations. A positive potential of a few hundred volts is applied to the grid while the plate receives a small potential which is negative with respect to the cathode. The waves generated in the oscillator are measured by means of a Lecher parallel wire system with a detector as indicator.

As the bridge *B* is moved along the wire, there occurs at certain points a sudden rise in the plate current which after passing through a maximum gradually returns to its original value. According to the results of Scheibe a rise in the oscillation intensity with resonance of the external oscillation circuit is to be expected; this must be followed by similar changes in the plate current since according to Barkhausen and Kurz this is entirely due to the oscillation of the electrons. The abrupt rise in the anode current as well as its lack of resemblance to a resonance curve leads us to suspect a complicated process.

As a matter of fact wavelength measurements showed that a considerable increase in frequency is associated with the rise of I_a, and also that with the gradual decrease in the anode current the wavelength again returns to its old value. The curves in Fig. 2b show this change in wavelength. They were taken by measuring the wavelengths for different positions of the bridge on the parallel wire system, and under different operating conditions. The anode potential was constantly -20 volts, while that of the grid was varied. As the heating current fluctuated slightly with a change in I_a and I_g, the emission current I_a+I_g was kept constant instead.

The curves indicate two completely different oscillation regimes, in region *A* one of lower frequency, and in region *B* one of higher frequency. The transition from *A* to *B* is always abrupt, but gradual in the reverse direction. We also see that it takes place sooner, that is, with a shorter natural wave of the tuning system the shorter the wave in the previous *A* region.

If the first transition takes place at a wire length *d* of 50–53 cm, the process repeats at $d=80$–100 cm. Evidently in the later *B* regions we have overtones of the parallel wire system.

The distinguishing characteristics of the two oscillation phenomena reside less in the different frequency regions than in the fact that the *A* oscillations are not influenced at all by the external system and their frequency depends only on the electrode potentials, whereas the frequency of the *B* oscillations is completely fixed by the oscillation system. The wavelengths of the *B* oscillations lie on straight lines, dotted in Fig. 2b, which represent the calibration curves of the parallel wire system.

It is clear from previous experiments that the *A* oscillations are pure electron oscillations as explained by Barkhausen and Kurz, that is, oscillations in which the electrons move to and fro in the stationary retarding field of the electrodes. The *B* oscillations likewise appear to be associated with electron oscillations. This is indicated primarily by the direct connection between the starting of the oscillation and the previous *A* frequency. However, the frequency itself is prescribed by the natural frequency of the operating system and is not affected by variations in the operating conditions. Here it is a case, therefore, of Gill-Morrell oscillations[4] which are due to the alternating potentials induced in the oscillation circuit and superimposed on the d.c. potentials of the electrodes. An alternating field of the frequency of the oscillations is thus superposed upon the stationary retarding field and results in an electron movement differing from the pure Barkhausen-Kurz oscillations.

2. Intensity Variation

In Fig. 2 the anode potential was kept constant while the grid potential was varied. A similar diagram is obtained if the grid potential is kept constant and the anode voltage is changed. In Fig. 3a is a corresponding diagram for a fixed grid potential of 240 volts. The effect of the anode voltage on the frequency of the *A* oscillations as well as on the incipience of the *B* region is similar to that of the grid potential. The lower Figs. 3b and 3c indicate how the anode current I_a and the high-frequency current in the bridge I_h vary with the detuning of the oscillation circuit. First of all the abrupt rise of the anode current with the incipience of the *B* oscillations is clearly shown. It will also be noticed that the maximum of I_a coincides with the maximum of I_h (oscillation intensity). This may therefore serve as an index in the tuning process. With increasing negative anode potential the anode current decreases, and its maximum is displaced toward the shorter wavelengths or higher frequencies. If the anode is strongly negative, at about 100 volts, there is no longer any noticeable anode current. Gill and Morrell obtained oscillations with the anode completely insulated. Electrometric measurements of the potentials of the free anode showed it to be strongly negatively charged. When the *B* oscillations start its potential increases. The curves of oscillation intensity I_h with a strongly negative anode and with a free anode are almost identical (cf. curves 5 and 6 in Fig. 3).

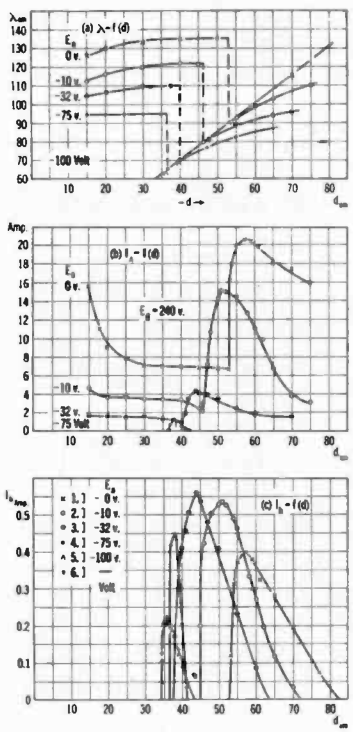

Fig. 3

The intensity of the *B* oscillations is considerable in comparison with type *A*; high-frequency currents up to 0.6 amperes occur in the thermo-element. The *A* oscillations could not be detected with the thermo-element even with the aid of a rather sensitive galvanometer. This shows that the external oscillation system is almost non-participating in the *A* oscillation process.

In the above case the maximum oscillation intensity was obtained at $E_p = -32$ volts, $d = 44$ cm, and $E_g = 240$ volts. The current in the thermo-element was 0.56 ampere, the wavelength 78 cm. Accordingly it can be assumed that there is a most favorable wave for each grid potential, at which the intensity can be maximized by the proper choice of the anode potential and tuning.

3. The Transition

In the first experiments it seemed as if the Barkhausen-Kurz (A range) oscillations changed abruptly into the Gill and Morrell oscillations in the B region. More careful experiments, however, have shown that the A oscillations decrease slowly, and that the B oscillations appear suddenly with great intensity. Therefore,

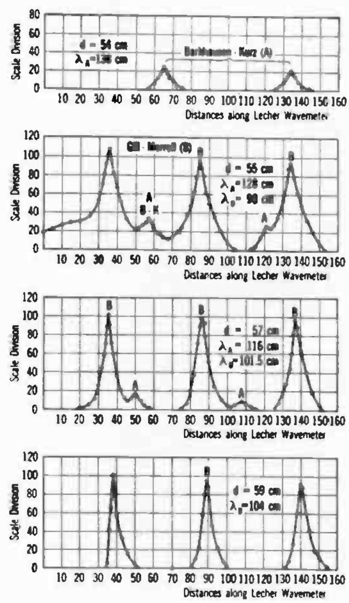

Fig. 4

in a narrow region both oscillations can be present simultaneously. Measurement of the differing frequencies by means of the Lecher system is generally impossible because the maxima of the Gill and Morrell oscillations mask the far weaker Barkhausen-Kurz oscillations. However, it was observed that the A frequencies no longer remained constant after the B oscillations started, but drifted upward a few per cent. Fig. 4 shows the conditions at the transition in a particularly clear case; here the intensity distribution on the Lecher system was recorded over several maxima for lengths d differing slightly. At $d = 54$ cm only Barkhausen-Kurz electron oscillations are present; on displacing the bridge but one centimeter, however, the Gill and Morrell oscillations predominate. At $d = 57$ cm, the A oscillations are weak and at the same time their wavelength has decreased from 138 to 101.5 cm. Finally, with $d = 60$ cm, only the short-wave B oscillations of Gill and Morrell are present.

4. Influence of the External Oscillation System

For the production of the highest possible frequencies it is important first to generate "pure" electron oscillations of the A type which serve to initiate the shortest possible waves. According to the theory of Barkhausen-Kurz and the investigations of Scheibe this may be brought about by increasing the field strength between the electrodes. Secondly, the beginning of the Gill and Morrell oscillations is to be displaced as far as possible toward the short waves.

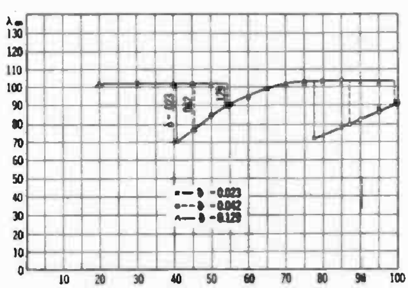

Fig. 5

Fig. 5 shows that the starting point of the Gill-Morrell oscillations is primarily determined by the character of the external oscillation system. The curves represent the effects of the high-frequency systems of different damping. This was done without shifting the tuning, by using materials with different conductivity for the parallel wires. As shown, the beginning of the Gill and Morrell oscillation is delayed by increasing the damping.

It is to be expected that a reduction of the damping must cause an extension of the B region. Experiments in this direction were not made, but this is explained by an observation given by Scheibe.[14]

[14] Scheibe, *Jahrb. d. Draht. Tel.*, **27**, 1, 1926.

In order to obtain Barkhausen-Kurz oscillations of greater intensity with parallel wires, Scheibe coupled two tubes together by their resonance circuits. Each of these tubes was able to set up independent oscillations, and he found that with external excitation by one, the other tube started to oscillate with 30–40 per cent lower potentials than was required for spontaneous oscillation. If we assume that the oscillations produced by Scheibe were of the Gill-Morrell type, as is probable on account of his arrangement, the extension of the oscillation range is easily explained as a reduction in damping by the external excitation which naturally acts in the opposite sense to the increase in damping previously described.

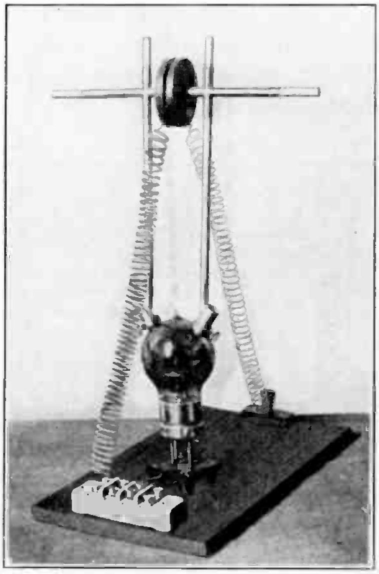

Fig. 6

During the course of the investigations a large number of widely different types of tubes were tried. Powerful electron oscillations of the A and B types were obtained with the French short-wave tube of the TMC type. This tube gave a 38-cm wave with a high-frequency current of 0.12 ampere. Fig. 6 shows a short-wave oscillator connected with this tube. To increase the radiation the tuning bridge has been extended on both sides to form a symmetrical antenna.

5. Theoretical Observations

(a) **Calculation of the Electron Vibration Frequency in the Alternating Field.** Barkhausen-Kurz gave a formula for the calculation of the electron frequency under simplified conditions. If in the calculation we consider instead of the stationary retarding field an alternating field, we may expect theoretically a change in the electron motion and with it a change in the frequency. The complete period is given as the time spent by an electron in the grid-anode and grid-cathode spaces. It will be assumed that both spaces are equal so that the calculation need be carried out for one space only.

Assumptions:

(1) Anode and cathode potentials are zero.

(2) The grid is located midway between anode and cathode; that is, $d_a = d_k = d$ (see Fig. 7).

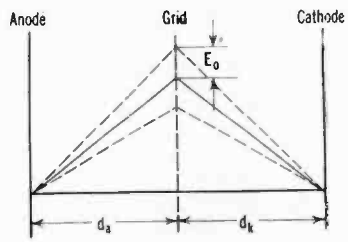

Fig. 7

(3) The electrodes are plane.

(4) Space charges are disregarded.

(5) The oscillation produces an alternating potential only at the grid.

Then the electric force in the inter-electrode space is:

$$E = \frac{E_g + E_0 \sin(\omega t + \phi)}{d}$$

and the equation for the motion of an electron is:

$$-e \cdot \frac{E_g + E_0 \sin(\omega t + \phi)}{d} = m \frac{d^2 x}{dt^2} \cdot \tag{1}$$

If we assume that the electron enters the inter-electrode space at time $t = 0$ with the velocity v_0 and that at this instant also $x = 0$, by double integration of (1) we get:

$$x = -\frac{e}{m\cdot d}\cdot E_a\cdot\frac{t^2}{2}+v_0\cdot t+\frac{e}{m\cdot d}E_0\left(\frac{\sin(\omega t+\phi)}{\omega^2}-\frac{t\cdot\cos\phi}{\omega}-\frac{\sin\phi}{\omega^2}\right) \quad (2)$$

A simple solution of this equation for t is possible with the following assumptions:

(1) The frequency of the alternating field is determined by the electron frequency. Therefore, if we make $x=0$ in (2), we get the time t, which corresponds to a half period; this is:

$$\omega\cdot t=\frac{\omega T}{2}=\pi$$

(2) At the time $t=0$, that is, at the instant of the passage of the electron through its meshes, the grid receives a negative charge, as most of the electrons strike the grid. At this instant, therefore, E_0 has its maximum negative value, that is, ϕ must be equal to $-\pi/2$.

Then the duration of a semi-period is

$$\tau=\frac{v_0}{\frac{e}{m\cdot d}\left(\frac{E_a}{2}-\frac{2E_0}{\pi^2}\right)} \quad (3)$$

In which

$$v_0=\sqrt{\frac{e}{m}2(E_a-E_0)}$$

If in (3) we substitute the value of v_0 and of the potential in volts, we get the wavelength

$$\lambda=c\cdot 2\cdot\tau$$

$$\lambda=\frac{4000\sqrt{E_a-E_0}\,d}{E_a-\frac{4E_0}{\pi^2}} \quad (4)$$

Concerning this equation (4) we notice first that when $E_0=0$ it changes into the equation given by Barkhausen-Kurz. It is only necessary to replace the electrode space d by the "anode diameter" d_a which is equal to $4\cdot d$, *ex hypothesi*.

There is a shortening of the wavelength due to the alternating potential, E_0, as is seen from the following table:

$E_a=500$ volts, $d=0.5$ cm

E_0	λ
0	89.5 cm
100	86.9
200	82.5
300	74.6
400	58.8

If in (2) we substitute various values for t starting from $\omega t=\pi$ we get the curve of Fig. 8 for the path described by the electron. In this figure different values of E_0, alternating potential, are illustrated. We see how the excursions of the electron continually diminish with increasing E_0 and the electrons always return sooner to the grid. Due to the simplifications, the curves give the course of the electron motion only qualitatively, and a hopeful comparison with experimental values cannot be attempted.

(b) Mechanism of the Gill and Morrell Oscillations. On the basis of the above observations the mechanism of the Gill and Morrell oscillations can be explained as follows:

Assume that the oscillation circuit tuning is changed from shorter to longer waves. At first its natural wavelength is far below the excitation frequency of the Barkhausen-Kurz oscilla-

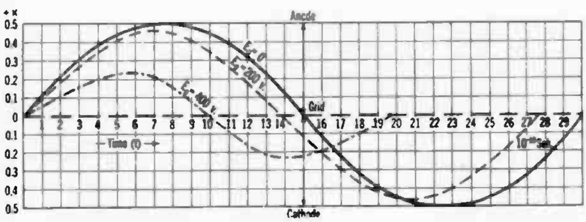

Fig. 8

tions; hence the alternating potential between the electrodes is very small and we have "pure" electron oscillations in the sense of Barkhausen and Kurz. (See Fig. 8, $E_0=0$.) As the oscillation circuit is tuned to the exciting wave, the alternating potential at the electrodes rises, corresponding to the resonance curves, until a certain potential, designated as the "ignition potential," is reached.

Now the alternating field superposed upon the stationary retarding fields causes a shortening of the exciting wave so that the attainment of resonance is accelerated; the result is an additional strengthening of the alternating field and so on, that is, a building-up process takes place which rapidly affects both the alternating potential and the frequency. The final stable state is reached when the frequency of the electron vibrations coincides with the proper frequency of the oscillating system, for an increase in the excitation frequency above resonance would cause a weakening of the alternating field. From this it is evident that the frequency of the Gill and Morrell oscillations is determined only by the natural wavelength of the tuning system.

On the other hand, if the natural wavelength of the oscillation circuit is above the Barkhausen-Kurz wavelength, an alternating field also arises on approaching resonance, but as this reduces the exciting wave its action is opposite to the above and causes a displacement from resonance. A building-up process is therefore discouraged. Only on approaching the exciting wavelength from longer natural wavelengths are the Barkhausen-Kurz waves gradually shortened, and thus we get the explanation of the course of the frequency curves in Fig. 2b as well as Fig. 3a, with its abrupt start on one side and its gradual transition on the other.

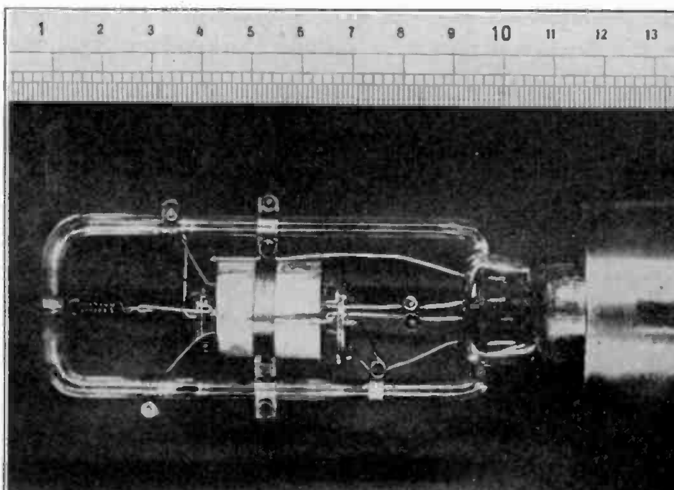

Fig. 9

It is proper to find analogies between the above building-up process and the building-up of oscillations in regenerative circuits. Therefore it is not to be wondered at that the well known "pulling" (ziehen) of the usual transmitter was also observed in the oscillations present here. The starting point of the Gill and Morrell oscillations is not determined uniquely, but is displaced somewhat according to whether the tuning of the system is varied from longer to shorter waves or the reverse. For this reason the data of the curves in these illustrations were all taken in the same direction, namely from shorter to longer waves.

Summarizing the results of Part I it may be stated that two different kinds of electron oscillations occur with the retarding field connection of Barkhausen-Kurz, those of a pure Barkhausen-Kurz type, in which the frequency is determined by the stationary interelectrode fields, and those of Gill and Morrell whose frequency agrees exactly with the natural frequency of the tuning system. Between these types of oscillations there is a region of steady transition, in which there is a departure from the true wavelength of the oscillation circuit and a gradual transition toward the Barkhausen-Kurz frequency. Therefore if "pure" oscillations of one or the other form are desired, the external system must be carefully and properly tuned. Carelessness in this tuning process may be suggested as the cause of many disagreements between different authors. In this connection it must be observed that the leads or tube electrodes may themselves comprise an oscillation system and be capable of producing Gill and Morrell oscillations.

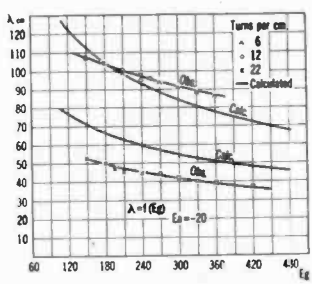

Fig. 10

II. Electron Oscillations of Similar Type of Different Frequencies

1. Shorter and Longer Barkhausen-Kurz Oscillations

Scheibe[4] has already reported Barkhausen-Kurz oscillations of two different frequency ranges. In the course of my own investigations I likewise observed a higher frequency wave range with the lower range, but could not reproduce it with certainty. I later substituted for the Schott tube a specially constructed electrode system supported by a small glass frame which by means of a ground joint could be placed in a glass vessel connected with a vacuum pump.

For the generation of highest frequencies it is naturally of the greatest importance to ascertain the conditions under which the higher frequency range of the Barkhausen-Kurz oscillations occurs. On account of the large number of frequency-determining factors the variables must be restricted as much as possible and for this reason the radial dimensions of the tube, that is, the grid and anode diameters, were first maintained constant. The only variable factor which remained was the pitch of the grid.

As the Schott tubes have an average of 8–10 turns per cm of grid length, a grid with 15 turns per cm was made and inserted in the anode cylinder of a Schott tube. The study of this system showed that under the same operating conditions strong "shorter" waves were readily obtained, whereas even at higher grid potentials no "longer" Barkhausen-Kurz oscillations could be detected. For more systematic investigation of the distribution of the two frequencies over the entire oscillation range under different operating conditions, the same grid system was successively provided with windings of 8 different pitches and with different grid potentials. In order to exclude Gill and Morrell oscillations with certainty, the electrodes, as can be seen in Fig. 9, were bridged on both sides by two blocking condensers and the external oscillation system, or leads, are carried off in this way.

The results of the investigation are shown in Fig. 10, in which both frequencies are given as functions of the grid potential. With 6 windings per cm the "longer" waves are present exclusively; with 12 windings per cm the short waves appear at $E_g=260$; with 22 windings per cm these start at $E_g=150$ volts. Then the longer waves appear with the shorter up to 180 volts. It is clear from the measurements that the shorter waves start the more readily the smaller the pitch of the grid helix. If we compute the ratio of the wavelengths, the resulting values differ from 2. Thus, as Scheibe has emphasized, it is demonstrated that the shorter waves cannot be regarded as overtones of the longer.

2. Mechanism of the Shorter and Longer Barkhausen-Kurz Oscillations

The observation that the permeability of the grid is determinative for the appearance of the longer or shorter Barkhausen-Kurz oscillation leads to the following explanation: In their theory Barkhausen and Kurz assume that the electrons swing about the grid; this view is fortified by the experimental agreement. In Fig. 11 on the left there is shown the motion of an electron which swings twice about the grid. As the anode has a negative potential the electron reverses in front of it at the position of the surface of zero potential.

A large part of the electrons emitted by the cathode, of course, do not penetrate the grid after their first journey through the grid-cathode space, but strike the grid wires. These electrons impart all their energy to the grid, and need not be considered in the generation of oscillations. With the electrons which are reversed in the grid-anode space but which do not pass through the grid the second time, however, the case is quite different. On account of their vibratory motion in the retarding field of the anode, they are capable of supplying energy for the production of oscillations. The frequency of these oscillations is determined

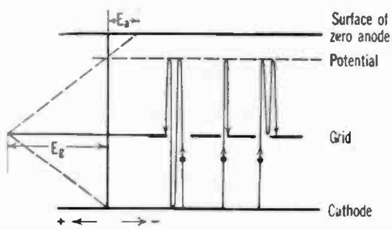

Fig. 11

only by the stay of the electrons in the grid-anode space and must therefore be higher than the electron oscillations in the anode-cathode space itself. At the right of Fig. 11 is shown the motion of an electron which moves to and fro once in the grid-anode space. It must also be remembered that the electrons can swing to and fro several times in the grid-anode space, chiefly under the influence of the alternating potentials at the grid. We are then dealing with Gill and Morrell oscillations, which will be discussed later.

On the basis of this picture of the mechanism of the higher frequency electron oscillations, it is comprehensible that the permeability of the grid influences the appearance of one frequency range or the other. The narrower the mesh of the grid the lower will be the percentage of electrons that pass through it until finally no vibratory motion at all takes place about the grid, it being then entirely confined to the grid-anode space. This could be shown experimentally.

Scheibe has already considered the stay of the electrons in the grid-anode and in the grid-cathode spaces separately. His formula for the calculation of the wavelength permits the direct calculation of the two partial frequencies separately. This calculation is carried out for the system used and the result is shown in Fig. 10. It is seen that the longer waves show a considerable deviation from the experimental values; neither is a quantitative agreement of the shorter waves to be expected, so that a proof could not be adduced from this.

3. Experimental Tests

The wavelength of the Barkhausen-Kurz oscillations involving the entire interelectrode space will be designated by λ_{AC}; as has already been explained they consist of two partial oscillations, namely the electron motion in the grid-anode space whose wavelength is λ_{GA} and in the grid-cathode space of wavelength λ_{GC}, so that

$$\lambda_{AC}=\lambda_{GA}+\lambda_{GC}$$

Only the waves λ_{AC} and λ_{GA} can be excited; λ_{GC} cannot appear physically, at least not in a normal three-electrode tube. It can be determined indirectly experimentally from the difference between the longer and the shorter waves, for according to the above equation:

$$\lambda_{GC}=\lambda_{AC}-\lambda_{GA}$$

Constant operating conditions being assumed, λ_{GC} must depend entirely on the dimensions of the grid-cathode space. If the dimensions of the grid-anode space are changed by increasing the anode diameter, only a change in λ_{AC} and λ_{GA} can be expected, while their difference, λ_{GC}, must remain constant.

I have previously discussed the dependence of both waves upon the grid potential (Fig. 10) for an anode diameter of 18 mm. With the same grid inserted in an anode with a diameter of 26 mm, and with a variation of the grid pitch, the longer and shorter waves are found over the same potential range. The new system is shown in Fig. 12. It is also shunted by two blocking condensers so that only "pure" electron oscillations can occur. The curves in Fig. 13 show the frequency; in spite of the great change in grid winding, both waves could be obtained over a small potential range. At low grid potentials the frequency deviates from the direction expected from theory, increasing with the potential; only above 180 volts does the course correspond with the theory. This abnormal action is probably due to the fact that at low potentials the space charge is accentuated

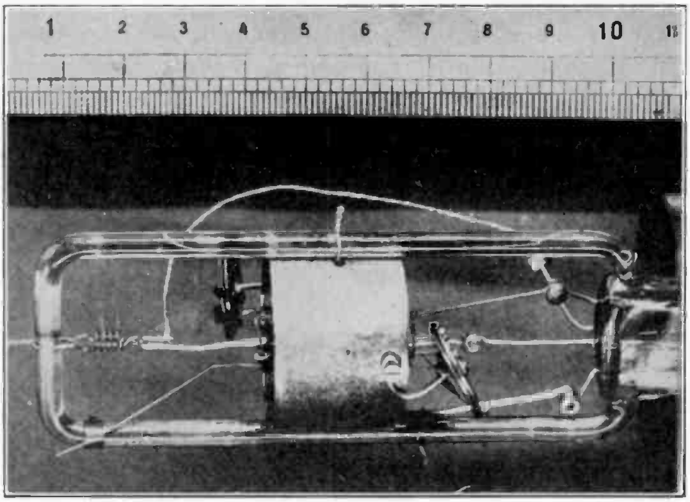

Fig. 12

and its effect on the frequency, especially in the large grid-anode space, may become of importance.

The difference in both curves gives λ_{GC} as a function of the operating potentials for each system. The course of this curve

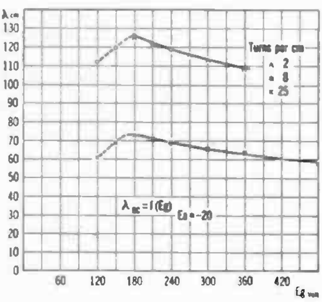

Fig. 13

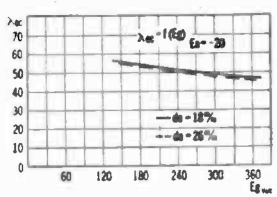

Fig. 14

is shown in Fig. 14, the solid curve for the 18 mm system, and the dotted curve for the 26 mm system. The close correspondence of the grid-cathode waves of the two systems furnishes excellent support for the above view of the mechanism of the shorter electron waves.

4. Gill-Morrell Oscillations of Higher Frequency

If the shorter electron oscillations are a special form of the Barkhausen-Kurz oscillations one may expect that, like the longer waves, they are affected by the alternating field between the electrodes. This follows from the theoretical observations of Section 5 of Part I which hold as well for the grid-anode space as for the whole anode-cathode space.

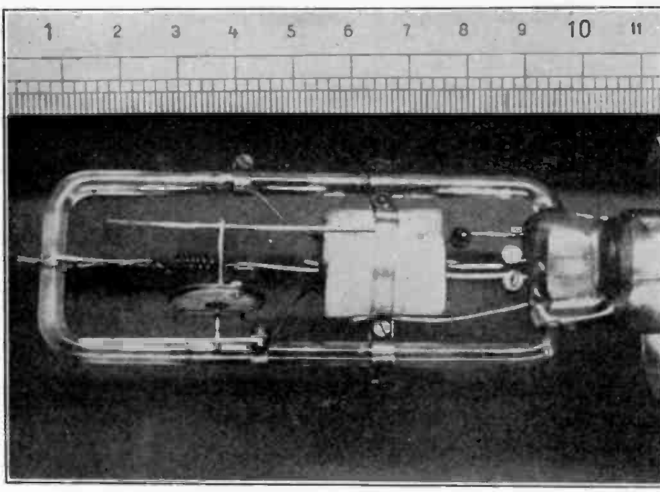

Fig. 15

In the effort to produce the highest possible frequencies, a system was constructed in which the electrodes were excited in antiresonance, and in which the oscillation-circuit was partially inside the tube and partially outside. If one oscillation node is fixed by the position of the internal condenser (see Fig. 15), the

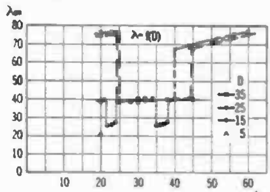

Fig. 16

other can be adjusted by shifting the bridge on the external parallel wires. Fig. 16 shows the frequency curves for constant operating potentials of 360 and −20 volts. *D*, the length of the wire system within the tube, is taken as a parameter. Gill-Morrell oscillations of lower or higher frequency are set up depending upon the tuning condition. With D equal to 5 mm, a wave of 21.7 cm can be produced whose intensity is so great that measurement of wavelength could be made at a distance of 2 meters.

In order to produce still shorter waves a system of low damping was constructed in which only the half-wave oscillation was permitted. As shown in Fig. 17, for this purpose grid and anode are equipped *on both sides* with longitudinal wires along which capacity bridges can be shifted. The entire oscillation system is thus within the tube.

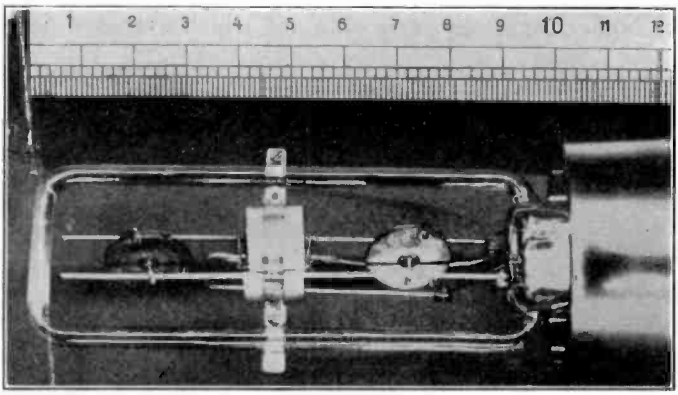

Fig. 17

With this system two frequency ranges could be obtained by varying the operating conditions, first the Barkhausen-Kurz higher-frequency oscillations (as a comparison with earlier measurements showed) and second, the Gill-Morrell higher

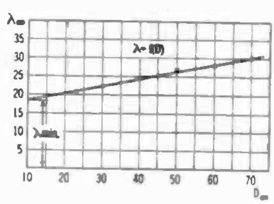

Fig. 18

frequency oscillation, whose wavelength is given by the distance of the two condensers in the tube. If the Gill-Morrell wave is plotted as a function of this distance, we get the calibration curve of the oscillator (Fig. 18).

The smallest adjustable length, D, was 23 mm; for this there was obtained a 20.8-cm wave. Extrapolating the curve to smaller D, which is permissible on account of its linearity, we find for $D=10$ mm, the length of the electrodes themselves, a wavelength of 18.5 cm; *this represents the shortest wave that can be obtained with the present system.*

During all previous investigations the radial dimensions of the electrodes were kept constant. These are potential additional means for reducing the Barkhausen-Kurz frequency. The system shown in Fig. 19 was therefore constructed, in which the anode was reduced to a diameter of 11 mm, and the grid to a section of 3×3 mm. With this system Barkhausen-Kurz oscillations of the higher frequency were obtained which averaged an *octave*

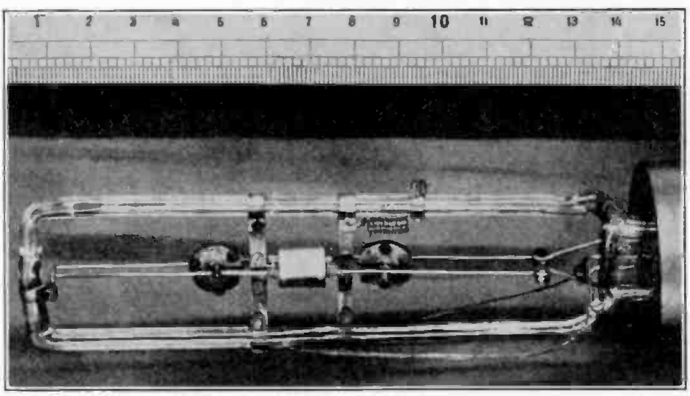

Fig. 19

higher than the previous system. By suitable tuning of the oscillation system Gill-Morrell oscillations of higher and lower frequency could be produced. The shortest tuning-system length was fixed by the fastening-pieces of the heating filament at 36 mm; for this there was a natural wave of 22.5 cm. With a grid potential of 360 volts Barkhausen-Kurz oscillations of 21.4 cm were still obtained; accordingly, with a proper tuning system Gill-Morrell oscillations down to 15 cm might be obtained with safety.

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A CONVENIENT METHOD FOR REFERRING SECONDARY FREQUENCY STANDARDS TO A STANDARD TIME INTERVAL*

By

L. M. Hull and J. K. Clapp
(Research Laboratory, General Radio Co., Cambridge, Mass.)

Summary—*A method is described for obtaining a convenient low frequency from a high-frequency standard by means of harmonic control of distorted wave oscillators (multi-vibrators). In the equipment described two such oscillators are employed, for convenience in arriving at a harmonic series of frequencies, based on 10-kc fundamental, as well as a frequency of 1 kc for operation of the clock motor, from a standard frequency of 50 kc. The conclusions based upon an experimental investigation of harmonic control of the distorted wave oscillators are given. Representative curves of the variations in frequency of the 50-kc standard for 10-day intervals are shown.*

1. Introduction

In recent years the technique of frequency standardization has been refined and improved through the development of oscillating systems which are so nearly invariant that their average frequency is practically independent of the time interval for which the average is determined. Such a system permits the determination of frequency, by counting successive cycles, to a degree of precision which is usually limited only by the patience of the observer. Extensive use of this "absolute" method of frequency measurement has suggested the concept of the earth's rotational frequency, or a time interval defined thereby, as a fundamental standard for frequencies of all magnitudes.

Practical methods for maintaining various types of oscillators as the intermediate links between the desired frequency and a standard time interval have been described in the literature. It is the purpose of the present paper to call attention to an application of the same general principle to a specific problem of laboratory calibration. The results described have a general significance insofar as they relate to the establishment of a harmonic series of standard frequencies.

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At the present time there are approximately 600 radio broadcast stations operating more or less continuously in the United States. The transmitters of these stations have an irregular geographical distribution and to each transmitter is assigned as a carrier one of the 96 decimal frequencies between and including the limits of 1500-550 kc. The universal problem of interference is not solved by accurate maintenance of the carrier frequencies on their assigned values, but such maintenance is obviously a primary essential. Although piezo-electric oscillators, or their equivalent, will ultimately be incorporated as integral frequency control elements of all broadcast transmitters, such arrangements are not now in general use. Small portable piezo oscillators are widely employed in the broadcasting stations as comparison standards, by means of which their carrier frequencies may be occasionally corrected. The problem of accurate adjustment of quantities of these control and comparison standards to the even decimal frequencies employed by these stations has suggested the calibration method which we are to describe.

2. Experimental Method

The general technique of "direct" calibration from a single-frequency source, which in turn is compared with the earth's rotational frequency, is based upon an assumption which may be stated as follows: if oscillations of constant periodicity occur in two independent systems and either system be so adjusted that the fundamental constituent of its oscillation coincides constantly with a harmonic constituent of the oscillation in the other system, then the fundamental frequency of either system is uniquely determined in terms of the other. The propriety of this assumption will not be discussed here although other investigators have given attention to specific phases of it, as for instance by examining the validity of identifying with integral harmonic ratios the various emission frequencies from a multivibrator[1]. If, therefore, our fundamental oscillation occurs at a low frequency and is associated with a substantial series of real harmonics, the physical comparison of an unknown frequency with the frequency of one of these harmonics is considered to be an absolute comparison with the fundamental. Since the ultimate standard of frequency is very low indeed compared with what

[1] D. W. Dye, "A Self-Contained Harmonic Wavemeter", *Phil. Trans.* Roy. Soc. of 224, A, London, 279.

we normally have to measure, it is customary to employ the intermediate source of oscillations previously mentioned as the basis of the harmonic series from which calibrations are actually made.

An electro-mechanical oscillator such as a tuning fork has been extensively used as an intermediate source. A tube-driven fork can be coupled directly to the counting mechanism, but the fundamental frequency is inconveniently remote from the frequencies at which calibrations are required in the present case; also the fork is subject to the familiar limitations arising from its damping and its temperature coefficient. The use of a piezo-electric oscillator as the intermediate source has been described in an important paper read before the Washington meeting of U.R.S.I. in October, 1927.[2] The quartz oscillator exhibits small perturbations under reasonable operating conditions, and permits the establishment of a harmonic series from a fundamental of higher frequency. It requires, however, a very stable and reliable electrical system for converting its fundamental frequency into a lower frequency suitable for operating the counting mechanism.

In the system to be discussed in the present paper, a piezo-electric oscillator is employed as the intermediate source. But for the step-down frequency converter, a system describing "relaxation oscillations"[3] is adopted because it has been found to offer many advantages for the specific problem in hand. An intermediate frequency of 50 kc is chosen for convenience. The relaxation oscillator can be adjusted to yield a highly stable fundamental at 1 kc or even less, directly controlled by the 50-kc source, which operates a small impulse motor, running in air, for the counting mechanism. Furthermore, the controlled relaxation oscillator, if properly designed, is an extremely simple piece of equipment.

Two general types of relaxation oscillator have been suggested for use as step-down frequency converters. The simplest is a neon discharge tube, used in virtue of its so-called cut-off characteristic to sustain oscillations in a circuit comprising a direct-current source, a resistance, and a condenser. The oscil-

[2] J. W. Horton and W. A. Marrison, "Precision Determination of Frequency." Paper read at meeting of the International Union of Scientific Radiotelegraphy, Washington, D. C., Oct. 13, 1927. Proc. I.R.E., 16, 137; February, 1928.

[3] Balth. van der Pol, Jr., "On Relaxation Oscillations," *Phil. Mag.*, 978, Nov., 1926, also *Zeits. f. Hochfreq. Technik*, 29, 114, 1927.

lations of such a circuit are rich in harmonics and if a second oscillation of a higher frequency is injected into the circuit, the relaxation oscillation may assume a frequency which is an integral submultiple of the injected frequency. It is possible to maintain an oscillation in such a circuit whose fiftieth harmonic coincides with the injected frequency. But the behavior of this control effect on such high orders is relatively unsatisfactory, as minute variations in either the injected voltage or the direct voltage are likely to destroy the control. The other type of oscillator is the Abraham-Bloch multivibrator. Doubtless there are many other forms of relaxation oscillator which present similar characteristics, but the modified multivibrator has proved so satisfactory that no others have been investigated.

The multivibrator consists essentially of an aperiodic circuit in which oscillations of irregular waveform are sustained by a triode excited by a second triode which provides the proper phase relation for maintenance. It was first used as a rich source of harmonics, in which the fundamental constituent is determined by and probably maintains a constant phase relation with a low-frequency injected oscillation. In its usual form it therefore constitutes a frequency multiplier or step-up frequency converter suitable for comparing high frequencies with a low-frequency standard. Some time ago one of the present authors[4] suggested the possibility of controlling a multivibrator at a fundamental frequency which is an integral submultiple of the injected control frequency, thus making it a step-down frequency converter. In this case the control oscillation coincides with a harmonic of the multivibrator. When this condition is established experimentally, the multivibrator is definitely controlled by the injected high-frequency oscillation, and the multivibrator oscillation has certain characteristics which distinguish this "controlled" state of oscillation. Since this system appears to be well adapted for driving a counting mechanism at frequencies as low as one one-hundredth of the constant frequency source, we shall examine in detail its behavior in the role of a controlled submultiple generator.

[4] J. K. Clapp, "Universal Frequency Standardization from a Single Frequency Source," *Jour.* Opt. Soc. Amer. and Rev. Scien. Instr., 15, 25, July, 1927.

3. Harmonic Control of Multivibrator

No adequate mathematical analysis of the action of the multivibrator has been given. In the practical case large voltage amplitudes are encountered, carrying the operating points of the vacuum tubes far beyond the region of the characteristic which might be treated under limited series solutions for small amplitude oscillations such as van der Pol's.[3] But large amplitudes are required to obtain an extended series of harmonic frequencies of useful amplitude. It thus appears that the only feasible investigation of multivibrator oscillations, meeting our physical requirements, demands sufficiently complete experimentation to permit valid generalizations to be drawn.

The basic assumption of integral ratios between the harmonics and fundamental *emitted* by a multivibrator allows frequency comparisons to be made, but does not necessarily imply that the controlled multivibrator has a harmonic frequency, of value equal to the frequency of the controlling source, and of invariant phase with respect to this source. A regular, or irregular, variation in phase might conceivably exist.

We define the "controlled" state for the multivibrator as that state of operation in which one of the multivibrator harmonics does not change phase with respect to the fundamental frequency of the controlling source by more than 360 electrical degrees in any arbitrary time interval. Then, by the assumption of integral harmonic frequencies, the fundamental frequency of the multivibrator is an exact submultiple of the controlling frequency. We shall next examine the experimental conditions for permanently establishing this controlled state on the desired harmonic order.

All the experiments have been carried out with no inductance, save that of strays, in the circuits of the multivibrators because it seemed logical to reduce the electrical inertia to a minimum when searching for the most stable and unambiguous control condition.

Preliminary studies indicated that the control stability in any multivibrator was increased by the injection of the control voltage into the common plate lead of the multivibrator tubes, as compared with injection into the circuits of either tube individually. Such increased stability is obtained, however, only if the multivibrator tube circuits are unsymmetrical so that an appreciable resultant plate current exists in the common lead.

The use of the unsymmetrical multivibrator in this connection resulted from earlier work by one of the authors with the symmetrical arrangements. Dissymmetry was first attained through the use of tubes of different type, having widely differing filament emissions. Later work showed the same result was obtained with tubes of the same type, if the circuit constants were made dissymmetrical, as by the use of resistances 20 to 50 times as great in one tube plate circuit as in the other.

As a result of this dissymmetry, one tube operates as a "throttle," the operating point sweeping across the entire characteristic from saturation to cut-off, while the other operates substantially as a normal amplifier. This type of operation, no matter how attained, is characteristically favorable for stable control.

A tube is employed between the controlling source and the multivibrator which does not function primarily as an amplifier, but as a one-way relay, whose purpose is to isolate the standard source from reactions due to the multivibrator. This is not only generally desirable, but it is a necessity if the standard is to maintain its frequency with extreme constancy. An ordinary vacuum-tube amplifier circuit is not sufficient when the controlling frequency is as high as 50 kc. Either a tetrode, utilized as a "screened grid" amplifier, or a carefully neutralized triode, must be employed.

In this work the choice of the standard was made in favor of a very heavy quartz bar whose frequency lay close to 50 kc. For convenience, only, the order of frequency division was taken as 50, so that the controlled multivibrator fundamental would be close to 1 kc. The impulse motor and clock train were designed to keep correct time when the frequency applied was 1,000 cycles per second, so that with this division the timing of the high-frequency source is reduced to observations of small time errors indicated by the clock.

The schematic circuit arrangement for a single controlled multivibrator is given in Fig. 1. In the absence of the tetrode and the resistance R, the uncontrolled multivibrator remains. In this work the resistances were fixed and variation of the fundamental frequency of the multivibrator attained by simultaneous variation of the capacities C. In the uncontrolled state, variation of these capacities produces a corresponding smooth variation of fundamental frequency of the multivibrator.

The application of a controlling voltage, e_c, to the grid of the isolating tube results in the injection of a voltage of the same frequency into the multivibrator circuit. Provided that the voltage is sufficiently great, that is, greater than a few hundredths of a volt, variation of the condensers C no longer produces a smooth and continuous change of fundamental frequency of the multivibrator. For certain separated ranges of values for C, the fundamental frequency of the multivibrator assumes discreet values, these being submultiples of the control frequency and invariant over the particular range of C. Between the ranges of C at which these discreet frequencies exist there are "transition" ranges in which no stable value exists for the fundamental frequency of the multivibrator. In these transition ranges the multivibrator attempts to oscillate in the controlled and uncontrolled states simultaneously, or, if the control voltage be

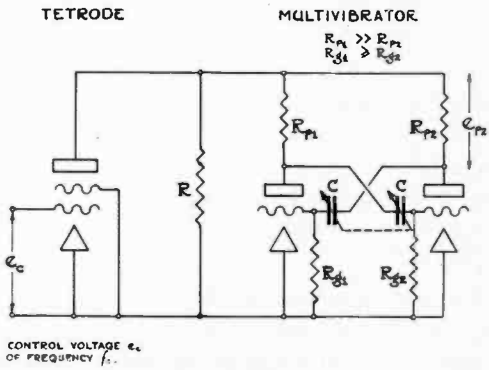

Fig. 1—Multivibrator of Fundamental Frequency f_c/n, where n is order number of harmonic at which control takes place.

sufficiently great, in adjacent controlled states. Observed by oscillograph, it is seen that in the latter case the fundamental oscillation of the multivibrator consists of a number of cycles of one discreet frequency and then one or more cycles at another discreet frequency. The oscillations alternate between the two frequencies in an irregular manner, there being no definite number of cycles in sequence of either of the two frequencies.

With small values of controlling voltage the operation is generally such that in the transition ranges of capacity, between those at which discreet and invariant values of fundamental frequency are obtained, the multivibrator oscillates in the uncontrolled condition. The regions in which control is attained are well defined and the controlled state is fairly stable. This mode of operation may be termed the "under-voltaged controlled state." If the control voltage be increased in magnitude, the transition ranges of capacity over which uncontrolled or unstable operation is obtained, are reduced progressively as the control voltage magnitude is raised, until finally the multivibrator fundamental frequency changes abruptly from one mode to the next upon varying the capacities through a very small transition range. This state of control may be termed the "over-voltaged controlled state" and represents that condition of operation which is believed to be the most stable and most desirable. In this condition, variation of capacity from maximum to minimum results in the production of a series of discreet fundamental frequencies for the multivibrator, the number of such frequencies which may be obtained being dependent upon the range of variation of capacity and upon the order of control which has been set up.

If the control voltage be increased beyond the point at which the transition ranges first shrink substantially to zero an effect which may be termed "drawing" is produced. This drawing effect consists of a progressive increase in the multivibrator fundamental frequency through discreet steps representing the successive harmonic control orders, as the amplitude of the control voltage is increased. In other words, the fundamental frequency is drawn toward the control frequency as the control influence is made stronger. If a particular control order is desired this drawing effect must be compensated by an increase in the multivibrator capacity or resistance. But this in turn tends to move the time constant of the multivibrator circuit away from the fundamental frequency at which it is being forced to oscillate, which ultimately reduces the control stability on its own account. The net result of these effects is the appearance of a fairly definite *optimum value* of control voltage for any given harmonic order.

We come now to somewhat more generalized conclusions concerning the dependence of the control range, the control order and the wave-form of the multivibrator oscillations upon variations of control voltage magnitude, variations in multivibrator fundamental frequency as produced by *symmetrical* variations in the multivibrator circuit constants, and by variations in the *degree of dissymmetry* of the multivibrator circuit produced by a non-symmetrical variation of the circuit constants. These considerations are best indicated by an outline:

1.	Variations in control voltage, e_c	Small e_c	Limited control range, though control is stable within range. Waveform of same type for various orders of control.
		Large e_c.	Maximum control range; further increases of control voltage produce changes in order of control by "drawing," with narrowing of control range if e_c be made large enough.
2.	Variations in multivibrator fundamental frequency due to symmetrical changes in circuit constants. (C, C, varied).		Control order changes in successive steps to lower orders as fundamental frequency of multivibrator is raised toward control frequency. Control range narrow with small e_c; control range wide with large e_c. Waveform of same type for the various contro orders.
3.	Variations in degree of dissymmetry of multivibrator circuit. (R_p; only varied)		Control order remains constant. Waveform alters in discreet steps to forms in which the "positive" and "negative" portions of the wave are in *integral ratios of time*, the total time remaining constant.

Based upon the experimental evidence, an important generalization may be drawn. *Symmetrical changes* in *circuit constants* produce effects similar to *symmetrical changes* in *control voltage magnitude*. (Changes in control voltage are inherently symmetrical in the arrangement used, since the same voltage is injected in the plate circuits of both tubes.) These effects are characterized by control at successive orders with no major changes in the waveform. Changes in the *degree of dissymmetry* of the circuit result in *discreet changes of waveform*, the control order usually remaining fixed. These changes of waveform are characterized by discreet changes in the ratio of time of "negative" to time of "positive" portion of cycle, the total time remaining constant.

It is worthy of note that under no operating condition was the stability of control improved by distorting the waveform of the control voltage to produce harmonics which might "lock in" with the higher harmonics of the multivibrator. This appears to be due to the fact that the effect which we term "control" occurs at a certain definite phase relation between the control wave and the corresponding multivibrator harmonic, and the phase differences between the higher harmonics of both sources, although invariant, may either aid or oppose the control which is already established.

The oscillograms of Fig. 2 indicate the phenomena observed. These oscillograms are line drawings of experimental observations of the plate current in that multivibrator tube which operates through the low external plate resistance. This current has in all cases the same general waveform as the relatively minute plate current in the throttle tube and also the current in the common plate circuit which includes the control tube.

Diagram A indicates the changes resulting from variations of the magnitude of the control voltage, all other factors remaining constant. The waveform remains essentially the same. As the control voltage is raised, the fundamental frequency of the multivibrator is also raised, being "drawn" toward the frequency of the controlling voltage. These observations give a reasonable explanation of the fact that in a symmetrical multivibrator a strong preference is evidenced for control to take place at even ratios, because of the symmetry of the "positive" and "negative"

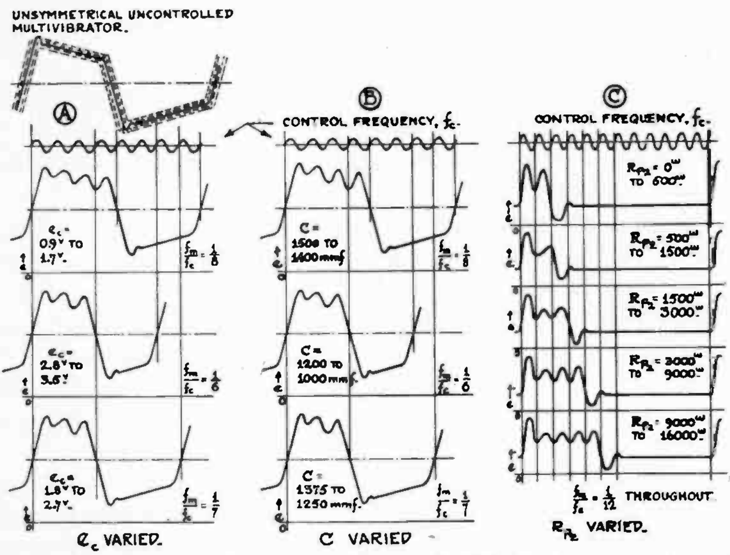

Fig. 2—Waveform and Frequency of Controlled Multivibrator, as shown by e_{p2} (Fig. 1), as functions of e_c, Cm, and R_{p2} as variables.

portions of the wave. With the dissymmetrical multivibrator such a preference does not appear to exist, which is explainable on the basis that the inherently unsymmetrical waveform readily adjusts itself so that one portion of the wave occupies a time interval equal to that of an odd number of cycles of the controlling frequency, while the other portion of the wave occupies a time interval corresponding to an even number of cycles of the control frequency. These conditions are illustrated in particular by the lower diagram for the frequency ratio of 1/7. Here the "positive" portion of the wave occupies a time interval of 3 cycles of the control frequency, while the "negative" portion occupies a time interval of 4 cycles. If the waveform is more nearly symmetrical, the multivibrator must be forced to operate dissymmetrically, by the controlling influence, in order that odd frequency ratios may be obtained, consequently requiring a greater controlling influence than in the unsymmetrical case.

In diagram B are shown sketches of the waveforms resulting from variation, in a symmetrical manner, of the capacities C of the multivibrator circuit, all other factors remaining constant. The waveform is seen to remain essentially constant but the frequency assumes three discreet values, bearing the ratios 1/6, 1/7, and 1/8 to the controlling frequency. The most stable adjustment of capacities is at a value slightly below the middle of the control range.

These two diagrams, 2A and 2B, illustrate clearly the equivalent effects of symmetrical changes of controlling voltage and of circuit constants upon the waveform and frequency of the multivibrator oscillations.

Diagram 2C indicates the changes in waveform as the lower of the two plate circuit resistances of the multivibrator is varied. In this case R_{p1} was 400,000 ohms, R_{p2} variable between 0 and 18,000 ohms, and R was 10,000 ohms. The multivibrator remained in control at all values of R_{p2}, with a fundamental frequency which was one-twelfth of the controlling frequency. The successive discreet waveforms were obtained over considerable ranges of values for R_{p2}, as indicated by the marginal notations in the figure. The integral ratios of the times for the two portions of the cycle are clearly indicated, as well as the manner in which the serrations due to the controlling frequency shift by single units as R_{p2} was continuously varied.

The assembly finally employed consists of two multivibrators in cascade, the choice of ratios being determined by the available standard frequency, the normal frequency of the clock motor and the fact that a series of harmonics based on 10-kc fundamental frequency was desired for calibration purposes. While frequency ratios with a single multivibrator of the order of 50 may be obtained with stable operation, it would have been necessary to utilize a second multivibrator in this case to obtain the desired harmonic series. Therefore the multivibrators were compounded, the first operating with a frequency ratio of 5, whose fundamental frequency was therefore 10 kc, and the second with a frequency ratio of 10, whose fundamental was consequently 1 kc. Provision was made for a small coupling coil to be inserted in the first multivibrator circuits so that the series of 10-kc harmonics could be utilized external to the assembly. The complete wiring diagram of the assembly is given in Fig. 4, on which the values of the various circuit constants are shown as well as the normal ranges of supply voltages, through which stable control was obtained.

With careful adjustments of the condensers C, *within the control range*, stable operation has been attained for the following ranges of supply voltages, the frequency ratio being 50 for the entire assembly:

e_c from 1.8 to 2.8 volts, at 50 kc
E_f from 3.5 to 6.0 volts
E_p from 90 to 135 volts (or more)

While ranges of supply voltages such as given above have been attained many times, the ranges shown on the diagram are more indicative of nominal conditions, in which no excessive care has been used in making the adjustments. Under these conditions no difficulties are encountered in operation while "trickle charging" the filament and plate batteries from the commercial 60-cycle power line.

The foregoing observations indicate that by injecting a 50-kc voltage into the circuits of the compound multivibrator it is possible to establish a highly stable condition in which the fundamental and all the harmonics of both multivibrator elements are constant in frequency to the same degree that the injected voltage is constant. Moreover, the fundamental output at 1 kc bears every indication of being precisely controlled by the injected voltage. However, the exact mechanism of this control effect has not been analyzed, and we still have no direct experimental evidence that, when the multivibrator operates in a condition which has the physical characteristics of being "controlled," its fiftieth harmonic actually bears a constant phase relation to the control oscillation.

As a final critical test of this point, simultaneous observations of the control oscillation and the fundamental constituent of the multivibrator were made by means of a Braun tube. The circuit arrangement is indicated in Fig. 3A. Two tetrode amplifiers were used, one coupled loosely to the quartz oscillator furnishing the control voltage, and the other to the ouptut of multivibrator *A* or multivibrator *B*. The output circuits of the amplifiers were carefully tuned so as to emphasize the fundamental frequencies of the standard oscillator and the multivibrator. When the circuits were not accurately resonant, the patterns indicated by the Braun tube were highly distorted. For the operation of the compound multivibrator, with a frequency ratio of 5 in the first and of 8 in the second, the pattern given by the Braun tube appeared as sketched in Fig. 3B.

The form of the pattern clearly indicated the frequency ratio of 40, remaining entirely stationary. As the tuned circuits were changed, the phase of one frequency component could be shifted with respect to the other, of course, but with the circuits

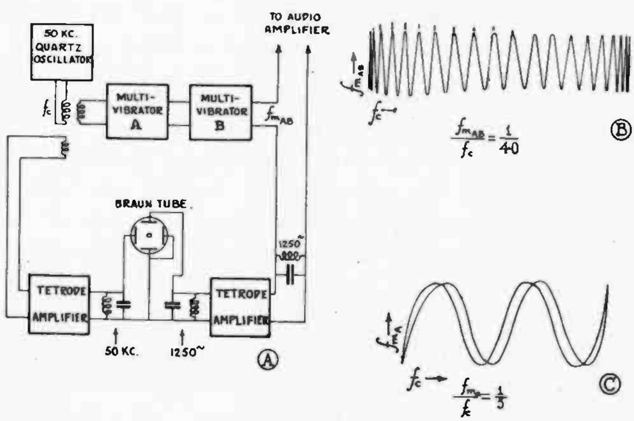

Fig. 3

fixed no visible fluctuations in frequency, phase, or amplitude took place. Upon changing either the resistances or the capacities of the multivibrator circuits, *keeping within the control range*, a change of phase was indicated by the pattern. This change of phase was "permanent,"—that is, varying the multivibrator constants displaced the phase and produced a new form of pattern, which form persisted until some further change was made.

A similar test, made between the frequency of the quartz oscillator and the output of the first multivibrator, gave a pattern which indicated the frequency ratio of 5, with no visible changes in phase or amplitude, indicated by Fig. 3C. As before, changing the multivibrator constants *within the control range* had the effect of changing the phase between the two voltages acting on the Braun tube, but for fixed values of the constants the phase remained fixed over indefinite periods. In both of these cases, variation of the circuit constants to a point where the control order changed indicated that a maximum phase variation of not more than 30 deg. (between the control wave and the synchronized harmonic of the multivibrator) could take place before the multivibrator jumped out of control at the desired order. In this test the multivibrator was brought to the edge of a transition range and the amplifiers were then adjusted to obtain the "in phase" pattern on the Braun tube screen. The multivibrator constants were then changed across the control range to the opposite transition point, the resulting change in phase being noted.

When in control, the controlling frequency may be varied by slight amounts, the multivibrator remaining in control. As the variation in frequency is made, the multivibrator suffers a slight phase displacement with respect to the standard frequency, but as long as the phase is not displaced beyond the limit of 30 deg., approximately, the phase takes up the new value and remains constant.

Continuous observations over periods of several hours have not indicated any perturbations of frequency, phase, or amplitude of sufficient duration to register on the Braun tube. After making the amplifier circuit adjustments necessary to obtain a distinctive pattern, on which any slight drift in phase would be readily observable, intermittent observations over periods of days have failed to show any change whatever in the phase of the synchronized multivibrator harmonic with respect to the control voltage.

We conclude from these observations that this state of oscillation of the multivibrator in the presence of a high-frequency injected oscillation which we have called the "controlled" oscillation throughout the discussion, is in fact an oscillatory state wherein the appropriate multivibrator harmonic bears a constant and permanent phase relation to the injected oscillation. Moreover, the frequency of this harmonic and hence of the fundamental follows any drift or perturbation in the frequency of the control oscillation without appreciably altering the phase relation or dropping out of synchronism even for a single cycle. The unsymmetrical multivibrator therefore constitutes a stable submultiple generator suitable for converting the vibrations of the piezo oscillator into the rotation of a clock train.

4. The Piezo Oscillator

The circuit of the 50-kc piezo oscillator is shown in Fig. 4. The piezo element *P* is a normal-cut quartz bar, vibrating in the direction of its length; its approximate dimensions are 5.6 x 2.2 x 1.4 cm. Fig. 5 shows the complete assembly of the measuring equipment. On the lower shelf are the batteries and chargers. On the upper, left to right, are the 50-kc temperature-controlled crystal, crystal oscillator, first isolating amplifier, 10-kc multivibrator with coupling coil, second isolating amplifier, 1-kc multivibrator, 1-kc amplifier, and synchronous motor with clock train. The mounting is shown in Fig. 6. The bar is balanced on a metal support *C* having a felt wedge on its upper surface and confined between heavy metal condenser plates *B* by the screw *E*, carrying a felt pad, the whole assembly being mounted on a base of insulating material *D*. The holder and quartz bar are mounted in a constant-temperature chamber *Q*, Fig. 4. This constant-temperature unit consists of an aluminum casting of high heat capacity containing a cavity for the quartz condenser; the casting is surrounded by an insulating layer, which is in turn enclosed in a closed metal box *R*. Heat is supplied from coils distributed around this second metal box, the heating circuit being closed through a bimetallic thermostat positioned, with the heating coils, outside the second container. The whole is enclosed in a heavily insulated outer box of wood, *W*.

The quartz bar is maintained in oscillation by the triode *T*, operated from plate circuit batteries which are independent of the multivibrator batteries.

The quartz oscillator circuit is one of the general class in which sustained oscillations are impossible without the quartz bar. This type of circuit was chosen because experience with numbers of piezo oscillators used as secondary standards has indicated that the frequency of the oscillator is in general more precisely determined by the mechanical vibration of the quartz if no other periodic element is present which can support sustained oscillations independent of the mechanical vibrator. For this reason the regenerative element *G*, which is added to augment the regeneration through the grid-plate capacity of the triode, is a pure resistance. The load in the plate circuit, consisting of inductance L shunted by capacity C, is far removed from resonance at the frequency of the quartz bar. It merely represents a convenient method of obtaining the high inductive reactance required to sustain oscillations. The coil has an inductance of 20 mh and the capacity is normally set at about 0.0003 μf, thus providing an inductive reactance of approximately 15,000 ohms. This reactance is still approximately proportional to the frequency in the operating region. The capacity C was added, not to augment the voltage available from the oscillator, since an excess of amplification is provided in order to avoid the necessity of large amplitudes in the oscillator circuit, but merely to provide a means for making fine corrections on the oscillator frequency.

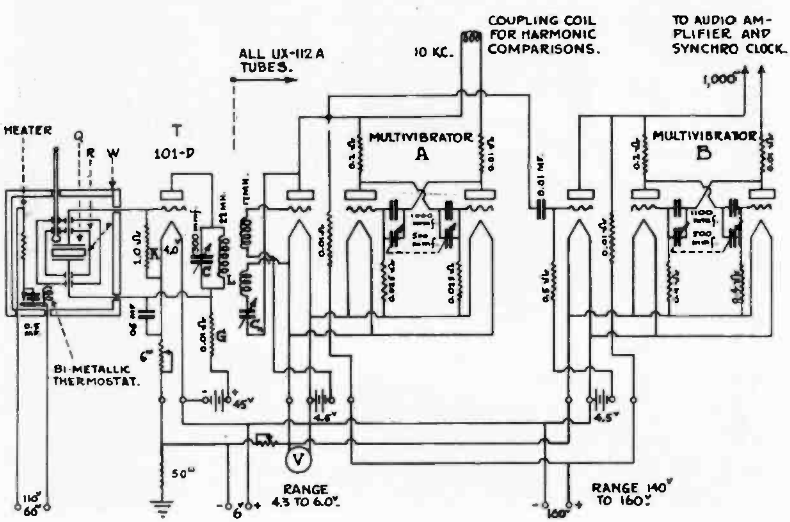

Fig. 4—Schematic Wiring Diagram of Controlled Compound Multivibrator.

The damping coefficient of the quartz bar itself has not been determined, but the characteristics of the whole oscillator in the actual operating region are as given in the following table. The figures here are the fractional change in frequency, positive or negative, for an increase of two per cent in the specified voltage or circuit element.

Two Per Cent Increase in:	Fractional Change in Frequency
Plate Voltage	-0.02×10^{-6}
Filament Voltage	-0.04×10^{-6}
Resistance R_p	$+0.8\ \times10^{-6}$
Resistance R_g	$-0.1\ \times10^{-6}$
Capacity	$-3.0\ \times10^{-6}$

The temperature coefficient of the quartz bar and holder combined is -5.4×10^{-6} per deg. C.

5. Determination of an Unknown Frequency Between 500 and 1500 kc

The specific purpose of this development was to provide means for accurately adjusting a secondary standard oscillator to one of the even decimal frequencies between 500 and 1500 kc. These secondary standards are piezo oscillators including quartz plates mounted in holders with adjustable air gaps. The plates are adjusted to within one-tenth of one per cent by grinding, and the final setting is made by adjusting the air gap with a screw

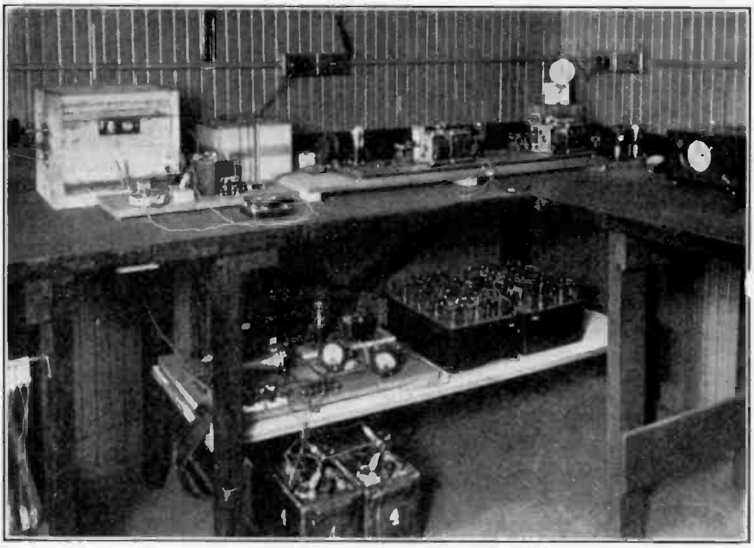

Fig. 5

provided for this purpose which is then locked. The process of referring the adjusted frequency to a standard time interval is carried out by coupling the oscillator under test loosely to the calibration coil of multivibrator A (see Fig. 4) which has a fundamental of 10 kc, then coupling a third oscillator and detector loosely to the combination, and listening with telephone receivers to the output of this detector as the oscillator under test is adjusted to beat zero with the appropriate harmonic of the multivibrator. This harmonic can be identified without difficulty with a subsidiary wavemeter, although there is usually not the slightest ambiguity as to the order of the harmonic, since the preliminary adjustment of the test oscillator by grinding the quartz is carried out with the aid of a heterodyne wavemeter calibrated to an accuracy of 0.1 per cent and the beat note between the oscillator under test and the multivibrator harmonic is normally well within the audible range before final process is begun. Since a third frequency is superposed on the output of the multivibrator and the test oscillator, the frequency of the test oscillator can be adjusted aurally with a precision defined by about one beat in five seconds between it and the multivibrator.

When this adjustment has been made, the frequency of the oscillator under test is known directly in terms of the frequency of the quartz bar at the time of the measurement, and thus in terms of the standard time interval, to a precision which depends upon how closely the frequency of the quartz is maintained at its average value.

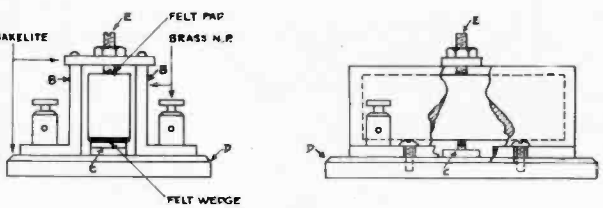

Fig. 6—Quartz Bar Mounting.

The standard time interval now employed in this work is one mean solar day, as indicated by radio time signals from the U. S. Naval Observatory. The oscillation counter consists of a small synchronous motor geared down through a clock train to a large second hand making one rotation per minute. The synchronous impulse motor has a toothed rotor of 120 teeth, and is driven by two U-shaped magnets around which the drive coils are wound. The gearing is such that the second hand indicates true solar time when the driving current has a fundamental frequency of 1 kc per second. The dial reading is compared visually with the standard time signals with a probable error of not over 0.05 second, the mean being derived from ten consecutive observations. Thus the mean frequency over each 24-hour interval is observed with a probable error slightly over one-half part in a million. The use of a longer standard time interval to reduce the error in observing the mean frequency will not be justified until substantial reductions are made in the probable short-period perturbations from the mean frequency.

The fluctuation of the frequency during the first ten days in August, 1928, is shown graphically in Fig. 7. Each point in this curve indicates the average frequency during the preceding 24-hour interval except in two cases, where the interval was 48 hours. The horizontal line marked "Mean Frequency 50,007.6 cycles, per second" is the average for ten days. The maximum deviation from the 10-day mean is 0.36 cycle during the fourth and fifth days.

The limitations at present imposed upon the commercial standards to be calibrated require that they be adjusted to an accuracy of ten parts in one hundred thousand or better. The

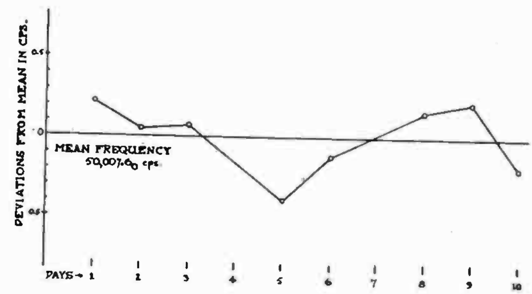

Fig. 7

24-hour mean frequency of the calibration oscillator is known to well within one part in one million. The battery voltages are held within such limits that the short-period fluctuations attributable to them are probably negligible compared with the fluctuations due to temperature. A definite correlation was observed between the temperature and the long-period fluctuations of Fig. 7. It is reasonable to assume that the momentary perturbations from the 24-hour mean are not greater than the maximum fluctuation from the mean observed in a long interval, which is of the order of eight parts in one million. Thus if the calibration is performed with a precision defined by 0.2 cycle in 500,000 or more, the normal accuracy at present attainable in adjusting the oscillator under test is about eight parts in one million which is safely within the current requirements. The piezo oscillator has recently been adjusted to a mean frequency of 50000.15 cycles per second, and in most oscillator calibrations the correction to 50 kc is neglected, allowing the adjustment to zero beat of the oscillator under test.

The variations in mean frequency of the piezo oscillator as indicated in Fig. 7, while insignificant as regards present requirements, are of course much too large for a precision standard, and are not representative of the possible performance of the system under the most favorable conditions which modern technique of temperature and circuit control may afford.

In an effort to improve the performance of the standard, a new temperature control box was built, though use was still made of the bimetallic thermo-regulator. The fluctuation of the frequency during the ten-day interval from November 4, 1928, is shown graphically in Fig. 8. The improvement over the performance shown by Fig. 6 is apparent, but the results still leave

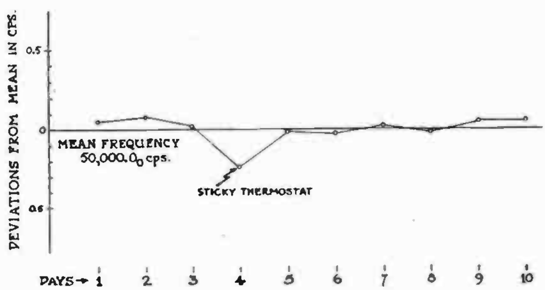

Fig. 8

room for considerable effort. Outstanding lines of development to be pursued are (1) provision of temperature control within one or two hundredths of a degree; (2) replacement of present insulation in the quartz-bar holder by a vitreous material showing negligible distortion with age at constant temperature; (3) improvements in the clock mechanism and in methods of checking against the time signals to provide a higher degree of accuracy in measuring the time interval.

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A SYSTEM FOR FREQUENCY MEASUREMENTS BASED ON A SINGLE FREQUENCY*

By

E. L. Hall

(Radio Section, Bureau of Standards, Washington, D. C.)

Summary—*The method here described is suitable for calibration of either piezo oscillators or frequency meters in terms of an accurately standardized temperature controlled piezo oscillator. A radio-frequency generator is adjusted and maintained at 10 kc in terms of the standard piezo oscillator. The correct adjustment of this generator is shown by a special form of beat indicator which gives both visible and audible indication of the frequency difference between the piezo oscillator and a harmonic of the 10-kc generating set. A second auxiliary generator may then be set to a harmonic of the 10-kc generating set, such as some frequency assignment in the broadcasting band. The correct setting of the auxiliary generator is also determined by means of the special beat indicator. The piezo oscillator to be tested will usually be found to give an audible note in the telephone receivers connected to it. The audible note represents the difference in frequency between the piezo oscillator and the auxiliary generating set. The frequency difference is determined by comparison with a calibrated audio-frequency generator.*

A SIMPLE method of measuring the frequency of a piezo oscillator is to adjust a radio-frequency generator so that it has the same frequency as the piezo oscillator to be tested, and then measure the frequency of the generator in terms of a calibrated frequency meter. This measurement can be repeated at several harmonics if desired, and the fundamental frequency of the piezo oscillator determined in terms of several coils of the frequency meter. In the method here described, the frequency of the piezo oscillator is determined in terms of an accurately known frequency standard, and the result is not dependent upon the calibration of a frequency meter as an intermediate step in the measurement.

At the time work was begun on the development of this system, in March, 1928, the procedure followed in testing a piezo oscillator for a broadcasting station consisted in determining when the quartz plate was adjusted to the frequency assigned to the station. As the frequency assignments end in

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tens between 550 and 1500 kc, it is apparent that an electrical system giving harmonics spaced 10 kc apart could be used to give the desired frequencies without final measurement upon a frequency meter of extreme precision and accuracy.

A radio-frequency generator for 10 kc can readily be constructed using a vacuum tube. The output of such a generator has a large number of harmonics which may be used in measurements at frequencies many times the fundamental. A generator of this type, however, cannot be relied upon to maintain its frequency adjustment accurately enough for use as a standard, in terms of which standards for broadcasting stations are to be adjusted. A piezo oscillator, however, will maintain its frequency with a high degree of precision if it is maintained at a constant temperature. A temperature-controlled piezo oscillator is therefore a desirable standard to use in maintaining a 10-kc generator accurately upon its frequency.

The system herein described is best explained by reference to Fig. 1 which shows schematically the apparatus as used.

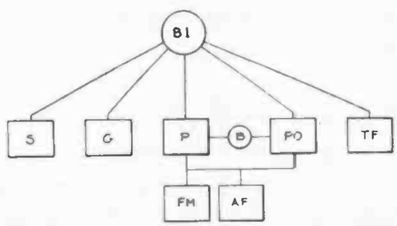

Fig. 1—Schematic Diagram of Apparatus Used in Frequency Measurements. *S*—temperature-controlled piezo oscillator; *G*—10-kc generator; *P*—adjustable frequency power generator; *PO*—piezo oscillator to be tested; *TF*—electrically driven tuning fork; *FM*—frequency meter; *AF*—audio-frequency generator; *BI*—aural and visual beat indicator; *B*—beat indicator.

The system includes: a temperature-controlled piezo oscillator *S*, the frequency of which is accurately known, in terms of which the measurements are made; a 10-kc vacuum-tube generating set *G*, capable of very precise frequency control and giving a very constant frequency; a radio-frequency vacuum-tube generating set *P*, covering the range of frequencies desired; the piezo oscillator to be calibrated, *PO*; an electrically-driven tuning fork of known frequency, *TF*; a calibrated frequency meter, *FM*, for determination of the order of the harmonic of generator *P* in terms of generator *G*; a calibrated audio-frequency generator *AF*, capable of precise adjustment; a special form of combined beat indicator and generating detector *BI*, by means of which the operator maintains the various elements of the system in the frequency relationship desired; a beat indicator, *B*, which can be used for adjusting the piezo oscillator *PO* precisely to the frequency or harmonic of generator *P*.

In brief, the general method of operating the apparatus is as follows. The frequency of the 10-kc generator *G* is controlled manually by changing the setting of a very small variable condenser and is held at a frequency a harmonic of which produces "zero beat" with the fundamental or a harmonic of the standard piezo oscillator *S*. This condition is determined by means of a beat indicator *BI*. The power generator *P* is then set so that it is at zero beat with a suitable harmonic of the generator *G*, as indicated by beat indicator *BI*. After this adjustment generator *P* is operating at the required frequency. The piezo oscillator to be tested is then adjusted to give zero beat with generator *P* as indicated by beat indicator *B*. The frequency meter *FM* is used only to determine the order of the harmonics or to measure frequency differences. The adjustment of the piezo oscillator is based directly on a standard piezo oscillator.

The low-frequency generator *G* is a Bureau of Standards Type O auxiliary generator described in Letter Circular 187,[1] with the addition of a number of fixed condensers and a small two-plate variable air condenser. This generator employs a type 201-A tube. It was found that by using one of the largest coils and an additional capacity of about 0.012 μf, a frequency of 10 kc could be obtained. After enclosing the generator in a box to eliminate sudden temperature changes due to air currents, it was found to hold its frequency adjustment very satisfactorily. The beat indicator circuits give audible indication of the slightest frequency shift. Three different piezo oscillators with frequencies of 25, 100, and 200 kc have been used at different times as the standard in terms of which the 10-kc generator was adjusted.

In order to emphasize some of the merits of the method, reference is made to Fig. 2, which gives a diagram of the connections for a simple beat indicator where the outputs of the two generators *G* and *H* are led into the indicator circuits having

[1] Specifications for portable auxiliary generator, Bureau of Standards Type O, Bureau of Standards Letter Circular 187.

coupling coils *C* and *E*, crystal detector *D*, and milliammeter *A*. Whenever generators *G* and *H* have nearly the same frequency or a simple harmonic relation exists between their frequencies, the pointer of the milliammeter *A* will vibrate, indicating the frequency difference or error in the harmonic relationship. Fre-

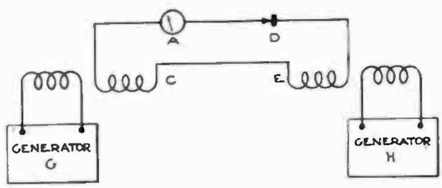

Fig. 2—Circuit for Simple Beat Indicator which Operates When Generator *G* Is Nearly Adjusted to Frequency or A Harmonic of Generator *H*.

quency differences up to a few cycles per second can be observed with the milliammeter of the beat indicator, Fig. 2. A careful adjustment of the frequency of one of the generators will then cause the pointer to vibrate more and more slowly until its deflection does not change. While very precise frequency adjustments can be made in this manner, there may be some tendency for the frequency of the standard to be changed slightly, because of the presence of the coupled indicating circuit. Efforts were made to eliminate the possibility of errors of this type, which lead to the beat indicator shown in Fig. 3.

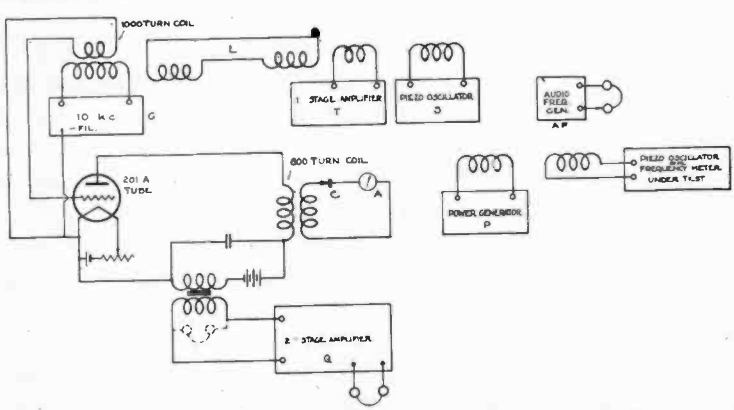

Fig. 3—Diagram of Special Aural and Visual Beat Indicator with Associated Apparatus Used in Frequency Measurements.

This indicator gives visible indications on the milliammeter *A* with very loose coupling between the standard frequency generator S and the low-frequency generator G. When a shielded piezo oscillator is employed at S, Fig. 3, a one-stage radio-frequency amplifier and the link circuit L are required. The use of this form of coupling does not affect the frequency of the piezo oscillator S.

Essentially the beat indicator circuit shown in Fig. 3 is a generating detector circuit. A 1000-turn honeycomb coil in the grid circuit couples to the generator G and to the output of the piezo oscillator S. A 600-turn coil in the plate circuit couples to a coil of 8 or 10 turns in series with a crystal detector and d.c. microammeter A, which gives the visible indication. An audio-frequency transformer is connected in the plate circuit as shown. Telephone receivers or a two-stage amplifier are connected to the secondary winding of the transformer.

To operate the beat indicator the approximate setting of the 10-kc generator is first found by listening in the telephones connected in the plate circuit of the 10-kc generator or as shown by the dotted lines, in which case the two-stage amplifier would be disconnected, or by listening in telephones in the output of the amplifier. After adjusting the 10-kc generator to zero beat, with the telephones connected to the output of the amplifier, a rather faint high note may be heard, or, if the power generator, P, Fig. 3, is operating near some harmonic of the 10-kc generator rapid pulsations in the beat note between the two generators are heard. This renders it much easier to adjust the 10-kc generator correctly. By careful adjustment of the tuning control of the 10-kc generator this high note will fluctuate in intensity, and when the fluctuation becomes slow enough the pointer of the milliammeter A will be found to vibrate in step with the fluctuation. By further adjustment of the tuning control the pointer can be kept from moving. At this point the 10-kc generator G can be shifted by further slight adjustment so that there is either a high-pitched faint note or absolute quiet. The latter condition occurs when the generator is correctly adjusted with respect to the frequency standard used. If the generator gets out of adjustment by as much as one cycle in a minute, the operator will detect it by hearing the high-pitched note appear in the telephones. Slight errors in this setting may not be important unless a high order harmonic is to be used. Errors in adjustment of the order of a few hundredths of a cycle are immediately apparent and the generator can be readjusted.

The sensitiveness of the visible beat indicator is shown by the fact that when the coils of an unshielded piezo oscillator and 10-kc generator were coaxial and from two to three feet apart, visible beats of the beat indicator milliammeter *A* (Fig. 3) were readily obtained. Type 201-A tubes were used in both the piezo oscillator and the 10-kc generator. The link circuit *L* was not used in this case. In other words, the coupling to the piezo oscillator *S* was extremely loose, which meant that there was no danger of the frequency of the piezo oscillator (which was used as the standard) being changed by the other circuits. Thus the beat indicator system could not affect the frequency of the standard.

Having adjusted the 10-kc generator accurately, the next step is to adjust the third generator, *P*, in Fig. 3,called the power generator, to the desired harmonic of the 10-kc generator. A vacuum-tube detector and amplifier and coil system coupling the two generators would give beat notes in telephones connected in the output circuit of the amplifier, but it has been found unnecessary to add this extra circuit when working up to 1500 kc using a 10-kc fundamental. The telephones in the two-stage amplifier *Q* in Fig. 3 will also serve to indicate the correct adjustment of the power generator. When working at the higher frequencies in the broadcast band, beats which otherwise would not be heard are made audible by placing a pick-up coil and the phone leads in the vicinity of the power generator.

The frequencies of the power generator ending in tens of kilocycles give a zero beat which cannot be set accurately by aural methods because of the uncertainty of the setting for true zero beat. The frequencies ending in 2 1/2, 5, and 7 1/2 can be set accurately by matching beat notes in the telephone receivers, which are beats of beats.

Although the 10-kc points cannot be set precisely, they can be accurately determined by matching a beat note in the phones with a corresponding note from a tuning fork. If the frequency of the fork is known, it is only necessary to make the measurement on one side of zero beat. If this method is adopted, great care is necessary to insure that the desired fork frequency is used, as the measurement may be made by matching the generator beat note with a submultiple or harmonic of the tuning fork. In order to guard against this error it has been found convenient to make an approximate zero beat measurement first.

Then a measurement is made with the tuning fork by detuning the generator so that the frequency heard in the telephones due to the beat note between the 10-kc generator and the generator is the same as that of the tuning fork. It is necessary that the frequency meter used have sufficient resolving power to show an easily read difference in the condenser settings for these two frequencies, and the curves should be large enough to allow a correct determination of what harmonic of the fork is used.

If it is desired to prepare an enlarged frequency meter curve, it is evident from the above paragraph that a large number of points can be obtained for the calibration over as small a frequency range as 10 kc. Assuming that a 400-cycle electrically-driven tuning fork is available, in the range from 200 to 210 kc, for example, the following points could be very accurately determined using the beat notes 199.200, 199.600, 200.400, 200.800, 202.500, 204.600, 205.000, 205.400, 207.500, 209.200, 209.600, 210.400, and 210.800. These thirteen points permit the drawing of a large accurate curve.

The circuits described above have been used for calibration purposes between 40 kc and 1500 kc, and beat notes have been obtained up to 2500 kc, spaced 10 kc apart. In the lower part of the frequency range beat notes were obtained 2500 cycles apart when using a 25-kc piezo oscillator and a 10-kc fundamental from generator *G*.

The measurement of the frequency of a piezo oscillator can be made by the method described with the addition of an accurate audio-frequency generator. For example, suppose the frequency of a piezo oscillator is to be accurately measured, and it is known that its frequency is approximately 600 kc. The power generator *P* of Fig. 3 is set to the sixtieth harmonic of the 10-kc generator as determined by a common type of frequency meter. Then when listening in the phones of the piezo oscillator a beat note will, in general, be heard, which is the difference between the frequency of the piezo oscillator and the generator. This beat note is matched by a similar note from the audio-frequency generator, from the calibration of which the audio frequency is determined. Whether this audio-frequency is to be added to or subtracted from 600 kc is determined by listening in the piezo oscillator telephones and noting whether the frequency of the power generator must be increased or decreased to produce zero beat. Check measurements can be made at other harmonics or subharmonics of the piezo-oscillator frequency.

It has been found desirable for checking purposes to take a reading on a frequency meter of high resolving power at the time that the audio-frequency beat note measurement is made. A certain frequency value is obtained from the calibration of the frequency meter. Generator P is then adjusted to zero beat with the piezo oscillator under test, and another reading taken on the frequency meter. These two frequency meter readings give a frequency difference which serves as a check on the frequency measured with the audio-frequency generator, and also indicates whether the frequency as determined by the latter generator is to be added to or subtracted from the frequency as given by generator P, which, as previously stated, is based on the standard piezo oscillator.

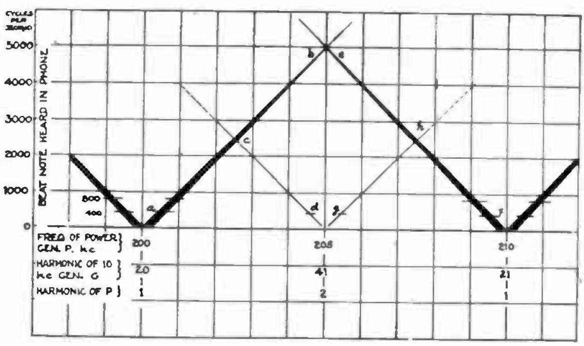

Fig. 4—Chart Showing Example of Relations between Beat Notes Heard in the Special Beat Indicator Telephones, with the Harmonics of Generators G and P Indicated.

The frequency of a quartz plate mounted in an adjustable mounting can readily be altered to the desired value by setting the power generator to the correct harmonic and then adjusting the quartz plate mounting to give zero beat. A beat indicator can be brought into use for a more accurate adjustment than is possible by use of the telephones alone.

Fig. 4 illustrates how the various beat notes heard in the telephones can be interpreted, i.e., whether the frequency to which the power generator is set ends in 10 or in some other number. The diagonal lines represent the beat notes heard in the phones, and the breadth of the line its relative strength. Thus it is seen that the beat notes occurring near the frequencies ending in 10 are very loud or strong. Zero beat being indefinite, a setting is taken by matching with the note from an electrically-driven tuning fork as described above. Increasing the frequency of the power generator from 200 kc, for example, as indicated in Fig. 4, the beat note gradually gets higher and fainter, line *ab*, until at a frequency of the order of 204.5 kc the beat note is quite high, but a lower note, line *cd*, will also be heard which becomes lower as the other becomes higher. Soon after the low note has become silent the high note will be heard to pulsate, and, with critical adjustment of the generator, beats with another beat note will be heard which permit a very accurate setting of the generator. This setting gives one of the 5-kc points and is readily distinguished over a range of several hundred kilocycles by the aural method described. If the frequency of the power generator is increased further the other high-frequency beat note, line *ef*, becomes lower and stronger, and a weaker beat note, line *gh*, becomes higher and fainter. Further change in the power generator brings in a loud beat note and finally silence or zero beat in the telephones. This is another 10-kc point. The beat note represented at *b* is 5000 cycles when a 10-kc fundamental is used. When changing the frequency of the power generator from *a* to *b*, if conditions are right, another beat note may be heard at *c* and a precise setting made at this point which will be midway between the 10-kc point and the 5-kc point. This setting is more difficult to obtain than the 5-kc point because it is indicated by the beats between two 2500-cycle notes of much different intensities. When using a piezo oscillator having a fundamental frequency of 25 kc, the 2500-cycle notes are readily distinguishable. Other beat notes may be heard and can be set accurately but may not be readily usable. Points 5 kc apart are much closer than are usually required.

An explanation of the various beat notes heard in the phones can be given by further reference to Fig. 4 where harmonic numbers are indicated for the three zero-beat frequencies given. The 200-kc frequency or fundamental of the power generator beats with the twentieth harmonic of the 10-kc generator, and the 210-kc fundamental of the power generator beats with the 21st harmonic of the 10-kc generator. When the frequency used is 205 kc, the second harmonic of the power generator, or 410 kc, beats with the 41st harmonic of the 10-kc generator.

As previously explained, the 202.5-kc and 207.5-kc points are obtained by listening to the beats between the beat notes heard in the phones. Beat notes other than those indicated in Fig. 4 may be heard in the phones, but those described are usually more than are required for calibration purposes.

While this system or method was intended chiefly for calibration of piezo oscillators, it has been found to be ideal for frequency meter calibration work. The system has the following advantages:

(1) Extreme accuracy and precision obtainable.
(2) Large number of calibration points possible.
(3) Flexibility and ease of operation.
(4) Ease of working up final results from experimental data.

With reference to (1), the accuracy with which the frequency of the standard is known and maintained will determine the accuracy with which the measured frequencies can be determined, because of the extreme precision possible in working with beat notes. This precision is far in excess of even specially built frequency meters. While the 10-kc generator is not held automatically to its frequency, yet the characteristics of the circuit are such that the operator is immediately aware of any change in its frequency and can readjust it.

Item (2) above is important because the system permits calibration of piezo oscillators for broadcasting stations for any frequency in the broadcasting band. It extends the usefulness of a standard piezo oscillator since it allows the exact setting of many harmonics which could not be obtained from the piezo oscillator alone. This fact is of great importance in the calibration of special frequency meters capable of extreme precision and having high resolving power such as one used in the radio laboratory of the Bureau of Standards. The large number of calibration points possible can be seen from the example cited above. This fact also adapts the circuits to the measurement of the frequency of any piezo oscillator by final measurement of an audio frequency by means of an audio-frequency generator.

With reference to (3), after the apparatus is correctly adjusted it is very easy to operate and adapts itself to any frequency calibration. The frequencies to be adjusted can be spaced uniformly as desired by setting what has been called the 10-kc generator to frequencies other than 10 kc, such as 5 kc, 15 kc, 20 kc, 25 kc, etc. For example, it may be that points spaced 100 kc apart are desired. The 10-kc generator in that case would be set to 100 kc.

With reference to (4), data on frequency meter calibration can be worked up and results obtained almost by inspection if the frequency of the piezo oscillator used as a standard is such as to permit the setting of the 10-kc generator accurately on 10 kc, and provided the experimental data is carefully taken and the same procedure is used in matching the fork frequency. Ten-kc and 5-kc points are readily distinguished experimentally, and existing curves can be checked rapidly or corrected by adding 10 kc to successive 10-kc points starting from a known point. By adding 10 kc to successive 5-kc points starting at a known 5-kc point these frequencies are rapidly determined.

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RECEIVING SETS FOR AIRCRAFT BEACON AND TELEPHONY*

By

HARADEN PRATT[1]† AND HARRY DIAMOND[2]

([1]Formerly Radio Engineer, Bureau of Standards, Washington D. C.; [2]Associate Radio Engineer, Bureau of Standards, Washington, D. C.)

Summary—*Special light-weight radio receiving sets of high sensitivity have been designed adapted for the reception on aircraft of signals from radio beacons and ground radio telephone stations. A statement of the important considerations involved in the aircraft reception of signals from Government-operated beacons and telephone stations along the civil airways in this country, such as frequency range, tuning means, volume control, vacuum tubes, amplifier characteristics, size, weight, and performance specifications, is given.*

Design details for three receiving sets of slightly different types, with numerous characteristic and performance curves, are discussed, with a brief discussion of the results of practical flight tests.

INTRODUCTION

THE conditions surrounding the use of radio receiving equipment installed on aircraft impose many special considerations of both design and performance. The specifications for aircraft sets, therefore, differ materially from those employed for other purposes.

Such sets at this time are of three types. One type fulfills military requirements where a rather broad band of frequencies must be covered while in flight. Another type is required to receive the signals from beacons and ground radiotelephone stations which the United States Government will supply along the civil airways under the air commerce act. The third type is adapted to the high-frequency (above 1,500 kc) field. Most of the activities of the past have, in this country, been confined to sets of the first type mentioned. The other two fields are new.

European radio developments applied to commercial aviation have been under way for several years. Resulting practice has centered around a narrow band of frequencies between 315 and 350 kc for aircraft services. Radio beacons, particularly those

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† Chief Engineer, Mackay Radio and Telegraph Company, New York City.

erected in the United States for aiding marine navigation, have adopted adjacent frequencies near 300 kc. The 1927 International Radio Conference recognized these established practices and the International Convention allocated the bands 285 to 315 kc for all radio beacon services and 315 to 350 kc for aircraft communication services, 333 kc being adopted as an international calling and distress frequency for aircraft. The directive radio beacons and weather broadcasting services planned for our civil airways will, therefore, be confined to these two bands.

These bands being adjacent and covering a total range of only 65 kc permit of simplifications in receiving set specifications for this type of service. As part of its research and development work[1] on directive radio beacons and radiotelephony for aircraft under the air commerce act, the Bureau of Standards has developed such receiving sets, the design elements of which are treated in this paper.

Requirements for Beacon and Telephone Receiving Set

The dominating requirement in design is that of small weight and physical dimensions. This refers not only to the set, but also to the batteries requisite for its operation. Every design element must be considered in the light of this requirement.

The set must be capable of installation anywhere on the aircraft with simple remote controls when placed at an inaccessible point. The tuning system should, therefore, be of a uni-control type over the entire 285 to 350 kc band for the sake of convenience.

As an aircraft is in very rapid motion, the distance over which signals are traveling is constantly changing. A simple and rugged volume control having a uniform action from zero to maximum signal strength and capable of remote installation from the receiving set is consequently necessary.

Sufficient shielding of the set must be provided to limit interference from the engine ignition system to that induced in the antenna structure.

Rugged construction to withstand the continuous mechanical vibration on aircraft must be provided. Vacuum tubes must be used that do not introduce microphonic noises, which, if present, effectively increase the already prevailing noise level.

[1] J. H. Dellinger and H. Pratt, "Development of Radio Aids to Air Navigation", Proc. I. R. E., **16**, 890–920; July, 1928.

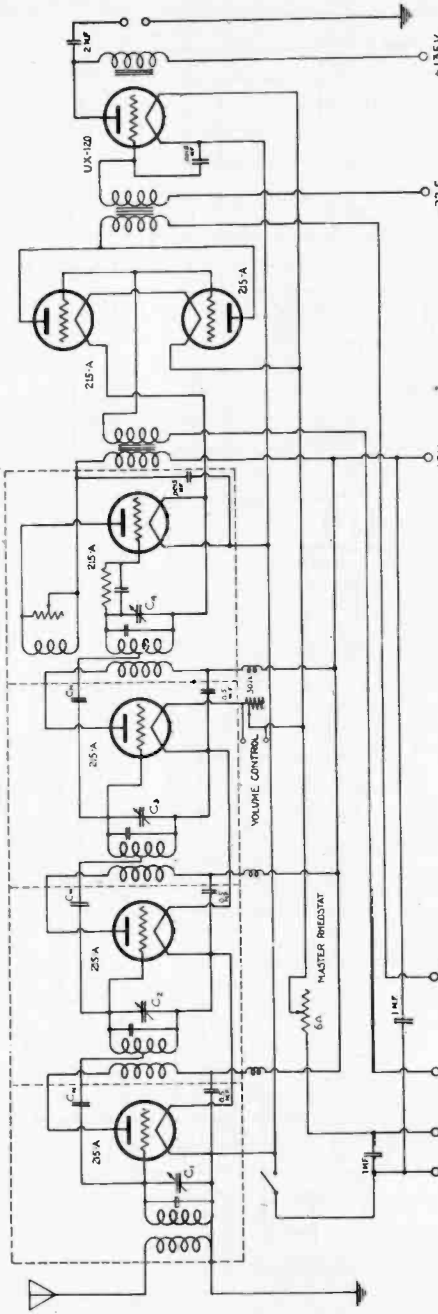

Fig. 1—Circuit Diagram of Completely Shielded Receiving Set.

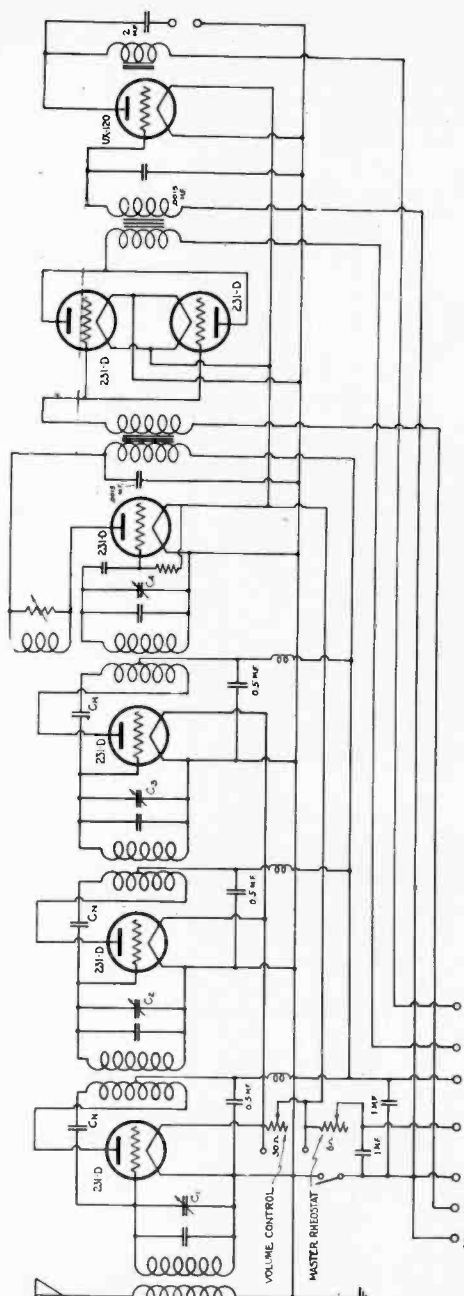

Fig. 2—Circuit Diagram of Partially Shielded Receiving Set Using 3-volt Tubes.

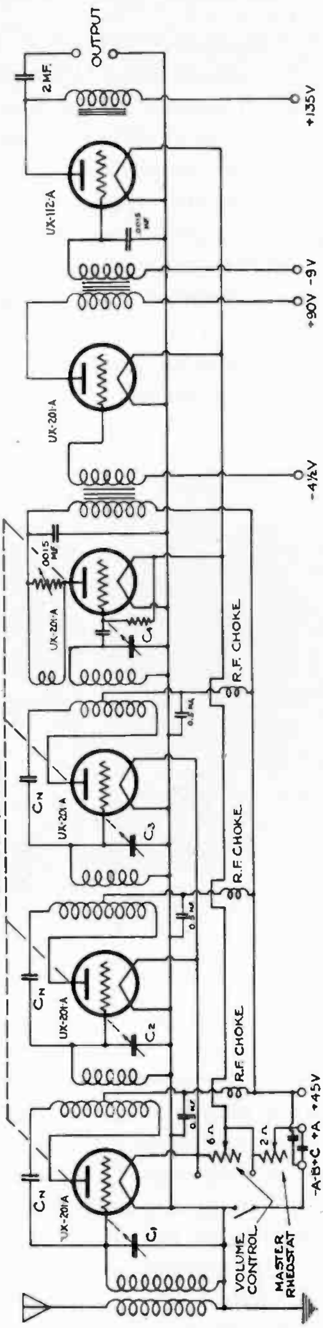

Fig. 3—Circuit Diagram of Partially Shielded Receiving Set Using 5-volt Tubes.

The set must be capable of operating a beacon course indicating instrument. Its output must efficiently supply a maximum of 10 volts of audio-frequency power into a load impedance of from 4,000 to 7,000 ohms, to provide an ample margin of signal strength. The direct-current plate supply must be isolated from the output circuit to prevent the observer from receiving shocks when wearing headphones, and to prevent the setting up of a polarization in the course indicator's magnetic system.

The audio amplifiers and output circuit must provide a nearly uniform amplification over a frequency range of 40 to 3,000 cycles. The beacon signals are modulated in the range 40 to 120 cycles, and a voice frequency range from 200 to 3,000 is needed for good intelligibility.

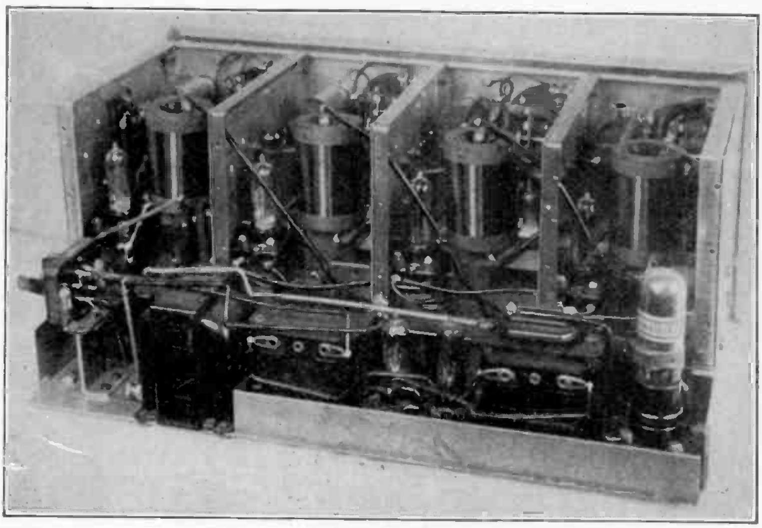

Fig. 4—Interior of Completely Shielded Receiving Set.

Both the sensitivity and selectivity must be high. High sensitivity is desirable to permit short vertical pole antennas to be employed because they eliminate direction errors and remove the danger of weighted trailing wires. Good selectivity is necessary to enable close spacing of frequency channels.

The choice of vacuum tubes has an important bearing on the total weight and bulk of the equipment. Three types of receiving sets will be described; two of which employ 1-volt and 3-volt tubes, respectively, where the total *A* battery current is of the order of 0.5 ampere; the third using 5-volt tubes where the *A* battery must supply 1.5 amperes. By employing the lower voltage types both the *A* storage battery can be reduced in size and the advantages of vacuum tubes having small physical dimensions secured. The standard low-voltage tubes offered on the market, however, are not sufficiently sturdy for aircraft use, being often mechanically weak and usually causing microphonic noises. Special tubes having these disadvantages to a less degree were employed, but owing to their special nature the third set

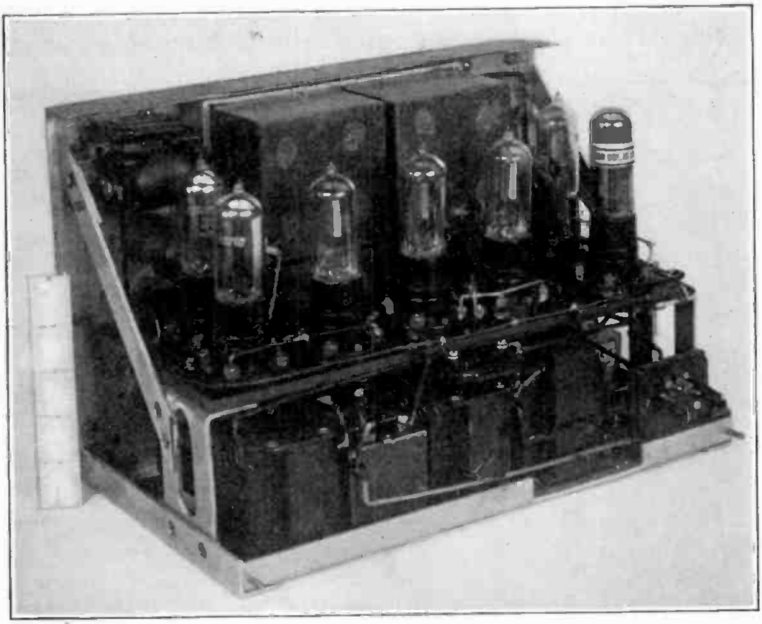

Fig. 5—Interior of Partially Shielded Receiving Set Using 3-volt Tubes.

equipped with the standard 5-volt tubes was also designed. There is much need for better tubes for aircraft sets, and it is hoped that commercial designs will be improved.

Receiving Set Design

Figs. 1, 2, and 3 show the circuit diagrams of the three receiving sets designed to meet the above requirements, and Figs. 4, 5, and 6 are corresponding photographs. Their physical characteristics are tabulated below.

TABLE I

WEIGHTS AND DIMENSIONS OF RECEIVING SETS WITH ASSOCIATED BATTERIES

Receiving Set No.	Weight of Set (pounds)	Dimensions (inches)	Weight of batteries (pounds)
1	18	9½ by 7¼ by 16½	19
2	16	8½ by 9 by 14	19
3	17	6¼ by 8 by 13	25

Sensitivity, selectivity, and unicontrol features are met by employing three or four tuned radio-frequency amplifier circuits with gang variable air condensers. Some regeneration in the detector circuit is provided. After many trials superheterodyne sets were eliminated from consideration owing to the presence of two tuning

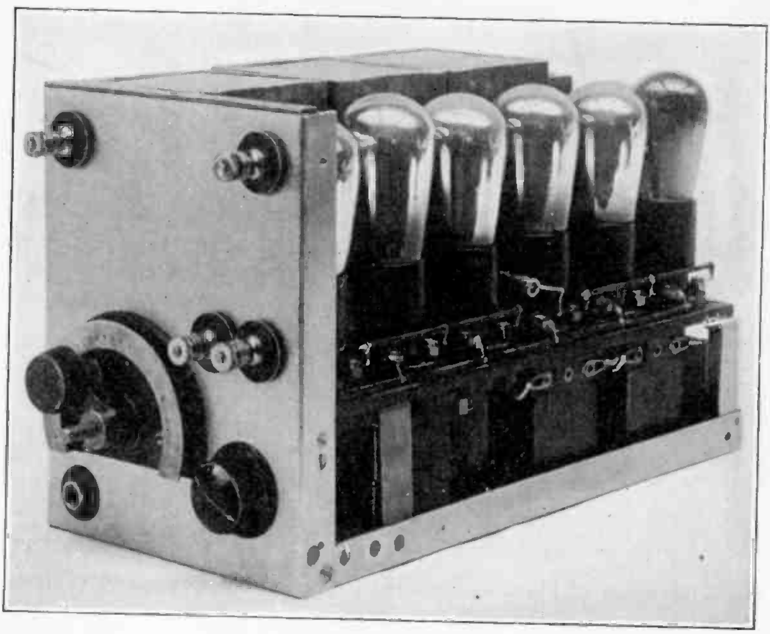

Fig. 6—Interior of Partially Shielded Receiving Set Using 5-volt Tubes.

positions for each frequency. A few hours of flight would usually bring the aircraft within close range of stations having frequencies such as to create interference by beating with the intermediate frequency or its harmonics. A pilot operating a radio-equipped airplane in bad weather, when the radio services are needed, must devote all his attention to piloting, and the radio set must convey the information it receives without attention and without interference from other signals or it will lose usefulness and possibly become a hazard.

1. Design of Radio-Frequency Transformer (Interstage). Radio-frequency transformer theory has for some time been reduced to mathematical treatment.[2] A brief consideration of the analysis involved is necessary to a clear understanding of the actual design procedure employed. Fig. 7 shows the commonly accepted equivalent circuit corresponding to a stage of radio-frequency amplification employing tuned transformer coupling between stages; R_1 represents the tube plate resistance plus the radio-frequency resistance of the transformer primary, L_1 the primary inductance, M the mutual inductance between windings, and R_2, L_2, and C_2 the constants of the secondary circuit.

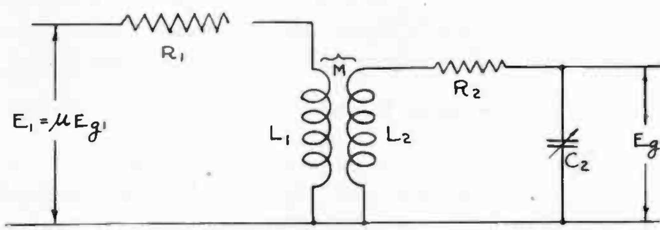

Fig. 7—Accepted Approximate Equivalent Diagram of a Stage of Radio-Frequency Amplification.

In transformer design those values of the three-circuit parameters X_1, X_2, and M which will make the voltage gain per stage a maximum at the desired frequency are of interest. If X_2 alone is varied (by means of the tuning condenser C_2) the voltage gain is a maximum when

$$\frac{X_2}{X_1} = \frac{\omega^2 M^2}{Z_1^2} \tag{1}$$

If, now, either X_1 or M is varied, X_2 being simultaneously varied in order to satisfy the condition expressed by (1) for each value of X_1 (or M), the voltage gain will reach an optimum value when

$$\frac{R_2}{R_1} = \frac{\omega^2 M^2}{Z_1^2} \tag{2}$$

Equations (1) and (2) may be combined into the simpler relationship

$$\omega^2 M^2 = Z_2 Z_1 \tag{3}$$

[2] Victor G. Smith, "A Mathematical Study of Radio-Frequency Amplification," Proc. I. R. E., 15, 525–536; June, 1927. G. W. Pierce, "Electric Waves and Oscillations", McGraw Hill Book Co. (Inc.).

When this relationship is fulfilled the voltage gain K is given by (4)

$$K_{\text{optimum}} = \frac{1}{2} \frac{\mu}{\sqrt{R_1}} \frac{\omega L_2}{\sqrt{R_2}} \tag{4}$$

and is the maximum amplification that can be obtained with a given tube and coil.

The actual transformer design closely followed the above theoretical considerations.

Two secondary coils (single-layer solenoids) were chosen, having suitable values for $\dfrac{\omega L_2}{\sqrt{R_2}}\cdot$ The first coil consisted of 135 turns of No. 28 single silk enameled wire wound on a 2¾-in. diameter coil (winding length 2¾ in.) and having a self-inductance of 1 millihenry, a radio-frequency resistance of 18 ohms at 290 kc, and a value for $\dfrac{\omega L_2}{\sqrt{R_2}}$ at the same frequency equal to 430. The second coil was of 250 turns of No. 34 enameled wire wound on a 1¾-in. diameter coil (winding length 1¾ in.) and having a self-inductance of 2 millihenries, a radio-frequency resistance of 70 ohms at 290 kc, and a value for $\dfrac{\omega L_2}{\sqrt{R_2}}$ equal to 435 at this frequency. It should be noted that to cover the necessary frequency range of 285 to 350 kc with a comfortable margin on each side a tuning condenser of 0.00035 μf maximum capacity is required with the first coil, and a condenser of 0.00015 μf maximum capacity with the second coil.

Since $\dfrac{\omega L_2}{\sqrt{R_2}}$ is very nearly the same for each coil, the same optimum voltage amplification should be obtained using either coil in conjunction with a given tube. The second coil, however, has the considerable advantage of compactness. Furthermore, since its selectivity is less than that of the first coil (due to its greater resistance), it becomes somewhat more adaptable for use in a unicontrolled multistaged amplifier.

To determine the optimum value of mutual inductance between windings for each transformer suitable primary forms were chosen and test amplifiers constructed comprising the particular transformer under test together with a given tube. A known voltage of the desired frequency was applied to the input terminals of the test amplifier and the transformer secondary circuit tuned by condenser C_2 for maximum output voltage (as indicated by a tube voltmeter); these measurements being taken for a series of values of mutual inductance. For each measurement, therefore, (1) above is satisfied; while at the particular value of M which yields the optimum output voltage, (2) is, in addition, fulfilled. This is the point of optimum mutual inductance.

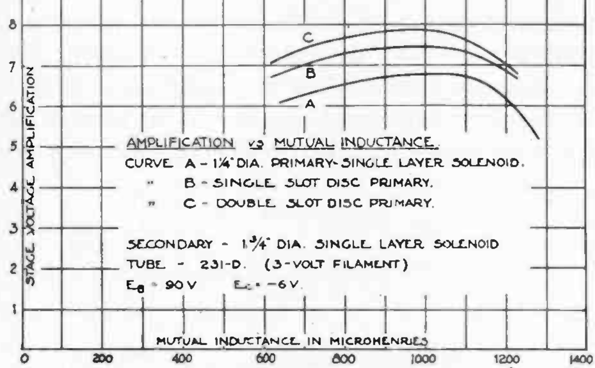

Fig. 8—Amplification vs. Mutual Inductance, 1¼-in. Diameter Primary Single-Layer Solenoid.

Curves A of Figs. 8 and 9 show the variation of voltage amplification with mutual inductance for the 1¾ and 2¾ in. diameter

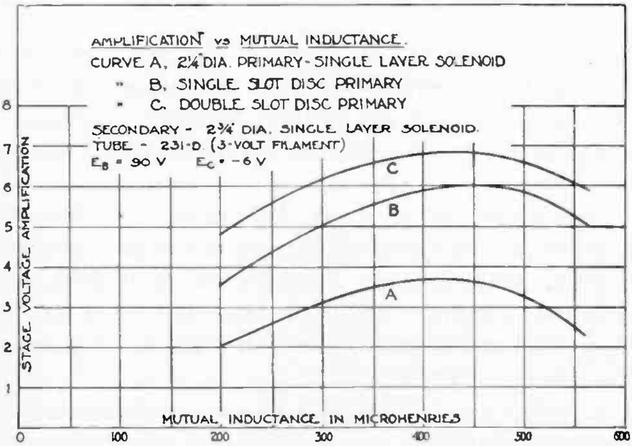

Fig. 9—Amplification vs. Mutual Inductance, 2¼-in. Diameter Primary Single-Layer Solenoid.

transformers, respectively. The amplification obtained with the two transformers is seen to be decidedly different despite the nearly equal values of $\frac{\omega L_2}{\sqrt{R_2}}$ in each case, and the exactly similar conditions of test; the superiority of the smaller transformer residing in the relatively lower capacitive coupling existing between its windings by virtue of their smaller physical dimensions. That this is true may be seen from a consideration of curves B and C of Figs. 8 and 9. Curves B were obtained with the primaries wound in single slotted disks inserted at the low potential ends of the secondaries, thereby reducing the effective capacitive couplings. The resultant increase in amplification is greater for the $2\frac{3}{4}$-in. diameter transformer because of the greater reduction in its capacitive coupling. Distributing the primary windings

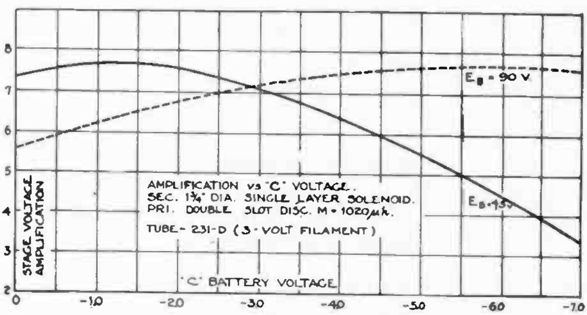

Fig. 10—Amplification vs. C Voltage.

in two slots in the disk form causes a still further improvement in amplification, the percentage increase being again in proportion to the decrease in capacitive coupling between windings.

The curves of Figs. 8 and 9 were taken for 90 volts on the plate with a grid bias of -6 volts. It was decided to reduce the plate voltage sufficiently to preclude the use of a C battery and thereby to remove the need of filtering in the grid return leads. Fig. 10 shows that the same amplification as in Figs. 8 and 9 can be obtained using 45 volts on the plate and $-1\frac{1}{2}$ volts on the grid. A further reduction of the C voltage to zero results in but a small loss in amplification.

Fig. 11 gives the amplification-frequency characteristics for the test transformers; the optimum mutual inductance as found from the curves of Figs. 8 and 9 being provided in each case. Curve A is for the $2\frac{3}{4}$-in. diameter transformer with double-slotted disk primary; curve B for the 1¾-in. diameter transformer with single layer primary; and curve C for the 1¾-in. diameter transformer with double-slotted disk primary. All the important

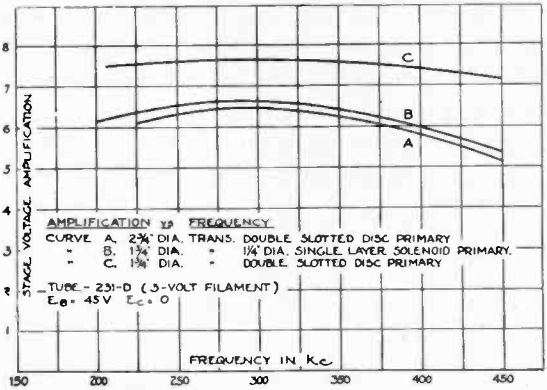

Fig. 11—Amplification vs. Frequency.

constants of these transformers are collected in Table II. A study of the amplification-frequency characteristics shows that they are highly satisfactory over the desired frequency range (285 to 350 kc).

TABLE II
TRANSFORMER CONSTANTS

Transformer No.	Primary Winding	Outside Diameter	Primary Turns	Size of Wire	Secondary Winding	Outside Diameter	Secondary Turns
		Inches				*Inches*	
1	Double-slot disk..	2⅝	135	No. 38 S. C. C.	Single-Layer solenoid	2¾	135
2	Single-Layer solenoid	1¾	250	No. 34 enamel.	..... "	1¾	250
3	Double-slot disk..	1½	250	No. 38 S. C. C.	..... "	1¾	250

TABLE II (cont'd.)
TRANSFORMER CONSTANTS

Transformer No.	Size of Wire	L_1 (μh)	L_2 (μh)	R_1 at 290 kc	M (μh)	Capacitive Coupling between Winding μμf	$\omega^2 M^2$ at 290 kc	Z_1Z_1 at 290 kc, approximately
1	No. 28 S. C. C. enamel	2,930	950	18	450	18	0.67×10^6	0.34×10^6
2	No. 34 enamel..	1,100	1,990	70	1,065	18	3.77×10^6	1.27×10^6
3	....do........	4,290	1,990	70	1,020	6	3.44×10^6	1.37×10^6

In Fig. 12 are shown the resonance curves of the above three transformers, used in conjunction with a 3-volt tube. Curves A, B, and C correspond to the transformers of curves A, B, and C, respectively, of Fig. 11. It is interesting to note the mutual inductance effect upon the selectivity. Fig. 13 indicates how the selectivity varies for four values of M, using the $1\frac{3}{4}$-in. diameter transformer with double-slotted disk primary.

2. Effect of Capacitive Coupling. It was shown above that the presence of capacitive coupling between windings adversely affects the efficient operation of a transformer. When capacitive coupling exists, the approximate equivalent circuit diagram be-

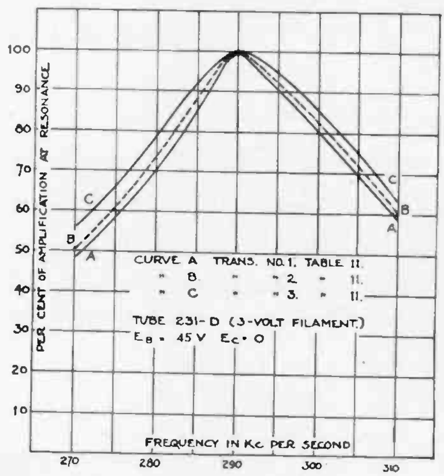

Fig. 12—Amplification vs. Frequency.

comes as shown in Fig. 14 rather than that of Fig. 7. The conditions for optimum voltage gain and the actual magnitude of this optimum gain as found from a solution of this circuit are also different.

A close agreement between experimental results and the theoretical equations based on the circuit of Fig. 7 cannot, therefore, be expected unless the effect of capacitive coupling is minimized. As computed from (3) the optimum mutual inductance for transformer Nos. 1, 2, and 3 of Table II should be 315, 615, and 615 μh, respectively. Actually the values obtained in the laboratory are 450, 1,065, and 1,020 μh, respectively.

The effect of capacitive coupling between transformer windings is thus not only to reduce the magnitude of the optimum voltage gain but also to require considerably more mutual inductive coupling in order to obtain the best operating condition.

3. Method of Neutralizing. The system of tube grid-plate capacity neutralization adopted depends considerably upon the type of transformer employed. The neutralizing schemes in common use are essentially bridge circuits in which two coils, the grid-plate tube capacity, and a neutralizing condenser constitute the four bridge arms. The bridge maintains a balanced adjustment for all operating frequencies, provided the two coil arms are tightly coupled. When the coupling between the two coils is small a leakage inductance is introduced into one of the bridge arms, thus making the balance dependent upon the frequency.

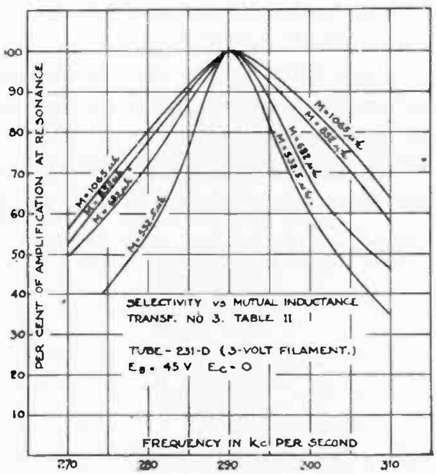

Fig. 13—Selectivity vs. Mutual Inductance.

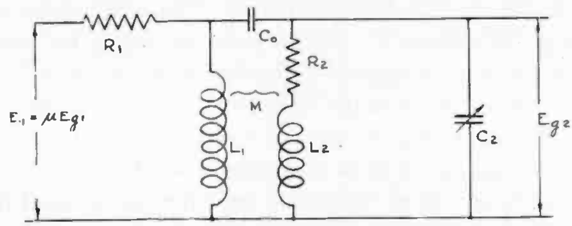

Fig. 14—Closer Approximate Equivalent Diagram of a Stage of Radio-Frequency Amplification.

With several neutralizing systems the primary and secondary windings of the transformer may be used as the two coil arms, provided there is sufficient coupling between them. In the case of the 1¾-in. diameter transformer with single-layer primary, a coupling coefficient of the order of 70 per cent obtains. The system employed can, therefore, be that indicated in the circuit diagram of Fig. 1.

In the case of the 1¾-in. diameter transformer with disk primary (used in the receiving set of Fig. 2), however, the coupling coefficient between windings was about 35 per cent. The use of an auxiliary coil coupled either to the primary or secondary winding is therefore necessary. A simple solution consisted in placing the auxiliary coil in a third slot in the primary disk form thus obtaining nearly unity coupling to the primary. An alternative method, due to Miller, is also applicable in this case. In this method the four bridge arms are made up of the tube grid-plate capacity, a neutralizing condenser, and two auxiliary condensers, one of which may be the tube grid-filament capacity. The closeness of coupling between primary and secondary windings is here obviously immaterial.

4. Design of Antenna Transformer. No mention has as yet been made concerning the design of the antenna coupling transformer.

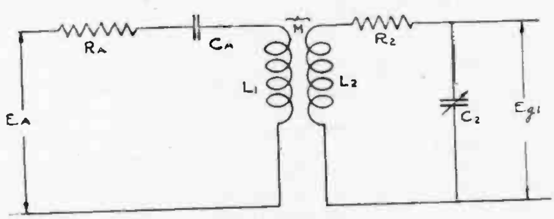

Fig. 15—Approximate Equivalent Diagram of Antenna Tuning System.

The theory and procedure of design is essentially the same as for interstage transformers. The equivalent circuit is that of Fig. 15, where C_A is the antenna capacity and R_A the equivalent antenna resistance. Receiving set sensitivity of such order was desired as to make possible the use of a rigid antenna of about 10 feet in length, with the airplane itself as a counterpoise. The capacity of such a system is of the order of 25 to 50 $\mu\mu$f. Using these values, the same transformer as used for interstage coupling was found highly satisfactory.

5. Design of Regenerative Coil. To obtain the order of amplification desired, it was decided to use three stages of radio-frequency amplification with regeneration on the detector. The circuit arrangement of Fig. 16 was used in the design of the regenerative coil. A known modulated voltage E_1 (of the desired radio-frequency) was applied to the input circuit and the voltage

E_2 measured first without regeneration and then with regeneration. The ratio of the second to the first value of E_2 is the amplification due to regeneration. These measurements were taken for a series of values of regenerative coil turns, adjusting the shunting resistance R to just below oscillation at each point (where necessary). For the particular transformer considered, it was found that the best value of turns was approximately 40 per cent of the number of turns in the secondary winding.

6. Shielding and Arrangement of Parts. Both complete and partial shielding methods were employed in the tuned amplifier stages. Reference to Figs. 4, 5, and 6 will indicate the shielding arrangements. Completely shielding the stages as in Fig. 4 was found to be difficult owing to the involved mechanical assembly. Aluminum was selected to reduce weight, but resulted in objectionable eddy-current losses. Partial shielding using copper containers housing the radio-frequency transformers, with proper care taken to make all high potential leads as short as possible, resulted satisfactorily. This latter method was applied as shown in Figs. 5 and 6.

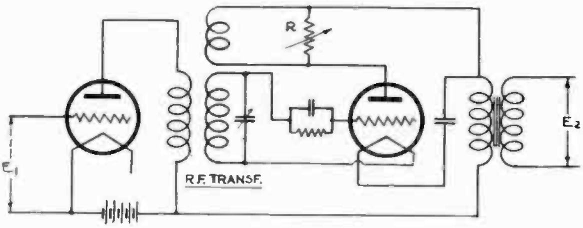

Fig. 16—Test Circuit Arrangement for Measuring Amplification Due to Regeneration.

The copper containers maintain an effective clearance of ⅞ in. around the coil windings, and are of square cross section. The coil inductances are decreased by about 20 per cent without any material increase in resistance.

To confine the radio-frequency currents as far as possible to the shielded portions of the set, battery leads were filtered where necessary. The selection of a low-plate voltage served not only to reduce the demand on the B battery, but to permit the radio-frequency stage grid returns to be connected directly to the negative filament terminals, eliminating a common C battery impedance. All other precautions in arrangements and wiring used in good design practice were observed. A multiple conductor battery cable, equipped with a detachable plug connector, was provided to allow a rapid change of sets to be made when in practical operation.

A small separate panel containing a combination *A* and *B* battery voltmeter, filament switch, head telephone jack, and volume control was provided for mounting near the pilot, preferably on the instrument board. The volume control consists of a rheostat connected in the radio-frequency amplifier tube filament supply circuit.

The unicontrol tuning feature makes it possible for the operator to tune to any station frequency within the two bands. In this case an antenna coupling arrangement is needed which makes the setting of the antenna tuning condenser at a given frequency independent of the antenna length. This is taken care of by an antenna trimming condenser.

Adjustable stops to allow the unicontrol handle to be rapidly turned from one predetermined frequency in the 285 to 315 kc band to one in the 315 to 350 kc band were provided.

An alternative tuning arrangement was incorporated in one of the receiving sets built, applicable for the case where only one beacon and one telephone frequency would be encountered on any one airplane run. Two sets of condensers connected in parallel were used in each tuned amplifier stage. Normally with both condensers in circuit the tuning adjustment is made for the beacon station frequency. When a switch is closed by the operator four small automatic relays respond, removing one condenser from each circuit, leaving the set tuned to the telephone station frequency which is in the higher band.

7. Detector. In general, grid rectification has been found considerably more sensitive than plate detection, and was, therefore, used. Laboratory measurements gave 3 megohms as the optimum value for the grid leak and 0.00035 μf as the best value of blocking condenser at the frequencies used. These values are not at all critical.

Sensitivity, however, is not always the deciding factor in the choice of the type of detection to be employed. When used in an airplane not excellently shielded, the ignition interference picked up by the antenna structure may be sufficient to block the detector, even though the incoming signal is small. By using plate detection a considerably greater interference would be necessary for blocking the detector. Since the present form of course indicator suggested for the beacon signals is practically immune to interference, it will operate satisfactorily so long as the tubes employed are not blocked. It should be noted, however, that plate detection precludes the use of a transformer in the detector output due to the large tube plate resistance with this form of detection.

8. Audio Amplifier. The low amplification factor of the 3-volt tubes used in the first two models (Figs. 1 and 2) necessitated the use of transformer coupling to provide a degree of audio amplification sufficient to supply the desired voltage to the power tube without overloading the detector. Due to the relatively high plate resistance of these tubes, however, the low-frequency pitches are not amplified in the same magnitude as those of higher frequencies. This is indicated in Curve *A* of Fig. 17

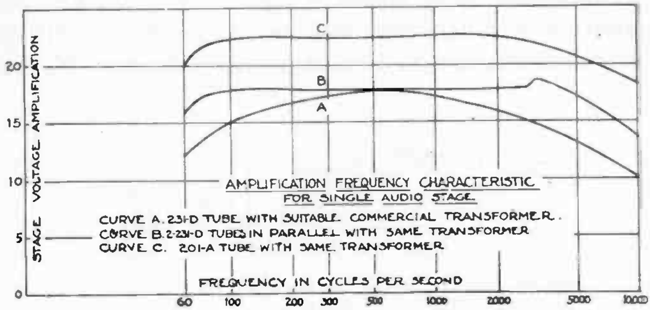

Fig. 17—Amplification-Frequency Characteristic for Single Audio Stage.

which represents the amplification-frequency characteristics for a 3-volt tube used together with a commercial transformer. A large number of commercial transformers were tested, the one finally adopted being chosen for its lightness as well as its electrical characteristics. A much more uniform gain over the desired frequency range is obtained by using two tubes in parallel in the first audio stage in order to reduce the net plate resistance. The amplification-frequency characteristic for this condition is given by curve *B* of Fig. 14. This arrangement was, therefore, used in both the 3-volt receiving sets. Curve *C* shows that when 5-volt tubes are used a satisfactory characteristic is obtained with one tube in the first audio stage.

Using 5-volt tubes a resistance-coupled audio amplifier immediately suggests itself. The third model (Fig. 3) employing 5-volt tubes uses transformer coupling, however, chiefly for the sake of flexibility. In this set it is feasible to use either 3 or 5 volt tubes throughout, requiring only a change in the size of the filament control rheostats. The operator can, therefore, decide as to whether the cost of tubes or weight of battery is the chief consideration.

Quite fortunately, the impedance of the visual course indicator is suitable for use with any of the power tubes available. The choice of the power tubes is therefore controlled by the filament voltage and plate battery consumption. A filter arrangement consisting of a choke and condenser is used in the output circuit for reasons previously mentioned. In the case of a resistance-coupled amplifier this filter circuit would also tend to prevent low-frequency oscillation.

9. Over-All Performance. The performance of a receiving set may be indicated by means of three graphs; the first showing the variation of sensitivity with carrier frequency; the second showing the degree of its selectivity at several carrier frequencies;

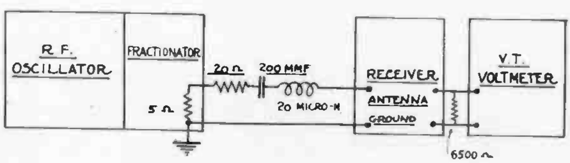

Fig. 18—Test Circuit Arrangement for Measuring Receiving Set Performance.

and the third giving a picture of its fidelity or response to audio frequencies which modulate the carrier to which it is tuned. The conditions under which these tests should be taken have been worked out by the Standardization Committee of the Institute of Radio Engineers for broadcast sets. Those conditions apply here also in large measure, but some departures were found desirable in order to approximate actual use more closely.

The sensitivity curve (Fig. 19) was obtained with the circuit arrangement of Fig. 18. For convenience, 100 per cent modulation at an audio-frequency of 60 cycles was employed. This more nearly approximates actual conditions as low modulating frequencies of the order of 60 cycles are used in the beacon, and the modulation is constant and as near 100 per cent as can be obtained. The curve shows the relationship between the radio field intensity in microvolts per meter necessary to obtain 6 volts across the 6,500-ohm load and the carrier frequency to which the receiving set is tuned.

The value of 6 volts was chosen, since this is the maximum voltage needed across the present type of course indicator.

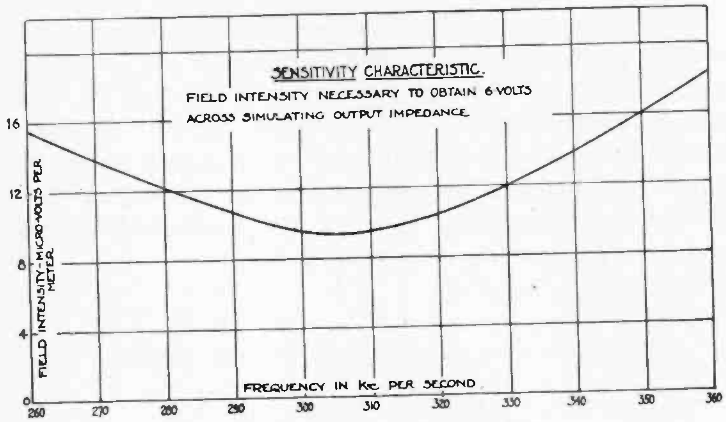

Fig. 19—Sensitivity Characteristic.

Fig. 20 shows the selectivity curves for the set when tuned to 290 and 330 kc. These are, respectively, the approximate beacon and phone frequencies. These curves represent the field intensity in microvolts per meter to give 6 volts across the simulating impedance referred to the carrier frequency.

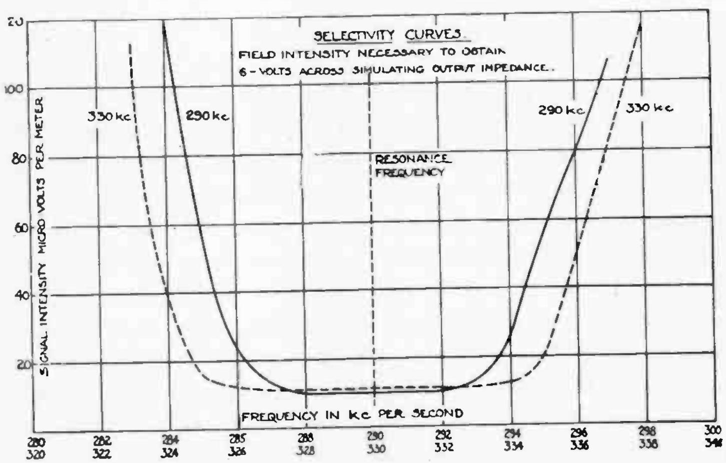

Fig. 20—Selectivity Curves.

Fig. 21 shows the fidelity curve. This was obtained using the audio portion of the receiving set only, and is plotted as the ratio of voltage across the simulating output impedance to that between the grid and filament of the detector tube against audio-frequency.

Twenty volts across the simulating output impedance was obtained before tube overloading occurred.

It should be noted that the curves of Figs. 19, 20, and 21 were obtained with one of the 3-volt sets. When 5-volt tubes are used, measurements indicate that the performance of the radio-frequency amplifier is but slightly altered. The voltage gain in the audio amplifier will, however, be considerably increased due to the greater tube amplification factors.

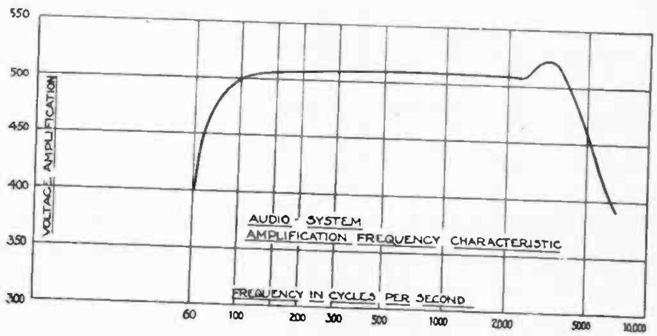

Fig. 21—Audio System Amplification-Frequency Characteristic.

Conclusion

These sets were developed concurrently with the double modulation type of directive radio beacon, the reed type of visual indicator for aircraft, and the airplane vertical pole antenna, all of which are parts of the bureau's research program on aids to air navigation. The development of these receiving sets was found essential because of the special receiving requirements on airplanes peculiar to these air navigation aids. Without receiving sets of high sensitivity, the elimination of the dangerous trailing wire antenna, and the reduction of night beacon course shift errors obtained with the short vertical antenna, would not be possible.

The second set described was extensively used in flight tests for a period of six months. Satisfactory daytime beacon signals were received on an airplane at a distance of 100 miles with an antenna 10 feet high, the engine ignition system being adequately shielded.

The writers wish to acknowledge the assistance of L. L. Hughes and D. O. Lybrand in constructing these sets, including several preliminary models.

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AN AIRCRAFT RADIO RECEIVER FOR USE WITH RIGID ANTENNA*

By

F. H. Drake

(Aircraft Division, Radio Frequency Laboratories, Inc., Boonton, N. J.)

Summary—*An outline is given of the physical and electrical requirements of an aircraft radio receiver suitable for the reception on a rigid antenna of radio beacons and weather service. The design of a special unicontrol receiver calculated to fulfill these requirements is described. Quantitative performance data are presented, with particular attention to the problem of detector overloading when operating a visual indicator from a beacon of the Bureau of Standards type. The corroboration of these data by practical tests is briefly discussed. The paper is concluded with quantitative discussion of the problem of ignition shielding on a particular type of airplane motor.*

THE rapid growth in air transportation of freight, mail, and passengers, together with the development of radio aids to air navigation in the forms of beacon and weather service, has created an urgent need for suitable aircraft radio receiving equipment to operate over a limited band of frequencies. In the 1927 International Convention the 285–315-kc band was allocated for radio beacon services, and the 315–350-kc band for aircraft communication. It is the purpose of this article to describe the physical and electrical characteristics of an aircraft radio receiver designed for the above frequency ranges. This receiver is of practical commercial design, and has successfully withstood the severe conditions of regular operation on mail planes.

The design of the receiver was undertaken over a year ago at the request of the Bureau of Standards. The primary requirement of such a receiver was that it should be suitable for the reception of both visual and aural beacons with a consistent range of 150 miles on signals from a 2-kw ground transmitter, using a rigid pole as antenna not more than two meters in height and the bonded metal members of the airplane structure as a counterpoise. The design of a receiver to meet this unusual sensitivity requirement was undertaken as a progressive step in that program of aircraft receiver development so ably initiated in the work described elsewhere by Messrs. Pratt and Diamond.[1]

* Dewey decimal classification: R521. Original manuscript received by the Institute, January 2, 1929.

[1] This issue of the Proceedings, page 283.

The severe conditions of vibration and shock, and the great diversity of climatic conditions under which an airplane receiver is used, present certain novel problems of design. Above all else, an airplane receiver must be rugged, light, and compact. The receiver as a whole must be adequately shock-proofed from its mounting to prevent destruction of tubes and wiring by vibration; supplemental shock-proofing of the vacuum tubes within the receiver is also advantageous since the tube elements have natural frequencies of mechanical vibration which may be excited at certain engine speeds. Lightness and compactness in the power supply are highly desirable. A further important requirement of present airplane receivers is that they be capable of installation wherever a vacant spot may be discovered in the plane. For simplicity of tuning and remote-controlling the receiver should have few controls and these should be of ample dimensions and shaped so that a pilot may easily operate them even though wearing heavy gloves. Further, it is highly important that all controls be equipped with some form of locking device to prevent creeping caused by vibration. The problem of rendering radio receiver performance sensibly independent of moisture is acute in designing a receiver for airplane use. A plane flying through fog encounters extremes of humidity seldom met indoors on the ground. The combination of receiver and associated antenna should have sufficient sensitivity to reach the average atmospheric noise level, assuming, of course, that ignition interference from the motor has been reduced to insignificance by shielding. A brief discussion of ignition shielding will be given later in this paper.

It was formerly considered necessary to employ a trailing wire as a receiving antenna. Mechanical objections to this arrangement are obvious, particularly in the case of mail or freight airplanes where the pilot must handle the antenna. Experience with reception of beacon signals has disclosed the further disadvantage of a trailing wire receiving antenna which results from the interception by the antenna of any horizontal components of electric force in the incoming wave. Two types of error are associated with these horizontal components, the "airplane effect" and the "night effect." "Airplane effect"[2] is produced by the inclination of

[2] By the Sommerfeld reciprocity theorem this situation is equivalent to a direction-finding loop on the ground and transmission from a partly horizontal antenna. The error so produced has been recognized for many years. See Ballantine: "Yearbook of Wireless Telegraphy," (London, 1921).

the antenna when the airplane flies at any angle greater than zero with respect to the beacon course. It can occur irrespective of the presence of a downcoming or reflected wave, and thus may be troublesome in the daytime. It would be entirely eliminated by a wholly vertical antenna system. "Night effect" is produced by the downward wave reflected from the Heaviside-Kennelly layer and may be modified by the ground wave. In general it is increased by increasing length of the transmission path. It manifests itself, in the case of directional transmission to a non-directional receiver, as an apparent wandering of the transmitted beam. When the directional transmitter is a loop system, "night effect" cannot be eliminated by the use of a vertical antenna, although it may be somewhat reduced thereby.

To summarize, the use of a rigid vertical antenna is justified by the elimination of physical hazards and burdens of maintenance on the pilot, as well as by a substantial reduction in the normal beacon errors. Accordingly, such an antenna was recommended by Radio Frequency Laboratories at the outset and the receiving equipment was designed with this in view. Subsequent experience has tended to justify this line of approach.

Owing to constantly changing conditions in the allocated frequency band it is difficult to specify just what minimum degree of selectivity the aircraft receiver should possess. At the present time, the problem of congestion in the 285–350-kc band is not acute, and for satisfactory discrimination the aircraft receiver requires somewhat less selectivity than is required in a modern broadcast receiver. Experience indicates that the order of selectivity corresponding roughly to three tuned circuits in cascade, none of which has a power factor greater than one per cent, is sufficient for present requirements.

A severe demand on fidelity of the aircraft receiver is imposed by the visual type of beacon. In the visual beacon the radio-frequency excitations of the two crossed coils are modulated respectively with two different audio frequencies instead of with dots and dashes as in the aural type beacon. These two audio frequencies differ by about 20 cycles and are usually both between 50 and 120 cycles per second. The aircraft receiver should therefore respond uniformly to modulation frequencies down to 50 cycles per second. For radio-telephonic reception of weather reports the aircraft receiver should maintain substantial fidelity up to frequencies of 3000 cycles per second and cut off rather sharply above this frequency in order to reduce noise.

Fig. 1 is from a photograph of a mail plane equipped with the RFL receiver and vertical pole antenna. In this installation the receiver was mounted in the front of the pilot's cockpit so that a remote tuning control was not required. The dash control panel (which will be described later) contains the volume control and is mounted on the instrument board. The front face of the receiver, dash control panel, and reed indicator are visible in Fig. 2, which is from a photograph taken looking forward into the pilot's cockpit. The antenna is a 7/8-in. diameter duralumin tube triangularly guyed to the upper wing and fuselage. In later installations the

Fig. 1—Installation on Pitcairn PA-5 Mail Plane Showing Short Rigid Vertical Antenna.

tube has been stream-lined to reduce wind resistance. The height of the tube above the fuselage is about two meters. With most planes, this will permit entrance to hangars having a clearance of 14 feet, so that the antenna need never be taken down.

Fig. 3 shows the receiver in its shock-proof mounting, the dash control panel, and a set of reeds for use on the visual indicating type of beacon. These reeds are held in a shock-proof mounting and are accurately tuned with temperature compensation to the two modulation frequencies of the beacon transmitter. They vibrate with equal amplitudes when the airplane is on its course. If the plane strays to the left of its course, the left reed develops a greater amplitude of vibration than the right, and vice versa. The dash control panel includes a volume control, which varies the filament current of the radio-frequency amplifier tubes, a double-range voltmeter for indicating the condition of plate and filament batteries, a switch for turning the receiver filaments "on" and "off," two cables, one making connection with the receiver and the other with the power supply, and a jack for

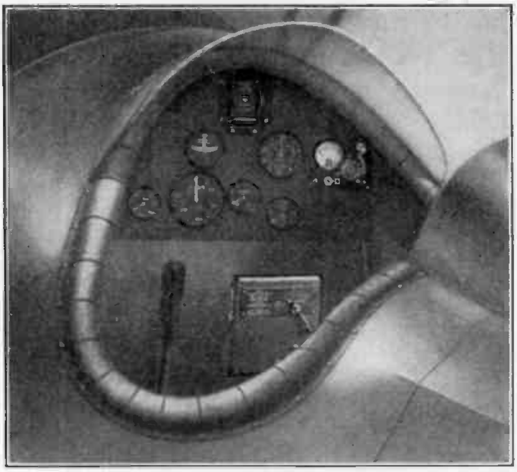

Fig. 2—View of Control Cockpit in Pitcairn Plane and Receiver Installation; Reed Indicators on Instrument Panel, upper center; Receiver Control Panel at right; Receiver Below.

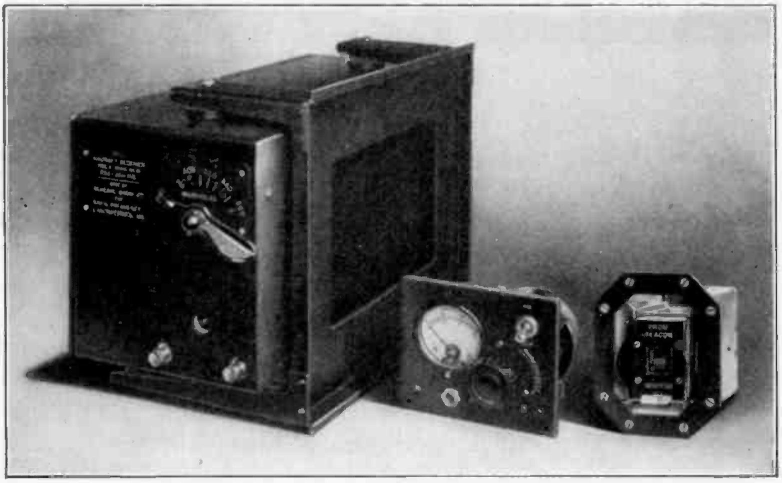

Fig. 3—Model B Unicontrol Aircraft Receiver, Dash Control Panel and Reed Course-Indicator.

the connection of telephones or reeds. The cable from the control panel to the receiver makes connection with the receiver through a multiple point plug. The mounting frame which supports the receiver is constructed to slide and lock in a track which is permanently fixed in the plane. These arrangements simplify the removal of the receiver from the plane. The front face of the receiver contains antenna and ground binding posts and the single tuning control, about which a direct calibration in kilocycles is engraved. The tuning control is the only control physically associated with the receiver itself, and is therefore the only control for which remote operation need ever be specially provided.

Fig. 4—Special Light-Weight Combined *B* and *C* Dry-Cell Battery for Use with Aircraft Receiver.

A special combined *B* and *C* battery has been developed for use with this receiver. One of these batteries is shown in Fig. 4. This battery is designed to give a life of 100 operating hours at a current drain of 10 milliamperes, which is the maximum drain of the receiver. Dimensions and weight of this battery, together with the dimensions and weights of the various units of the receiving equipment are as follows:

DIMENSIONS

Overall dimensions, receiver in mounting	15 in. × 9 in. × 6 in.
Dash control panel	$4\frac{1}{2}$ in. × 4 in. × $3\frac{1}{2}$ in.
Reed box in holder	$4\frac{1}{2}$ in. × 4 in. × 6 in.
Combination *B* and *C* battery	10 in. × $7\frac{3}{4}$ in. × 3 in.

WEIGHTS

Receiver with tubes and shock-proof frame	11 lbs.	0 oz.
Dash contol panel and cables	1 "	14 "
Reeds in shock-proof mounting	1 "	9 "
Combination *B* and *C* battery	10 "	6 "
Antenna pole and fittings	5 "	0 "
Total	29 lbs.	13 oz.

In almost all instances the airplane carries a storage battery for navigating or landing lights. This battery may be used for the filament supply of the radio receiver. Otherwise, a small 6-volt lead storage battery is provided. Such a battery, especially designed for airplane use, is now made by a number of manufacturers and weighs about ten pounds.

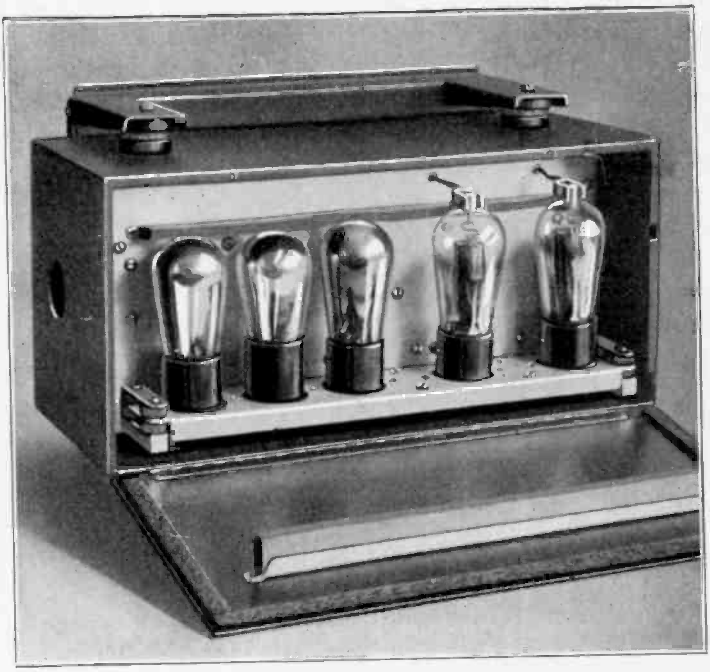

Fig. 5—Tube Compartment of Model B Receiver, Showing Shock Insulation of Tube Girder and Auxiliary Shock-Proof Support at Top of Tubes.

The left side of the receiver is hinged to permit access to the tube compartment. In Fig. 5 this side has been opened to show the arrangement and method of shock-proofing the tubes. The tubes are all mounted on a rigid aluminum channel, which is insulated from the receiver chassis by sponge rubber washers. Two horizontal sponge rubber strips, one fitted to the hinged side and the other to the receiver chassis, bear lightly against the tops of the tubes above the points of maximum diameter of the bulbs, and prevent the tubes from creeping out of their sockets or moving laterally in case appreciable vibration reaches them.

Felt strips along the edges of the hinged side serve to protect the chassis from dust and moisture.

Five tubes are employed as follows: two shielded tetrodes as radio-frequency amplifiers; two high-μ triodes as detector and first audio amplifier; and a power triode as the second audio amplifier.

There are three tuned circuits operated by a single control. In place of the conventional gang condenser a gang variometer is used. This gang variometer is a rigid unitary assemblage of three approximately cubical cells, each of which houses a variometer. The design of the radio stage is simple and merely involves determining for a chosen size of shield the variometer of required inductance range which has the lowest power factor. Two general types of construction have been successfully employed for the variometer gang: first, a welded construction using pure aluminum sheet; and second, a cast construction using an aluminum alloy coated with copper by the Schoop spray. Fixed condensers molded in bakelite are used in conjunction with the variometers in the tuned circuits.

Tuning by variable inductance is not desirable where a wide frequency range must be covered, but it possesses marked advantages when a frequency band defined by limits of two to one or less is required. It results in a very obvious saving of weight and space. Further, when used in conjuction with an antenna short in comparison with the wavelength, it permits exact single control with no trimming adjustments, when the antenna capacity is made equal to the fixed capacities employed in the other tuned circuits. Identical variometers may then be employed in the radio stages and antenna input circuit.

The use of tetrodes in the radio amplifier results in a radio gain of approximately 50 per stage instead of the gain of 10 or 15 per stage ordinarily obtainable with triodes. A high radio-frequency gain per stage is a more important consideration in an aircraft receiver than in ground sets where lightness and compactness are not vital. For the first audio amplifier, or at least for the detector, the tetrode is not so acceptable because of microphonic noises inherent in the structure of tubes now commercially available.

A detector capable of withstanding considerable overloading is provided in the receiver by the use of plate rectification with automatic grid bias.[3] This feature is particularly important when using the receiver in conjunction with the visual type of beacon indicator. When closely approaching the beacon a pilot may neglect to keep his reed amplitude down to the proper level by adjustment of the volume control; if this condition persists the detector may be so overloaded as to cause the reed amplitude to pass through a maximum and then fall to normal levels. Under

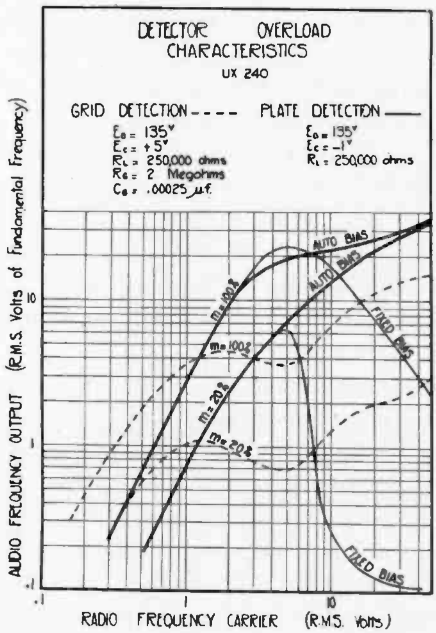

Fig. 6—Comparison of Ordinary Grid and Plate Rectifying Detectors and Special Detector with Automatic Grid Bias Used in Model B Receiver; Fundamental Component of Audio Output Voltage for Modulations of 20 and 100 per cent Against Carrier Voltage.

such conditions the indicated course may be reversed, i.e., the reed of lesser amplitude will indicate the side to which the airplane is off course.

Fig. 6 depicts the overload characteristic of three detector arrangements for two degrees of modulation (m). The curves show audio output of *fundamental* modulation frequency produced by a wide range of radio inputs. Examination of the curves

[3] This arrangement was devised by Stuart Ballantine.

shows that a somewhat higher output may be obtained with plate rectification than with grid rectification before overload. With either type, however, reversal or lack of course indications may result if the input voltage is increased through values corresponding to the maxima of the curves into the regions of negative slope. This matter is of considerable importance as in the practical operation of the receiver without automatic bias near a beacon, many instances of detector overloading have been observed. The method of preventing overloading employed in this receiver has been found quite satisfactory since it meets the essential requirement for correct indication of course, that the output curve shall always have a positive slope. While the audio amplifier may "overload" before the detector (plate rectification), the necessity for avoiding a negative slope in the detector response is in no way affected thereby, since the amplifier "overloading" at these voltages merely involves a falling-off in response which does not attain a negative slope.

Plate detection with automatic grid bias has other advantages. It is commonly supposed that grid rectification results in greater sensitivity than plate rectification. As a matter of experimental fact the microvolt sensitivity of this tetrode receiver is twice as great with plate detection as with grid detection, although the small-signal detection factor for the latter is about three times as great as for the former. Plate detection exceeds grid rectification in this case because it leaves unaffected the radio gain of the preceding stage, whereas due to electronic conductance grid detection reduces this gain by a factor greater than two to one. For the same reason the selectivity is considerably greater with plate rectification than with grid rectification. The higher output impedance resulting from plate detection does not impair the uniform transmission of low modulation frequencies essential for the visual beacon provided the coupling between detector and audio amplifier is properly designed.

A resistance coupled audio amplifier is employed in the receiver. In addition to mechanical advantages of compactness and light weight, such an amplifier has electrical characteristics desirable in the airplane receiver. Among these are low *B* battery drain and uniform transmission of the required modulation frequencies.

Fig. 7 gives the overall performance characteristics of the receiver. Sensitivity is expressed as the 400-cycle, 30 per cent modulated microvolt input required in the standard antenna to produce 15 rms volts audio-frequency output across a 4000-ohm resistive load. The standard antenna here considered has 100 $\mu\mu$f capacity and negligible resistance. These constants are characteristic of the rigid antenna for which the receiver was designed. If the effective height of the rod antenna is one meter the receiver will yield an output of 15 volts for field intensities ranging from 5 to 8 microvolts per meter.

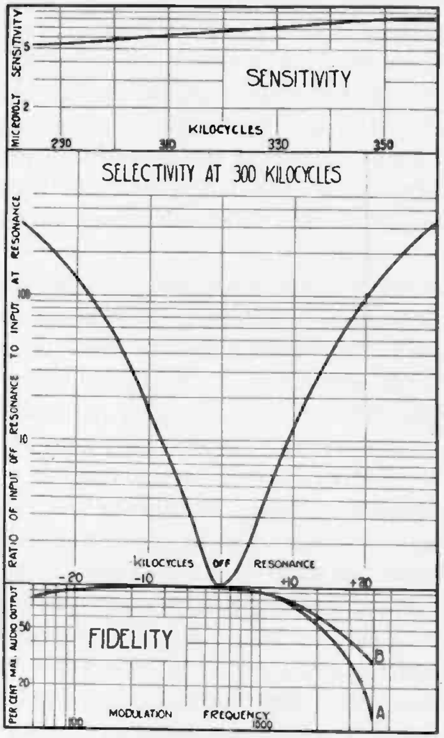

Fig. 7—Performance of Model B Receiver as Shown by Standard Measurements of Sensitivity, Selectivity at 300 kc, and Fidelity (A, of entire receiver; B, with radio-frequency amplifier out).

Selectivity is expressed as the ratio of the radio input voltage off resonance to the radio input voltage at resonance required for the normal audio output. It is important that this ratio be plotted over a range of several frequency channels on either side of resonance in order to present a sufficient basis for judging the discrimination of the receiver. The "skirts" of the selectivity curve are as important as the region immediately about resonance.

Fidelity is expressed as the percentage of the audio output at any modulation frequency to that at a reference frequency of 400 cycles per second, and is obtained with fixed radio input by measuring the audio output at various modulation frequencies. The overall fidelity thus measured includes all frequency distortions produced in normal operation of the receiver. Curve *B* represents fidelity in the absence of sideband attenuation in the radio-frequency amplifier. It is obvious from a comparison of this curve with the actual fidelity curve (curve *A*) that a major reduction in high audio-frequency response occurs in the radio amplifier. Fidelity and selectivity were measured at a carrier frequency of 300 kc per second; the results are not substantially different at other carrier frequencies in the range.

In addition to the standard laboratory measurements of performance represented by the sensitivity, selectivity, and fidelity curves, extensive tests of the receiver have been made under a wide variety of service conditions with the rigid antenna.

Tests were made early in 1928 with the receiver installed in an airplane belonging to the Bureau of Standards. With the receiver operating at less than full sensitivity, aural beacon signals from Bellefonte, Penna., were received at College Park, Md., a distance of about 140 miles. An installation on a Pitcairn Mailwing affords a satisfactory indication of the range on visual beacon service by producing at Hadley Field, N. J., normal amplitudes of the reed indicators from signals transmitted from College Park, Md., when the College Park transmitter is supplying a current of only 5 amperes to each loop. The two loops are triangular in shape, 70 feet high and 300 feet along the base. Repeated tests on an airplane belonging to the Radio Frequency Laboratories have demonstrated a reliable reception range of at least 150 miles for telephone weather broadcasts from Hadley Field, N. J. In the course of a number of flights, the Hadley beacon and weather services have been heard from Boonton, N. J., to Washington, D. C.

Similar ranges of reception of beacon and weather transmissions in different parts of the country have been observed in the course of a test flight from Boonton to Los Angeles and return.

Perhaps the most severe tests of the receiver by operators relatively unskilled in radio have been made on airplanes of the National Air Transport, Inc., in flights over the New York-Chicago air mail route. In the course of tests under all sorts of night and day flying conditions a reliable range of 150 miles on weather and beacon service has been reported. In a number of cases satisfactory beacon service has been obtained at a distance of 250 miles over mountainous territory. The airplanes used in this service are powered by Liberty motors with battery ignition. In such planes the level of ignition noise cannot readily be reduced to the negligible amount attainable in magneto ignited air-cooled motors.

Obviously any highly sensitive radio receiver imposes severe demands on the degree of ignition shielding. On planes equipped with battery ignition systems it is possible by careful shielding to reduce the disturbance for the receiver sensitivity herein described to a satisfactory low level. On planes using air-cooled motors with magneto ignition we have found that ignition disturbances may be completely eliminated by suitable shielding. Such shielding includes complete covering of magneto terminal blocks and all low-tension and high-tension wiring. Shielded spark plugs or their equivalent must be used.

In order to determine the relative contributions of the various parts of the ignition system to the general ignition disturbance a simple test has been devised. A receiver of the type described above is calibrated for sensitivity in microvolts against the position of the volume control rheostat. This receiver is then installed in a plane with a completely shielded ignition system. The test consists in removing various elements of the shielding and observing the points to which the receiver sensitivity must be reduced in order to make the interference just audible. Results of this test for a particular installation in an airplane equipped with a single Wright Whirlwind motor (Type J-5) are exhibited in Fig. 8, which shows for various amounts of shielding the corresponding maximum receiver sensitivities permissible with negligible ignition interference. These data may be translated directly into attainable reception range if the law of attenuation is known for wave propagation between ground and plane. Of course, information of the sort given in Fig. 8 must be applied with caution to other installations, where the various elements of the ignition system may be differently disposed with respect to the antenna; the only points of general interest are the relative magnitudes of the various factors contributing to the total disturbance. Of particular interest in Fig. 8 is the effect of shielded spark plugs in reducing ignition disturbance; they permit the use of a receiver forty times as sensitive as may be used with unshielded plugs.

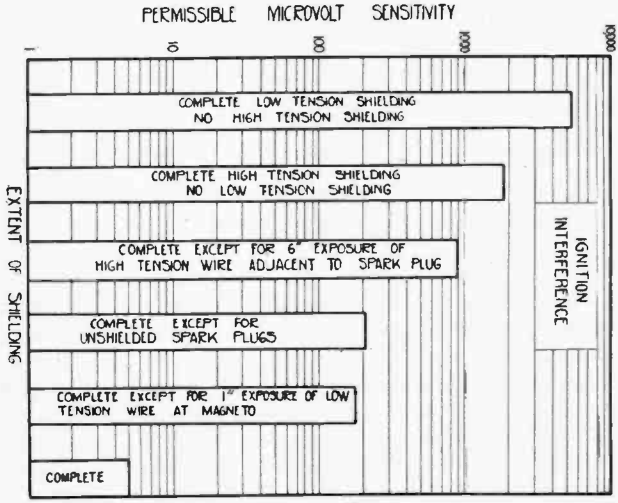

Fig. 8—Analysis of Ignition Disturbance from Wright Whirlwind Motor; Relation between Permissible Receiver Sensitivity and Degree of Shielding.

In conclusion the author wishes to acknowledge his indebtedness to Mr. W. D. Loughlin of the Engineering Division of Radio Frequency Laboratories for the structural design of the receiver.

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CIRCUIT ANALYSIS APPLIED TO THE SCREEN-GRID TUBE*

By

J. R. Nelson

(Engineering Dept., E. T. Cunningham, Inc., New York City.)

Summary—*General radio-frequency circuit theory is discussed in this paper. The theoretical work and discussion is divided into two parts, amplification and stability.*

General amplification equations for impedance and transformer coupling, using an untuned primary whose period is above the highest frequency considered, are derived and discussed for the case of a screen-grid tube such as Cunningham type CX-322.

Feedback through the mutual capacity plate to grid capacity is also considered. A general expression for the limit of stable amplification per stage is inferred for n stages from the expressions found for one and two stages. This general expression $Av. < \sqrt{2gm/nwc_0}$ *is in terms of the mutual conductance, total grid to plate capacity, frequency and number of stages.*

CIRCUIT analysis is very simple with the perfect screen-grid tube. The mutual capacity between the input and output circuits is eliminated by the screening action of the fourth element, the screen grid. All other tube capacities are merely circuit constants. The plate impedance is infinite but the mutual conductance is finite. Under operating conditions the plate current is independent of plate voltage. The alternating component of the plate current is a function of only one tube constant, the mutual conductance. As the tube impedance is infinite the amplification is simply the product of the mutual conductance and external impedance.[1]

The mutual capacity, the control grid to plate capacity, of commerical screen-grid tubes such as Cunningham type CX-322 has been reduced to as small a value as is practical, less than 0.025 $\mu\mu$f. Although this capacity is very small it is large enough to affect the stability of an efficient radio-frequency amplifier at the higher broadcast frequencies particularly if care is not taken in the shielding. Neutralization of the mutual capacity is not necessary as stable amplification, four or five times that given

* Dewey decimal classification: R1320. Original manuscript received by the Institute, October 22, 1928.

[1] A. W. Hull and N. H. Williams, Characteristics of Shielded-Grid Pliotrons, *Phys. Rev.*, 27, 432–438; April, 1926.

by circuits using three-element tubes, is obtainable. The plate impedance is not infinite but has a value of about 800,000 ohms under operating conditions. The amplification, because of this finite plate impedance, is only approximately equal to the mutual conductance times the external impedance.

Hull[2] has discussed the circuit theory of the perfect screen-grid tube using impedance coupling. He made the assumption that the amplification is equal to the product of the mutual conductance and external impedance. An external plate impedance of around 150,000 ohms may be obtained in the broadcast band. This is appreciable compared with about 800,000 ohms plate resistance and its effect is to reduce the mutual conductance of the tube and load to some value less than the mutual conductance of the tube alone. The grid resistor across the tuned impedance also reduces its value. This effect is usually small unless the grid resistor has some value less than one megohm. Hull[2] found no indication of regeneration in making his overall amplification measurements as his calculated and measured values of amplification agreed. His measured amplification should have been about fifteen per cent lower than his calculated value as he assumed the amplification was equal to the product of the mutual conductance and external impedance. The regeneration factor for his circuit was also about fifteen per cent so that this would just about make up for the reduction of mutual conductance.

The purpose of this article is to consider the general theory of radio-frequency amplifiers and to apply this theory to the commercial screen-grid tubes. Feedback due to the mutual capacity, control grid to plate capacity, is considered in the general theory for n stages, and an expression for the limit of stable amplification is found. This paper will only cover, for steady conditions the circuits using impedance coupling, or transformer coupling with an untuned primary whose natural period is above the highest frequency considered.

Transformer Coupling

Where

μ is the amplification factor of the tube

r_p is the internal resistance of the tube

[2] A. W. Hull, Measurements of High Frequency Amplification with the Shielded-Grid Pliotron, *Phys. Rev.*, 27, 439–454; April, 1926.

L_1 is the primary inductance
L_2 is the secondary inductance
M is the mutual inductance
C_2 is the secondary capacity
R_1 is the primary resistance
R_2 is the secondary resistance

Fig. 1 shows the equivalent circuit of a three-element tube and transformer. The screen-grid tube with a radio-frequency transformer may be represented by the same equivalent circuit

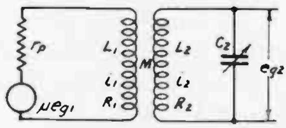

Fig. 1

if the d.c. bias of the screen grid is constant and if there is no external radio-frequency impedance in the screen-grid circuit. The equivalent values of μ and r_p vary with the d.c. voltages, but have definite values as soon as all the d.c. conditions are specified.

The voltage amplification Av. defined as the ratio of eg_2 to eg_1, is

$$\text{Av.} = \frac{\mu \quad \underline{|\delta}}{\omega C_2\left[\dfrac{Z_1{}^2 Z_2{}^2}{x^2{}_m} + x_m{}^2 - 2(x_2 x_1 - r_p R_2)\right]^{1/2}} \tag{1}$$

where

$$z_1 = \sqrt{r_p{}^2 + x_1{}^2}$$

$$z_2 = \sqrt{R_2{}^2 + x_2{}^2}$$

$$x_m = \omega m$$

$$\delta = \tan^{-1}\frac{r_p x_2 + R_2 x_1}{x_2 x_1 - x_m{}^2 - R_2 r_p}.$$

(For the derivation of the equations in this section see Appendix A.).

Equation (1) may be used to calculate the resonance curve. If we are interested only in the amplification at resonance simpler formulas may be derived. When any two of the three reactances X_1, X_2, or WM are varied so as to make I_2 a maximum, the maximum possible amplification for a given coil and tube is obtained.

The value of this amplification is

$$\text{Av.} = \frac{1}{2}\frac{\omega L_2}{\sqrt{R_2}}\frac{\mu}{\sqrt{r_p}} \tag{2}$$

When the secondary reactance X_2 alone is varied to make I_2 a maximum the amplification is

$$\text{Av.} = \frac{\mu}{r_p}\frac{\omega L_2}{R_2\left[1+\dfrac{x_m^2}{R_2 r_p}\right]}x_m. \tag{3}$$

When X_2 is varied the frequency at which the circuit will resonate may be found from (7), Appendix A.

$$[r_p^2+\omega^2 L_1^2]\left[\omega L_2 - \frac{1}{\omega C_2}\right] = \omega L_1 \omega^2 M^2. \tag{4}$$

If the primary is open rp is almost infinite and as $\omega L_1\omega^2 M^2$ is finite, the value of $(\omega L_2 - 1/\omega C_2)$ must approach zero so the resonant frequency is determined by the values of the secondary capacity and inductance.

$$\omega_r = \frac{1}{\sqrt{L_2 C_2}} \tag{5}$$

where ω_r is 2π times the resonant frequency.

When r_p is not infinite ω_r becomes

$$\omega_r^2 = -\omega_0^2\left[\frac{r_p^2 L_2 C_2}{2L_1^2} - \frac{1}{2}\right] + \sqrt{\left(\frac{r_p^2 L_2 C_2}{2L_1^2} - \frac{1}{2}\right)^2 \omega_0^4 + \frac{\omega_0^2 r_p^2}{L_1^2}} \tag{6}$$

where

$$\omega_0^2 = \frac{1}{C_2 L_2 (1-K^2)}$$

K = coefficient of magnetic coupling.

If r_p in (6) is made zero by shorting the primary on itself ω_r becomes

$$\omega^2_r = \tfrac{1}{2}\omega_0^2 + \tfrac{1}{2}\omega_0^2 = \omega_0^2 \tag{7}$$

The values of ω_r given by (5), (6), and (7) are independent of the secondary resistance. Under the usual circuit conditions if the value of r_p is small, the coupling is weak so that (5) may be used without introducing much error.

Impedance Coupling

Impedance coupling will be considered next. This should be the equivalent of transformer coupling when both inductances are equal and the coefficient of magnetic coupling is unity.

Fig. 2-A shows the circuit using impedance coupling. When

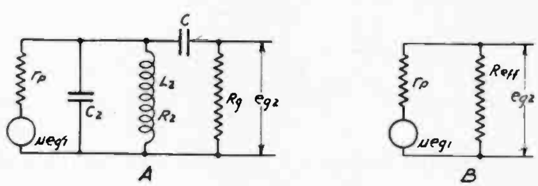

Fig. 2

resonance is defined as the frequency at which the circuit has zero reactance the combination L_2, R_2, C_2 acts a resistance whose value is

$$R \text{ equivalent} = \frac{L_2}{R_2 C_2} = \frac{\omega^2 L_2^2}{R_2} \tag{8}$$

If the coupling condenser C is large enough to have a small reactance at the frequency considered, Fig. 2-A may be replaced by Fig. 2-B. The effect of Rg is to add a resistance to the secondary resistance R_2 and the value of this added resistance is given by

$$R = \frac{\omega^2 L_2^2}{R_g}. \tag{9}$$

Calling the total effective resistance of the tuned circuit $R^1{}_2$ the value of R eff. is now

$$R_{eff} = \frac{\omega^2 L_2^2}{R_2 + \dfrac{\omega^2 L_2^2}{R_g}} = \frac{\omega^2 L_2^2}{R_2'} \tag{10}$$

$$i_p = \frac{\mu e_{g1}}{r_p + R_{eff}} \tag{11}$$

$$e_{g2} = \frac{\mu e_{g1} R_{eff}}{r_p + R_{eff}} \tag{12}$$

$$\text{Av.}=\frac{\mu}{p}\,\omega L_2\frac{\omega L_2}{R_2'\left(1+\dfrac{\omega^2L_2^2}{R_2'r_p}\right)} \tag{13}$$

Equation (13) is of the same form as (3) with ωL_2 in (13) replacing ωM of (3). $R^1{}_2$ of (13) includes the effect of the grid resistor R_g being shunted across the tuned circuit. The above expression for Av. will only give the amplification at resonance. The response at any other frequency may be found from (14) derived in Appendix B.

$$\text{Av.}=\frac{\mu}{r_p}\,\frac{\omega^2L_2^2+R_2'^2}{\sqrt{[C_2\omega(\omega^2L_2^2+R_2'^2)-\omega L_2]^2+\left(\dfrac{\omega^2L_2^2}{r_p}+R_2'\right)^2}} \tag{14}$$

Hull gives the same value of impedance with the exception that his term does not contain $\omega^2L_2^2/r_p$. With a perfect screen-grid tube this term would be zero because of the high plate impedance, but with the commercial screen-grid tubes it should be considered as its value is appreciable.

The amplification neglecting R_2^{12} where it is added to $\omega^2L_2^2$ is

$$\text{Av.}=\frac{\mu}{r_p}\,\frac{\omega^2L_2^2}{\sqrt{[\omega L_2(\omega^2L_2C_2-1)]^2+\left[\dfrac{\omega^2L_2^2}{r_p}+R_2'\right]^2}} \tag{15}$$

at resonance $\omega^2L_2C_2=1$ and (15) becomes

$$\text{Av.}=\frac{\mu}{r_p}\,\frac{\omega^2L_2^2}{R_2'+\dfrac{\omega^2L_2^2}{r_p}} \tag{16}$$

which is equivalent to (13).

Discussion of the Amplification Equations

In the usual multistage set it is not convenient to vary L_1. A maximum value of secondary current may be obtained by varying any two of the three reactances x_1, x_2, and x_m. Varying both x_m and x_2, the amplification obtained is given by Eq. (2). This will give the maximum amplification obtainable. The selectivity given by

$$S=\frac{\omega^2L_2^2}{R_2'} \tag{17}$$

(where R_2' includes the secondary resistance R_2 and the resistance transferred from the primary ω^2M^2/r_p) is only one-half that of the coil alone.

The condition for optimum coupling is given by (9), Appendix A, as

$$x_m{}^2 = R_2 r_p + x_1 x_2$$

As x_2 is small this is approximately

$$x_m{}^2 = R_2 r_p \tag{18}$$

With the screen-grid tube the plate resistance r_p is around 800,000 ohms under operating conditions. R_2 will be at least 5 ohms so (18) cannot be satisfied. Eq. (2) will not apply to this tube. To obtain the maximum amplification impedance coupling should be used.

As the condition for optimum coupling cannot be realized the selectivity of the tuned circuit will always be greater than one-half that of the coil alone.

The value of x_2 is varied so (3) will apply to this tube using transformer coupling. Almost as much amplification may be obtained with a properly designed transformer coupled circuit as with an impedance coupled circuit. The primary inductance may be slightly larger than the secondary inductance. If the primary is made small physically by winding it with small wire it may be coupled rather closely magnetically to the secondary without adding excessive dielectric losses to the tuned circuit.

There is some gain in selectivity by using a carefully designed transformer over that obtained by using impedance coupling. The effective resistance added to the secondary when a transformer is used is ω^2M^2/r_p. The resistance added by impedance coupling is $\omega^2L_2{}^2/r_p$ plus a term $\omega^2L_2{}^2/R_g$. The resistance added to the secondary by transformer coupling is less than that added by impedance coupling, and if the transformer is well designed the transformer coupled circuit will have better selectivity than the impedance coupled circuit.

All the amplification equations show that the amplification will decrease with a decrease of mutual conductance. Fig. 3 shows μ, r_p and gm plotted against the screen-grid voltage, all other voltages being left constant. The screen-grid voltage may be used as a volume control. The value of r_p will increase as gm is decreased so the selectivity will increase as the volume is decreased. Tuning of the circuits will not be affected by the volume control.

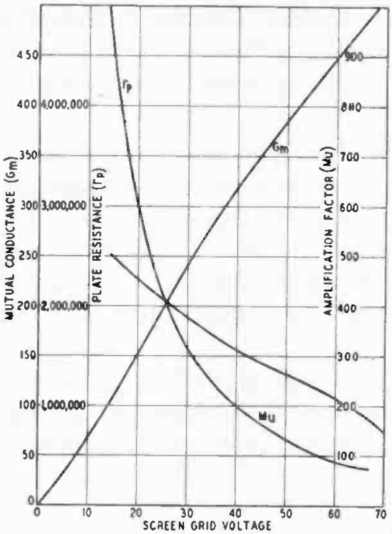

Fig. 3—CX-322 Screen-Grid Connection. Plate resistance, mutual conductance, and mu vs. screen-grid voltage. $Ec_1 = 1.5$ volts, $Eb = 135$ volts.

STABILITY

The control grid to plate capacity couples the input and output circuits in an amplifier. The output circuit contains a source of energy and will react on the input circuit because of the mutual reactance. The voltage fed back from the output to the input may be either in phase or out of phase with the input voltage. If the amount of energy fed back is enough to supply the input losses the circuit will oscillate. The circuit conditions necessary to cause oscillation through the mutual tube capacity depend upon the number of stages. The amplification increases geometrically and the capacity coupling the input and output decreases arithmetically so that a stage is reached where the output will supply enough energy to supply the input losses causing the amplifier to oscillate.

Hull[3] has discussed stability by assuming that maximum regeneration will be obtained when all the circuits are in exact

[3] loc. cit.

resonance. He does not consider n stages, but derives results for only two stages.

Any discussion of stability should make no assumptions as to values of the phase angles. The theory should also show what factor multiplies the amplification value if oscillations do not occur. The final results in order to be general should be for n stages.

Beatty[4] has found the circuit conditions that will cause one stage to oscillate. His limit as to the value of certain circuit constants that will cause oscillations to occur is found graphically without any mathematical proof. This article will show that his limit is correct, and will in addition carry out the work to two stages. A general law for n stages may be inferred from these results.

Fig. 4 shows the circuit that will be considered. Any transformer coupled circuit may be reduced to this circuit so the results will be general. The coils are assumed to be similar. The

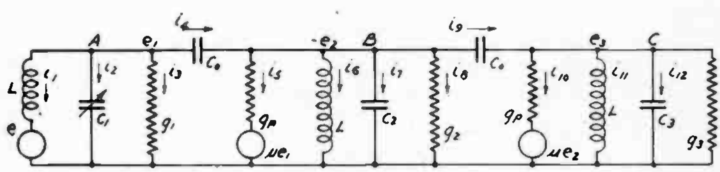

Fig. 4

resistances in the coils have been replaced with conductances g_1, g_2, g_3 in parallel with the tuned circuits. The value of this conductance is

$$gc = \frac{R_2}{\omega^2 L^2}. \tag{19}$$

The plate to filament conductance is g_p. A voltage e is induced in the circuit. The ratio of the voltage e_1 and e is

$$\frac{e_1}{e} = \frac{\omega L}{R} \quad \text{or} \quad \frac{1}{gc\omega L}. \tag{20}$$

For the derivation of equations in this section see Appendix C.

[4] R. T. Beatty, The Stability of the Tuned-Grid Tuned-Plate, H F Amplifier, *Exper. Wireless and Wireless Engineer*, V, 3; January, 1928.

Beatty[5] by means of (1C) and (2C), Appendix C, without the last term $(-e_3 j\omega C_0)$ in Eq. (2C), found the following relation between e_2 and e.

$$\frac{e_2}{e} = jAF \tag{21}$$

where

$$\frac{1}{F} = (1+j\tan\theta_1)(1+j\tan\theta_2)+jH \tag{22}$$

$$H = \frac{gmC_0\omega}{gc(gp+gc)} \tag{23}$$

$$A = \frac{gm}{\omega L_1 gc(gp+gc)} \cdot \tag{24}$$

The factor F will be infinity when

$$-jH = (1+j\tan\theta_1)(1+j\tan\theta_2) \tag{25}$$

H is a circuit constant depending on ω, gm, C_0 and the coils.

Beatty[6] shows graphically that $1/F$ is the distance between H and a parabola. When H is 2 this vector will be zero and the circuit will oscillate.

The smallest value of H that will cause the circuit to oscillate can be shown mathematically to be 2 as follows:

$$-jH = (1+jx)(1+jy) \tag{26}$$

or

$$-jH = 1+jy+jx-xy. \tag{27}$$

Equating reals and imaginaries,

$$xy = 1$$

$$-H = y+x. \tag{29}$$

From (28) x and y must have the same sign.

From (29) x and y must be negative.

This means that the phase angles are negative or both circuits will have inductive reactance.

From (28) $x = 1/y$.

Substituting this in (29)

[5] loc. cit.
[6] loc. cit.

$$y+\frac{1}{y}+H=0 \quad \text{or} \quad y^2+1+Hy=0 \tag{30}$$

$$y=\frac{-H\pm\sqrt{H^2-4}}{2} \tag{31}$$

y must be real so that H^2 must be equal to or greater than 4, hence the smallest value that H may have that will cause the circuit to oscillate is 2. When H is 2 each circuit must have its effective resistance equal to the effective inductance. If H is larger than 2 this restriction no longer holds. As we are only interested in the smallest value of H that will cause oscillations to occur the values of the phase angles when H is greater than 2 are of no interest in this discussion.

If H is smaller than 2 the regeneration factor for one stage may be found as explained in Beatty's[7] article. This phase of the question will not be further discussed here.

The relation between e_3 and e_1 is from (6) Appendix C.

$$\frac{e_3}{e}=\frac{gm^2}{j\omega L_1 g_1 g_4 g_5}\,\frac{1}{A+B} \tag{32}$$

where

$$A=(1+j\tan\theta_1)(1+j\tan\theta_2)(1+j\tan\theta_3) \tag{33}$$

$$B=j\omega C_0 gm\left(\frac{1}{g_4\,g_5}+\frac{1}{g_4\,g_1}\right) -\omega C_0 gm\left(\frac{\tan\theta_1}{g_4\,g_5}+\frac{\tan\theta_3}{g_4\,g_1}\right). \tag{34}$$

Let $g=g_1=g_4=g_5$ as would be the case in a multistage set. Eq. (34) becomes

$$B=j\frac{2\omega C_0 gm}{g^2}-\frac{\omega C_0 gm}{g^2}(\tan\theta_1+\tan\theta_3) \tag{35}$$

The quantity $\omega C_0 gm^2/g^2$ is a circuit constant. Denote its value by H'. Writing x for $\tan\theta_1$, y for $\tan\theta_2$, z for $\tan\theta_3$ we have the following expression:

$$A+B=(1+jx)(1+jy)(1+jz)+2jH'-H'(x+y).$$

The condition for the circuit to oscillate is that $A+B=0$.

[7] loc. cit.

In the single stage the conditions to cause oscillations were that H should be 2 and the phase angles negative and 45 deg. Assuming that the phase angles will be equal and the tangents equal to -1, we find that H' for two stages is equal to 1.

TABLE I

X	Y	Z	H'
	0	(Impossible as H' is imaginary)	
−4	−0.25	−4	2.125
−2	−0.50	−2	1.25
−1	−0.75	−3	1.25
−1.33	−0.75	−1.33	1.04
−1	−0.78	−2	1.11
−1	−0.84	−1.5	1.04
−1	−1	−1	1.00
−0.5	−1.44	−1	1.11
−0.667	−1.5	−0.667	1.08
−0.5	−2.0	0.5	1.25
−0.5	−2.6	0.3	1.50
−0.333	−3.0	0.333	1.67
−0.0	−3.0	−1	2.0
−0.1	−3.0	0	2.0
−0.25	−4.0	−0.25	2.125
	+1	(Impossible as H' must be -1.)	

Table I shows how the value of H' necessary to cause the circuit to oscillate varies with the phase angles. From this table it is seen that any value of H equal to or greater than 1 will cause the circuit to oscillate. Oscillations cannot occur if H is less than 1. From the table y must always be some negative value. When either x or z is made zero, H is 2, the same value as found for one stage. The phase angles are different, which is to be expected. Assume one stage was oscillating and another stage was added. If the phase angle was zero this would merely act as an amplifier, and to cause the whole circuit to oscillate it would be necessary to change the tuning on the first two condensers.

The results to be general should be expressed in terms of Av. Divide both the numerator and denominator of (13) by ω^2L^2

$$\text{Av.} = \frac{\mu}{r_p} \frac{1}{\dfrac{R_2}{\omega^2 L^2} + \dfrac{1}{r_p}} = \frac{gm}{gc+gp} = \frac{gm}{g}. \tag{36}$$

Refer to (32). As was previously mentioned, $1/j\omega Lg_1$ is the ratio between e_1 and e. The remaining part of (32) $gm^2/g_4 \cdot g_5(A+B)$ is the ratio of the voltage e_3 to e_1. If regeneration is not taken into account, $gm^2/g_4 \cdot g_5$ is the amplification of two stages as the amplification of one stage is gm/g as was shown in (66). When the value of the vector $1/(A+B)$ is known the total amplification considering reaction is the amplification per stage squared times the regeneration factor $1/(A+B)$.

For one stage we found

$$H=2=\frac{\omega C_0 gm}{g^2} \tag{37}$$

for two stages

$$H'=1=\frac{\omega C_0 gm}{g^2} \tag{38}$$

multiply (38) by 2

$$2=2\omega C_0 gm/g^2 \tag{39}$$

or

$$2=\frac{n\omega C_0 gm}{g^2}=\frac{n\omega C_0 \mathrm{Av.}^2}{gm} \tag{40}$$

where n is the number of stages or the number of tuned circuits minus 1.

$$\mathrm{Av.}^2=\frac{2gm}{n\omega C_0}\cdot \tag{41}$$

For the circuit to be stable,

$$\mathrm{Av.}<\sqrt{\frac{2gm}{n\omega C_0}}\cdot \tag{42}$$

Hull[2] ound an expression for two stages,

$$\mathrm{Av.}<\sqrt{\frac{gm}{\omega C_0}+1}\quad +1.$$

When n is 2, Eq. (42) becomes

$$\mathrm{Av.}<\sqrt{\frac{gm}{\omega C_0}}\cdot \tag{43}$$

Hull's expression agrees with this expression as to the circuit values and differs in adding 1 under the square root sign and 1 to the whole expression. As he assumed exact resonance there is no way of telling from his equations whether there would be any feedback if the circuit did not oscillate. In other words according to his results there would be no regeneration until the point of oscillation was reached, when the circuit would suddenly start to oscillate.

Eq. (42) shows the importance of shielding the input from the output in using a tube with a low control grid to plate capacity if stability is to be obtained without neutralization or adding losses.

The tube should be operated so that the value of gm is as high as is consistent with good tube life as the stable value of Av. in (42) will be raised with an increase of gm.

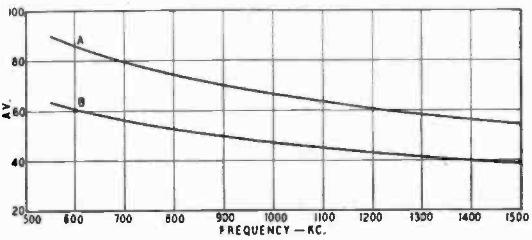

Fig. 5—Limit of Stable Amplification per Stage for Type CX-322 Tubes. $gm = 350$ micro-mhos, $Co = 0.25$ $\mu\mu$f, A—one-stage amplifier, B—two-stage amplifier.

Fig. 5 shows the limit of stable amplification plotted for a CX-322 tube. The value of C_0 is taken as 0.025 $\mu\mu$f. This value is the extreme limit and most of the tubes have less than 0.02 $\mu\mu$f control grid to plate capacity. The value of gm is taken as 350×10^{-6} mhos, which is an average value.

Eq. (42) is general and will apply to any tube where the only feedback is through the mutual capacity. Most three-element tube circuits are stabilized in some manner. Limits of stable amplification using three-element tubes are of more theoretical than of practical interest and will not be given here.

Note: Since this article was written Rockwood and Thompson[8] published a formula for the maximum load impedance which may be used with a tube without oscillations occurring.

$$R_p = \frac{1.41 r_p}{r_p(2\pi f C_{p-g} gm)^{1/2} - 1.41}.$$

This expression is the same as (42) when n is 1, as can be seen from the following algebraic transformation.

Clearing of fractions,

$$Rprp\sqrt{\omega C_0 gm} - \sqrt{2}Rp = \sqrt{2}r_p$$

$$\sqrt{2}(Rp + rp) = Rprp\sqrt{\omega C_0 gm}$$

[8] Rockwood and Thompson, Discussion on Application of the Four Electrode Receiving Tube (UX-222), *Proc.* of the Radio Club of America, 5, 69; Jun e, 1928.

$$\sqrt{29M}(Rp+rp)=Rprpgm\sqrt{\omega C_0}$$

$$\frac{\mu}{Rp+rp}R_p=\sqrt{\frac{2gm}{\omega C_0}}$$

$$\text{Av.}=\frac{\mu}{Rp+rp}R_p=\sqrt{\frac{2gm}{\omega C_0}}$$

which is the same as Eq. (42) when n is 1.

Appendix A

Derivations of Equations for Fig. 1

$$I_2=\frac{j\omega M\mu eg_1}{r_pR_2+\omega^2M^2-x_1x_2+j(r_px_2+\omega L_1R_2)} \tag{1}$$

R_1 has been neglected because it is small compared to r_p where

$$x_2=\omega L_2-\frac{1}{\omega C_2}$$

$$x_1=\omega L_1$$

$$eg_2=\frac{I_2}{j\omega C_2} \tag{2}$$

$$\text{Av.}=\frac{eg_2}{eg_1}=\frac{\mu\omega M}{\omega C_2\{r_pR_2+\omega^2M^2-x_1x_2)+j(r_px_2+x_1R_2)} \tag{3}$$

The absolute value of Av. is

$$\text{Av.}=\frac{\mu\ \underline{\lfloor\delta}}{\omega C_2\left[\frac{Z_1^2Z_2^2}{\omega^2M^2}+\omega^2M^2-2(x_1x_2-r_pR_2)\right]^{1/2}} \tag{4}$$

$$Z_{12}^2=\frac{eg_1^2}{I_2^2}=\frac{Z_1^2Z_2^2+\omega^4m^4-2\omega^2M^2(x_1x_2-r_pR_2)}{\omega^2M^2} \tag{5}$$

The value of I_2 will be a maximum when Z_{12} is a minimum. To find when I_2 will be a maximum, as any of the reactances are varied, it is only necessary to differentiate Z_{12}^2 with respect to that reactance and to set the derivate thus found equal to zero and to solve for the desired reactance in terms of the other impedances.

Doing this we find that I_2 will be maximum when

$$\omega L_1 = \frac{\omega^2 M^2 x_2}{R_2^2 + x_2^2} \tag{6}$$

$$x_2 = \frac{\omega^2 M^2 x_1}{r_p^2 + x_1^2} \tag{7}$$

$$\omega^2 M^2 = (r_p^2 + x_1^2)[R_2^2 + x_2^2). \tag{8}$$

Relations to satisfy (8), (9), and (10) may be found by equating values of $\omega^2 M^2$ from these equations. Doing this we find

$$\frac{x_2}{R_2} = \frac{x_1}{r_p}. \tag{9}$$

When (9) is satisfied $\omega^2 M^2$ from (8) becomes

$$\omega^2 M^2 = R_2 r_p + x_1 x_2 = \frac{R_2}{r_p}(r_p^2 + x_1^2) \tag{10}$$

If we satisfy any two of (6), (7), and (8), the third will also be satisfied. It is not usually convenient to vary x_1. Varying both x_2 and ωM, Z_{12} becomes

$$Z_{12} = 2\sqrt{R_2 r_p}\frac{\sqrt{r_p^2 + \omega^2 L_1^2}}{\omega L_1 + j r_p} \tag{11}$$

I_2 from Eq. (1) becomes

$$I_2 = \frac{1}{2}\frac{\mu}{\sqrt{R_2 r_p}}\frac{x_1 + j r_p}{\sqrt{r_p^2 + x_1^2}}. \tag{12}$$

If r_p is large compared to ωL_1,

$$\text{Av.} = \frac{1}{2}\frac{\omega L_2}{\sqrt{R_2}}\frac{\mu}{\sqrt{r_p}}. \tag{13}$$

If only x_2 is varied Z_{12} is

$$Z_{12} = \frac{R_2\sqrt{\omega^2 L_1^2 + r_p^2} + \dfrac{\omega^2 M^2 r_p}{\omega L_1 + j r_p}}{\omega^2 M^2}. \tag{14}$$

If r_p is large compared to (ωL_1) Av. is

$$\text{Av.} = \frac{\mu}{r_p}\frac{\omega L_2}{R_2}\frac{\omega M}{1 + \dfrac{\omega^2 M^2}{R_2 r_p}}. \tag{15}$$

Appendix B

General Equations for Fig. 2-A.

Let Z be the impedance of the tuned circuit.

$$Z=\frac{\omega L-jR_2}{R_2'\omega C_2+j(\omega^2L_2C_2-1)} \tag{1}$$

where R_2' includes the effect of the grid resistor.

Let Z_1 be the total impedance of the circuit.

$$Z_1=Z+r_p=\frac{r_pR_2'\omega C_2+j(\omega^2L_2C_2-1)+\omega L-jR_2'}{R_2'\omega C_2+j(\omega^2L_2C_2-1)} \tag{2}$$

$$i_p=\frac{\mu eg_1}{Z_1} \tag{3}$$

$$e_p \text{ or } eg_2=\frac{\mu eg_1}{Z_1}Z \tag{4}$$

$$\text{Av.}=\frac{eg_2}{eg_1}=\mu\frac{\dfrac{\omega L_2-jR_2'}{R_2'\omega C_2+j(\omega^2L_2C_2-1)}}{\dfrac{R_2'r_p\omega C_2+\omega L_2+j(r_p\omega^2L_2C_2-r_p-R_2')}{R_2'\omega C_2+j(\omega^2L_2C_2-1)}}. \tag{5}$$

The absolute value of Av., neglecting the phase angle and R_2'/r_p, is

$$\text{Av.}=\frac{\mu}{r_p}\frac{\omega^2L_2^2+R_2'^2}{\sqrt{[C_2\omega(\omega^2L_2^2+R_2'^2)-\omega L_2]^2+\left[\dfrac{\omega^2L_2^2}{r_p}+R_2'\right]^2}} \tag{6}$$

Appendix C

Analysis of Fig. 4

Applying Kirschoff's first law to the points A, B, and C, we have

$$i_1+i_2+i_3+i_4=0 \tag{1}$$

$$\frac{e_1-e}{j\omega L}+e_1j\omega C_1+e_1g_1+[e_1-(e_2)]j\omega C_0=0 \tag{1a}$$

$$e_1\left[g_1+j\left\{\omega(C_1+C_0)-\frac{1}{\omega L}\right]-\frac{e}{j\omega L}+e_2j\omega C_0=0 \tag{1b}$$

Let $\tan\theta_1$ be $\dfrac{\omega(C_1+C_0)-\dfrac{1}{\omega L}}{g_1}$; then Eq. (1B) becomes

$$e_1 g_1(1+j\tan\theta_1)-\frac{e}{j\omega L}+e_2 j\omega C_0=0 \tag{1c}$$

$$-i_4+i_5+i_6+i_7+i_8+i_9=0 \tag{2}$$

$$-(e_1+e_2)j\omega C_0+(-e_2+\mu e_1)gp-\frac{e_2}{j\omega L}-e_2 gc-(e_2+e_3)j\omega C_0=0 \tag{2a}$$

$$-e_2\left(j\omega(C_2+2C_0)-\frac{j}{\omega L}+gp+g_2\right)-e_1 j\omega C_0+gme_1-e_3 j\omega C_0=0 \tag{2b}$$

$$gm=\mu gp$$

$$\text{Let}\quad gp+g_2=g_4$$

ωC_0 may be neglected in comparison with gm. Eq. (2B) then becomes

$$-e_2 g_4(1+j\tan\theta_2)+e_1 gm-e_3 j\omega C_0=0 \tag{2c}$$

$$-i_9+i_{10}+i_{11}+i_{12}+i_{13}=0 \tag{3}$$

$$(e_2+e_3)j\omega C_0+(e_3-\mu e_2)g_p+\frac{e_3}{j\omega L}+e_3 j\omega C_3+e_3 g_3=0 \tag{3a}$$

$$e_3\left[j\omega(C_0+C_3)-\frac{j}{\omega L}+gp+g_3\right]+e_2 j\omega C_0-e_2 gm=0 \tag{3b}$$

$$e_3 g_5(1+j+\tan\theta_3)-e_2 gm=0 \tag{3c}$$

Multiply (1C) by gm and (2C) by $g_1(1+j\tan\theta_1)$.

Subtract (2C) from (1C) after multiplying by the above quantities and we have

$$e_2\{j\omega C_0 gm+g_4 g_1(1+j\tan\theta_1)(1+j\tan\theta_2)\}$$
$$+e_3 j\omega C_0 g_1 gm(1+j\tan\theta_1)=\frac{egm}{j\omega L} \tag{4}$$

Eliminating e_2 between (4) and (3C) we have

$$e_3(j\omega C_0 gm g_1(1+j\tan\theta_1)+j\omega C_0 gm g_5(1+j\tan\theta_3)$$
$$+g_5 g_4 g_1(1+j\tan\theta_1)(1+j\tan\theta_2)(1+j\tan\theta_3)=\frac{eg^2m}{j\omega L} \tag{5}$$

$$\frac{e_3}{e} = \frac{gm^2}{j\omega L_1 g_1\ g_4\ g_5} \cdot \frac{1}{A+B} \tag{6}$$

where

$$A = (1+j\tan\theta_1)(1+j\tan\theta_2)(1+j\tan\theta_3) \tag{7}$$

$$B = j\omega C_0 gm\left(\frac{1}{g_4\ g_5} + \frac{1}{g_4\ g_1}\right) - \omega C_0 gm\left(\frac{\tan\theta_2}{g_4\ g_5} + \frac{\tan\theta_3}{g_4\ g_1}\right) \tag{8}$$

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THE EFFECT OF REGENERATION ON THE RECEIVED SIGNAL STRENGTH*

By

BALTH. VAN DER POL

(Physical Laboratory, Philips' Glowlamp Works, Ltd., Eindhoven, Holland)

Summary—*It is the purpose of this paper to give a theory of the effect of regeneration using the solution of a non-linear differential equation, and to present experimental verification of the theory.*

It is shown that: (a) as a first approximation, detection has no effect on the radio-frequency grid voltage developed under the influence of an incoming signal, (b) the amplification obtained through regeneration equals the two-thirds power of the ratio of the "grid space" to the amplitude obtained with zero regeneration. It is apparent from (b) that much greater gain is obtained through regeneration with weak signals than with strong signals.

The verification of the theory is made with a circuit arrangement operating at 500 cycles per second. Application is made to radio frequencies using this as a model of a high-frequency system, following a theorem for model systems which is stated.

THE considerable increase in signal strength obtainable through the use of regeneration is well known. It is also common knowledge that this increase in signal strength is considerably greater when the incoming signal is weak than when it is strong. It is the purpose of this paper to provide a non-linear theory of this effect of regeneration and the experimental verification thereof.

Suppose a triode system of one degree of freedom, as in Fig. 1 where L, C, r form the tuned circuit and the mutual induction M provides the regeneration.

Let further an emf, $E \sin \omega_1 t$, representing the "incoming signal," to be applied to the oscillatory system. Neglecting the grid current and calling the current in the LCr circuit i, and the deviation of the anode current from its steady value i_a and the alternating grid P.D. V_g, we have

$$L\frac{di}{dt}+ri+\frac{1}{C}\int idt-M\frac{di_a}{dt}=E\sin\omega_1 t \qquad (1)$$

$$\frac{1}{C}\int idt=v_g \qquad (2)$$

* Dewey decimal classification: R134.4. Corrected copy; reprinted from August, 1928, issue of the PROCEEDINGS. Paper read at the meeting of the International Union of Scientific Radio-telegraphy, October, 1927.

In the "tuned grid" circuit, the variations of anode potential are usually small compared with the variations of the grid potential, so that as a first approximation, the anode current variation i_a is a function of the grid potential variation only, i.e.,

$$i_a = f\ (v_g) \tag{3}$$

In order to avoid complexity we neglect the grid current and we therefore imagine a negative grid bias to be provided. Our system of coördinates in the i_a, v_g plane is therefore as shown in Fig. 2. This differs from the usual notation in so far as the zero point of the coördinate system is shifted towards the steady d.c. position round which the oscillations occur.

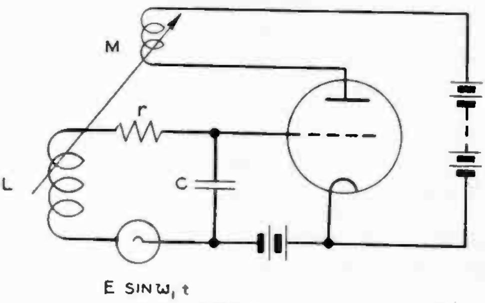

Fig. 1

We approximate equation (3), representing the curved plate current-grid voltage characteristic of the triode, by the cubic:

$$i_a = S_1 v_g + S_2 v_g^2 - S_3 v_g^3 \tag{3a}$$

where S_1 is the usual "mutual conductance" for infinitesimal grid-potential variations. S_2 and S_3 are further determined by the form of the characteristic. As the latter bends round both at the top and at the bottom we write in (3a) $-S_3 v_g^3$ instead of $+S_3 v_g^3$.

The elimination of i and i_a from (1), (2), and (3a) results in:

$$\frac{d^2 v_g}{dt^2} + \left\{\left(\frac{r}{L} - \frac{MS_1}{LC}\right) - \frac{2MS_2}{LC} v_g + \frac{3MS_3}{LC} v_g^2\right\}\frac{dv_g}{dt} + \omega_0^2 v_g = \omega_0^2 E \sin \omega_1 t. \tag{4}$$

where

$$\omega_0^2 = \frac{1}{LC}\cdot$$

Calling further

$$\frac{r}{L} - \frac{MS_1}{LC} = \alpha,$$

$$\frac{MS_2}{LC}=\beta,$$

$$\frac{MS_3}{LC}=\gamma,$$

(4) becomes

$$\ddot{v}_g+(-\alpha-2\beta v_g+3\gamma v_g^2)\,\dot{v}_g+\omega_0^2 v_g=\omega_0^2 E\sin\omega_1 t. \tag{4a}$$

which is a non-linear inhomogeneous differential equation of the second order with non-linear resistance terms. The usual elementary approximation of a linear characteristic would make $\beta=\gamma=0$ and for resonance (i.e. $\omega_0^2=\omega_1^2$) and critical regeneration (i.e. $\alpha=0$) the developing grid voltage v_g would become infinite. In order to obtain a satisfactory theory of the response of a regenerative triode system to an expressed emf a non-linear problem must therefore be solved.

The equation (4a) was fully considered in a former paper[1]; from the results obtained there it follows that the steady state solution in the neighborhood of resonance ($|\omega_0-\omega_1|\ll\omega_1$) can be written:

$$v_g=b\sin(\omega_1 t+\phi)$$

where the amplitude b of the resulting grid potential variation is given by:

$$b^2\left\{4(\omega_0-\omega_1)^2+(\alpha-\tfrac{3}{4}\gamma b^2)^2\right\}=\omega_0^2E^2 \tag{5}$$

a cubic equation in b^2. It is further seen from (5) that the symmetrical term in (4a) (with β) (which determines the detection) has, in the first approximation here considered, no influence on the resulting v_g. *Therefore as a first approximation detection has no effect on the h.f. grid potential difference developing under the influence of an incoming "signal."*

Further if (5) is compared with the usual linear case as represented by

$$b^2\left\{4(\omega_0-\omega_1)^2+\alpha_0^2\right\}=\omega_0^2E^2,$$

it follows that in the non-linear case $(\alpha-\tfrac{3}{4}\gamma b^2)$ is substituted for α_0, i.e. the system behaves as if for the resistance r a new resistance r' were substituted of a value given by

$$r'=r-\frac{M}{C}\left(S_1-\frac{3}{4}b^2S_3\right) \tag{6}$$

[1] *Phil. Mag.* **3**, 65, 1927.

which therefore depends upon the amplitude b already present in the system.[2]

When the regeneration is pushed so far that

$$MS_1 > rC \tag{7}$$

it is seen from (6) that the first order differential resistance of the system becomes negative and therefore the system has the tendency to oscillate spontaneously. However, as was shown in the *Phil. Mag.* paper quoted above, the forced oscillations may suppress the development of the free oscillations. The phenomenon manifests itself through the presence of a "silent region" extending at both sides of resonance. As was shown by Professor Appleton[3] the width of this silent region is determined by the amplitude of the incoming signal. For strong signals this width is given by

$$\frac{\omega_0 - \omega_1}{\omega_0} = \pm E\sqrt{\frac{3\gamma}{\alpha}} \tag{8}$$

It is of interest to investigate the resultant grid amplitude b if (a), the system is tuned exactly to resonance, i.e.,

$$\omega_0 = \omega_1$$

and (b), it is brought on the verge of free oscillation, i.e., when

$$S_1 M = Cr. \tag{9}$$

For this case we at once obtain from (5)

$$\tfrac{3}{4}\gamma b^3 = \pm\omega_0 E, \tag{10}$$

which expression will now be considered in detail.

First, it follows from (10) that *the resulting grid amplitude is proportional to the cube root of the emf applied to the system.*

Further, it is easy to assign an approximate value to γ directly from the triode characteristic. Referring to (3a), (4a) and Fig. 2, and taking the symmetrical case for which $S_2 = 0$, it follows that a good approximation to S_3 can be found from the maximum and minimum value of the cubic (3a), for $i_a = i_{a\ max}$ when

$$v_g^2 = \frac{S_1}{3S_3}.$$

[2] This property of the non-linear system was first derived in 1920. See Balth. van der Pol, *Radio Review*, Nov., Dec., 1920.

[3] E. V. Appleton, *Proc.* Camb. Phil. Soc., 23, 231, 1923.

Calling, therefore, the grid voltage change necessary to bring the anode current from zero to its saturation value the "grid space" and designing it by V_{g0} (see Fig. 2), we obtain

$$V^2_{g0} = \frac{4S_1}{3S_3} \tag{11}$$

hence

$$\gamma = \frac{M}{LC} \cdot \frac{4S_1}{3V^2_{g0}} \cdot$$

But, when the system has critical regeneration (9) obtains, i.e.,

$$M = \frac{Cr}{S_1}$$

hence

$$\gamma = \frac{r}{L} \cdot \frac{4}{3V^2_{g0}} \cdot$$

When we further call v_{g1} the grid-voltage amplitude which would be obtained with no regeneration at all, (M = 0), or (which is the same thing) with reduced filament current, we find from (5)

$$v^2_{g1} = \frac{\omega^2{}_0 L^2}{r^2} \cdot E^2,$$

hence we obtain from (10):

$$b^3 = v_{g1} \cdot V^2_{g0}$$

or

$$b = \sqrt[3]{v_{g1} \cdot V^2_{g0}} \tag{12}$$

reading in words:

The grid amplitude developing in resonance and with critical regeneration equals the cube root of the product of the grid amplitude which would be obtained with no regeneration at all, and the square of the "grid space," as defined by Fig 2.

(12) can further be written:

$$\frac{b}{v_{g1}} = \left(\frac{V_{g0}}{v_{g1}}\right)^{2/3} \tag{13}$$

which means that: *the amplification obtained through regeneration equals the two-third power of the ratio of the "grid space" V_{g0} into the amplitude obtained with zero regeneration.* The much greater gain obtained through regeneration with small signals than with stronger ones is at once apparent by (13).

Some measurements provided a very satisfactory experimental verification of the theory outlined above.

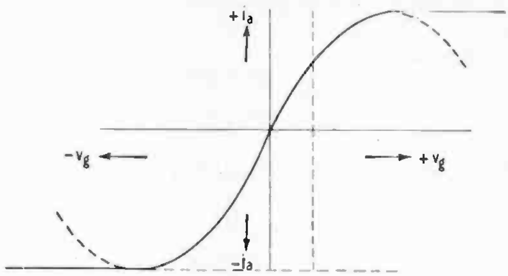

Fig. 2

The measurements were taken with relatively low frequency, thus avoiding obvious errors. That the results are, however, equally applicable to high-frequency circuits follows at once from the following theorem:

If a model is made of a high-frequency system consisting of linear inductances, linear capacities and non-linear resistances (e.g. triodes) and if the values of all the inductances (self and mutual) and capacities in the model are made n times these values in the original high-frequency system, but if the resistances (linear and non-linear) in the model are made equal to the resistances in the original circuit, the currents and potentials occurring in this model will be exactly equal in magnitude to the currents and potentials in the original high-frequency system but, considered as a function of the time, they will vary n times slower.[4]

Therefore, the natural periods of the model will be n times those of the original system, and the building up and decay of currents in the model will also occur n times slower.

This simple theorem at once follows from a dimensional consideration of the coefficients such as r, ωL, $1/\omega C$, $M\omega$, occurring in the differential equations, because, e.g.

$$\omega L = \frac{\omega}{n} \cdot nL, \text{ etc.}$$

[4] Incidentally, in this model system the stray capacities are reduced n times in magnitude. Therefore, in order to investigate the effect of a specified stray capacity in the original high-frequency circuit it is only necessary to insert at its place in the model circuit a capacity n times the original stray capacity.

In fact, with a system having a natural frequency of 500 Hertz (cycles per second) and with critically adjusted regeneration it often took a minute for the free oscillation of the triode system to reach its final value, hence it was not easy to decide, before the external emf was applied, whether the regeneration was exactly critical or not. Therefore, an intermediate way was chosen and a model of a receiving set was made having a natural frequency of 15000 Hertz.

Four sets of readings were taken[5] with a standard triode (tungsten filament $V_f=4.0$ volts, saturation current 11 milliamperes, anode potential varying between 100 and 200 volts, negative grid bias varying between -4.5 and -7.5 volts).

The applied emf $E \sin \omega_1 t$ was varied between 2.10^{-5} and 10^{-2} volts, the resulting alternating grid voltage V_{g1} with no regeneration varied between 100.10^{-6} and 0.5 volts, and the resulting alternating grid voltage b with critical regeneration varied between ca. 0.2 and 4 volts. The exponent s in the formula

$$b=(V^2_{g0}\cdot v_{g1})^s$$

which, according to the above theory, should be

$$s=0.33$$

was found from the four sets of measurements to be

$$\begin{array}{r} s=0.36 \\ 0.36 \\ 0.36 \\ 0.32 \\ \hline \text{mean: } s=0.35 \end{array}$$

while the "grid space" V_{g0} calculated from these measurements was

$$\begin{array}{r} V_{g0}=26 \text{ volts} \\ 25 \phantom{\text{ volts}} \\ 20 \phantom{\text{ volts}} \\ 15 \phantom{\text{ volts}} \\ \hline \text{mean: } V_{g0}=22 \text{ volts,} \end{array}$$

[5] The experiments and calibration of the necessary amplifiers were performed by Messrs. K. Posthumus and R. Veldhuyzen. Care was taken that the receiver did not react on the transmitter.

as compared with the value

$$V_{g0} = 27 \text{ volts}$$

obtained directly from the characteristic.

The experimental value for the exponent s fits in well with the theoretical value; provided only those parts of the characteristic are considered for which no oscillation hysteresis occurs. The representation of the i_a, v_g characteristic, by a cubical parabola, with three constants S_1, S_2, and S_3 only, obviously cannot yield very accurate values for V_{g0} obtained with a voltage swing of not more than 0.35 volts.

Therefore, there is no doubt that for all practical purposes the theory given above fits in well with experiment.

Finally, we give some practical figures calculated from formula (12) with the following data:

$$\frac{\omega L}{r} = 40,$$

$$V_{g0} = 27 \text{ volts}.$$

Electromotive force E working in grid circuit	Resulting alternating grid voltage with no regeneration (v_{g1})	Resulting alternating grid voltage b, with critical regeneration	Amplification obtainable through critical regeneration (b/v_{g1})
10^{-6} Volts	0.04×10^{-3} Volts	0.31 Volts	7700
10^{-5} "	0.4×10^{-3} "	0.66 "	1600
10^{-4} "	4.0×10^{-3} "	1.4 "	360
10^{-3} "	0.04 "	3.1 "	77
10^{-2} "	0.4 "	6.6 "	16

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RECEPTION EXPERIMENTS IN MOUNT ROYAL TUNNEL*

By

A. S. Eve,[1] W. A. Steel,[2] G. W. Olive,[3] A. R. McEwan,[4] and J. H. Thompson.[5]

([1] Director of Physics Dept., McGill University, Montreal, Canada; [2] Chief Technical Officer, The Royal Canadian Corps of Signals, Ottawa, Canada; [3] Director of Radio Dept., Canadian National Railways; [4] Chief Engineer, Canadian Marconi Company, Montreal, Canada; [5] Radio Engineer, Radio Dept., Canadian National Railways.)

Summary—*This paper deals with certain experiments carried out in the Mount Royal Tunnel of the Canadian National Railways at Montreal, Quebec, in order to determine how radio waves reach the receiving set. Preliminary experiments in 1926 indicated that the penetration of radio waves into the tunnel was a function of the frequency, short waves below 100 meters dying out within a few hundred feet of the mouth of the tunnel. More exact experiments and measurements were planned and carried out in 1928 to bring out the part played by the wires, cables, and rails leading into the tunnel. The tunnel mouths were blocked and the cables grounded, and the results so obtained indicated that the effect of cables and rails was also dependent on the frequency. It was also evident that more actual energy entered via the tunnel itself than was at first suspected. The effect of rails and cables is not a simple one but involves loop action, wave-antenna effects, and re-radiation.*

Curves are attached showing graphically the results obtained. A map of the area and an elevation of Mount Royal itself give general details of the geology of the region.

Introduction

EARLY in 1926 the problem of reception of radio transmission in the Mount Royal Tunnel in Montreal came up for consideration, and at the request of Dr. A. S. Eve arrangements were made to carry out a series of tests in order to determine the general nature and strength of radio waves under Mount Royal.

Information has been published from time to time in the *Technical Press* dealing with experiments of this nature both in England and America, but so far as is known, these experiments were confined to a judgment of signal strength by the ear and not to a measurement of the actual signal strength existing within the tunnel. It is a well known fact that the human ear is a very indifferent measuring instrument and it was therefore determined that in the proposed tests accurate measurements should be made.

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Before considering the experimental work it might be well to spend a little time in a discussion of the tunnel itself. The attached section of Mount Royal, Fig. 1, shows the outline of the mountain as well as the location of the tunnel, and in addition an indication is given of the nature of the rock of which the mountain is composed. The tunnel is $3\frac{1}{2}$ miles long and extends from Dorchester Street to the C. P. R. tracks on the northwest side of the mountain. The diagram reproduced herewith has been taken from a report on the "Essexites of Mount Royal," by Bancroft and Howard, of McGill University. It will be noted

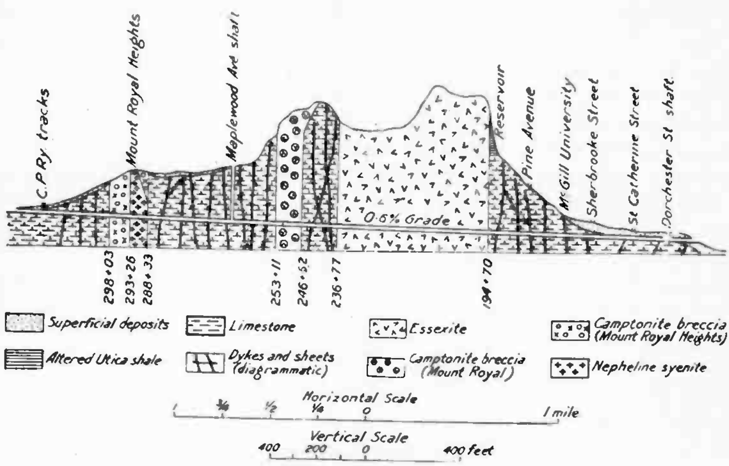

Fig. 1—Section of Mount Royal Showing Outline of Mountain and Location of Tunnel.

that the greater part of the mountain is composed of limestone, but the entire center section, that is, where the mountain is the highest, is composed of essexite. In addition to this there are one or two shafts of camptonite breccia in the northern part of the mountain. Attention is also drawn to the ventilation shaft at Maplewood Avenue.[1] This is the only ventilation shaft in the tunnel from Dorchester Street to the C. P. R. tracks. The tunnel is wide enough to carry a double track, and, in addition to the rails, the following conductors are carried through the tunnel from end to end:

(a) A lead covered conductor for 60-cycle, 12000-volt, 3-phase power.

[1] The tunnel also passes directly under the reservoir at Outremont not far from the ventilation shaft.

(b) Two phosphor bronze trolley wires over each track. These trolley wires are suspended from a 7/8-in. diameter phosphor bronze messenger cable, the messenger cable itself being thoroughly grounded.
(c) A lighting circuit for the tunnel carrying 60-cycle single phase at 120 volts.
(d) The negative return wire for the ground connection.
(e) 20 or 30 telegraph conductors, lead covered and insulated for 240 volts.
(f) Between Dorchester Street and the grotto there is an automatic track circuit 1650 feet long. There are numer-

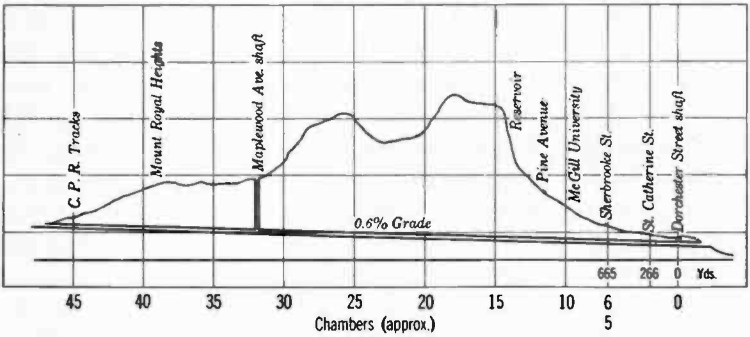

Fig. 2—Elevation, Mount Royal Tunnel, Montreal, Quebec, Canada.

ous signal and power wires required for this circuit, but the detail of the conductors is not definitely known.
(g) All four rails comprising the double track are thoroughly grounded with 250,000 circular mil stranded cable.

The tunnel is conveniently divided into chambers. By referring to Fig. 2 it will be noted that these chambers are not in all cases of uniform length. Chambers 5 and 6, for example, are both located at the same point in the tunnel, while Chambers 7 and 8 are very short. Again at the northwest end there are several chambers somewhat shorter than those in the central part. In general the chambers are 400 feet long. In plotting the curves for this report the base line has been taken as the subdivision in chambers, and an attempt has been made to take into consideration the essential differences mentioned above. This point, however, is not of very great importance in considering the results obtained.

The map attached to this report shows in general the direction taken by the tunnel as it passes under Mount Royal. It will be noted that the actual direction of the tunnel is nearly east and west and the Montreal end is referred to as the "east portal" and the Mount Royal Heights outlet as the "west portal."

Preliminary Tests

In June, 1926, arrangements were made by McGill University, the Canadian National Railways, The Canadian Marconi Company, and The Royal Canadian Corps of Signals, for a series of reception tests in Mount Royal Tunnel. It was thought that tests should be made on different wavelengths in order to determine whether the frequency affected the penetration of the signals into the tunnel, and 40 meters, 411 meters and 1300 meters were finally chosen, these being representative wavelengths in their respective bands. Signals from station WIZ were to be used for the short-wave band, as this station was nearly always in operation and the signals were remarkably clear and constant in Montreal. The Canadian Marconi Company were good enough to arrange for a series of programs from their Montreal station, gramaphone records were used, and care was taken to maintain a constant output throughout each test. For the 1300-meter signals, the 500-watt station of The Royal Canadian Corps of Signals in Ottawa was employed. The tests began at midnight and an hour's transmission on each of the three wavelengths was required to make a series of tests from the Dorchester Street station to the west end of the tunnel. No attempt was made to measure the field strength in these tests, but a carefully calibrated audibility meter, connected as a shunt across the telephones in the output circuit, was employed for each of the three bands of wavelengths. It is appreciated that tests of this nature are not highly accurate, but they are at least an improvement over any attempt that might be made to judge the signals by the ear alone. The order in which these tests were made was as follows:

(a) From midnight until 1 A.M., Eastern Standard Time, 411-meter signals were received.

(b) Between 1 and 2 A.M. signals on 1300 meters were received from Ottawa.

(c) From 2 to 3 A.M. signals from WIZ were used.

For each of these wavelengths a standard car aerial was employed. This aerial was of the box type and was mounted on top of the observation car used for the tests. For the short-wave reception the receiving set consisted of a detector and two steps of audio amplification. Marconi V.24 valves were used and regeneration was controlled by inductive coupling. A standard broadcast receiver of the neutrodyne type was used for the 411-meter wave. This was a 5-valve set operated without a power tube. An R. C. C. S. standard receiver was used for the long-wave band. This set had two stages of radio-frequency

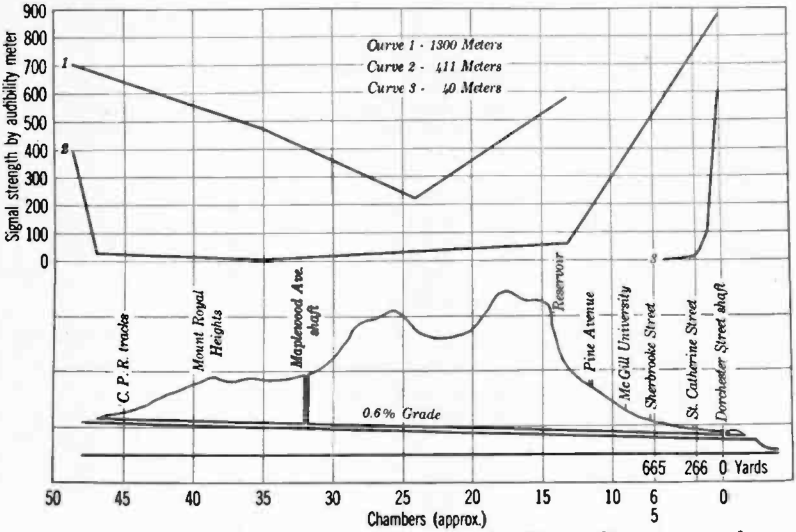

Fig. 3—Signal Strength, Measurements under Mount Royal Tunnel, Montreal, Quebec, Canada, June 16, 1926.

amplification and one stage of audio. On the radio-frequency side the coupling between valves was by means of the well known rejector circuit. In this receiver English "R" type valves were employed.

The procedure in taking all these tests was to start at the Dorchester Street terminus and to proceed by definite steps through the tunnel. In the case of the broadcast and long-wave bands, these steps were each approximately 10 chambers long. In the case of the short-wave signals, very much shorter steps had to be employed as the signals died out within a few hundred feet of the entrance to the tunnel.

The results of these experiments can best be followed by a study of Fig. 3. In this diagram the profile of Mount Royal itself has been laid down to the same base as is used for the plotting of the results. In this way it was possible to identify a given signal strength with the depth of rock existing over the tunnel at that point. The most noteworthy result was the rapid attenuation on the short-wave band. It will be observed that the signals were reduced to zero approximately 1500 feet from the mouth of the tunnel, that is, before Chamber 5 was reached. This result was so unexpected that repeated checks were made, the signal being traced step by step into the tunnel. The net result, however, in each case was as shown by Curve 3. It will be observed

TABLE I

42.5-Meter Tests—April 14, 1928

Position in Tunnel	Control Panel Reading in Current Ratios	Per Cent Signal Strength (5)	Remarks
300 ft. N of Chamber 45	13	100	
Chamber 45	10	76.8	
100 ft. S of Chamber 45	5	38.4	
150 ft. S of " 45	2.5	19.2	
200 ft. S of " 45	2.0	15.4	
300 ft. S of " 45	1.6	12.3	
Chamber 44	6.2	47.7	
100 ft. S of Chamber 44	2.0	15.4	
Chamber 43	1.0	7.7	
" 42	0.39	3.0	
" 41	0	0	

TABLE II

42.5-Meter Tests—April 15, 1928

		(8)	
300 ft. N of Chamber 45	3.1	100	
100 ft. S of " 45	1.25	40.4	
100 ft. S of " 45	1.6	51.5	
200 ft. S of " 45	0	0	

that there were only 48 feet of rock over the tunnel at the point where the short-wave signals disappeared. These tests were only made at the east or Dorchester Street end of the tunnel.

The 411-meter signals were audible throughout the entire length of the tunnel, but were very weak from Chamber 11 to Chamber 44. At Chamber 11 the strength was reduced to one twelfth of its maximum strength outside the tunnel. At this point the rock was approximately 275 feet deep. It should also be noticed that at Chamber 33, where the broadcast signals are weakest, the depth of rock is only 210 feet.

The 1300-meter signals were also audible throughout the tunnel, but were very much stronger than the broadcast signals although the two stations were of approximately the same power output. It should be noted, however, that the station from which the

1300-meter signals were received was about 100 miles from the tunnel, while the broadcast station was only about 3 miles away. In this case the signals reached their lowest value at Chamber 22, which is approximately the center of the tunnel.

From a consideration of these curves it would appear that the strength of signal at any point within the tunnel was a very definite function of the frequency employed. It still remained to be determined, however, whether these signals came through the open tunnel or along the conductors and rails running through the tunnel. Considerable discussion arose over this

TABLE III
55-Meter Tests—April 14, 1928

Position in Tunnel	Meter Reading	Per Cent Signal Strength (3)	Remarks
300 ft. N of Chamber 45	400	100	
Chamber 45	200	50	
100 ft. S of Chamber 45	10	2.5	Fading
150 ft. S of " 45	20	5.0	
200 ft. S of " 45	15	3.75	
300 ft. S of " 45	9	2.25	
Chamber 44	15	3.75	
100 ft. S of Chamber 44	12	3.0	
Chamber 43	10	2.5	
" 42	5	1.25	
" 41	1	0.25	

TABLE IV
55-Meter Tests—April 15, 1928

		(6)	
100 ft. S of Chamber 45	100	100	
200 ft. S of " 45	15	15	
300 ft. S of " 45	50	50	
Chamber 44	40	40	
" 43	30	30	
" 42	8	8	
" 41	0	0	

point and two general theories were evolved as a result of the work done. The first theory was that the greater part of the energy reached any given point within the tunnel by passing through the rock. It was appreciated that a certain percentage must come along the wires and perhaps through the tunnel itself. In support of this theory the following figures are of interest. From Zenneck's treatise on "Radio Telegraphy," page 248, the following figures with regard to conductivity are obtained,—

Conductivity of sea water, $\delta = 1$ to 5×10^{-11} E. M. Units
Dryish earth, $\delta = 10^{-14}$ E. M. Units

Ratio = 1 to 1000

From the same authority we learn that one meter of sea water reduces the strength of signals by 10 per cent. Therefore, it is assumed that it would require 1000 meters of rock to produce the same reduction in strength.

The other theory advanced considered that the greater part of the energy entered the tunnel via the tunnel mouth, although a proportion undoubtedly came along cables and rails. The curves of Fig. 3 afford the best support for this theory. From these curves it will be noted that the short wave dies out within a few feet of the mouth of the tunnel although there exists only a thin layer of rock at this place. Again, it will be noted that the 411-meter signal continues to diminish in strength as we go from the Dorchester Street shaft towards the west end of the tunnel and at Chamber 33 the lowest signal was received, although the rock is considerably shallower at this point than it is in the center of the tunnel. From Chamber 33 on, the signal increases in strength, but this might be accounted for by energy coming in via the west portal.

The 1300-meter signal is the only one that seems to follow the rock penetration theory; here the weakest signal was received under the center of the mountain, that is, under the greatest depth of rock.

On account of shortage of time it was not possible to carry out further experiments during 1926, in order to determine which theory was the more nearly correct.

During the summer of 1927 Dr. Eve was in a position to carry out further tests in the Caribou Mine of the American Mining and Prospecting Company at Caribou, Colorado.[2] The following extract from this report will indicate quite clearly the tests made and the results obtained—

> The experiments participated in by Dr. Eve were conducted with a superheterodyne set with nine electron tubes in the Caribou mine of the American Mining and Prospecting Company, at Caribou, Colorado. The first test was held at a depth of 220 feet, where, by means of a loop, a strong and clear reception was obtained of a musical concert given at Denver, 50 miles distant. The evidence pointed strongly to the conclusion that this clear reception was due to the penetration by the radio waves of the solid rock strata, although there was a remote possibility that the reception was obtained through shafts and crosscuts, toward which, however, the loop did not point. The nearest metal conductors, iron rails, were 66 feet away.

[2] A resumé of these tests appeared in the *Jour.* A.I.E.E., November, 1927. See also Technical Papers 434, pages 37-40, Dept. of Commerce, Bureau of Mines, Washington, D. C., U. S. A.

The next series of experiments was conducted at a depth of 550 feet, when "mushy" reception was obtained from Denver. This type of reception was, however, as good as could be obtained above ground at the time of making the test, the night being unfavorable for general radio reception. This series of tests was conducted at the end of a cross-cut reached with many turns, and 200 feet from the main shaft. A pipe came down the shaft and followed the tunnel up to 80 feet from the point of observation.

In previous experiments conducted by the Bureau of Mines at its Experimental Mine near Pittsburgh, Pa., it was at first concluded that radiation and induction would penetrate rock for considerable depths. Subsequent investigations have shown that in every case the transference of radiation was by some conductors in the mine, electric wires, pipes or rails, all of which abound in modern mines.

It is felt that further investigations should include a comparison of the penetration of radio waves greater and less than a wavelength.

It is pointed out that in making these tests no attempt was made to assist the ear in determining the relative signal strengths. It is a known fact that the unaided human ear is a relatively poor test instrument. The following facts in this connection should be kept in mind when comparing the results of the two

TABLE V
411-Meter Tests—April 14, 1928.

Position in Tunnel	Meter Reading	Per Cent Signal Strength (1)	Control Panel Reading in Current Ratios	Per Cent Signal Strength (2)
Dorchester Station	42	100	12.5	100
Chamber 5	11	36.2	6.2	49.6
" 10	7	16.7	2.0	16.0
" 15	5	11.9	2.0	16.0
" 20	2	4.75	1.25	10.0
" 30	4.5	10.7	2.0	16.0
" 40	22	52.4	12.5	100
" 45	18	42.9	0.39	3.12
300 ft. N of Chamber 45	45	107.0	10.0	80

TABLE VI
411-Meter Tests—April 17, 1928.

		(10		(11)
Chamber 1	3000	100	13.9	96.5
" 5	500	18.3	13.0	90.3
" 10	350	11.65	13.0	90.3
" 15	80	2.67	5.0	34.7
" 20	60	2.0	4.0	27.8
" 25	300	10.0	6.2	43.0
" 29	500	16.7	13.0	90.3
" 30	800	26.7	6.2	43.0
" 31	600	20.0	0.8	5.55
" 32	350	11.65	0	0
" 35	550	18.3	3.2	22.2
" 40	400	13.3	7.9	54.8
" 44	500	16.7	14.4	100

separate series of tests. Fig. 4 is a curve taken by means of a Shaw fading recorder and represents the actual carrier waves being received at the receiving station. Particular attention should be paid to the curve representing the carrier wave from WEAF on the night of April 10th, 1928. The signal in the loud-speaker due to the carrier wave, and represented by the height from the zero line 0, to the dotted line *A*, was a good average

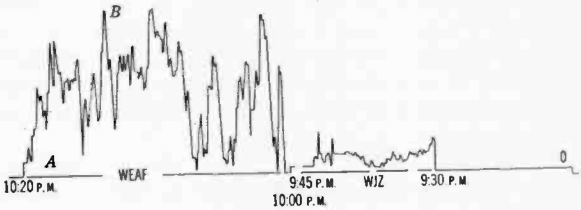

Fig. 4

signal, loud enough to be heard throughout an ordinary room. The signal equivalent to the peaks of the wave, while much stronger, did not sound loud enough to bear the proportion indicated by the relative heights of the corresponding lines. In fact, it is doubtful if the human ear would notice that the signal had changed by more than 50 per cent between the point *A* and the point *B*. Again referring to the tables of results in Appendix 1, attention is called to column 10, Table VI. In Chamber 1 the strength

TABLE VII
1400-METER TESTS, APRIL 14, 1928.

Position in Tunnel	Meter Reading	Per Cent Signal Strength (4)	Remarks
Dorchester Station	1000	100	
Chamber 11	300	30	
" 19	200	20	
" 31	400	40	
" 39	800	80	

TABLE VIII
1400-METER TESTS, APRIL 15, 1928.

		(7)	
Dorchester Station	1000	100	
Chamber 1	300	30	
" 11	150	15	
" 19	200	20	
" 31	400	40	
" 39	1000	100	
" 45	1500	150	
300 ft. N of Chamber 45	5000	500	

of signal is represented by a meter reading of 3000 while at Chamber 20 this had dropped to a reading of 60. To the human ear this did not appear to be a change of more than 100 per cent, when estimating the relative strengths of the signals coming from the loudspeaker.

In judging the results of the Colorado tests from the point of view of the theories already advanced, it would appear that they seem to strengthen the impression that the energy reached the receiver by passing through the ground or rock. It must be remembered, however, that no attempt was made to measure the strength of the carrier wave and that the ear alone has been shown to be a very poor measuring instrument. The differences have already been enumerated. It would, therefore, be inter-

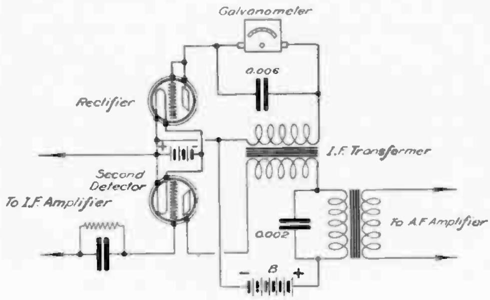

Fig. 5

esting to compare the results obtained in these two tests, insofar as the relative thickness of the covering is concerned. According to the Colorado figures radio waves should penetrate earth and rock to a depth of at least 550 feet. In the first tunnel experiments it was found that 48 feet of rock were sufficient to reduce the 40-meter wave to zero. In the case of the 411-meter wave, which was approximately the wavelength used in Colorado, the signal was reduced to one twelfth of its maximum strength by 275 feet of rock. In addition to this, the curves of Fig. 3 show that the strength of the 411-meter wave was of a very low order from Chamber 11 to almost the west portal of the tunnel although from Chamber 28 onward the thickness of rock is steadily decreasing. It is hardly fair to compare the 1300-meter results under Mount Royal with the Colorado tests as it is evident from the work already carried out that frequency has a very important bearing on the rate of attenuation in the tunnel.

Tests of April, 1928

In view of the conflicting results obtained in the preliminary tests it was decided to carry out a series of experiments, during the spring of 1928, in the Mount Royal Tunnel and to endeavor to settle definitely the various problems which arose as a result of the first work undertaken. Assuming a given strength of signal at any point within the tunnel three problems had to be solved—

(1) The percentage of the signal arriving by conduction through the rock.

(2) The percentage of the signal being conducted along rails and cables leading into the tunnel.

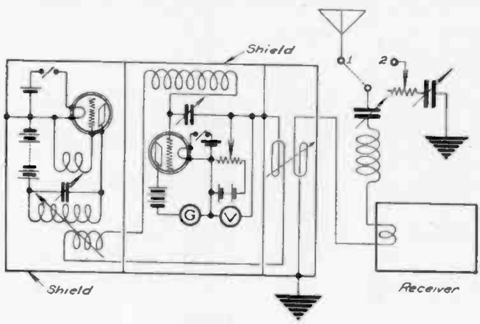

Fig. 6

(3) The percentage of signal entering the tunnel mouth and being conducted as in the case of waves passing over the surface of the ground.

From the description already given of the tunnel it will be appreciated that the separation of these three effects presented a difficult problem. It was not possible to take out the cables completely and while a section of the rail could have been removed, it is doubtful if any small section would, in itself, show an appreciable effect in view of the comparatively small dimensions of the tunnel and the large number of electrical conductors passing through it. After careful consideration it was decided to carry out the following three series of tests—

(1) With all rails, wires, and power conductors as for normal operation of the tunnel, tests were to be made on four wavelengths. These wavelengths were to be located in the following bands—

(a) Short wavelength band below 100 meters.
(b) Broadcast band.
(c) Medium wave band between 1000 and 2000 meters.
(d) Long-wave band above 10,000 meters.

(2) On the second night the same series of tests were to be run, but steel coaches were to be placed in each end of the tunnel so as to block both portals as far as possible.

(3) On the third night the same series of tests were to be made with the tunnel blocked and with all power, light,

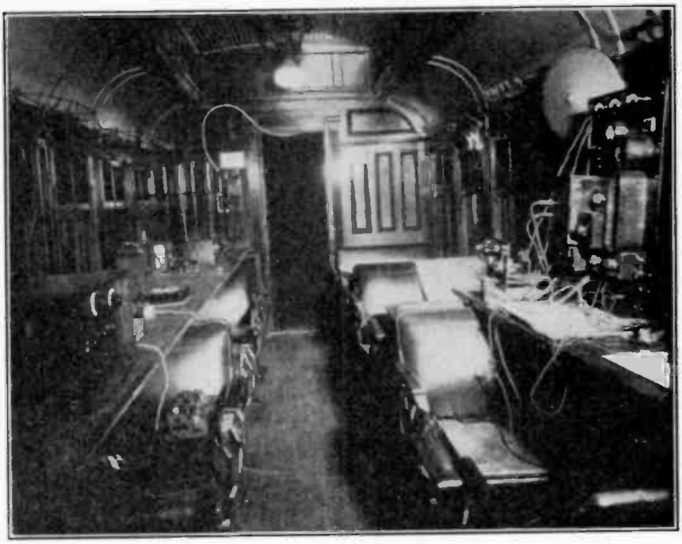

Fig. 7

and signal cables thoroughly grounded. In addition to this the overhead trolley wires were to be grounded at two points, one at either end of the tunnel.

It is appreciated that these tests would not constitute a complete solution of the problems, but it was hoped that they would, at least, indicate the general nature of the transmission. A wooden coach was used to carry the equipment, the standard pullman car aerial being used for the medium and long-wave results, but a special short-wave antenna was mounted inside the car for the short-wave signals.

In assembling the equipment for this work it was kept in mind that in each case the actual strength of the carrier wave was to be measured. This was not possible in the case of the short-wave receiver, as a field strength measuring set was not available for this band. In place of this a thermionic voltmeter was used to measure the actual voltage developed across an inductance coil in the output circuit. In addition to this method of measurement, an audibility meter was available so that the receiving set could be thrown from one to the other and check results rapidly obtained. For the broadcast reception a superheterodyne set with a loop receiver was employed, the strength of carrier wave being measured by the method described by Dr. G. W.

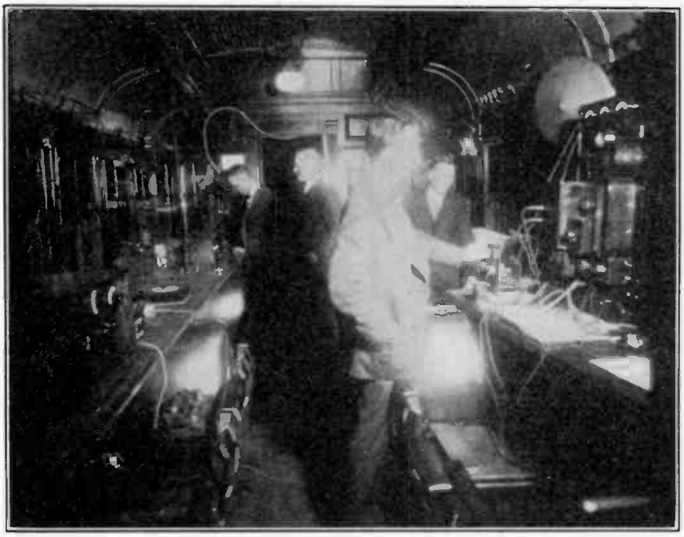

Fig. 8

Pickard.[3] This scheme consists essentially of introducing an additional intermediate frequency transformer into the second detector circuit of the superheterodyne. The output from this transformer is rectified and measured on a sensitive galvanometer. Fig. 5 shows the general principle of the method.

In addition to these two methods of measurement for the short and broadcast waves, a standard Western Electrical control panel was set up by the Canadian National Railways Radio Department and used with standard receivers to check the strength of the received signals. These results were measured in transmission units, and transferred to current ratios, for plotting in the form of curves.

[3] G. W. Pickard, *Radio Broadcast*, August, 1926.

The medium-wave signals were measured by using a thermionic voltmeter across an inductance in the detector circuit, the current in this inductance being proportional to the strength of the carrier wave. In this case the audibility meter was also available for purposes of checking.

For the long wave the Canadian Marconi Company supplied a signal strength measuring set similar to the type used by Captain Round during his expedition to Australia in 1922 and 1923. Fig. 6 is a schematic diagram of the equipment used.[4]

Figs. 7 and 8 attached will show the set-up used in the tunnel tests. These figures are from photographs taken inside the car while the measurements were being made. Referring to Fig. 7, the medium wave set may be observed in the foreground on the

TABLE IX
1400-Meter Tests, April 17, 1928.

Position in Tunnel	Meter Reading	Per Cent Signal Strength (12)	Per Cent Signal Strength (12a)	Remarks
Dorchester Station	1000		100	
Chamber 1	500	100	50	
" 5	800	160	80	
" 10	800	160	80	
" 15	150	30	15	
" 20	800	160	80	
" 25	1000	200	100	
" 29	200	40	20	
" 30	150	30	15	
" 31	200	40	20	
" 32	800	160	80	
" 35	600	120	60	
" 40	500	100	50	
" 44	2000	400	200	

left hand side of the photograph. Immediately behind it is the loop and broadcast receiver while on the right of the photograph may be seen the Western Electric control panel and one of the short-wave receivers. On account of its size the Marconi field strength set had to be mounted in the after compartment of the car.

In order to minimize the effects of fading and changes in power output at the transmitting station, it was decided to utilize, as far as possible, transmitting stations over which control could be exercised. Short and medium-wave signals were available from the Ottawa stations of The Royal Canadian Corps of Signals. As previously stated these stations were located about 100 miles from Montreal. Previous tests had shown that 55 meters was a

[4] A complete description of the method of operation will be found in the *Jour.* I.E.E., October, 1925.

very satisfactory wave for Montreal and this was therefore decided upon for these tests. The medium-wave transmission was on a wavelength of 1400 meters. In both of these cases the power output to the antenna was maintained at a constant value of 750 watts both for the short wave and for the medium wave signals. The broadcasting station of the Northern Electric Company in Montreal was secured for broadcast transmission on 411 meters. Arrangements were made to hold the output and modulation of this station very constant. The modulation was reduced

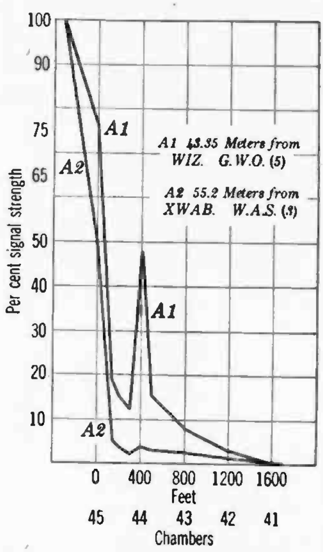

Fig. 9—Comparison of Signal Strengths. Short-Wave Tests, April 14, 1928; Tunnel Open.

from that normally employed and was approximately 15 per cent during the actual tests.

As there was no long-wave station in Canada that could be used for the 10,000-meter results, it was decided that one of the Radio Corporation stations on Long Island would be copied. The stations actually used for these tests were WCI on 16,700 meters, WSS on 16,120 meters, and WQK on 16,465 meters. The tabulated results from these stations will be found in Appendix 1, Tables X, XI, XII, and XIII.

The program followed for each of the three nights of the test was as follows; from midnight to 1 A.M. broadcast reception was measured. From 1:15 to 2:15 A.M. short-wave signals were received. From 2:30 to 3:30 both medium-wave signals and long-wave signals were checked as it was possible to carry out both of these measurements without interfering one with the other

In order to check possible errors due to natural fading, observers were stationed in and around Montreal for the purpose of recording the approximate strength of the various stations throughout the time of the test in the tunnel. Fortunately, it was found that there was very little natural fading during the three nights. As the medium and long-wave signals are not so

TABLE X
LONG-WAVE TESTS, APRIL 14, 1928.

Position in Tunnel	Reading in Microvolts	Per Cent Signal Strength (13)	Station and Wavelength	Remarks
Dorchester Station	233	100	WSS	Power on
Chamber 39	160	68.7	16,120	Moving North
200 ft. N of Chamber 45	1530	656	meters	
		(14)		
Dorchester Station				
Chamber 11	600	257	WSS	Power off
" 19	435	186	16,120	Moving North
" 31	870	373	meters	
" 39	870	373		
200 ft. N of Chamber 45	1530	656		
		(15)		
Dorchester Station	1100	472	WSS	Power on
Chamber 11	600	258	16,120	Moving South
" 19	260	111.5	meters	
" 31	2050	880		
		(16)		
Dorchester Station	600	258	WSS	Power off
Chamber 11	1300	558	16,120	Moving South
" 19	1300	558	meters	
" 31	2500	1071		

much affected by fading as are the broadcast and short-wave bands, no check record was kept of the signal strengths either for the 1400-meter or the 16,000-meter stations.

The following brief description of the method of preparing the tunnel for the tests on the second and third nights may be of interest. The steel cars were placed just inside the mouth of the tunnel where the tube has its smallest diameter. It was estimated that approximately 80 per cent of the area of the tunnel mouth was blocked by the two steel cars. The power cables were treated as follows,

(a) The 110-volt a.c. lighting circuit was opened at both ends by means of switches, but was not grounded.

(b) The 2400-volt d.c. feeder was opened at the sub-station. From the sub-station to the tunnel mouth there is a feeder approximately 1000 yards long. This feeder was not opened at the tunnel. The 2400-volt trolley wire terminates in the Dorchester Street station at the East end of the tunnel. This trolley wire, together with the supporting cable, was grounded to the bonding cable at Chambers 5 and 43. These points were not quite at the two ends of the tunnel, but it was convenient to make a thoroughly satisfactory ground at these two points.

(c) The 12,000-volt 3-phase power cable which supplies power to the sub-station and to the village of Mount Royal enters at the Dorchester Street Station and runs through the tunnel. It is carried in a duct between the two halves of the tunnel. This cable is covered with a lead sheath and the lead sheath was grounded at several points, but the power was not cut off during any part of the tests.

(d) All telegraph and signal wires are carried through the tunnel in lead covered cable. These wires were not opened at any time during the test, but the lead sheath was at all times thoroughly grounded in several places.

Complete tables of the results of the April tests are shown. As already pointed out more than one system of measurement was used. In view of this difference in methods all data have been reduced to the basis of "Per Cent Signal Strength," taking as the unit the strength of signals received in the Dorchester station, with each different system of measurement; that is, the signal received at this point was considered as a 100 per cent signal. The columns marked "Per Cent Signal Strength" are numbered from 1 to 21 inclusive, and the same numbers will be found on the curve sheets attached. Tables I to IV, inclusive, cover the short-wave tests, Tables V and VI the broadcast tests, Tables VII to IX, inclusive, are for the 1400-meter signals, and Tables X to XIII, inclusive, are the long-wave results. Thirteen curve sheets, Figs. 9 to 21, have been drawn based on these tables. In these curves,

A represents short-wave results,
B broadcast or 411 meters,
C 1400 meters,
D 17000-meter readings.

TABLE XI

Long-Wave Tests, April 15, 1928.

Position in Tunnel	Reading in Microvolts	Per Cent Signal Strength (17)	Station and Wavelength	Remarks
Dorchester Station	325	100	WCI 16,700 meters	Power on Moving South
Chamber 1	163	50		
" 11	455	140		
" 19	130	40		
" 31	490	151		
" 39	130	40		
" 45	360	111		
300 ft. N of Chamber 45	1170	360		
		(18)		
Dorchester Station	500	154	WCI 16,700 meters	Power off Moving South
Chamber 1	910	280		
" 11	1170	360		
" 19	780	240		
" 31	780	240		
" 39	780	240		
" 45	1050	323		
300 ft. N of Chamber 45	1170	360		

TABLE XII

Long-Wave Tests, April 16, 1928.

		(19)		
Chamber 1	65	100	WCI 16,700 meters	Power off Going North
" 11	160	246		
" 19	130	200		
" 31	520	800		
" 39	65	100		
" 45	110	169.5		

TABLE XIII

Long-Wave Tests, April 17, 1928.

Position in Tunnel	Readings in Microvolts	Per Cent Signal Strength (20)	Station and Wavelength	Remarks
Chamber 1	130	100	WCI 16,700 meters	Tunnel blocked cables grounded.
" 5	32	24.6		
" 10	0	0		
" 15	1170	900		
" 20	780	600		
" 25	1800	1382		
" 29	1800	1382		
" 30	2000	1540		
" 31	2000	1540		
" 32	1800	1382		
" 35	[illegible]	424		
" 40	1170	900		
" 44	910	700		
		(21)		
Chamber 1	65	50	WQK 16,465 meters	Tunnel blocked cables grounded.
" 5	30	23.1		
" 10	65	50		
" 15	540	415		
" 20	230	177		
" 25	540	415		
" 29	780	600		
" 30	980	754		
" 31	980	754		
" 32	780	600		
" 35	260	200		
" 40	260	200		
" 44	160	123		

Discussion of Results of 1928 Tests

The results obtained in the above mentioned tests can most easily be studied by means of the curve sheets mentioned above.

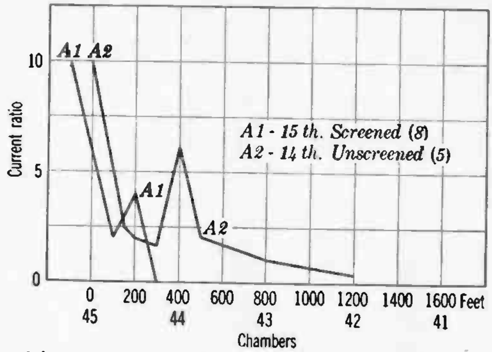

Fig. 10—Comparison of Signal Strengths.
42.5-meter Tests, April 14-15, 1928.

Fig. 9 is a preliminary comparison of the short-wave results on 43 and 55 meters. Curve A1 was taken by means of the control panel and Curve A2 by the vacuum-tube voltmeter. Atten-

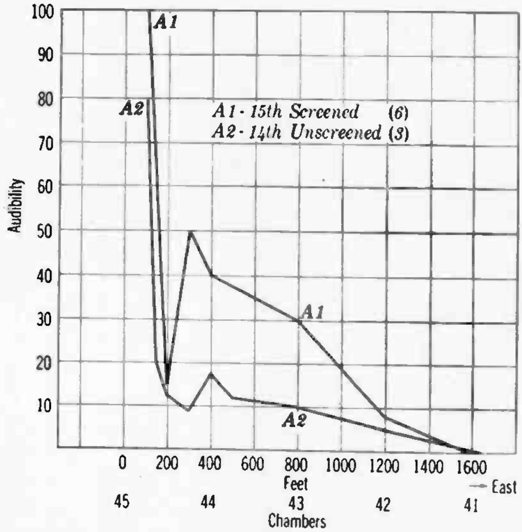

Fig. 11—Comparison of Signal Strengths.
55-meter Tests, April 14-15, 1928.

tion is called to the peculiar jump which occurred in both curves at Chamber 44, that is, 400 ft. inside the west portal. This station was immediately under the C.P.R. tracks. It is rather interesting to note that both waves act similarly and that both disappear at approximately the same point, although the signals were from stations located at widely distant points and varying greatly in power output. It might, therefore, be assumed that the attenuation on all waves below some certain frequency was approximately the same.

Figs. 10 and 11 have been drawn to show the effect on the short wave of screening the tunnel mouth by means of steel cars. In Fig. 10, it would appear from the 43-meter results that the screening had produced a decided decrease in strength. Fig. 11

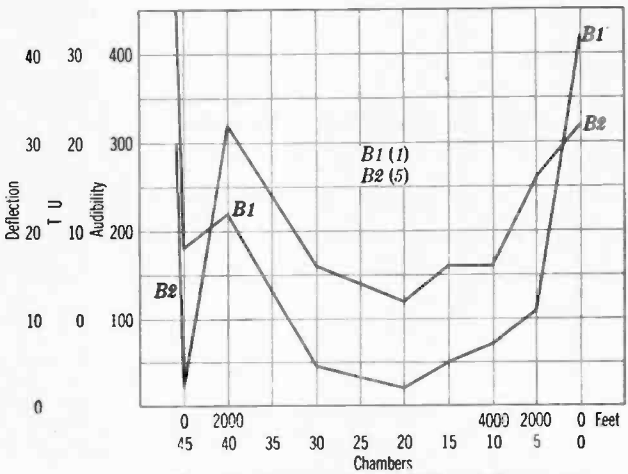

Fig. 12—Comparison of Signal Strengths. 411-meter Tests, April 14, 1928.

would seem to contradict this statement. There is at present no satisfactory explanation for this difference unless it may be accounted for by experimental error and the fact that the decrease in strength on the short wave is very rapid. The records at the transmitting station indicate that there was no appreciable change in power output during these tests.

Fig. 12, has been drawn for the broadcast band to show that while the results of different methods of measurement may vary slightly numerically, the nature of the effect is the same in all cases. Curve B1 was taken using the loop and superheterodyne and is a measurement of the actual carrier wave strength. Curve B2 was taken using the Western Electric control panel and is, therefore, more nearly a measure of the audio output of the receiving set. From Figs. 9 to 12 it may, therefore, be assumed that the general comparison of results of different methods of measurements is justifiable.

Fig. 13 is plotted for the broadcast band and brings out markedly the loss in signal strength due to blocking the tunnel and grounding the cables. The effects plotted here are the combined effects of the two actions. This was unavoidable due to the fact that it was impossible to make a 411-meter test on the 15th when only the blocking of the tunnel was being carried out. Attention is called to the sudden jump in Curve B2 at Chamber 30. This will be commented on later when dealing with the effect

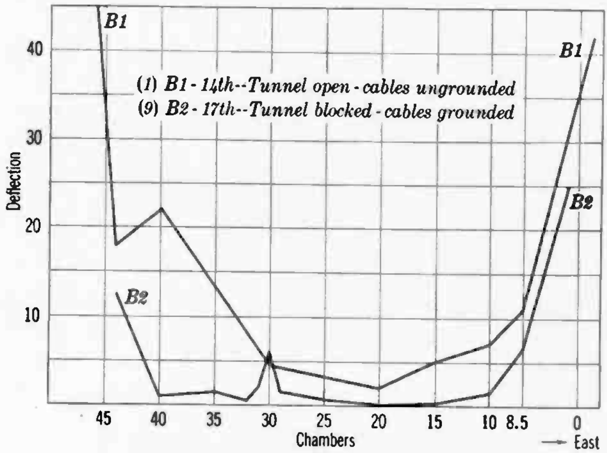

Fig. 13—Comparison of Signal Strengths. 411-meter Tests, April 14 and 17, 1928.

of the ventilation shaft at Maplewood Avenue. Figs. 13 and 14 are essentially the same, the only difference being that in Fig. 13 the actual deflection was taken as the ordinate, while in Fig. 14 the ordinate is the per cent signal strength. Curve B2 is the same in Figs. 13 and 14, but in Fig. 14 it is plotted from the audibility meter reading instead of the thermionic voltmeter reading.

An examination of the profile of the tunnel will show that the Maplewood Avenue ventilation shaft enters at about Chamber 30. This is the point at which the sudden jump occurs in the curves representing the results for the blocking of the tunnel and the grounding of the cables. It would, therefore, appear that sufficient signal comes in via this shaft to boost the curve very appreciably. This is a very interesting point as the shaft is only 30 ft. square and almost 300 ft. long. This fact considerably strengthens the theory that the signals enter the tunnel via the mouth rather than through the rock. The same effect will be noticed in the case of the long-wave results in Figs. 16, 17, and 18. In the case of the 1400-meter signals, Fig. 15, Curve C3, there would appear to be some additional action effecting the

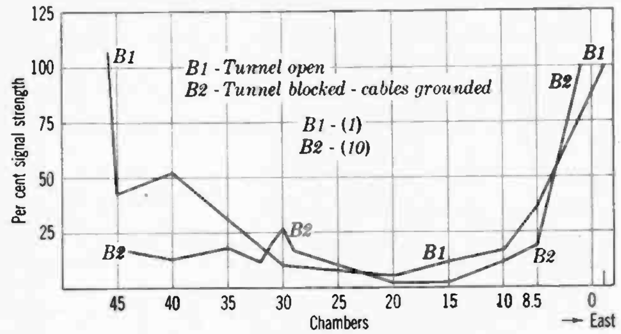

Fig. 14—Comparison of Signal Strengths. 411-meter Tests, April 14-17, 1928.

results. It is assumed that this is due to the loop action of the grounded cables and will be referred to in more detail when considering the long-wave results. It should be observed that the effect of the Maplewood shaft is only apparent when both screening and grounding are being employed. This is to be ex-

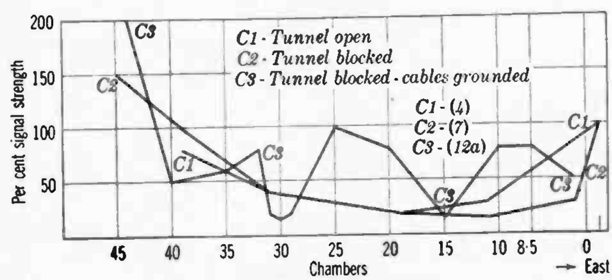

Fig. 15—Comparison of Signal Strengths. 1400-meter Tests, April 14, 15, 17, 1928.

pected as under ordinary conditions the signals enter through the main tunnel more readily than through the vertical shaft.

While studying the results of the tests, a suggestion was made that pipes, ladders, or steel work in the ventilation shaft might account for the conduction of signals down the shaft. At our request, a very careful search was made by Mr. Price, the electrical superintendent of the Mount Royal Tunnel, and he reports that there are no pipes, wires, hoist guides, or cables in the shaft, and in fact, no metallic substances of any nature of sufficient size to be worth considering.

Fig. 17 has been drawn from the long-wave results and demonstrates that the effect of the air shaft is similar in the case of the

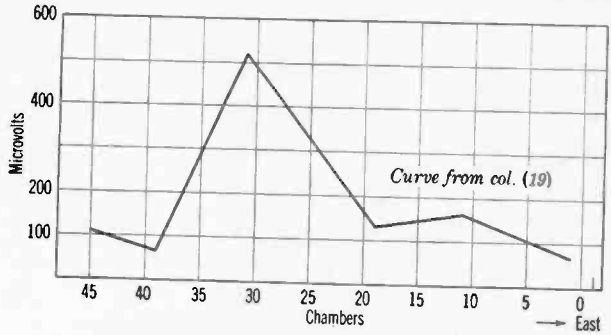

Fig. 16—WCI—17,000-meter Tests, April 16, 1928.

two long-wave stations used in the tests, although they have different output characteristics and different wavelengths. It would appear from the curves of Fig. 17 that the system of signal strength measurement used by the Marconi Company for the long-wave results was very reliable.

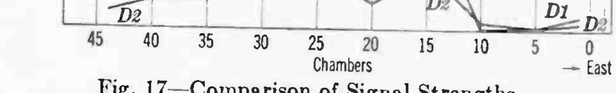

Fig. 17—Comparison of Signal Strengths.
17,000-meter Tests, April 17, 1928.

The relative effects of screening the tunnel and grounding the cables for long waves can readily be studied from Fig. 18. With the exception of Curve D3, the results are very consistent. In this curve the jump due to the air shaft is very marked. Attention is also called to the second peak at Chamber 15. This is a further evidence of the loop action to be discussed later.

In order to obtain a comparison on one sheet of the relative effects for all frequencies, Figs. 19, 20, and 21 have been drawn. Fig. 19 compares the results with the tunnel open, Fig. 20 shows the signal strengths with only the blocking in place in the tunnel mouth, while Fig. 21 shows the total effect of blocking and screening being carried out together.

It has already been mentioned that the 2400-volt power feeder was alive during part of the test. As a matter of experiment, the power was cut off on several occasions and it was discovered

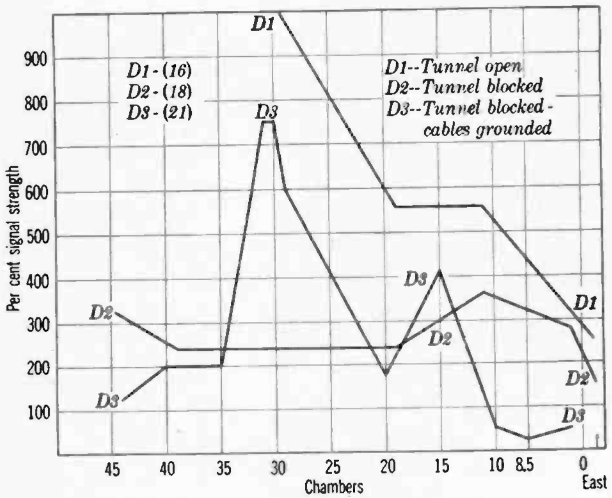

Fig. 18—Comparison of Signal Strengths. 17,000-meter Tests, April 14, 15, 17, 1928.

that there was a marked difference in signal strength with the power off, on certain frequencies. On the short waves there was practically no effect between power on and power off. On the broadcast wave the ratio of signal to noise was much better when the power was off. This, of course, was to be expected. On the 1400-meter and 17000-meter waves the signal strength actually seemed to be greater with the power off. This may be observed by comparing columns 13 and 14, 15 and 16, and 17 and 18, of Tables X and XI.

Referring again to Figs. 19, 20, and 21, with the tunnel open the per cent signal seems to be a direct function of the frequency, with the weakest signal in the case of each wavelength occurring at the center or about Chamber 20. The profile of the tunnel shows that this chamber is just about midway between the two highest points of the mountain. With the tunnel blocked the same relative effects occur but the long-wave signals reach their minimum at a different point from the 1400-meter signals. Fig. 21 has been drawn for both blocking and grounding On this sheet the 55-meter signals have been omitted for reasons already stated. The effect of the air shaft is clearly indicated. There is, however, evidence of a further effect. It will be observed that in both the 1400-meter and the 17000-meter curves, peaks occur at regular

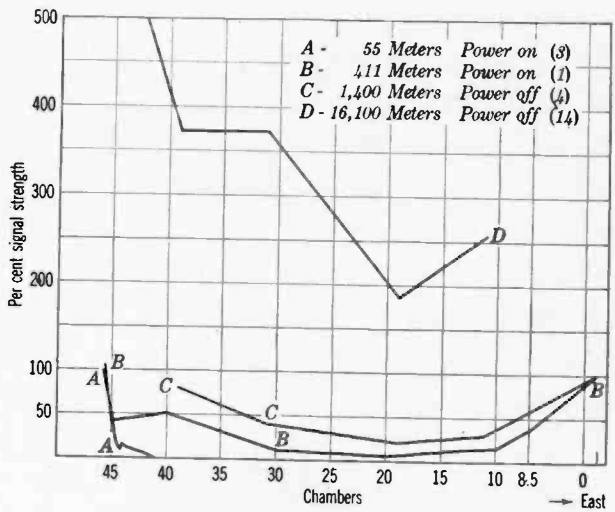

Fig. 19—Comparison of Signal Strengths, April 14, 1928. Tunnel Open.

intervals. These are not observable when the cables and trolley wires are ungrounded. It will be remembered that the grounding was carried out at Chambers 5 and 43, or approximately at the two ends of the tunnel. It would appear as if the grounding of the overhead trolley and cables had turned the system into a huge loop. It is rather difficult to calculate the characteristics of this loop on account of the nature of the construction. The messenger cable supporting the trolley wires is itself grounded. This brings the actual ground connection into close proximity with the overhead wires. There is, however, a 250,000 circular mil cable used to ground the tracks and this would, of course, constitute the other side of the loop. Whatever the characteristics of such a loop, it is quite conceivable that in a loop of this length, (approximately 3 miles) nodes and loops would occur for the different frequencies. From the curves it would appear that there were about twice as many loops for the 1400-meter signals as for the 17000-meter signals.

In connection with the long-wave signals it will be immediately evident that there is a very great difference between the results in the Dorchester station and those in the vicinity of the west portal. This would hardly be expected when it is considered that the Dorchester station is practically in the open at the east end of the tunnel, whereas the signals pass completely over or through the mountain to reach the vicinity of the west

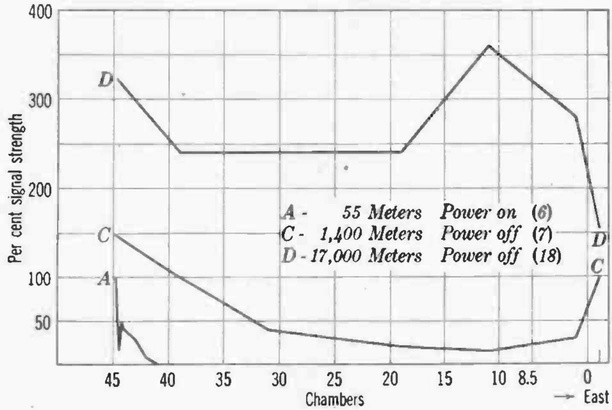

Fig. 20—Comparison of Signal Strengths, April 15, 1928. Tunnel Blocked.

portal. It is suggested that the action indicated in Tables X to XIII, and in the corresponding curve sheets, Figs. 18 to 21, is due to a wave-antenna effect. In other words, the overhead trolley wires which are open at the east end are acting as a Beveridge antenna with the east or open end pointing in the general direction of the transmitting station. If this assumption is correct the maximum effect will then occur when the receiving set is located at the west end of the feeder. It will, therefore, be seen that the results obtained fit in with the suggested theory. This effect would not be so noticeable on the 1400-meter signals since the transmitting station is located in a westerly direction from Montreal.

In this discussion little has been said by way of explanation of the peculiar jumps occurring in the curves in the vicinity of

Chambers 40 to 45. It is quite likely that these effects are in some measure due to reradiation from the numerous telegraph, telephone, and power wires which run along the C.P.R. tracks above this section of the tunnel. This would not explain all the effects observed, but must be given some consideration as the various lines extend for considerable distances across country.

Attention is called to a further fact observed during the broadcast reception. It will be remembered that a loop was used for these signals and it was found that upon entering the tunnel the loop had to be turned parallel with the axis of the tunnel in

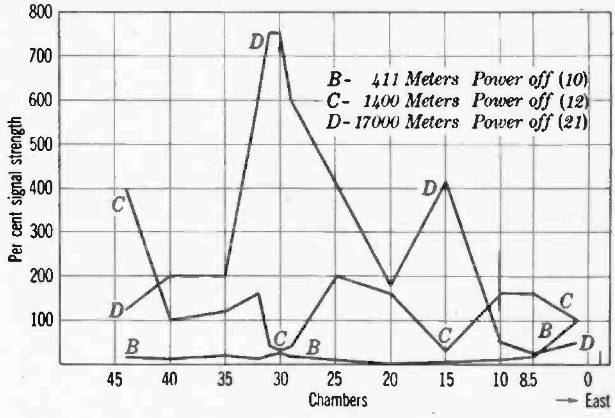

Fig. 21—Comparison of Signal Strengths, April 17, 1928. Tunnel Blocked, Cables Grounded.

order to obtain the maximum signal strength. Whether this indicates that the transmission was coming along the wires or was merely being confined to a general path by the walls of the tunnel, is difficult to state with any certainty.

Conclusion

The results considered above are very far from being conclusive, but time did not permit of further tests being made. From a consideration of the curves already discussed, it would seem that the following general conclusions might be drawn,

(1) The penetration of any wave into the tunnel is some definite function of the frequency of the wave.

(2) Waves below 100 meters do not penetrate rock or soil to any appreciable extent.

(3) Cables and rails do not conduct the shorter waves and broadcast waves to the same extent that they do long waves, although their effect on the broadcast band is appreciable.
(4) Wires and cables when ungrounded appear to act as wave antennas. When grounded they act as loops depending, of course, on the methods employed in carrying out the grounding.
(5) Much more energy appears to enter via the tunnel mouth than was at first suspected. This is indicated by the effect of the air shaft already discussed.

Dr. A. S. Eve, one of the co-workers in these experiments, has made the following comment,

> The most noteworthy point in the experiments is the fact that the short-wave signals (40 to 55 meters) did not penetrate well or far into the tunnel through any of the three possible channels:
>
> (1) through rock
> (2) through mouths and shaft
> (3) along conductors.
>
> The longer waves probably entered by all three methods, and Major Steel, who has prepared this report, is disposed from the evidence to attach considerable weight to the admission by air. If he is correct in this view there are interesting questions which arise connected with the nature of radio waves. Can they enter horizontal tunnels and also descend vertical shafts? Nor is it clear why the short waves should fail to go far into the mouth, while the long waves seem to penetrate the whole length of the tunnel.

Suggestions for Further Work

It would seem desirable to carry out further tests at some point where a tunnel exists without wires and rails leading into it. This may be difficult or, perhaps, impossible to carry out, but further experiments in the Mount Royal Tunnel would hardly seem advisable in view of the tremendous expense involved in completely screening or blocking the openings.

In future tests arrangements should be made to measure actual field strengths in microvolts per meter, some standard form of antenna being adopted for each particular wave. For the waves below 17000 meters it seems advisable to use a well constructed loop antenna.

In conclusion the writers wish to express their appreciation for the very great assistance given them by the President, Sir

Henry Thornton, and by all the officials of the Canadian National Railways System with whom we came in contact. We particularly wish to mention the following: Mr. R. C. Johnson, Superintendent, Canadian National Railways; Mr. C. P. Price, Electrical Superintendent, Canadian National Railways; Mr. W. Walker, District Engineer, Canadian National Railways.

Dr. D. A. Keys and Dr. E. S. Bieler, of the Physics Department, McGill University, have been particularly helpful both in the preliminary tests and in the discussions on the results obtained. Great credit is also due to Mr. Rushbrook of the Canadian Marconi Company for his assistance in making the tests, and to the officials of the Department of National Defence for permission to use much of the apparatus required for the experiments.

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A NOTE ON THE DIRECTIONAL OBSERVATIONS ON GRINDERS IN JAPAN*

By

Eitaro Yokoyama and Tomozo Nakai,

(Electrotechnical Laboratory, Ministry of Communications, Tokyo, Japan)

BEFORE the General Assembly of the U. R. S. I., 1927, a paper entitled "The Directional Observations on Atmospherics in Japan" was presented by the authors, and the following conclusions had been arrived at by the series of observations.

(1) No appreciable change had been found in the observed directions of atmospherics by the variation of wavelengths which lie between 10,000 and 20,000 meters.

(2) On the diurnal observations from 7 a.m. to 9 p.m. no appreciable regular change had been noticeable in the observed directions of atmospherics by the time of day. Though some expected variations had, in general, been noticed at sunrise, sunset, and night times, they had not been so remarkable as in the case of signal observations.

(3) The directions of clicks had been obviously different from those of grinders, which seems to indicate that the seat of clicks differs from those of grinders.

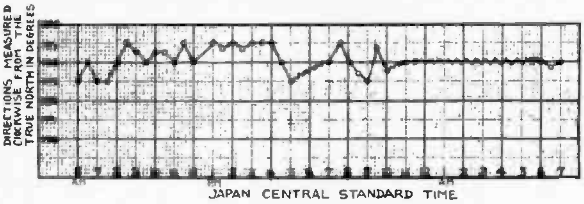

Fig. 1—February 17–18, 1928.
$\lambda = 10{,}000$ m.

(4) The diurnal variation had been much greater in the directions of clicks than in those of grinders. This seems to indicate that the seat of clicks lies nearer to the observed region than those of grinders, if the seats of both kinds of atmospherics might be assumed to have nearly equal areas. Judged from

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the great differences in the intensities, as well as in the nature of tones of clicks and grinders, this assumption may be considered to be probable.

(5) The directions of clicks had shifted from nearly South toward West as the season advanced from winter to summer, while those of grinders had remained unchanged.

(6) In summer, the clicks, which seem to originate from the mountain regions of Japan, seem to be most prevalent, while grinders probably come always from a tropical region such as the Dutch East-Indies.

Those were the results obtained from the observations over a period from February to July, 1927. The experiments were further continued and it was recognized that the same conclusions still hold good except in the night observations of grinders.

During the experiments, ten times of 24-hour observations were made covering all the seasons of the year. Though no other particular phenomena were noticed in the observations on clicks, it was remarkably found in those on grinders, on the contrary, that the directive property became indistinctive or disappeared at night. One example of directional variations (wavelength 10,000 meters) is shown in Fig. 1 in which the zig-zag lines indicate the durations of directivity fading, and in Fig. 2 which illustrates the change in the broadness of directivity in the same observation as shown in Fig. 1. In other 24-hour observations, the general tendencies of the variations were quite similar, though the times at both extremities of the fading bands of directivity shifted by a few hours according to the dates.

It is in general believed that the origins of long-wave atmospherics, as received in both England and the United States, are mainly in the lands situated in the Tropical Zone. [1, 2, 3] The grinders as received in Japan had also been indicated by the authors to be of a tropical origin such as the Dutch East-Indies. Taking the above into consideration and the daily movement of the illuminated and the darkened hemispheres into account the atmospherics originating at the tropical region in America seem very probable to be received in Japan as grinders early

[1] Lloyd Espenschied, C. N. Anderson, and Austin Bailey, "Transatlantic Radio Telephone Transmission," Proc. I.R.E., **14**, 7; February, 1926.

[2] R. A. Watson Watt, "The Directional Recording of Atmospherics," *Jour.* I.E.E., **64**, p. 596; May, 1926.

[3] L. W. Austin, "Direction Determinations of Atmospheric Disturbances on the Isthmus of Panama," Proc. I.R.E., **14**, 373; June, 1926.

at night and those in Africa late at night, while those in the Dutch East-Indies may be received for all the time of day and night. Thus the grinders coming from other parts than the Dutch East-Indies seem to have made them apparently non-directive at night, because the unidirectional combination of antennas was

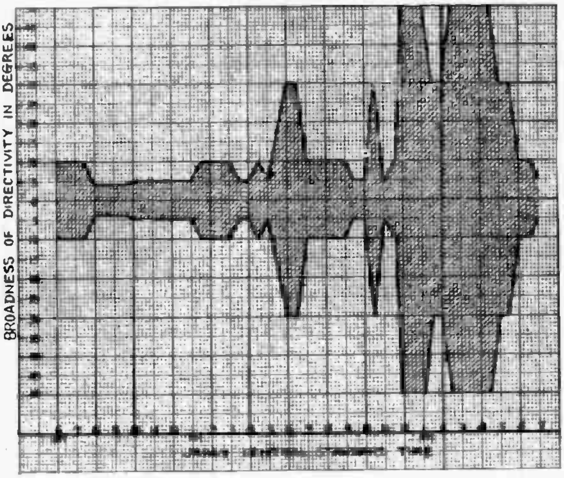

Fig. 2—February 17–18, 1928.
$\lambda = 10{,}000$ m.

not tried, a single loop being used as wave picking-up device in the observations. The observations are now still in progress, and it is expected that the phenomenon will be brought into light more definitely later on.

Acknowledgment is due to C. Asakawa and T. Takei of this aboratory for their assistance during the observations.

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ON THE BEHAVIOR OF NETWORKS WITH "NORMALIZED" MESHES*

By

E. A. Guillemin and W. Glendinning

(Department of Electrical Engineering, Massachusetts Institute of Technology, Cambridge, Massachusetts)

Summary—*The theory of normalizing meshes in electrical networks, as outlined in a recent article appearing in this journal,[1] is verified and illustrated by examples and figures relating to two- and three-mesh circuits. In the two-mesh circuit mesh No. 1 was normalized, thus confining the corresponding frequency to that mesh both for the transient and steady states as illustrated by Figs. 2 to 6 inclusive. For the three-mesh circuit meshes Nos. 1 and 3 were normalized, and the results are illustrated by Figs. 6 to 16 inclusive. In each case the theory was checked in every detail.*

In a recent issue of this journal[1] the mathematical theory concerning the interpretation of normal co-ordinates in dynamic systems as related to the electrical network was set forth in its more fundamental character, the stress being laid upon the analogy between a mechanical system of coupled mass points and its electrical network equivalent. At that time the interest lay chiefly in a deeper inquiry into this analogy which is so complete from the mathematical standpoint and yet so intangible from the physical angle.

The reason why such an investigation should present an unusual element of interest is due to the mechanical viewpoint rather than the electrical. For, in the mechanical system, the existence of so-called normal co-ordinates can easily be visualized, and these are known to possess distinct peculiarities; namely, of containing only one of the natural frequencies of the system. The projection of the oscillation of any mass point upon this particular direction in space will be simple harmonic, damped or undamped depending upon the presence or absence of frictional forces. This particular natural frequency is *isolated* in its corresponding normal co-ordinate in that it cannot be fed over into any other direction which is normal to it, or be brought into action unless the actuating force has a component in the direction of this normal co-ordinate.

* Dewey decimal classification: 621.319.2 Original manuscript received by the Institute, October 10, 1928.

[1] E. A. Guillemin. "Making Normal Co-ordinates Coincide with the Meshes of an Electrical Network," Proc. I. R. E., 15, 935; November, 1927.

In pursuing the analogy between the mechanical and electrical systems it was soon recognized that the counterpart of a physical dimension or direction (co-ordinate) is a mesh in the equivalent electrical system. The question then immediately arose as to whether it would be possible in the electrical system to make a mesh correspond to what we recognize mathematically as being a normal co-ordinate, just as it is possible in the mechanical problem to orient our arbitrarily chosen reference axes so that

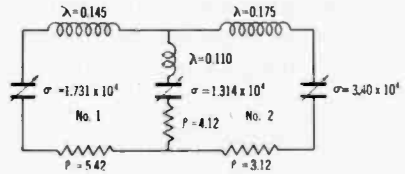

Fig. 1—Arrangement of Constants Used in Two-Mesh Circuit.

one of them will coincide with a normal direction. For, granting that this could be done, we should thus gain a means of isolating or confining one of the natural frequencies of the electrical network to that mesh. We should thus obtain a singular network which would show a distinctly different behavior, both in the

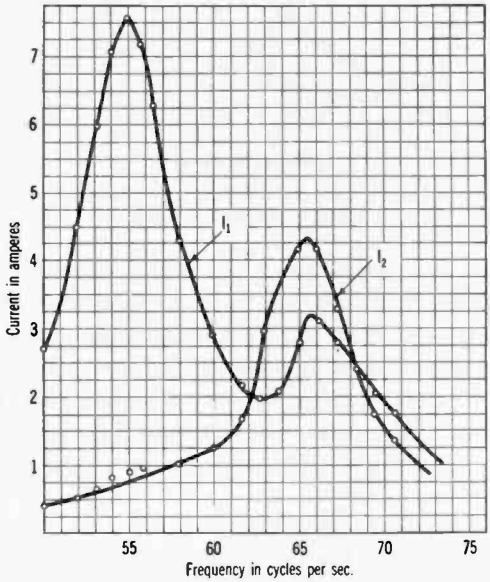

Fig. 2—Two-Mesh Circuit. Voltage in Mesh No. 1, Voltage Held Constant at 50.

transient and steady states, from that of the ordinary network having the same number of natural frequencies. Certain transient frequencies would be absent in some of the meshes, or would not appear at all, depending upon the location of the disturbing force. Furthermore, since resonance phenomena depend upon coincidence between natural and impressed frequencies, the network would exhibit isolation properties with regard to resonance peaks in the steady state.

These and similar considerations made the investigation seem worth while from the practical standpoint, even though the problem was basically purely of mathematical interest.

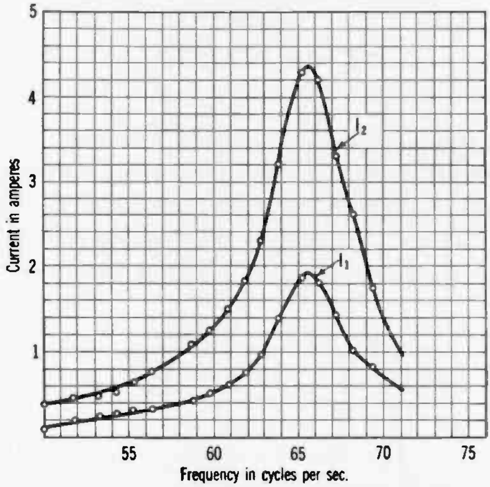

Fig. 3—Two-Mesh Circuit. Voltage in Mesh No. 2, Voltage Held Constant at 22.5.

It was for the purpose of providing concrete illustrations that an experimental investigation, of which the present paper is a report, was undertaken; and it is the hope of the writers that the following may prove interesting to those who have occasion to work with more intricate networks both theoretically and experimentally. For the fundamental theory as well as the notation used here we refer the reader to the paper already mentioned.

The first and simplest illustration will be given by the two-mesh circuit shown in Fig. 1. Each mesh contains independent inductance, resistance, and capacitance, and the two meshes are coupled by means of these three elements. Actually the common inductance shown in the center branch was a simple self-inductance of the air-core type, but it should be recognized that this might equally as well have been partly or wholly in the form of mutual inductance between two windings in proximity to each other, since mutual and self-inductance play exactly the same role both physically and mathematically.

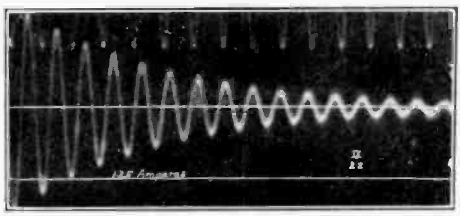

Fig. 4—Oscillogram for Two-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 1. Mesh 1 having been made a normal coordinate, the curve should be a damped double frequency. The beat shows this clearly. 60-cycle timing wave.

In this circuit the following constants were used:

$$\lambda_{11}=0.255 \text{ henries}$$
$$\sigma_{11}=3.045\times 10^4 \text{ darafs}$$
$$\rho_{11}=9.54 \text{ ohms}$$

$$-\lambda_{12}=0.110 \text{ henries}$$
$$-\sigma_{12}=1.314\times 10^4 \text{ darafs}$$
$$-\rho_{12}=4.12 \text{ ohms}$$

$$\lambda_{22}=0.285 \text{ henries}$$
$$\sigma_{22}=4.714\times 10^4 \text{ darafs}$$
$$\rho_{22}=7.24 \text{ ohms}$$

The system of differential equations for the case of no impressed forces becomes:

$$\lambda_{11}\frac{d^2x_1}{dt^2}+\rho_{11}\frac{dx_1}{dt}+\sigma_{11}x_1+\lambda_{12}\frac{d^2x_2}{dt^2}+\rho_{12}\frac{dx_2}{dt}+\sigma_{12}x_2=0$$

$$\lambda_{12}\frac{d^2x_1}{dt^2}+\rho_{12}\frac{dx_1}{dt}+\sigma_{12}x_1+\lambda_{22}\frac{d^2x_2}{dt^2}+\rho_{22}\frac{dx_2}{dt}+\sigma_{22}x_2=0$$

where:

$$x_1=\int i_1dt \quad \text{and} \quad x_2=\int i_2dt.$$

Assuming the solutions:

$$x_1 = Y_1 e^{pt} \quad \text{and} \quad x_2 = Y_2 e^{pt},$$

we get:

$$b_{11}Y_1 + b_{12}Y_2 = 0$$
$$b_{21}Y_1 + b_{22}Y_2 = 0$$

with:

$$b_{11} = 0.255p^2 + 9.54p + 3.045 \times 10^4$$
$$b_{12} = b_{21} = -0.110p^2 - 4.12p - 1.314 \times 10^4$$
$$b_{22} = 0.285p^2 + 7.24p + 4.714 \times 10^4.$$

From inspection we see that:

$$\frac{b_{12}}{b_{11}} = \alpha = -0.432.$$

The value of α is insignificant. The fact that such a relation exists is the necessary and sufficient condition for making mesh No. 1

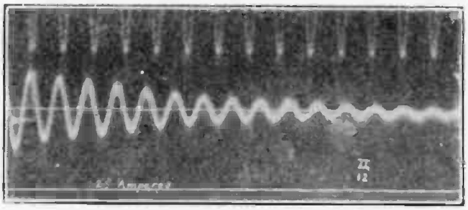

Fig. 5—Oscillogram for Two-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 2. Note that this is a damped single frequency. 60-cycle timing wave.

a normal co-ordinate. This is clear from the fact that the determinantal equation for this case becomes:

$$D = \begin{vmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{vmatrix} = 0$$

$$= \begin{vmatrix} b_{11} & \alpha b_{11} \\ \alpha b_{11} & b_{22} \end{vmatrix} = b_{11}b_{22} - \alpha^2 b_{11} = 0$$

or:

$$D = b_{11}(b_{22} - \alpha^2 b_{11}) = 0,$$

from which:

$$b_{11} = 0; \quad \text{or} \quad (b_{22} - \alpha^2 b_{11}) = 0.$$

The first of these determines that frequency which is confined to mesh No. 1. We shall call it the *normalized frequency*. The second of the above equations gives us that frequency which is not confined but still common to both meshes. We shall call it the *system frequency*.

Substituting numerical values, we get for the equation which determines the normalized frequency:

$$0.255p^2+9.54p+3.045\times10^4=0$$

Since the circuit is highly oscillatory, the frequency is given very nearly by neglecting the resistance term (linear term in p). We have:

$$0.255p^2+3.045\times10^4=0$$

or:

$$p=\pm j345=\pm j2\pi f,$$

so that the normalized frequency becomes:

$$f=55 \text{ cycles per second.}$$

In exactly the same way the system frequency is found to be 66.3 cycles per second. It should be borne in mind that although the above procedure of neglecting the resistance term is sufficiently accurate for the determination of the frequencies and a consequent discussion of steady-state phenomena, the damping constant which depends directly upon the resistance term becomes of vital importance in the consideration of transient phenomena. Since these facts are, however, well known in circuit theory, we shall not dwell upon them here.

In the actual experiment, the desired frequencies were chosen first and the constants of the circuit as well as the value of α adjusted so as to meet the available values of inductance and capacitance in the laboratory. Hence the odd figures given above. This method of reversing the calculations, which is essential in every problem of design, is given in a thesis[2] report filed in the library of the Massachusetts Institute of Technology and will not be reproduced here. We prefer rather to show the experimental results and give a discussion of them in the light of the theory upon which they are based.

Let us review the situation once more and see what results we should expect to obtain. We have made mesh No. 1 a normal co-ordinate to the frequency of 55 cycles per second. The remaining system frequency is 66.3 cycles per second. This

[2] "Experimental Verification of Theories Concerning Normalized Meshes in Electrical Networks," M. I. T. thesis, 1928, by W. Glendinning.

means that the frequency 55 is confined to mesh No. 1, but that the frequency 66.3 is still common to both meshes. If we impress an alternator in mesh No. 1 and vary the frequency over a range covering the above values, then the current in mesh No. 1 will show a peak for 55 cycles and another for 66.3 cycles. The current in mesh No. 2 will show a peak only for 66.3 cycles. It cannot show a peak for 55 cycles because this frequency is no longer a natural frequency to that mesh but is confined entirely to mesh

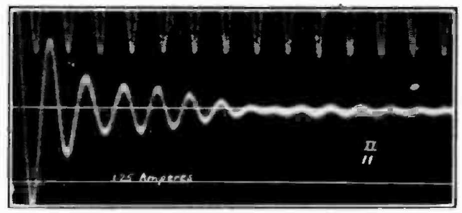

Fig. 6—Oscillogram for Two-Mesh Circuit. Voltage in Mesh 2, Current in Mesh 2. Note that this is a damped single frequency. 60-cycle timing wave.

No. 1. On the other hand, if we impress an alternator in mesh No. 2 and vary the frequency over the same range, we shall find that the currents in both meshes show peaks only at the system frequency of 66.3 cycles. The resonance peak corresponding to the normalized frequency of 55 cycles is suppressed even in the normal mesh (mesh No. 1)! This rather striking phenomenon is due to the fact already mentioned that the normalized frequency is so completely isolated within the corresponding normal mesh that it cannot be brought into action unless the actuating force is impressed in that mesh.

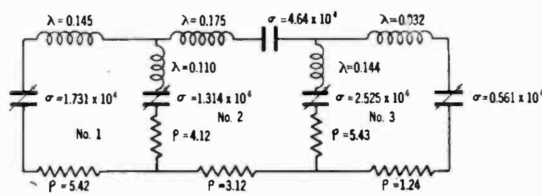

Fig. 7—Arrangement of Constants for Three-Mesh Circuit.

As to the transients that we have to expect we can predict that if we suddenly impress a constant emf in either mesh and take an oscillogram of the resulting current in the *other* mesh, then we should find that the result is a *simple damped harmonic* of system frequency. The same is true of the current in the non-normal mesh (mesh No. 2) with the emf impressed in No.2. However, if we examine the current in mesh No. 1 with the emf impressed there, then we shall find a superposition of two damped frequencies—the normalized and the system frequencies.

Figs. 2 and 3 show the experimental results of the steady-state conditions, and bear out the theory in every detail.

For the transient results given by the oscillogram Figs. 4, 5, and 6, the roman II designates the two-mesh circuit, while the 11, 12, and 22 indicate that current is measured in mesh No. 1 with voltage in No. 1, or current measured in No. 1 with voltage

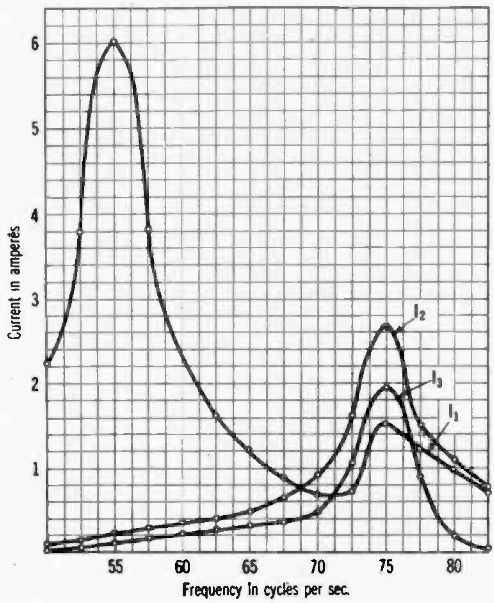

Fig. 8—Three-Mesh Circuit. Voltage in Mesh No. 1, Voltage Held Constant at 40.

in No. 2 or vice versa, or current measured in No. 2 with voltage in No. 2. With this in mind the figures need no further explanatory note. The same system of notation will be extended to the three-mesh circuit to be discussed later.

Close inspection of Figs. 4, 5, and 6 will show that the one designated 11 is the only one which contains two frequencies superimposed as may be detected from the beat. The other two are simple damped sinusoids.

A second study was made, using the three-mesh circuit shown in Fig. 7 with the constants so chosen that meshes Nos. 1 and 3 become normal co-ordinates. The values of the constants were fixed as follows:

$$\lambda_{11} = 0.255 \text{ henries}$$
$$\rho_{11} = 9.54 \text{ ohms}$$
$$\sigma_{11} = 3.045 \times 10^4 \text{ darafs}$$

$$-\lambda_{12} = 0.110 \text{ henries}$$
$$-\rho_{12} = 4.12 \text{ ohms}$$
$$-\sigma_{12} = 1.314 \times 10^4 \text{ darafs}$$

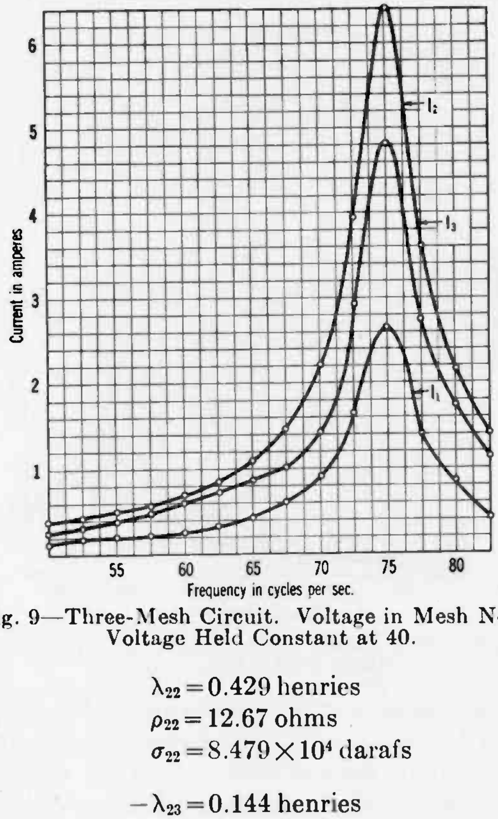

Fig. 9—Three-Mesh Circuit. Voltage in Mesh No. 2, Voltage Held Constant at 40.

$$\lambda_{22} = 0.429 \text{ henries}$$
$$\rho_{22} = 12.67 \text{ ohms}$$
$$\sigma_{22} = 8.479 \times 10^4 \text{ darafs}$$

$$-\lambda_{23} = 0.144 \text{ henries}$$
$$-\rho_{23} = 5.43 \text{ ohms}$$
$$-\sigma_{23} = 2.525 \times 10^4 \text{ darafs}$$

$$\lambda_{33} = 0.176 \text{ henries}$$
$$\rho_{33} = 6.67 \text{ ohms}$$
$$\sigma_{33} = 3.086 \times 10^4 \text{ darafs}$$

Thus we obtained for the three-mesh circuit:

$$b_{11}=0.255p^2+9.54p+3.045\times10^4$$
$$b_{12}=-0.110p^2-4.12p-1.314\times10^4$$
$$b_{22}=0.429p^2+12.67p+8.479\times10^4$$
$$b_{23}=-0.144p^2-5.43p-2.525\times10^4$$
$$b_{33}=0.176p^2+6.67p+3.086\times10^4.$$

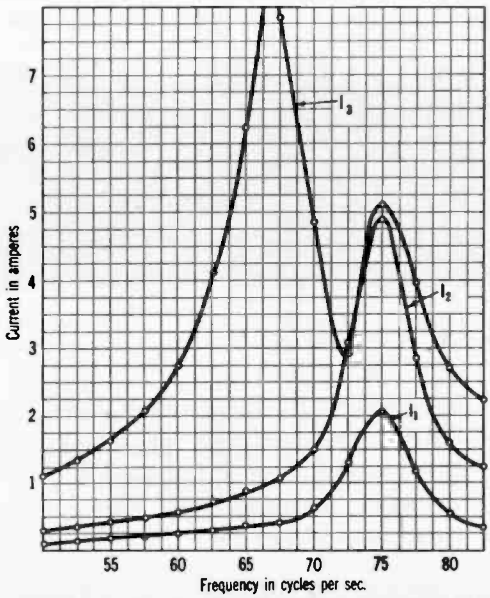

Fig. 10—Three-Mesh Circuit. Voltage in Mesh No. 3, Voltage Held Constant at 40.

Here we see that:

$$\frac{b_{12}}{b_{11}}=\alpha=-0.432; \quad \text{and} \quad \frac{b_{23}}{b_{33}}=\beta=-0.818.$$

Again the numerical values of α and β are not significant. The fact that the same ratio exists between inductance, resistance, and capacitance, is the necessary condition for making a mesh a normal co-ordinate. Here we have made meshes Nos. 1 and 3 normal co-ordinates.

The determinantal equation for this case becomes:

$$D(p)=\begin{vmatrix} b_{11} & \alpha b_{11} & 0 \\ \alpha b_{11} & b_{22} & \beta b_{33} \\ 0 & \beta b_{33} & b_{33} \end{vmatrix}=0$$

or:

$$b_{11}b_{33}(b_{22}-\beta^2 b_{33}-\alpha^2 b_{11})=0,$$

which is factorable into:

$$b_{11}=0; \quad b_{33}=0; \quad (b_{22}-\beta^2 b_{33}-\alpha^2 b_{11})=0.$$

The fact that b_{11} and b_{33} can be factored out in this way shows that meshes Nos. 1 and 3 are normal co-ordinates. The corresponding normal frequencies are given by the roots of the first two of the above equations. Incidentally this process of making

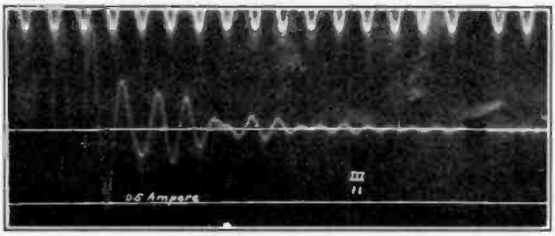

Fig. 11—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 1. Note that this is a damped double frequency as can be seen from the beat. 60-cycle timing wave.

meshes correspond to normal co-ordinates makes it easy to solve for the roots or natural frequencies. Ordinarily the three-mesh circuit discussed here would lead to a sixth-degree equation with three pairs of conjugate complex roots. This is true in our case also, but the equation is factorable into three quadratics which

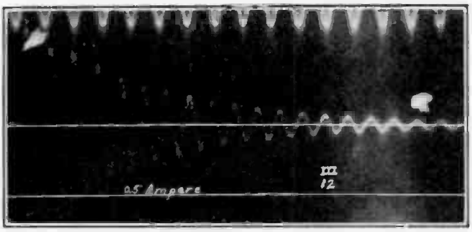

Fig 12—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 2. Note that this is a damped single frequency. The normal frequency of mesh 1 is suppressed and does not appear in mesh 2. 60-cycle timing wave.

may be solved by the simple formula. The investigation of networks with normalized meshes thus becomes much simpler than it is for ordinary networks. Also the evaluation of integration constants or transient current amplitudes by Heaviside's formula becomes very simple; but we cannot go into that here.

We find by solving the three quadratics given above that the normalized frequency belonging to mesh No. 1 is 55 cycles per second, that belonging to mesh No. 3 is 66.7 cycles per second, and that the remaining frequency is 75 cycles.

Here, as in the case of the two-mesh circuit, the values of frequency were chosen to start with and the constants as well as values of α and β adjusted to meet the available values. For this process of adjustment or "design calculation" we again refer to the thesis already mentioned above.

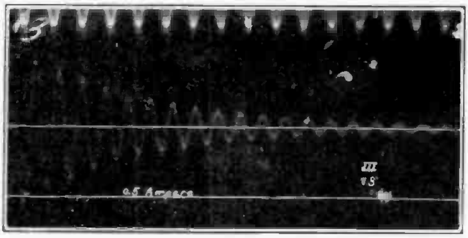

Fig. 13—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 3. Note that this is a damped single frequency. The normal frequency of mesh 1 is confined to that mesh. 60-cycle timing wave.

Let us now see what we should expect in the way of transient and steady-state results for this three-mesh network with meshes Nos. 1 and 3 normalized. Suppose we impress an alternator in mesh No. 1 and—keeping the voltage constant—vary the frequency so as to cover from 55 to 75 cycles per second. We should

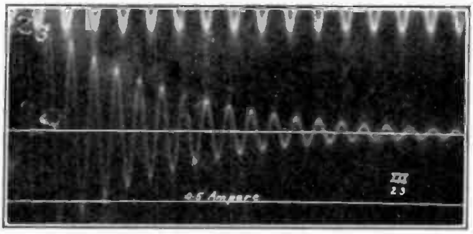

Fig. 14—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 2, Current in Mesh 2. Note that this is a damped single frequency. Mesh 2 is a non-normal mesh and only the system frequency will appear. The normal frequencies of meshes 1 and 2 do not come into action. 60-cycle timing wave.

expect the current in mesh No. 1 to show a peak at 55 cycles—its normal frequency—and another peak at 75 cycles—the system frequency. No peak should occur at 66.7 cycles because this is the frequency which is isolated in mesh No. 3. For the current in mesh No. 2 we should expect only one peak at 75 cycles, and for the current in mesh No. 3 we should expect only *one* peak also at 75 cycles. We do *not* get a peak for 66.7 cycles—the normalized frequency for mesh No. 3—because the driving force is not in that mesh but is located in mesh No. 1. This is a very interesting fact; namely, that a response to a normalized frequency can be brought about only by applying the emf in that mesh to which the normalized frequency in question is confined. With the generator in mesh No. 1, as we have here assumed, meshes Nos. 2 and 3 have only one resonance frequency. These facts were borne out very nicely in our experiments, and are shown in Fig. 8.

Suppose now we consider the generator in mesh No. 2 and again vary the frequency as before, keeping the voltage constant. In this case the currents in all three meshes should show only *one* resonance peak at 75 cycles—the system frequency. The

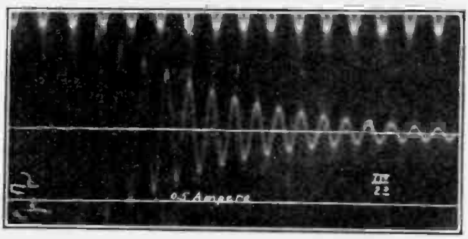

Fig. 15—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 2, Current in Mesh 3. Note that this is a damped single frequency. 60-cycle timing wave.

normalized frequencies of 55 and 66.7 cycles in meshes Nos. 1 and 3 do not respond, and this is again clearly shown by our experimental curves of Fig. 9 for this case.

Finally, suppose we impress the voltage in mesh No. 3 and make the same run as above. We expect in meshes Nos. 1 and 2 a response only to the system frequency of 75 cycles because the normalized frequency of 55 cycles in mesh No. 1 does not come into action. In mesh No. 3, however, we now expect to find a response to 66.7 cycles—its normalized frequency—and also to the system frequency of 75 cycles. These conditions were also verified experimentally and are illustrated by Fig. 10.

Now as to the type of transients we should expect to get. Here, as in the two-mesh case, we use the numbering 12, 13, etc., to designate runs, whereby the first figure designates the mesh in which the current is observed and the second, the mesh in which the constant emf is suddenly applied, or vice versa. Thus the runs 12, 13, 22, and 23 should show simple damped sinusoids of system frequency. The run 11, on the other hand, should show a double-frequency transient containing 55 and 75 cycles, and the run 33 should also show a double-frequency transient containing 66.7 and 75 cycles. The oscillogram for run 11 is given in Fig. 11. Those for runs 12, 13, 22, and 23 are given in Figs. 12, 13, 14, and 15, respectively, while the oscillogram for run 33 is given in Fig. 16. The double frequency of runs 11 and 33 is detected in each case by the beat, which is absent in the oscillograms for the remaining runs.

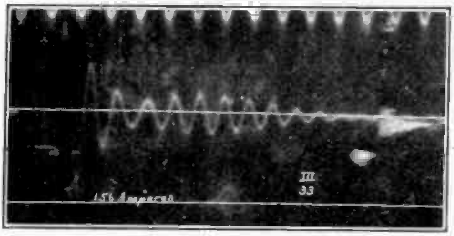

Fig. 16—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 3, Current in Mesh 3. Mesh 3 having been made a normal coordinate, the curve is a damped double frequency. The beat shows this clearly. 60-cycle timing wave.

We see that our mathematical theory is substantiated very nicely by these simple experiments, and it is safe to assume that the theory holds in its general form and applies to any type of network no matter how complicated. It seems reasonable that these properties of isolation and suppression of resonance effects should have some practical application in the design of circuits with special characteristics, and it is the hope of the authors that the principles involved have been made sufficiently clear by these illustrations so that more general interest will be aroused. In the latter event we earnestly invite criticism and co-operation.

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE*

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. Radio Communication

R007.1 Dellinger, J. H. Analysis of broadcasting station allocation. Proc. I. R. E., 16, 1477-85; November, 1928.

(A discussion of the technical aspects of the reallocation of November 11, 1928.)

R100. Radio Principles

R113 Diagramme des champs électriques mésures a Meudon pendant le deuxieme semestre 1927. (Diagrams of electric fields measured at Meudon during the second half of 1927). *L'Onde Electrique*, 7, 458-460; October, 1928.

(Curves taken on LY, UA, WSS, and GBL during second half of 1927.)

R113.1 Colwell, R. C. Fading curves along a meridian. Proc. I. R. E., 16, 1570-1573; November, 1928.

(Fluctuations in signal strength of KDKA, Pittsburgh, Pa. were observed through the sunset period of Morgantown, W. Va. Observations made covered twenty-one days. On bright clear days, the curve fluctuated considerably while on cloudy days the curve was fairly steady.)

R113.1 Merritt, E. and Bostwick, W. E. A visual method of observing the influence of atmospheric conditions on radio reception. *Proc.* Nat. Acad. Sci., 14, 884-88; November, 1928.

(A method is described which utilizes the cross-coil system and a cathode-ray oscillograph. It was possible to notice visually several successive rotations of the plane of polarization of the down-coming wave during sunset period.)

R113.4 Størmer, C. Short-wave echoes and the Aurora Borealis (letter). *Nature* (London), 122, 681; Nov. 3, 1928.

(Report on signals and echoes received at Bygdø, Oslo from the short-wave station at Eindhoven, Holland. The echoes arrived from 3 to 15 seconds later than the principal signal. The writer explains these belated echoes by the theory that radio waves penetrate the Heaviside layer passing then into empty pockets of large dimensions. The pockets are surrounded by walls of electrons from which the waves are reflected.)

* Original manuscript received by the Institute, December 14, 1928.

R113.4 Radio echoes and magnetic storms (letters). *Nature* (London), 122, 768; November 17, 1928.

(Two notes, one by S. Chapman and the other by E. T. Eckersley with respect to the note (*Nature*, p. 681, Nov. 3, 1928) by Størmer on the explanation for the long time interval echoes. Eckersley calls attention to his paper in the *Philosophical Magazine*, June, 1925, where he explained the whistles of lowering pitch by means of dispersion and the group velocity along the path in the Heaviside layer. Chapman calls attention to the fact that the positive ions are also present in addition to the electrons which were considered in Størmer's theory.)

R113.4 Beatty, R. T. Short wave signals which travel over a million miles and return. *Wireless World and Radio Review*, **23**, 722-23; Nov. 28, 1928.

(A picturization of electron vorteces of extremely large dimensions and their possible effects on radio transmission explaining long time interval echoes.)

R113.4 Schelleng, J. C. Note on the determination of the ionization in the upper atmosphere. PROC. I. R. E., 16, 1471-1476; November, 1928.

(Describes a method of estimating distribution of ionization in upper atmosphere. This method based on measurements on several frequencies of effective height as determined by interference or echo experiments.)

R116 Bergmann, L. and Holzlöhner, G. Über die Fortpflanzungschwindigkeit elektrischer Wellen an dunnen Drähten von verschiedenem Leitvermögen. (On the velocity of propagation of electric waves along thin wires of different conductivity.) *Ann. d. Phys.*, **87**, 653-676; No. 21, 1928.

(Brief review of the theories of propagation by Hertz, Rayleigh, Drude, Mie, Sommerfeld, Mercier, Hund, and experimental data with thin wires of copper, aluminium, brass, German silver, manganin, and constantan.)

R145.3 Bashenoff, V.I. Supplementary note to abbreviated method for calculating the inductance of irregular plane polygons of round wire. PROC. I. R. E., 16, 1553-1558; November, 1928.

(Additional discussion based on this paper.)

R200. RADIO MEASUREMENTS AND STANDARDIZATION

R210 Rangachari, T. S. A method of calibrating a low frequency valve generator with a single frequency standard source. *Experimental Wireless and Wireless Engineer* (London), 5, 633-634; November, 1928.

(The source of unknown frequency is connected across a series combination of resistance and an inductance and the former is varied until the voltage across it is the same as across the inductance. The same is done when an emf of known frequency is impressed. The unknown frequency can then be calculated from resistance settings and the known frequency.)

R214 Hitchcock, R. C. Quartz crystals: How to cut and grind them. *Radio Broadcast*, **14**, 85-87; December, 1928.

(Formulas for the thickness vibration are given for the 30-degree and the Curié cut. The method of cutting and grinding piezo-electric quartz plates is described.)

R214 Harrison, J. R. Piezo-electric oscillator circuits with four-electrode tubes. PROC. I. R. E., 16, 1455-1470; November, 1928.

(Description of two piezo-oscillator circuits which use screen-grid vacuum tubes. One circuit utilizes two pairs of electrodes and feedback takes place through the crystal. In the other circuit, the two electrodes of the crystal are connected between the plate and the control grid.)

R214 Terry, E. M. The dependence of the frequency of quartz piezo electric oscillators upon circuit constants. Proc. I. R. E., 16, 1486-1506; November, 1928.

(Mathematical theory of the piezo-electric quartz oscillator is given. It treats the case where the quartz element is between the grid and the filament as well as the case when between the grid and the plate. The analysis is made for the tuned-plate circuit, inductance and resistance load.)

R214 Strout, R. S. Temperature coefficient of quartz crystal oscillators. *Phys. Rev.*, 32, 829-831; November, 1928.

(The temperature coefficient of the frequency of a quartz plate 1.8×1.8×0.11 oscillating with the thickness frequency was found to decrease linearly from 22.7 parts per million per degree at 65 deg. C to 1.6 parts per million per degree at—189 deg. C.)

R280 Rybner, J. Note sur les experiences relatives aux proprietes dielectriques des gaz ionises de MM. Gutton et Clement. (Note on the experimental results of dielectric properties of ionized gas found by Gutton and Clement). *L'Onde Electrique*, 7, 428-436; October, 1928.

(Shows that the experimental facts observed originally by Gutton and Clement can be explained by the classical theory and that it is not necessary to assume that the ions are not free and their movement due to quasi elastic forces.)

R300. Radio Apparatus and Equipment

R330 Millen, J. The control grid glow tube. *Radio Engineering*, 8, 18-19; September, 1928.

(Description of this tube giving the breakdown characteristic and other useful data.)

R331 Van Dyck, A. F. and Engel, F. H. Vacuum tube production tests. Proc. I. R. E., 16, 1532-52; November, 1928.

(General description is given of methods and apparatus for testing tubes in large quantities.)

R343 Van Dyck, A. F. and Dickey, E. T. Quantitative methods used in tests of broadcast receiving sets. Proc. I. R. E., 16, 1507-1531; November, 1928.

(General classes of receiving set measurement used by authors are outlined as special engineering tests and production tests. New form of radio-frequency generator designed for this type of work is described.)

R343.7 Warren, A. G. Surges in eliminator smoothing circuits. *Experimental Wireless and Wireless Engineer* (London), 5, 601-606; November, 1928.

(A discussion of the effects of surges occurring in an eliminator circuit which is supplied from d. c. mains. It is shown that the effects are most serious when the set is not connected or the filaments of the set are not burning.)

R344 Aisberg, E. Utilisation des lampes de T.S.F. pour la production de musique electrique. (Use of vacuum tubes for the production of electric music). *L'Onde Electrique*, 7, 455-458; October, 1928.

(Reviews the various systems for producing music by means of vacuum-tube generators.)

R344 A tuning fork controlled audio oscillator. *Jnl. Scien. Instr.* (London), 5, 361-362; November, 1928.

(Description of audio oscillator made by General Radio Company.)

R385.5 Meyer, E. Über eine einfache Methode der automatischen Klanganalysis und der Messung der Nichtlinearitat von Kohlemicrophonen. (A simple method of automatic sound analysis and measurement of distribution in carbon microphones). *Elektrische Nachrichten Technik*, 5, 398-403; October, 1928.

(A bridge method is described for testing carbon microphones. The method is a simplified arrangement for sound analysis.)

R388 Coursey, P. R. A portable oscillograph equipment for observation of transient electrical phenomena. *Experimental Wireless and Wireless Engineer* (London), 5, 616-619; November, 1928.

(Description of the apparatus.)

R400. Radio Communication Systems

R470 Fuller, L. F. and Tolson, W. A. Power line carrier telephony. *Jnl.* A. I. E. E., **47**, 711-715; October, 1928.

(Outlines briefly modern communication requirements, provision of communication and different types of equipment available for this service.)

R470 Wolfe, W. B. and Sarros, J. D. Problems in power line carrier telephony and recent developments to meet them. *Jnl.* A. I. E. E., **47**, 727-731; October, 1928.

(Outline of difficulties met in application of high-frequency communication to power systems. Description of recent developments in equipment.)

R500. Applications of Radio

R520 Craft, E. B. Airways communication service. *Bell System Tech. Jour.*, **7**, 797-807; October, 1928.

(Two-way radio telephone communication between airplane and ground experiments by the A. T. and T. Co. Describes in addition the system used by the U. S. Government and between London and Paris. Suggests that further channels in the high-frequency band should be assigned to this work.)

R531.2 Radio stations of the world on frequencies above 1500 kc. Proc. I. R. E., **16**, 1575-1604; November, 1928.

(List of high-frequency stations throughout the World.)

R551 Bigourdan, G. Sur l'unification des signaux horaires radio telegraphiques. (On the unification of radio time signals). *Rev. Generale d'electricité*, **24**, 702-703; November 10, 1928.

(Chart for time signals.)

R582 Scheppmann, W. and Eulenhöfer, A. Der Bildfunk nach dem System Lorenz-Korn. (Picture transmission system of Lorenz-Korn). *Elektrische Nachrichten Technik*, 5, 373-381; October, 1928.

(Describes sending and receiving equipment as well as the latest development in synchronization.)

R582 Oglobinski, M. G. Derniers progress de la transmission Belinographique en France. (Latest progress on the transmission of pictures by the Belin system). *L'Onde Electrique*, **7**, 446-455; October, 1928.

(Description of the improved Belin system using photoelectric tubes.)

R584 Hermann, P. W. High frequency electric furnaces. *Radio* (San Francisco), **10**, 12-13; December, 1928.

(Methods and equipment used for melting precious metals.)

R800. Non-Radio Subjects

535.3 Bergstein, M. Applications of the photoelectric cell in industry. *Radio Engineering*, **8**, 28-30; September, 1928.

(Uses in commercial work.)

538 Maurain, Ch. Activite solaire et magnetisme terrestre. (Solar activity and terrestrial magnetism). *L'Onde Electrique*, **7**, 414-427; October, 1928.

(Review of the solar activity and its relation to the diurnal variations. Includes a list of references.)

621.314.6 Turner, H. M. The constant impedance method for measuring inductance of choke coils. Proc. I. R. E., **16**, 1559-1569; November, 1928.

(Description of a method of measuring at power frequencies the inductance of iron-cored choke coils of type commonly used in radio as filters.)

CONTRIBUTORS TO THIS ISSUE

Clapp, James K.: Born December 30, 1897 at Denver, Colorado. With Marconi Wireless Telegraph Company, 1914–16; U. S. Navy, 1917–19, foreign service, 1918–19; Radio Corporation of America, 1920, also 1922–23. Received B.S. degree, Massachusetts Institute of Technology, 1923; instructor in electrical communications, M.I.T., 1923–28; M.S. degree, 1926. Engineering department, General Radio Company, Cambridge, Mass., 1928 to date. Associate member, Institute of Radio Engineers, 1924; Member, 1928.

Diamond, Harry: Born February 12, 1900 at Quincy, Massachusetts. Received B.S. degree, Massachusetts Institute of Technology, 1922; M.S. degree, Lehigh University, 1925. Engaged in air compressor work with the General Electric Company for one and a half years; for four years instructor in electrical engineering at Lehigh University; summer work with the General Electric Company, the B. F. Sturtevant Company, and Boston Elevated Company. Radio section, Bureau of Standards, July, 1927 to date. Associate member, Institute of Radio Engineers, 1926.

Drake, F. H.: Born June 21, 1901 at Watertown, Massachusetts. Received A.B. degree, Harvard University in 1923; M.A. degree, Harvard, 1925, Ph.D. degree, Harvard, 1926. Since 1927 physicist with the Radio Frequency Laboratories, Inc., Boonton, N. J. Member of the American Physical Society. Associate member, Institute of Radio Engineers, 1925; Member, 1928.

Eve, Arthur Stewart: Born 1862 at Silsoe, Bedfordshire, England. Educated at Berkhamsted School; A.B. and M.A. degrees, Pembroke College, Cambridge, England. Assistant master and bursar, Marlborough College, England; lecturer, McGill University, Montreal, Canada, 1903; major, 148th battalion C.E.F., 1914; director of the Admiralty Experimental station, Anti-submarine Division, Harwich, England, 1917; Macdonald professor of physics and director of department of physics, McGill University, 1920. Author of various publications dealing with radioactivity, ionization of gases, and applied geophysics.

Glendinning, William: Born August 16, 1896 at Dumfried, Scotland. Received B.S. degree in E.E., Massachusetts Institute of Technology, 1923; instructor of electrical engineering, 1923–28. M.S. degree in E.E., M.I.T., 1928. Transmission and distribution department, Philadelphia Electric Company, 1928 to date. Associate member, Institute of Radio Engineers, 1926.

Guillemin, E. A.: Born May 8, 1898 at Milwaukee, Wisconsin. Received B.S. degree in E.E., University of Wisconsin, 1922; assistant in electrical engineering, Massachusetts Institute of Technology, 1922–24; M.S. degree in E.E., M.I.T., 1924; Saltonstall Fellow at University of Munich, Germany, 1924–26; Ph.D. degree with major in theoretical physics, University of Munich, 1926; instructor in physics, M.I.T. 1926–28; assistant professor in E. E., M. I. T., 1928 to date. Associate member, American Institute of Electrical Engineers.

Hall, E. L.: Born April 20, 1893 at Mansfield, Ohio. Received B.S degree in E.E., Ohio State University, 1918. Radio section, Bureau of Standards, April, 1919 to date. Associate member, Institute of Radio Engineers, 1928.

Hollman, H. E.: Born November 4, 1899 at Solingen, Germany. Educated at Humanistic High School of Solingen; passed examination, 1917. Enlisted in the army, March 3, 1918, and returned from France as a prisoner of war on February 20, 1920. In 1920 began study of electrotechnics at the Technical High School of Darmstadt. Laboratory engineer, Radio Works "Tonfrequenz," 1924–1926; passed examination and received prize offered by the General Department of the High School for an essay on acoustics. At present engaged in research work for thesis.

Hull, L. M.: Born February 27, 1898 at Great Bend, Kansas. Graduated from University of Kansas in 1917; received Ph.D. degree, Harvard University, 1922. Instructor in radio at military training camp 1917–18. Associated with Radio Frequency Laboratories, Inc., Boonton, N. J., and more recently with General Radio Company, Cambridge Mass. as well. Occasional contributor to the PROCEEDINGS. Junior member, Institute of Radio Engineers 1917; Associate, 1919; Member, 1927; Fellow, 1928.

McEwan, Alexander R.: Educated Brooklyn, N. Y. public schools; course in mechanical engineering, Brooklyn Polytechnic; course in electrical engineering, Cooper Union, New York City. Became associated with W. H. Swift, Jr., then director of radio, at the formation of the radio department of the Canadian National Railways in 1923, as assistant director of radio, with active charge of the technical side of the development of that department. Director of radio, Canadian National Railways, since June 1, 1925.

Nakai, Tomozo: Born March 10, 1901 at Fukuyama City, Japan. Graduated electrical department of Osaka Higher Technical School, 1922. Entered radio section of the Electrotechnical Laboratory, Ministry of Communications, Japan, engaged in radio researches in the laboratory. Since 1925 working on wave propagation.

Nelson, J. R.: Born October 27, 1899 at Murray, Utah. Power inspector, Western Electric Company, 1922–23; B.S. in E.E., University of Southern California, 1925; Engineering Record Office, Bureau of Power and Light, Los Angeles, Calif., 1925; radio test, Radio Development Laboratory and Tube Research Laboratory, 1925–27. M.S. degree in E.E., Union College, 1927. Engineering Department, E. T. Cunningham, Inc., July, 1927 to date. Associate member of the Institute, 1927.

Olive, Gordon W.: Born 1898. Graduate of McGill University, Canada. Served overseas during World War with McGill Siege Battery. Built and operated radio broadcasting station CFCO in 1923 for Semmelack and Dickson, while manager of radio department of that firm. Joined staff of radio department of Canadian National Railways as radio engineer, February 1, 1924. Appointed technical assistant to director of radio, Canadian National Railways, October 1, 1927.

Pratt, Haraden: Born July 18, 1891 at San Francisco, Calif. Graduated from College of Mechanical and Electrical Engineering, University of California, 1914. Radio operator and installer, United Wireless and Marconi Companies, 1910–13. Expert radio aide to Navy Department, 1915–20. Engineer with Federal Telegraph Company in charge of research factory and design and construction of chain of radio stations, 1922–27. In charge of development of radio aids to air navigation, Bureau of

Standards, Washington, D. C. At present chief engineer, Mackay Radio and Telegraph Company, New York City. Associate member, Institute of Radio Engineers, 1914; Member, 1917.

Steel, William Arthur: Born November 3, 1890 at Castleton, Ontario, Canada. Hydro-electric construction with Smith, Kerry, and Chase, Toronto, 1908–11. Received B.A. Sc. degree, Toronto University, 1915. Canadian Signal Service, France, 1915–19. Captain in charge radio services for Canadian Corps, 1918–19; chief technical officer, Royal Canadian Corps of Signals with rank of major, 1920 to date. Associate member, Institute of Electrical Engineers, London; Associate member, Engineering Institute of Canada; Member, Institute of Radio Engineers, 1919.

Thompson, John H.: Chief Engineer, Canadian Marconi Company, Montreal, Canada, 1921 to present date. Previously chief engineer with Radio Branch of the Department of Marine and Fisheries in Canada. During past seven years largely instrumental in the development of the beam system in Canada. Member, Engineering Institute of Canada; Associate member, Institute of Radio Engineers, 1916.

van der Pol, Balth.: Born January 27, 1889 at Utrecht, Holland. Received Doctor's degree, University of Utrecht, 1916. Research in radiotelegraphy and allied subjects, Cambridge University, England, 1917. Conservator of Physical Research Laboratory, Teyler's Institute, Holland, 1919. Research Laboratory, Philips' Glowlamp Works, Ltd., Eindhoven, Holland, 1928. Member, Institute of Radio Engineers, 1920.

Yokoyama, Eitaro: Born July 21, 1883 at Fukui City, Japan. Graduated Engineering College, Tokyo Imperial University, 1908. Entered Ministry of Communications, Japan, engaged in radio researches at the Electrotechnical Laboratory. One of the inventors of the T.Y.K. oscillation gaps for radiotelephony. Chief of the Radio Section, Electrotechnical Laboratory, 1920. Member, Institute of Radio Engineers, 1917.

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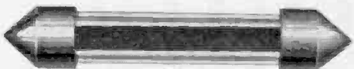

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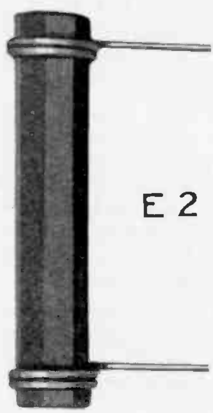

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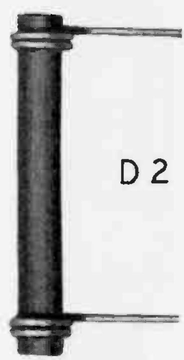

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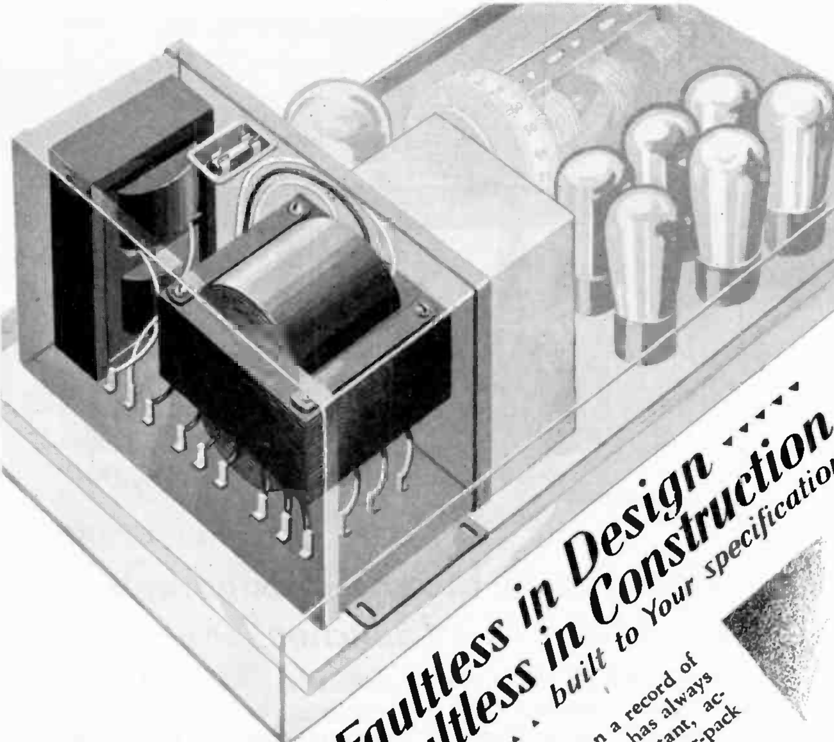

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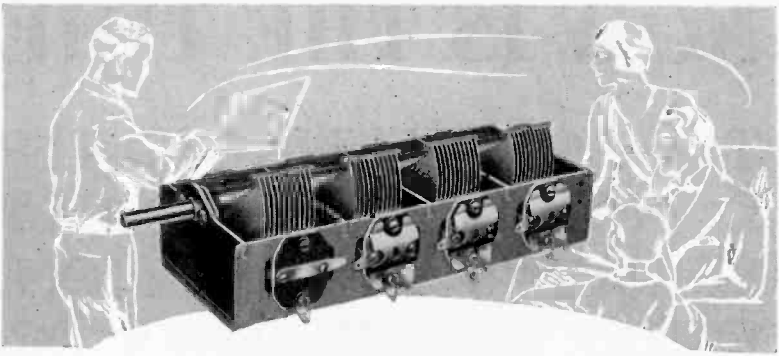

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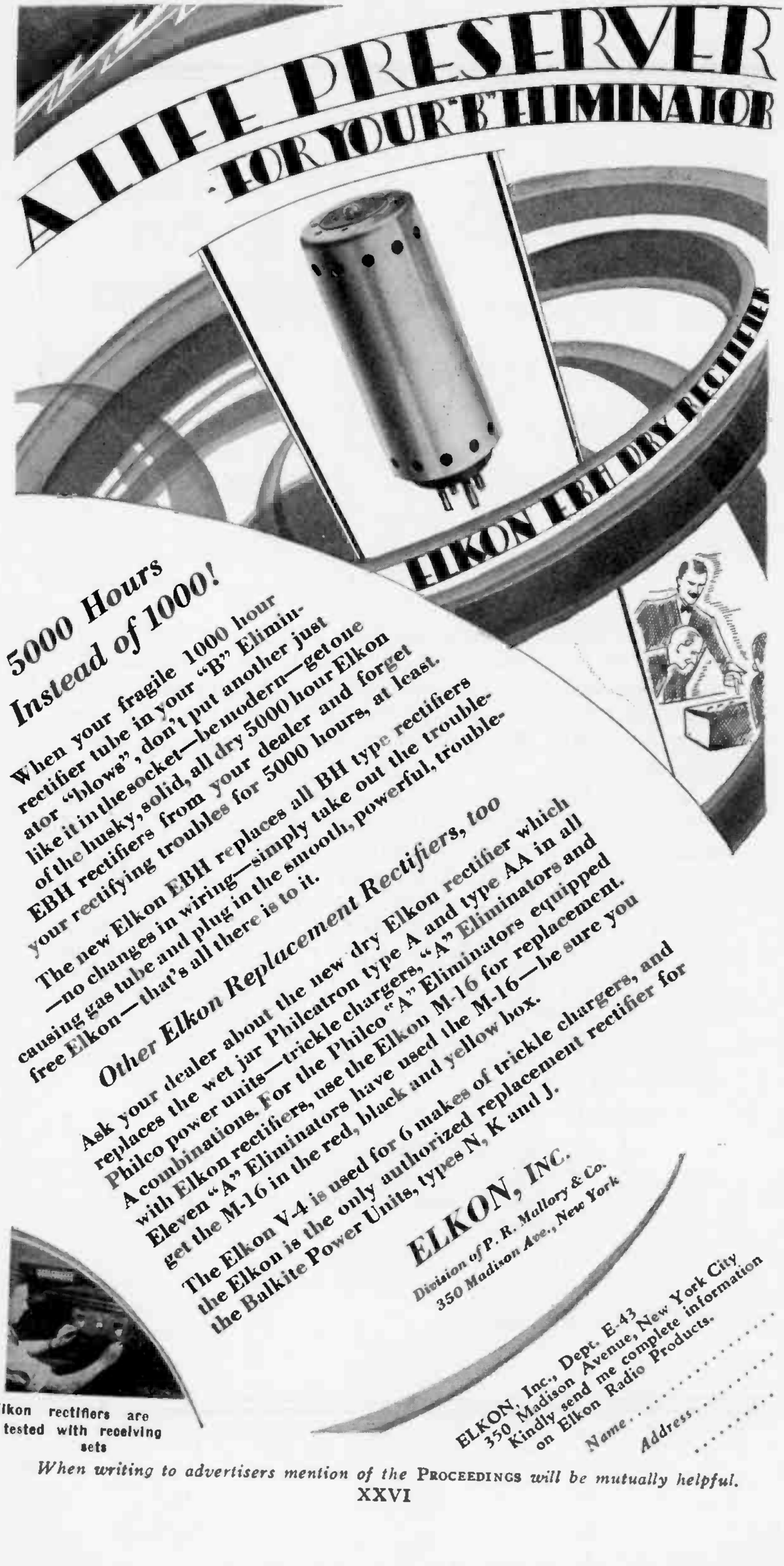
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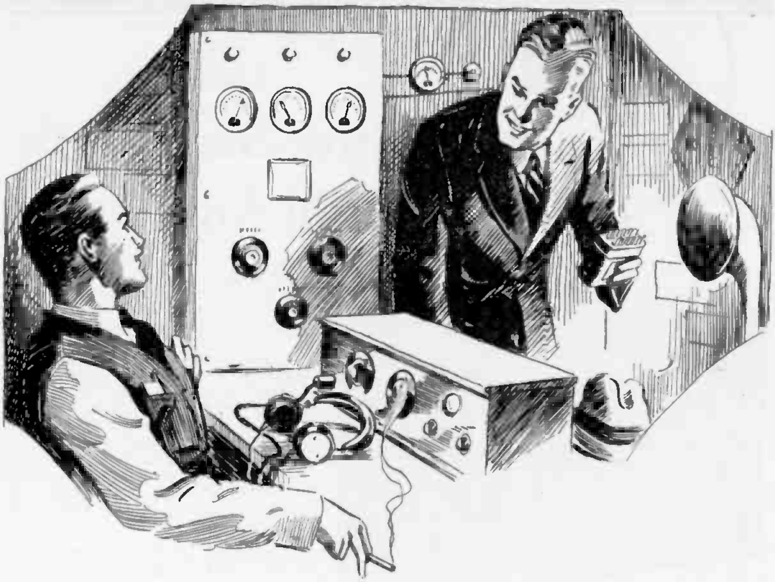

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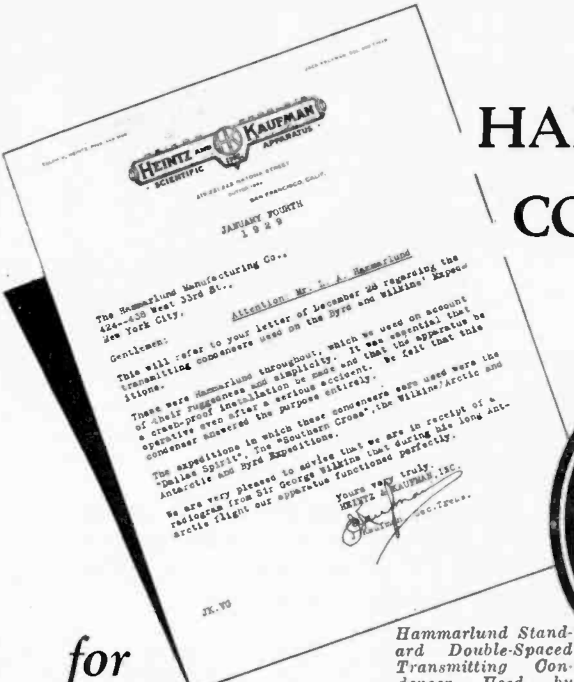

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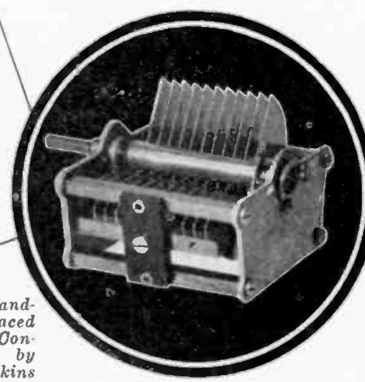

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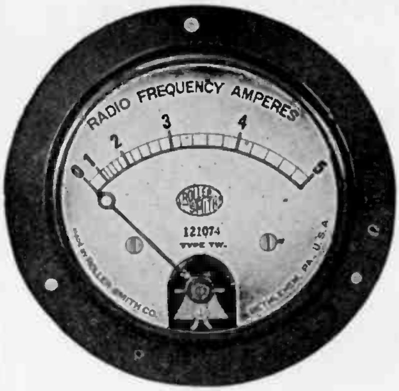

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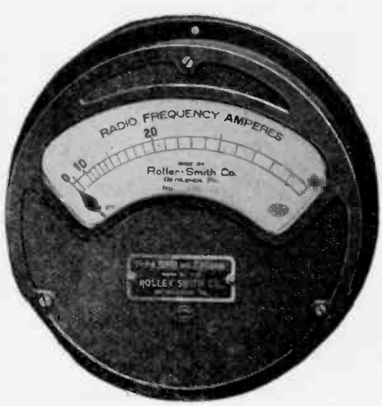

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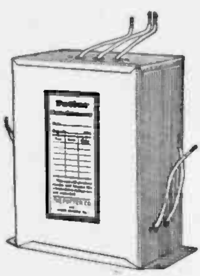

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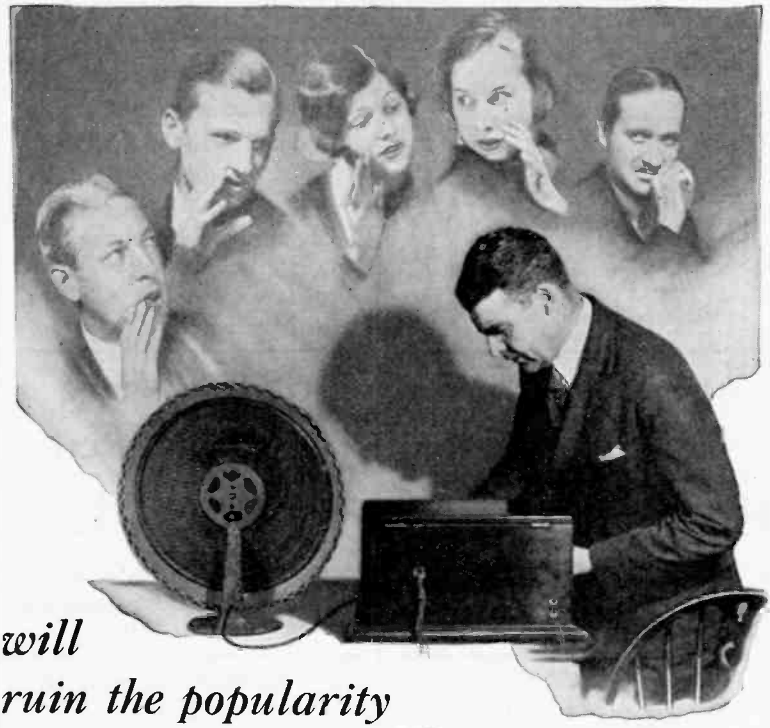

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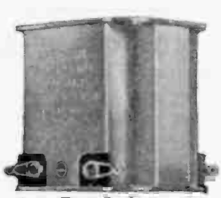

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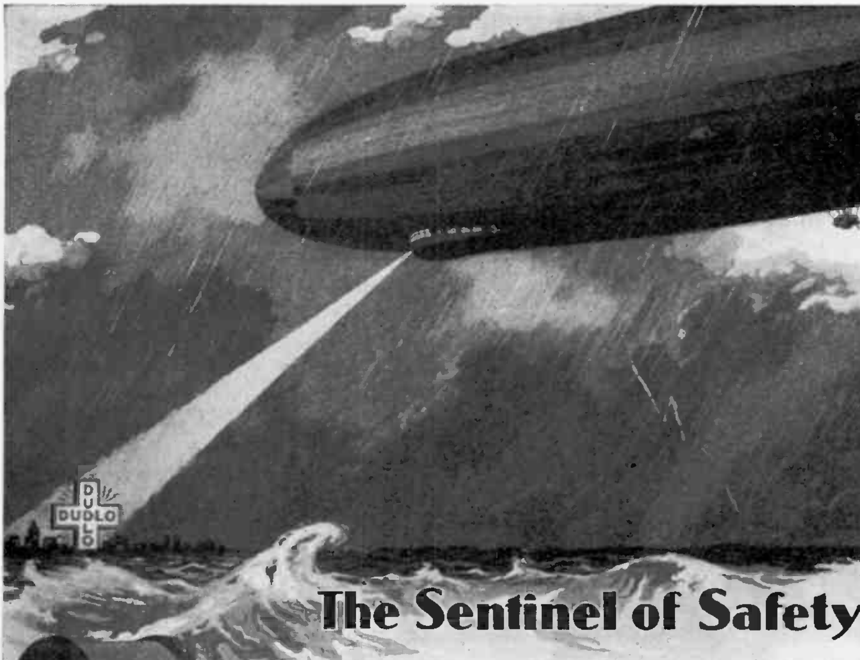

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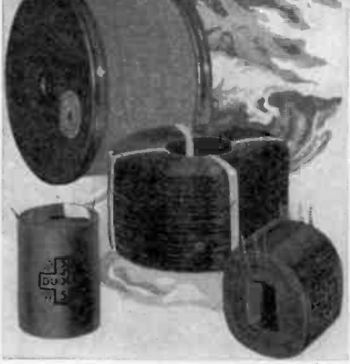

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