VOLUME 17 APRIL, 1929 NUMBER 4

PROCEEDINGS
of
The Institute of Radio Engineers

1929 Convention Program
In This Issue

General Information and Subscription Rates on Page 592

TENTATIVE PROGRAM OF FOURTH ANNUAL CONVENTION INSTITUTE OF RADIO ENGINEERS WASHINGTON, D. C.

MAY 13–15, 1929

May 12th—Sunday	
2:00 P.M.–6:00 P.M.	Registration at the Mayflower Hotel. Badges, trips and banquet tickets.
May 13th—Monday	
8:00 A.M.–10:00 A.M.	Registration at the Mayflower Hotel. Badges, trips and banquet tickets.
10:00 A.M.–12:00 M.	Sightseeing around Washington in private cars, to be arranged by ladies of Washington (ladies exclusively).
10:00 A.M.–12:30 P.M.	Opening session at the U. S. Chamber of Commerce Building, Connecticut Ave. and H St. N. W. Short speeches by President A. Hoyt Taylor, F. P. Guthrie, Chairman Washington Section, and C. B. Jolliffe, Chairman Convention Committee. Symposium on "Technical Problems of Radio Regulation" consisting of ten papers and discussions (see complete program in this issue).
12:00 M.	The ladies of Washington hostesses at luncheon to the ladies attending the Convention.
1:15 P.M.–5:00 P.M.	Trip No. 1 (ladies exclusively) to Mount Vernon by motor-bus.
1:15 P.M.–5:30 P.M.	Trip No. 2 (limited to American citizens only) Technical Inspection trip from Mayflower Hotel to the U. S. Naval Research Laboratory at Bellevue, D. C., including sightseeing trip and inspection of the Naval Research Laboratory.
8:00 P.M.	Popular lecture by Professor M. I. Pupin in U. S. Chamber of Commerce, Vice President A. Meissner presiding.
May 14th—Tuesday	
9:00 A.M.–12:00 M.	Technical Session in the U. S. Chamber of Commerce Building. Technical address by Vice President Alexander Meissner. Symposium on "Photo Radio and Television." Seven papers and discussions. (See complete program.) Demonstration of educational sound pictures dealing with the principles of modulation and filtering.
10:30 A.M.–12:30 P.M.	Trip No. 3. Sightseeing trip from U. S. Chamber of Commerce around Washington, with stops at Arlington Cemetery and Arlington Radio station.
1:15 P.M.–5:30 P.M.	Trip No. 4. Technical inspection trip from Mayflower Hotel to the U. S. Bureau of Standards, including sightseeing trip through Rock Creek Park. Inspection of the Bureau of Standards.
7:30 P.M.	Informal banquet at the Mayflower Hotel, including speeches by President A. Hoyt Taylor, and several other members of the Institute, presentation of the Institute annual awards, dancing and other entertainment.
May 15th—Wednesday	
10:00 P.M.–5:00 P.M.	Joint meeting of the Institute of Radio Engineers and the International Scientific Radio Union, American Section, at the National Academy of Sciences, 21st and B Sts. N. W. The program includes twenty minute abstract presentation of twenty papers (See complete program in this issue).
6:30 P.M.	Informal dinner of the Institute of Radio Engineers Section delegates followed by a meeting of the Committee on Sections. National Press Club, 14th and F Sts. N. W.

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 17 — April, 1929 — Number 4

CONTENTS

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 non-members 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 later 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 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 divisions. 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 proportion.

Abbreviations—Write a.c. and d.c., kc, μf, $\mu\mu$f, 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 galley proof a 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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Vice President of the Institute Comes From Germany to Attend Convention

ALEXANDER MEISSNER
Vice President of the Institute, 1929

The Board of Direction of the Institute is happy to announce that Dr. Alexander Meissner, Vice President of the Institute, is making the trip from Germany to attend the Fourth Annual Convention and to participate in the technical program thereof.

Alexander Meissner was born in Vienna, Austria, on September 14, 1883. He studied at the Vienna College of Engineering and received the degree of Doctor of Technical Science in 1909. In 1922 he was given the honorary degree of Doctor of Engineering by the College of Engineering in Munich. Since 1907 he has been associated continuously with the Telefunken Company of Berlin. He is now Director of the research laboratory of that organization. Dr. Meissner has made a number of fundamental contributions to the radio art through a series of investigations among which are the problems of losses in coils, transmitting antenna design which was especially confined to the study of reducing the resistance losses in the earth, machine transmitters, impulse excitation of quenched spark transmitters, and many vacuum-tube circuits. Dr. Meissner is well known for his active work in the field of vacuum-tube amplification systems, his investigations of quartz crystals in high-frequency work, his study of the physical structure of quartz in piezo and pyro-electricity, and the manufacture of piezo-electric plates from non-crystalline materials.

In 1925 Dr. Meissner was awarded the Heinrich Hertz gold medal in recognition of his many inventions in the radio field.

He became an Associate member of the Institute in 1914 and was transferred to the Fellow grade in 1915. He has been a frequent contributor to the PROCEEDINGS and other technical journals. The membership of the Institute elected him Vice President of the Institute in 1929.

INSTITUTE NEWS AND RADIO NOTES

Washington Convention of the Institute

The Institute is affording its members an unusual program of varied interest at its Fourth Annual Convention which is to be held in Washington, D. C., May 13th to 15th. As well as an excellent technical and inspection program, the Convention committee has arranged a number of special attractions of particular interest to the lady guests to assure all members that their wives are not only welcome but are urged to be in Washington for the entire Convention. The flowers in the many parks of Washington are at their height in May and it is doubtful if they are equalled anywhere in the United States.

Among the features of the Convention will be technical sessions, a popular lecture, technical inspection trips, sightseeing trips for the ladies, a banquet, presentation of the Institute annual awards, a joint meeting of the Institute and the American Section of the International Scientific Radio Union, and a reception by the President of the United States.

Technical Sessions

Thirty-seven papers on radio engineering subjects of timely importance will be presented during the technical sessions. Two half-days will be devoted to symposiums treating the technical problems of radio regulation, and photo radio and television. One full day will be occupied by the joint I. R. E.-U. R. S. I. meeting, at which twenty individual technical contributions will be offered The *tentative* program of technical papers follows:

May 13th, 10:00 a.m. to 12:30 p.m.

Speeches of introduction by President A. Hoyt Taylor, F. P. Guthrie,and C. B. Jolliffe.

Symposium on Technical Problems of Radio Regulation

General Introduction—by the Chairman of the Federal Radio Commission.

Engineering Aspects of the Work of the Federal Radio Commission—by J. H. Dellinger, Federal Radio Commission.

Receiver Developments Affecting Broadcasting Regulations—by L. M. Hull, Radio Frequency Laboratories.

Principles of Broadcast Frequency Allocation—by L. E. Whittemore, American Telephone and Telegraph Company.

Heterodyne Interference—by J. V. L. Hogan, Consulting Engineer.

Ship Radio Inspection and Frequency Monitoring—by Arthur Batcheller, Department of Commerce.

Harris & Ewing

The Capitol

Lincoln Memorial Through the Cherry Blossoms (Trip No. 3)

H. H. Rideout

Mount Vernon (Trip No. 1)

H. H. Rideout

The Japanese Cherry Blossoms—Washington Monument in Background (Trip No. 3)

Promenade of the Mayflower Hotel—Convention Headquarters

Grand Ballroom of the Mayflower Hotel where Annual Banquet Will Be Held

The Problems Centering About the Measurement of Field Intensity—by S. W. Edwards, Department of Commerce.

Transmitter Developments Affecting Broadcast Regulation—by E. L. Nelson, Bell Telephone Laboratories.

The Radio Engineer's Responsibility in Coping with Man-Made Interference—by Edgar Felix, National Electrical Manufacturers' Association.

Radio Coordination, by M. D. Hooven, National Electric Light Assciation.

The Development of United States Radio Broadcasting—by R. H. Marriott, Federal Radio Commission.

May 14th, 9:00 A.M. to 12:00 M.

Technical Address—by Vice President Alexander Meissner.

Symposium on Photo Radio and Television

General Introduction—by Alfred N. Goldsmith, Radio Corporation of America.

Electrical Transmission of Pictures and Images—by J. W. Horton, General Radio Company.

Commercial Radio Facsimile Communication—by R. H. Ranger, Radio Corporation of America.

High Speed Facsimile Picture Transmission—by V. Zworykin, Westinghouse Electric and Manufacturing Company.

The Drum Scanner in Radiomovies Receivers—by C. Francis Jenkins, Jenkins Laboratories.

The Selection of Standards in Commercial Radio Television—by Julius Weinberger, Theodore A. Smith, and George Rodwin, Radio Corporation of America.

Optical Systems for Television Image Reproduction—by Bernard Cioffari, Radio Corporation of America.

Following the symposium there will be a demonstration of educational sound pictures dealing with the principles of modulation and filtering, presented by John Mills, Bell Telephone Laboratories.

I. R. E.—U. R. S. I. Joint Meeting

May 15th, 10:00 A.M. to 5:00 P.M.

Current Developments in Radio Measurement Research—by J. H. Dellinger, Bureau of Standards.

The Inductance of Iron-cored Coils as Determined with Superimposed Alternating and Direct Current, Length of Air Gap, and Initial Magnetic State—by H. M. Turner, Yale University.

Push-Pull Piezo-oscillator Circuits—by J. R. Harrison, Wesleyan University.

The Calculation of the Inductance of Single-layer Coils and Spirals Wound with Wire of Large Cross Section—by F. W. Grover, Union College.

Current Developments in Interference Measurement—by E. F. W. Alexanderson, General Electric Company.

Radio Interference from Power Circuits—by Norman Snyder, General Electric Company.

Current Developments in Radio Wave Propagation Research—by L. W. Austin, Bureau of Standards.

Radio Wave Propagation Phenomena—by Major W. R. Blair, Signal Corps, War Department.

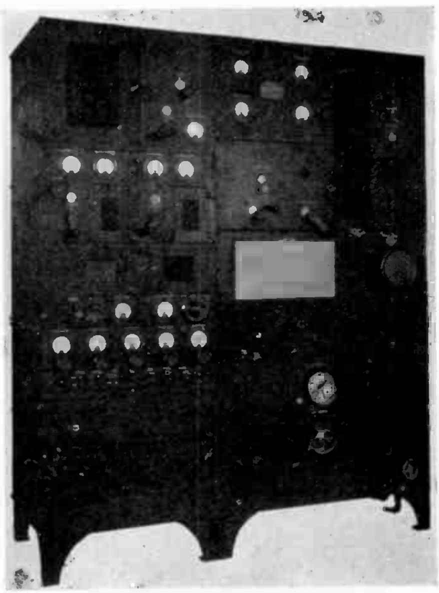
The Effect of Group Velocity on the Retardation of Radio Signals—by G. Breit, Carnegie Institution.

Studies of the Height of the Kennelly-Heaviside Layer by the Echo Method—by M. A. Tuve and L. R. Hafstad, Carnegie Institution.

Further Studies of Echo Signals—by A. Hoyt Taylor and L. C. Young, Naval Research Laboratory.

Ionization in the Atmosphere of Mars—by E. O. Hulburt, Naval Research Laboratory.

Current Developments in Radio Atmospherics Research—by H. T. Friis, Bell Telephone Laboratories.



Front View of Navy High-Power Transmitter
Designed at Bellevue (Trip No. 2)

Current Developments in Radio Research Cooperation—by A. E. Kennelly, Harvard University.

Input and Output Admittance of Triodes—by E. L. Chaffee, Harvard University.

Operating Characteristics of Triode Oscillators—by E. L. Chaffee and R. F. Field, Harvard University.

Reciprocal Theorems in Radio Transmission—by J. R. Carson, American Telephone and Telegraph Company.

Circuit Tuning by Wave Resonance with Application to Signal Reception—by L. Cohen, Signals Corps, War Department.

Application of Wave Resonance Tuning to Signal Transmission—by W. R. Blair and L. Cohen, Signal Corps, War Department.

Detection at High Signal Voltages: I. Plate Rectification—by Stuart Ballantine, Radio Frequency Laboratories.

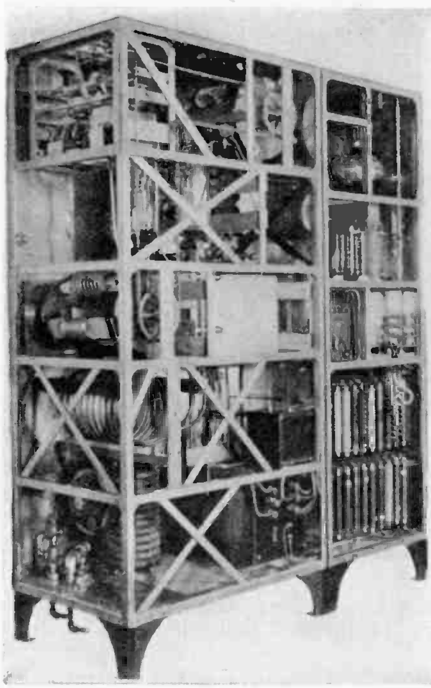
The Mutual Impedance between Adjacent Antennas—by C. R. Englund and A. B. Crawford, Bell Telephone Laboratories.

Reciprocity in Electromagnetic, Mechanical, Acoustical, and Interconnecting Systems—by Stuart Ballantine, Radio Frequency Laboratories.

Popular Lecture

May 13th, 8:00 p.m.

We are particularly fortunate in being able to announce that Professor M. I. Pupin, of Columbia University, past president of both the



Rear View of High-Power Transmitter—Case Removed (Trip No. 2)

Institute of Radio Engineers and the American Institute of Electrical Engineers, will deliver a lecture of general scientific interest.

Dr. Alexander Meissner will preside at this meeting and introduce Professor Pupin.

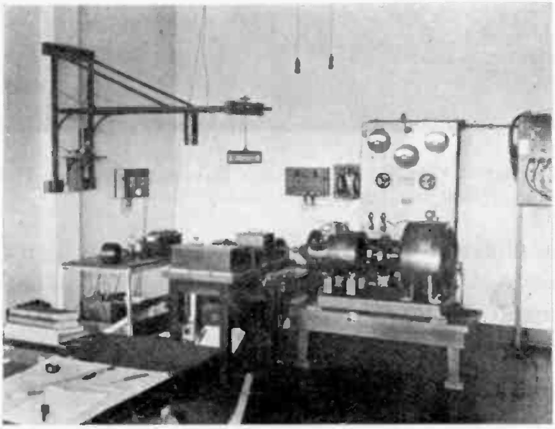
Inspection Trips

As the places to be inspected on all organized trips are at some distance from the center of Washington (and Convention Headquarters), these trips will constitute sightseeing tours as well as technical inspection trips, and consequently will be of much interest to the ladies. Trip No. 1 to Mount Vernon is exclusively for the ladies. The ladies of Washington will be glad to arrange special sightseeing trips for visiting ladies in private cars to points of interest.

Trip No. 2. U. S. Naval Research Laboratory

May 13th, 1:15 p.m. to 5:30 p.m.

This trip is limited to American citizens; the Naval Research Laboratory is a part of the nation's defense system, and as it is the universal policy among nations to limit visitors to such places to its own citizens, unfortunately this same policy will have to be followed in



One of the Several Forms of Frequency Standardization Equipment at the Naval Research Laboratory (Trip No. 2)

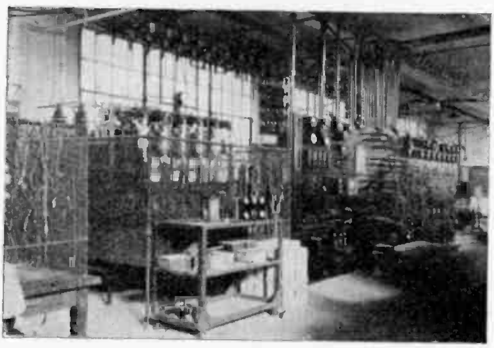
connection with this trip. Members of the Institute and their immediate families planning to take this trip will be required by the Government to register especially for it at the Convention registration desk in the Mayflower Hotel prior to 11:00 a.m. on May 13th.

The Naval Research Laboratory is located seven miles from the center of Washington, at Bellevue, D. C. Buses will start from the Mayflower Hotel and will proceed via the White House and Treasury Building along historic Pennsylvania Avenue past the Botanical Gardens, the Capitol, the Congressional Library, Navy Yard, and St. Elizabeth's Hospital to the Laboratory. Members will return in buses via The Mall, Smithsonian Institute, the National Museum, and the Department of Agriculture, arriving at Convention Headquarters.

At the Laboratory members will be conducted through the radio and sound divisions. The sections to be visited in the radio division are precision measurement, radio compass, transmitter, aircraft, re-

A Corner of the Naval Research Laboratory's Low-Power Transmitter Design Room (Trip No. 2)

ceiver and amplifier, and lastly the "field house" in which the extensive study of high-frequency transmission phenomena has been carried on for several years.



The High-Power Direct-Current Gallery at the Naval Research Laboratory (Trip No. 2)

The precision measurement section contains the Navy's standards of frequency, permanently installed bridges for inductance, capacity, and resistance measurements, and elaborate equipments for absolute calibration of quartz crystals which check back to Observatory time. Of special interest will be the apparatus for absolute calibration by means of which it is hoped an accuracy of one part in ten million will be realized.

The transmitter section will have several problems of general interest in progress. The high-power gallery, in addition to the high-voltage direct-current machinery supply, contains the transmitters with which the Kennelly-Heaviside layer measurements, using the spurt measure-

Bank of Four High-Power High-Frequency Transmitters at Arlington Radio Station (Trip No. 3)

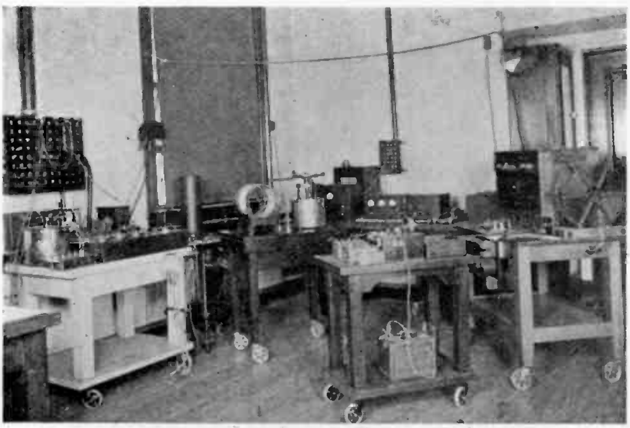
ment method, are being studied. It is expected that a high-power six-meter transmitter will also be in operation.

At the field house the visitors will see a high-power push-pull, high-frequency, crystal-controlled transmitter, as well as a half-kilowatt crystal-controlled transmitter which can be operated up to 35,000 kc. The receiver and oscillograph with which round-the-world signals and echoes are studied will be on display.

The aircraft section will have several sets of historic interest on exhibition. Such equipment as that used on the *Shenandoah*, the

PN7, and Byrd's *America* will be seen. Modern naval aircraft transmitters, receivers, generators, alternators, and dynamotors will be shown. A study of the trend of modern aircraft design, here and abroad, may be made through the comparison of equipment of British and American design on display.

The receiver and amplifier section has a well-equipped laboratory for all manner of receiving investigations. Unique specially screened booths, in which sensitive receiving measurements are made in the midst of high-power high-frequency transmitters, are provided. These booths (three in number) are doubly shielded, each complete screen



Bureau of Standards Laboratory for Testing Piezo-Oscillators for Broadcast Stations (Trip No. 4)

being insulated from the other. They are also arranged so that the center booth can house a driver while in the end one measurement data can be taken without external interference to the indicating instruments, or modulation of the driver frequencies.

Near the main Laboratory there will be found the radio compass building which is designed to approximate the standard shore station layout. In this building the radio compass section carries on its investigations and development work.

At the docks on the Potomac river will be seen the field stations of the sound division—consisting of two well equipped barges which are operated in Chesapeake Bay except during the winter months. The sound division will have on display a direct reading sonic depth finder. There will also be a demonstration of a few applications of piezoelectric activity of Rochelle Salt crystals.

Trip No. 3. Sightseeing with Stops at Arlington Cemetery and Arlington Radio Station

May 14th, 10:00 a.m. to 12:30 p.m.

This trip will parallel a portion of the symposium on photo radio. Leaving the Mayflower in buses, those participating in this trip will be taken on a two and a half-hour trip to many famous points of Washington. Lecturers will be provided on each bus. The buses will pass the War and Navy Buildings, Corcoran Art Gallery, Red Cross Building, Pan American Union, National Academy of Sciences, Lincoln Memorial, around Haynes Point from which an excellent view of the Naval Air Station, Bolling Field, and the War College may be had, past the

Laboratory for Study of Radio Wave Phenomena and Measurement of Intensity of Radio Waves at the Bureau of Standards (Trip No. 4)

public golf links, the Bureau of Printing and Engraving, the Tidal Basin surrounding which are the famous Japanese Cherry trees, to Arlington Cemetery and the Tomb of the Unknown Soldier. From Arlington Cemetery the buses will go to Arlington Radio Station, well known to all, where a stop will be made to inspect the latest low- and high-frequency naval transmitters in operation. The return trip will be through Fort Meyer across Francis Scott Key bridge, through historic Georgetown to the Mayflower Hotel.

Trip No. 4. U. S. Bureau of Standards

May 14th, 1:15 p.m. to 5:30 p.m.

En route the Bureau of Standards, buses with lecturers will leave Convention Headquarters, proceeding up 16th Street, passing many embassies and fine churches, through the finest residential section of Washington to one of the north entrances of Rock Creek Park. A four-mile ride through this famous picturesque park will bring the delegates out at Old Pierce Mill. From thence the route will lead direct to the Bureau of Standards. The return trip will be made along Connecticut Avenue, across the Million Dollar bridge, past the former home of President Hoover and the "S" Street home of Woodrow Wilson to the Mayflower hotel.

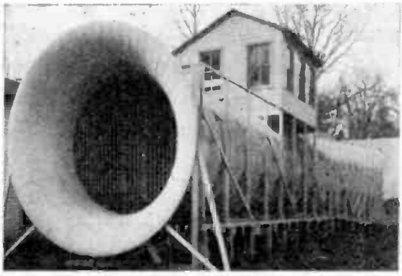
The Bureau of Standards affords a large number of interesting sights. The attention of the members will be concentrated on the radio and electrical buildings, but opportunity will be given for anyone

Radio Transmission Research (Low-Frequency) Laboratory of Dr. L. W. Austin at the Bureau of Standards (Trip No. 4)

interested to visit other laboratories. The work of the Bureau of Standards includes nearly all phases of physics, chemistry, and engineering and informal visits to any section will be welcomed. Information may be obtained from the laboratory members.

In the radio building there are many interesting developments to be observed. Members of the staff will be available for answering questions and discussing the various projects. Probably of most interest will be the following: the study of radio wave phenomena and measurement of radio field intensity showing apparatus for recording fading and illustrative records, the laboratory car and other measuring equipment, demonstration of testing of piezo-oscillators for broadcast stations and measurements on standard piezo-oscillators, demonstration of the visual radio beacon system and display of apparatus developed in connection with work on radio aids to air navigation, the laboratory of Dr. L. W. Austin for radio transmission research (low frequency) where the classic transmission measurement work of Dr. Austin, which the PROCEEDINGS has published from time to time, has been carried on for a number of years, the Signal Corps Laboratory where the work on resonance wave tunings will be demonstrated.

In the Electrical Building will be seen the life test equipment for incadescent lamps where several thousand lamps are tested annually, heavy current testing and tests of electrical measuring instruments, standard cell laboratory in which is maintained the United States standard of voltage and life testing of dry batteries, including radio batteries in which all switching is done automatically according to any predetermined time schedule.



Ten-Foot Wind Tunnel for the Study of Effect of Wind on Structures Such as Airplanes, Dirigibles, and Buildings, at the Bureau of Standards (Trip No. 4)

In the main laboratory of the Wind Tunnel Building there is one of the three wind tunnels used at the Bureau for investigating models of airplanes, dirigibles, bombs, buildings, and chimneys. This tunnel is fifty inches across at its smallest section, and in it a wind speed of 75 miles an hour can be maintained.

The work of the Bureau on radio aids to air navigation is carried on at an experimental station at College Park, Maryland. Members interested in this work are welcome to visit the station at any time between 9:00 A.M. and 4:30 P.M. Special arrangements for a visit by a group of members may be made at Convention Headquarters at registration time.

Reception By President Hoover

A reception by the President of the United States at the White House to all persons attending the Convention is being arranged. The date and hour will be announced later.

The Banquet

MAY 14TH, 7:30 P.M.

The annual banquet of the Institute, which will be informal, will be held in the grand ballroom of the Mayflower Hotel. Included in the program are speeches by A. Hoyt Taylor, toastmaster, and by several other members of the Institute, presentation of the Institute Medal of Honor and the Morris Liebmann Memorial Prize, and several forms of light entertainment. Dancing will be provided for those who desire.

Entertainment for the Ladies

From the foregoing paragraphs it may be judged that the lady guests are to be provided with a varied program of interest. In addition to these the Ladies' Committee of the Convention has arranged a number of added attractions.

On Monday morning the Ladies' Committee will be present at the registration desks to arrange informal sightseeing trips exclusively for the ladies. These trips will be made in private cars and will include any of the sights of Washington which are of interest to the individual persons.

At noon on Monday the ladies of Washington will be hostesses to the lady guests at luncheon at one of the Washington Women's Clubs. Following the luncheon the ladies will go on Trip No. 1 to Mount Vernon. This will be a four-hour tour by motor-bus and will include Georgetown, Alexandria, and Mount Vernon.

The Ladies' Committee will be on hand at other times to direct women visitors to any of the numerous interesting sights which they may care to take in.

Committee on Sections Meeting

May 15th, 6:30 p.m.

A meeting of the Committee on Sections, which it is believed will be attended by delegates from each Section of the Institute, will be held at an informal dinner at the National Press Club, 14th and F Streets, N. W. Those members of the Sections who attended this session of the 1928 Convention will recall the helpful and instructive meeting which took place. These meetings afford the only opportunity for the Officers of the Institute and the Committee on Sections with its various representatives to meet in a body to study Section plans and operation. The 1929 Committee on Sections is arranging a very interesting program for this meeting.

General Information

Convention Headquarters will be at the Mayflower Hotel which is centrally located (Connecticut Ave. and L St. N. W.), and very near the U. S. Chamber of Commerce building in which the technical sessions will be held.

The Mayflower Hotel has set aside the following rooms for Convention delegates: Sixty-five single rooms at $5.00 a day, twenty-five single rooms at $6.00 a day, twenty-five rooms at $7.00 a day, fifty twin-bedded rooms at $9.00 a day for two and $7.00 for one, fifty twin-bedded rooms at $10.00 for two and $8.00 for one. If more than one person occupies a single room all of which are provided with double beds the price will be $2.00 a day additional. To insure reservations desired the members are cautioned to fill in and return promptly the hotel reservation postcard which will be sent with the Convention programs during the latter part of April.

Members coming to Washington by auto will find a number of garages near the Mayflower. Among these are: Mayflower Garage, 1830 L Street, $1.50 a day; Standard Garage, 1815 L Street, $1.00 a day; L Street garage, 1705 L Street, $1.00 a day; and the Palace Garage, 1216 20th Street, $1.00 a day.

Also with the Convention programs there will be mailed a detailed map of the city of Washington showing the auto approaches to the city and Convention Headquarters.

Tickets for all inspection trips must be obtained at the time of registration either on Sunday, May 12th or Monday, May 13th, prior to the opening session. These tickets will be available free of charge to members of the Institute and their immediate families. Due to the limited facilities it will be impossible to take non-members, other than their families, on any inspection trip or sight-seeing tour.

The banquet tickets will be $5.00 a plate, and may be obtained either at the time of registration or in advance thereof from Mr. F. P. Guthrie, Radio Corporation of America, 1112 Connecticut Avenue, Washington, D. C.; or the Secretary, The Institute of Radio Engineers, 33 West 39th Street, New York City. Remittances must accompany orders for tickets. Particular table assignments will be made only when one or more complete tables (seating ten persons) are reserved.

March Meeting of the Board of Direction

The March meeting of the Board of Direction of the Institute was held on the 6th of the month in the office of the Institute. The following were present: R. H. Marriott, Acting Chairman; Melville Eastham, Treasurer; John M. Clayton, Secretary; Alfred N. Goldsmith, Junior Past President; Arthur Batcheller, W. G. Cady, R. A. Heising, and R. H. Manson.

The Board transferred and elected the following to the higher grades of membership in the Institute: transferred to the Fellow grade:

Haraden Pratt; elected to the Fellow grade: Guiseppe Pession; transferred to the Member grade: W. D. Loughlin; elected to the Member grade: Roland B. Bourne, G. A. Struthers, and Ralph Wyckoff.

One hundred and twenty-one Associate members and eight Junior members were elected.

Standard Frequency Transmissions by the Bureau of Standards

The schedule of standard frequency transmissions by the Bureau of Standards for the months of April and July (inclusive) appear below. 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

Institute Meetings

Atlanta Section

The Atlanta section held a meeting February 8th in the Federal Building, Atlanta, Georgia, presided over by W. Van Nostrand, chairman. There was an attendance of twenty-seven members.

J. K. Clapp, of the General Radio Company, presented a paper entitled, "A Convenient Method for Referring Secondary Frequency Standards to a Standard Time Interval." This paper was published in the February, 1929, issue of the Proceedings.

Buffalo-Niagara Section

The February meeting of the Buffalo-Niagara section was held at the University of Buffalo, Buffalo, N. Y., on February 14th. L. C. F. Horle, chairman, presided. Fifty members attended.

The following papers were presented: "Gain Measurements at Amateur Band Frequencies," by F. A. Lidbury; "Resistance Measurements at Broadcast Frequencies" by Alton Velia; and "Design and Construction of Beat Frequency Oscillator," by G. Edgar Stone. A general discussion followed the presentation of these papers.

Cleveland Section

The Cleveland section held a meeting in the Case School of Applied Science, Cleveland, Ohio, on February 28th. Bruce W. David, chairman of the section, presided. Fifty-five members attended.

O. T. McIlvaine presented a paper, "Vacuum Tubes—New Developments and Problems," describing a new rectifier tube which converts 110-volt alternating current to 220-volt direct current without a transformer. Two separate anodes and heater type cathodes are enclosed in the tube which fits a standard 227-type socket. The heater elements are supplied from a 6-volt source of alternating current or direct current. A general discussion followed the presentation of this paper.

Los Angeles Section

On February 18th the Los Angeles section held a meeting in the Elite Cafe, Los Angeles, California. T. F. McDonough, chairman, presided. Sixty-four members attended the meeting.

R. B. Parrish and G. D. Ellis presented short informal talks on "Sound Recording As Applied To Motion Pictures" and "Photophone System of Sound Recording." A general discussion followed.

New York Meeting

On March 6th a New York meeting of the Institute was held in the Engineering Societies Building, 33 West 39th Street. Alfred N. Goldsmith, Past President, presided in the absence of President Taylor.

G. L. Beers, of the Westinghouse Electric and Manufacturing Company, presented a paper by W. L. Carlson and himself entitled "Recent Developments in Superheterodyne Receivers."

Following the presentation of this paper, which was published in the March, 1929, issue of the Proceedings, it was discussed by Messrs. Beers, Goldsmith, MacDonald, Jones, Murray, and others. Four hundred and twenty-five members of the Institute and their guests attended this meeting.

Philadelphia Section

The Philadelphia section held a meeting on February 19th at the Franklin Institute, Philadelphia, Pa. J. C. Van Horn, chairman, presided. Twenty members of the section attended the meeting. J. K. Clapp, of the General Radio Company, presented a paper, "A Convenient Method for Referring Secondary Frequency Standards To A Standard Time Interval." This paper was published in the February, 1929, issue of the Proceedings.

Messrs. Miller, Synder, Van Horn, and others participated in the discussion which followed.

Pittsburgh Section

J. K. Clapp, of the General Radio Company, was the speaker at the Pittsburgh section meeting, which was held February 21st in the Fort Pitt Hotel, Pittsburgh, Pa. W. K. Thomas presided and twenty-two members of the section attended.

The paper, "A Convenient Method for Referring Secondary Frequency Standards to a Standard Time Interval," which was presented by J. K. Clapp, was published in the February issue of the Proceedings.

Messrs. McKinley, Hitchcock, Harmon, Mag, Thomas, and Allen participated in the discussion which followed.

San Francisco Section

The San Francisco section held a meeting February 13th in the Bellevue Hotel, San Francisco, Calif. Donald K. Lippincott, chairman, presided and fifty-one members of the Institute attended.

Haraden Pratt, Chief Engineer of the Mackay Radio and Telegraph Co., presented a paper, "Radio Communication in the Field of Aeronautics." The paper told of the development of beacon reception on airplanes, starting with the trailing wire antenna. The difficulties with this type of antenna were told, and the inaccuracies in direction pointed out. A description of the transmitters used to give beacon signals was given, and the final development of a two-reed indication receiver was explained. Field intensity curves for the beacon transmitters were shown, and the variation of this curve by variation of phase angle between two loops or transmitters was explained. The vertical pole antennas were described as being very true in directional qualities, and in utilizing the field of the propagated wave night effects were eliminated. The shielding of ignition, including spark plugs, was described. A general discussion followed the presentation of this paper.

Chairman Donald K. Lippincott announced the appointment of the Publicity, Meetings and Papers, and Membership Committees. Arthur H. Halloran was appointed chairman of the Publicity Committee, and Preston Allen and John L. Tustin, members; Benjamin E. Wolf, chairman of Meetings and Papers Committee, with George T. Royden and Philo T. Farnsworth, members; and Bernard H. Linden, chairman of the Membership Committee, with William P. Bear and Clifford J. Dow, members.

Toronto Section

On February 20th the Toronto section held a meeting in the Electrical Bldg., University of Toronto, Toronto, Canada. A. M. Patience, chairman of the section, presided. Fifty-four members attended.

J. H. Thompson, Chief Engineer of the Canadian Marconi Co., Montreal, presented a paper, "Marconi Beam System." The system of transmission and reception were explained, and methods of feeding the aerial and of concentrating and directing the beams were explained in detail. The illustrated slides showed the Beam Stations at Drummondville and Yamachiche, Province of Quebec. Messrs. Mott, Roseburgh, Patience, and others participated in the discussion which followed.

Washington Section

The Washington section held a meeting February 14th in the Continental Hotel, Washington, D. C. F. P. Guthrie, chairman, presided. Sixty-five members and guests attended the meeting.

J. K. Clapp, of the General Radio Company, presented a paper, " A Convenient Method of Referring Secondary Frequency Standards to a Standard Time Interval." This paper was published in the

February, 1929, issue of the PROCEEDINGS. August Hund and others entered into the discussion which followed.

Committee Work

Committee on Admissions

At the March 6th meeting of the Committee on Admissions the following members were present: R. A. Heising, chairman; H. F. Dart, A. F. Van Dyck, E. R. Shute, J. S. Smith, and B. W. David, chairman of the Cleveland Section. The Committee considered thirty-nine applications for transfer or election to the higher grades of membership in the Institute.

Committee on Membership

A meeting of the Committee on Membership was held in the Institute office on March 6th with attendance as follows: H. B. Coxhead, acting chairman; J. C. Stroebel, Jr., A. F. Murray, and F. R. Brick.

The Committee considered in detail a memorandum circulated by Mr. Coggeshall, chairman, outlining a very progressive program of work for 1929.

The Committee particularly solicits correspondence from out of town members to secure the benefit of the points of view of members throughout the entire world. It is felt that the growth of the Institute's membership can be materially assisted if members will endeavor to bring the Institute and its aims and activities before prospective members whenever an occasion presents itself.

Committee on Sections

The first meeting of the 1929 Committee on Sections was held in the Institute Office at 5:00 P.M. on Wednesday, March 6th. The following were present: E. R. Shute, chairman, Austin Bailey, L. A. Briggs, A. F. Murray, Q. A. Brackett (of the Connecticut Valley Section), B. W. David (of the Cleveland Section), and C. J. Porter, Assistant Secretary.

The Committee discussed the Section meeting program for the 1929 Convention. The Committee initiated the revision of the "Section Manual" which was prepared several years ago for the benefit of persons interested in the organization of Sections.

In detail the Committee considered correspondence with persons in five territories in which it was thought that Sections might be organized. It is anticipated that the majority of these prospective Sections will be in operation before the end of the year.

Committee on Meetings and Papers

A meeting of the Committee on Meetings and Papers was held at 9:30 A.M. March 6th in the Institute Office, 33 West 29th Street. Those present were K. S. Van Dyke, chairman; W. G. Cady, Wilson Aull, Ralph Batcher, Edgar Felix, Virgil Graham, J. W. Horton, W. H. Murphy, Greenleaf W. Pickard, R. H. Ranger, Julius Weinberger, W. Wilson, B. E. Shackelford, W. C. White, and H. F. Olson.

The Committee discussed the relation between the Board of Editors and the Committee on Meetings and Papers. A subcommittee to prepare a code of the Committee on Meetings and Papers was appointed. This Committee personnel is W. C. White, chairman; B. E. Shackelford, W. G. Cady, and K. S. Van Dyke.

To arrange for the programs of New York meetings of the Institute a subcommittee of the Committee on Meetings and Papers was appointed as follows: W. Wilson, chairman; F. H. Kroger and John M. Clayton.

The Committee read and approved the tentative program of the technical portion of the 1929 Convention.

Correction

Through typographical error in connection with the publication of the paper "Reception Experiments in Mt. Royal Tunnel" in the February, 1929, issue of the PROCEEDINGS, the positions held by two of the authors were incorrectly listed. A. R. McEwan should be listed as Director of Radio Department of the Canadian National Railways; G. W. Olive as Radio Engineer, Department of Radio, Canadian National Railways; and J. H. Thompson as Chief Engineer of the Canadian Marconi Company.

Personal Mention

William K. Aughenbaugh, formerly operating engineer at Station WFBG at Altoona, Pa., has recently become control room operator of Station WLW in Cincinnati, Ohio.

L. F. Jones, 3d, has been transferred from the Schenectady Radio Engineering Department of the General Electric Company to the same department of the General Electric Company at Oakland, Cal. Mr. Jones is engaged in short-wave development work.

Louis E. Kearney has recently resigned from the Photo Radio Laboratory of the Radio Corporation of America at New York City to become Radio Inspector for the Port of Philadelphia, replacing John E. Leitch, who is now Chief Engineer for Station WCAU of the Universal Broadcasting Company.

Paul J. Schwarzhaupt, formerly with the Radio Department of Spear and Company of New York City, is now in the Radio Engineering Department of the General Electric Company at Schenectady, N. Y.

Elmer L. Brown has resigned from the California Victor Distributing Company to assume a position with the Engineering Division of the Technical and Test Department of the Radio Corporation of America in New York City.

Leonard F. Fuller, member of the Board of Direction of the Institute in 1928 and formerly Chairman of the San Francisco Section, has resigned from the General Electric Company at San Francisco, where he was engaged in carrier current work and application of high-voltage high-frequency equipment in the light and power industry, to become executive vice-president of the Federal Telegraph Company in charge of that organization's affairs on the west coast. Dr. Fuller's headquarters are at Palo Alto.

Dan T. Fernandez has left the Stewart Warner Speedometer Company of Chicago to become audio engineer for the Day Fan Electric Company of Dayton, Ohio.

L. T. Newell, of the U. S. Lighthouse Service, has been transferred from Washington, D. C. to the Airways Communication Station at Bellefonte, Pa., as operator in charge.

O. H. Caldwell, formerly a member of the Federal Radio Commission, has returned to New York City to resume his position as Editor of *Radio Retailing* and *Electrical Merchandising*.

Robert Hertzberg, for the past two and one-half years managing editor of *Radio News* magazine, has resigned that position to join the staff of the Pilot Electrical Manufacturing Company of Brooklyn.

I. G. Maloff, formerly associated with John Minton at White Plains, N. Y. in consulting work, is now vice-president and in charge of engineering for Valley Appliances, Inc., of Rochester, N. Y.

Carrington H. Stone is now associated with Jenkins and Adair of Chicago, Ill., working on talking motion picture and public address system problems. Mr. Stone was formerly assistant chief engineer of the Radio Division, Stewart-Warner Speedometer Corporation, Chicago, Ill.

Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED MARCH 6, 1929

Transferred to the Fellow Grade

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

Elected to the Fellow Grade

Italy	Rome, Via Tevere No. 20	Pession, Guiseppe

Transferred to the Member Grade

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

Elected 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.

Elected 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 Ave.	Hernfeld, 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 Columbia	Washington, Radio Laboratory, Bureau of Standards	Doherty, William H.
	Washington U. S. Naval Research Laboratory	Wallace, James D.
Florida	Fort Myers, 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	Sosbee, 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 T.
	Wilmette, 1025 Sheridan Road	Freis, J. H.
Louisiana	St. Francis, P. O. Box 232	Harris, David H.
Maine	Houlton, P. O. Box 69	Willes, Emerson T.
	Portland, 56 Lafayette Street	Giro, Daniel J.
Maryland	Baltimore, 1611 Lexington Bldg.	Mathers, Earl S.
Massachusetts	Boston, 160 W. Brookling Street	Nelson, Robert Neil
	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.
Michigan	Ann Arbor, 304 S. Fifth Avenue	Allen, Roy B.
	Ann Arbor, 423 Fuller Street	Embury, Milburn S.
	Ann Arbor, 402 N. Main St.	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.
	Ann Arbor, 1108 Michigan Avenue	Wilber, Harold
	Birmingham, 33 Linden Road	Soeters, Raymond A.
	Detroit, 3403 Cicotte	Tomchuck, John
	Fordson, 5615 Horger Avenue	Knight, Donald M.
	Oxford	Capron, E. S.
Missouri	St. Louis, 3727 Juniata Street	Ruddy, Ralph P.
	University City, 8224 Fairham Avenue	Tevis, Graham L.
New Jersey	East Orange, 5 Eppirt Street	Somers, Richard M.
	Jersey City, 241 Van Vorst Street	Benin, Zolmon
	Keanburg, 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 (cont'd)	New York City, 411 Fifth Avenue	Frank, James, Jr.
	New York City, R. C. A., 70 Van Cortlandt Park So.	Georghegan, 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, 33 West 60th Street	Perryman, G. H.
	New York City, 40 Rector St., c/o British Marine Wireless Service	Veralls, 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	Allentown, 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 Shikwnjo	Namba, Shogo
Scotland	Gourock, Western House	Greenlees, Leslie M.
South Africa	Muizenberg, Cape, Main Road	Gilmour, Peter

Elected 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, Engineer's 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.

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 April 29th. These applicants will be considered by the Board of Direction at its May 1st meeting.

For Election to the Fellow Grade

England	London, SW7, 15 Rutland Mews	Slee, John A.

For Transfer to the Member Grade

Arkansas	Conway, State Teachers' College	Cordrey, E. E.
California	Pasadena, California Institute of Technology	Mackeown, S. S.
Dist. of Columbia	Bellevue, Naval Research Laboratory	Dallin, Edwin B.
	Washington, 6310 5th St. N. W.	Durkee, James D.
	Washington, 3100 Connecticut Ave., Apt. 105	Powell, Edwin L.
Illinois	Chicago, 1917 Warner Ave.	Schoenwolf, F. L.
	Peoria, 147 Cooper Ave	Shalkhauser, Eric George
Massachusetts	Springfield, 162 Springfield St.	Curtis, Leslie F.
New York	Schenectady, 9 N. Ferry St.	Andrews, Paul D.

For Election to the Member Grade

Arkansas	Fayetteville, 143 S. Hill St.	Barton, Loy E.
Dist. of Columbia	Washington, c/o Japanese Military Attaché, Portland Hotel	Kono, T.
Illinois	Chicago, 430 S. Green St.	Frankel, Mortimer
New Jersey	Cranford, 31 Spruce St.	Briggs, Loyd A.
New York	Brooklyn, 1244 East 35th St.	Dreyer, Harry W.
England	Hull, c/o British Broadcasting Corp., 26 Bishop Lane	Calver, Frank N.
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	Newcastle-on-Tyne, 5, Selborne Gardens, Jesmond	Nicholson, James K. A.
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Part II

TECHNICAL PAPERS

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FREQUENCY MULTIPLICATION BY SHOCK EXCITATION*

By

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Summary—*The fundamental principles involved in the theory of frequency multiplication by means of iron-core coupled circuits are briefly reviewed from the standpoint of Fourier analysis as well as that of recurring transients. Oscillograms and figures illustrating the effect of transformer inductance and shock duration upon wave form are given. Oscillations of the first and second kind are discussed in their relation to the analogous arc circuit problem. Efficiency and power output are considered in their dependence upon primary and secondary current amplitude, and the conditions for smoothest wave form and maximum efficiency are pointed out.*

THE method of frequency multiplication discussed in this paper makes use of the fact that if a non-harmonic voltage is impressed upon a simple oscillatory circuit, the response will reach a maximum whenever the natural frequency of the circuit coincides with one of the harmonics contained in the impressed voltage. If the latter is highly distorted, i.e., contains large harmonic components, then the response of the oscillatory circuit will of course also be appreciable. The circuit arrangement which produces this distorted voltage and impresses it upon an oscillatory mesh is given in Fig. 1. The transformer inductance L_t is small compared with L_1. Hence if L_1 and C_1 are tuned to the machine frequency, the primary circuit will be practically in resonance, and the current will be nearly sinusoidal regardless of the saturation effects in the transformer. The latter is so designed that saturation sets in at a current value considerably below the primary maximum, so that the induced voltage is highly distorted. This is the voltage which is impressed upon the secondary tuned circuit, thus giving rise to frequency multiplication.

This arrangement has been investigated both theoretically[1] and experimentally in the past, but the present paper proposes to throw light upon some of the features involved which have not yet been fully appreciated. Some of these are: the effect of the duration of shock upon response and wave form; the production of frequencies which are even multiples of the fundamental; oscillations of the first and second kind and their analogy to the case of frequency multiplication by means of arc circuits; the effect of shock duration upon efficiency; etc.

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A rigorous mathematical theory based upon assumptions elaborated below has been given[1] for this type of circuit, and consequently will not be reproduced in its entirety here. We shall rather limit ourselves to a brief review of the method of attack and some of its consequences for the purpose of gaining for the reader a better orientation in the present picture so that he may follow without difficulty the arguments which are presented to clarify the particular phenomena offered in this paper.

We have called this method of frequency multiplication (following the suggestion of other writers on this subject) the method of *shock excitation*. It might also be called the method of *filtering harmonics*, as will easily be appreciated by the reader from the remarks made in the opening paragraph. The difference in these two ways of designating the same phenomena lies in a fundamental difference in the mathematical method of attack, of course, and has no counterpart in the actual physical mechanism involved. Since each way of looking upon the situation possesses some distinct advantages in the nature of illuminating certain features of the general problem, both will be presented below, at least in fundamental outline.

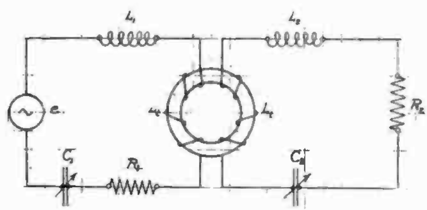

Fig. 1—General Circuit Arrangement.

The basic simplifying assumptions, by which the mathematical attack is made possible, are the same for both methods. First we idealize the magnetization curve as shown in Fig. 2, i.e., the slope is assumed constant between the saturation points, and zero above these. Although this rather strong idealization may seem to depart too radically from actual fact, it nevertheless retains (and even accentuates) those features which are the very essence of our problem, thereby amplifying the points of chief interest. Our second basic assumption is that the primary current varies linearly with time throughout the interval in which the transformer is below saturation. Since the primary current is sinusoidal, and its maximum large compared with i_{sat}, this is nothing more than replacing $\sin x$ by x for small values of the latter.

[1] "Zur Theorie der Frequenzvervielfachung durch Eisenkernkoppelung," *Archiv. für Elektrotechnik*, Band XVII, 1926, Heft 1, S. 17.

Our first assumption obviously causes the induced voltage in the transformer to be zero unless the flux is below the saturation value, and the second assumption gives this induced voltage a constant value during the period below saturation. This induced voltage therefore, which is the internally generated voltage for the secondary circuit, takes the

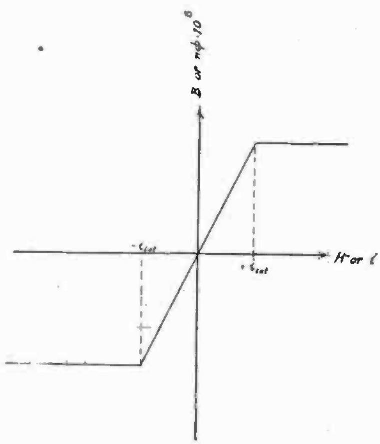

Fig. 2—Idealized B-H Relation.

form shown in Fig. 3. It is a periodic succession of rectangular impulses with the same period as that of the generator in the primary circuit.

This internally generated voltage is impressed upon a very peculiar circuit, however, and it is just this peculiarity which accounts for numerous anomalies to be pointed out later; namely, the inductance of the secondary circuit is not a constant, but equal to L_2+L_t so long as

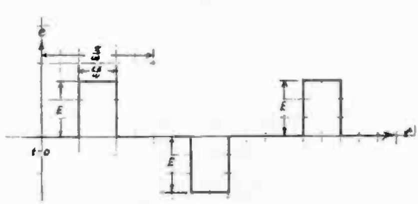

Fig. 3—Resulting Induced Transformer Voltage Plotted with Symmetrical Time Origin for Simplification of Fourier Expansion.

the transformer is below saturation, and L_2 above saturation, because here L_t is zero due to the zero slope of the magnetization curve. These discontinuities in the secondary circuit inductance coincide in time with those of the voltage of Fig. 3. From the mathematical standpoint they complicate the situation exceedingly. From the physical standpoint they cause the secondary circuit to have a periodically-changing natural frequency and hence a bad wave form. In order to get a fairly smooth secondary wave, L_2 should be large compared with L_t—as large as the rest of the design will allow. Incidentally this large secondary inductance reduces the damping constant also, and improves the wave form due to somewhat suppressed attenuation, as will be seen later. In order to gain a first approximation to our solution, let us neglect L_t in the secondary for the moment and assume the voltage of Fig. 3 impressed upon a circuit of constant R, L, and C. After having familiarized ourselves with this simpler problem, we can then introduce the discontinuous secondary inductance and discuss its general effect.

So far we have said nothing at all about method of attack. The first of these, which suggests itself immediately upon glancing at the voltage of Fig. 3, is to represent this voltage by a Fourier series and operate with it as the impressed electromotive force upon the secondary oscillatory circuit with R_2, L_2, and C_2. Viewed from this angle, the method of frequency multiplication becomes one of filtering out harmonics contained in the distorted voltage wave.

There is, however, another very logical method of attack which suggests itself from the analogy between an electrical oscillatory circuit and the mechanical pendulum. If we were to subject an ordinary pendulum to periodically repeated pushes of short duration, we could obviously keep the system in motion in spite of its attenuation resulting from frictional forces. If the direction and frequency of the pushes or shocks were judiciously chosen with regard to the pendulum's own natural period, then the resulting amplitude in the steady state could reach large values. If, on the other hand, the frequency and direction of the repeated shocks were injudiciously chosen, the result would be one of misdirected effort and the amplitude attainable would be correspondingly small.

In an electrical oscillatory circuit a suddenly applied force produces damped oscillations. The same are produced by a suddenly relinquished force. We call these *transient effects*. If we repeat the same applications periodically we shall eventually arrive at a steady-state phenomenon which consists of a series of periodically recurring transients. Viewed from this angle, the present problem of frequency multiplication appears to be due to *shock excitation*, which is probably the term more commonly used to designate this method.

Having thus oriented the reader in our problem, we shall proceed to sketch the mathematics involved in the two methods of attack given above.

Using the notation illustrated in Fig. 3, we assume for the impressed voltage the following Fourier series in complex form:

$$e(t) = \sum_{-\infty}^{+\infty} {}_k\, a_k e^{jk\omega t}, \tag{1}$$

in which $e(t)$ is the instantaneous secondary voltage at time t. Taking theorigin of time as indicated in the figure, we findaccording to Fourier:

$$a_k = \frac{2jE \sin k\epsilon \dfrac{\pi}{2}}{\pi k} \cdot (-1)^{(k+1)/2} \tag{2}$$

with k restricted to odd integer values. We shall call ϵ the *breadth factor*. It varies between zero and unity as the breadth of the shock varies between zero and a half-period of the fundamental. Putting $\epsilon = 1$ in (2) we get:

$$(a_k)_{\epsilon=1} = \frac{-2jE}{\pi k},$$

which is the well-known coefficient for the square wave. The differential equation for the secondary circuit is

$$L_2 \frac{di}{dt} + R_2 i + \frac{1}{C_2} \int i dt = e(t). \tag{3}$$

The steady-state solution (particular integral) must be of the form

$$i = \sum_{-\infty}^{+\infty} {}_k\, b_k e^{jk\omega t}. \tag{4}$$

Substituting (1), (2), and (4) into (3) and equating coefficients of like frequency, we see that

$$b_k = \frac{a_k}{\Omega_k} \tag{5}$$

where

$$\Omega_k = R_2 + j \left\{ L_2 k\omega - \frac{1}{C_2 k\omega} \right\}. \tag{6}$$

The Fourier solution for this simple case is thus complete. Equation (6) we recognize as being a resonance denominator. Resonance to a given harmonic occurs when

$$L_2 k\omega = \frac{1}{C_2 k\omega}.$$

If we define

$$q_{20}^2 = \frac{1}{L_2 C_2} \tag{7}$$

as the squared undamped secondary angular velocity, and

$$n_0 = \frac{q_{20}}{\omega} \tag{8}$$

as the undamped frequency multiplication, then the resonance condition obviously becomes

$$n_0 = k. \tag{9}$$

Hence we should get a maximum response in the secondary whenever it is tuned to an odd multiple of the primary, the resonance amplitude being given by

$$\frac{a_k}{R_2}.$$

For practical comparisons it is necessary to find the effective value of the secondary current. We have by the usual definition:

$$i_{\text{eff}}^2 = \frac{\omega}{2\pi} \int_0^{2\pi/\omega} i^2 dt. \tag{10}$$

In forming the square of the current we shall use (4) twice, with different summation indices, to avoid confusion. We get

$$i_{\text{eff}}^2 = \frac{\omega}{2\pi} \int_0^{2\pi/\omega} \sum_{-\infty}^{+\infty}{}_k b_k e^{jk\omega t} \cdot \sum_{-\infty}^{+\infty}{}_s b_s e^{jk\omega t} dt$$

$$= \frac{\omega}{2\pi} \sum_{-\infty}^{+\infty}{}_{ks} b_k b_s \int_0^{2\pi/\omega} e^{j(k+s)\omega t} dt.$$

Here all integrals vanish except those for which $k = -s$. Hence we get

$$i_{\text{eff}}^2 = \sum_{-\infty}^{+\infty}{}_k b_k b_{-k}. \tag{11}$$

But b_{-k} is the conjugate of b_k so that

$$b_k b_{-k} = |b_k|^2,$$

and thus

$$i_{\text{eff}}^2 = 2 \sum_1^{\infty}{}_k |b_k|^2. \tag{12}$$

But from the preceding:

$$|b_k|^2 = \frac{4E^2 \sin^2 k\epsilon \dfrac{\pi}{2}}{\pi^2 k^2 (R_2^2 + X_k^2)} \tag{13}$$

with the substitution

$$X_k = L_2 k\omega - \frac{1}{C_2 k\omega}. \tag{14}$$

Hence

$$i_{eff}^2 = \frac{8E^2}{\pi^2} \sum_{1,3,5\ldots}^{\infty} {}_k \frac{\sin^2 k\epsilon\frac{\pi}{2}}{k^2(R_2^2 + X_k^2)}. \tag{15}$$

The above summation can be carried out and the expression for the effective current put into a closed form which is more useful for purposes of calculation than the infinite sum. We shall close our discussion of the Fourier method of attack with this brief outline and pass on to the more interesting shock-excitation method.

This method we shall also merely sketch briefly here. It is discussed in detail in the paper already cited. Let us review the simple transient problem involved. We know that the damped oscillations which take place in a circuit with R_2, L_2, C_2 in the absence of driving forces may be represented analytically by the form:

$$i_b = Im\{Be^{mt}\} \tag{16}$$

where Im denotes that the imaginary portion be taken, B is a complex amplitude which takes care of phase relations, and

$$\left.\begin{aligned} m &= -\alpha + jq_2 \\ \alpha &= \frac{R_2}{2L_2} \\ q_2 &= \sqrt{q_0^2 - \alpha^2} \\ q_0^2 &= \frac{1}{L_2 C_2} \end{aligned}\right\} \tag{17}$$

Equation (16) includes the rest condition for which $B=0$. If a constant emf is impressed, or more generally, if a constant impressed force is suddenly changed by an amount ΔE, then the form of the oscillation is unaffected—its amplitude merely changes in magnitude and phase. The frequency and rate of damping must of necessity remain unaltered.

Suppose now that our circuit is oscillating according to (16) before a certain instant which we choose to call $t=0$. At $t=0$ we suddenly change the impressed force by ΔE. After $t=0$, the oscillation will be given by

$$i_a = Im\{Ae^{mt}\}. \tag{18}$$

At the instant $t=0$ we have to satisfy two conditions. First, the current must be continuous. This gives us

$$Im\{B\}=Im\{A\},$$

or if we write

$$\left.\begin{aligned} B&=B_1+jB_2\\ A&=A_1+jA_2\end{aligned}\right\} \tag{19}$$

we get

$$A_{20}=B_{20} \tag{20}$$

the subscript zero relating to the instant $t=0$.

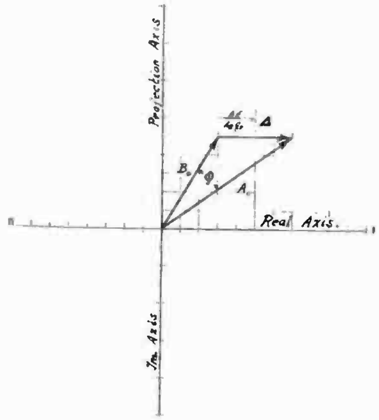

Fig. 4—Vector Relations in the Gaussian Plane for the Shock-Excitation Method of Attack.

Secondly, the discontinuity in the induced voltage must equal ΔE, i.e.

$$\left(\frac{di_a}{dt}\right)_{t=0}-\left(\frac{di_b}{dt}\right)_{t=0}=\frac{\Delta E}{L_2}.$$

Substitution gives

$$A_{10}=B_{10}+\frac{\Delta E}{L_2q_2}. \tag{21}$$

Combining (20) and (21) we have

$$A_0=B_0+\frac{\Delta E}{L_2q_2}. \tag{22}$$

Hence the amplitude after the disturbance equals the amplitude before the disturbance plus a real quantity which is proportional to the magnitude and direction of the sudden change in the constant applied force. Equation (22) is illustrated in the Gaussian plane by Fig. 4.

The instantaneous value of current equals the projection of these rotating and decaying vectors upon the vertical axis. At the instant ΔE occurs the current has some value. It cannot change instantly, and hence the instantaneous change in the vector amplitude can be only in a horizontal direction as shown. This will give rise to an instantaneous phase shift ϕ, which will be zero only when B_0 is in a

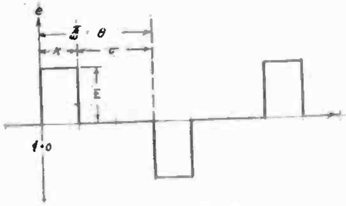

Fig. 5—Induced Transformer Voltage with Time Origin Chosen to Fit Shock-Excitation Method.

horizontal position. It is easy to see also that this phase shift may be in either direction, depending upon the direction (algebraic sign) of ΔE. Also A_0 need not always be larger than B_0 as a result of a suddenly applied (positive ΔE) emf. The amplitude can equally well be increased by suddenly removing an emf provided this is done at the proper instant. We shall let the reader illustrate these points for himself.

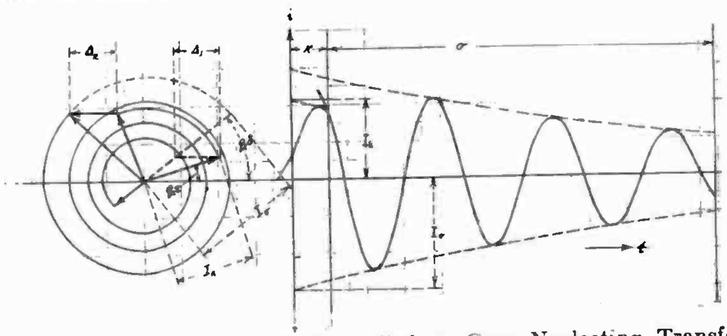

Fig. 6—Shock-Excitation Method Applied to Case Neglecting Transformer Inductance.
ϵ is chosen equal to 0.075; frequency multiplication equal to 7.0.

Our problem of handling the applied voltage of Fig. 3 for the oscillatory circuit R_2, L_2, C_2 is now very simple. During each fundamental period we have four sudden changes in voltage to consider, viz: The voltage jumps from zero to a constant value E and remains there for the duration of the shock; at the end of the shock duration it suddenly drops to zero and remains there until the advent of the negative shock; at this time it suddenly jumps to the constant negative value $-E$; and finally at the end of the negative shock it again drops back to zero and remains there until the beginning of the next fundamental period when the same sequence recurs.

We shall determine those amplitudes in magnitude and phase which obtain in the steady state. Here obviously the negative fundamental half-period must be an exact negative duplicate of the preceding positive fundamental half-period. Hence only one fundamental half-period need be considered. This half-period we shall divide into two parts. The duration of the shock we shall call the "coupling-period," because actually during this time the primary and secondary meshes are coupled with each other by the transformer mutual inductance. The remainder

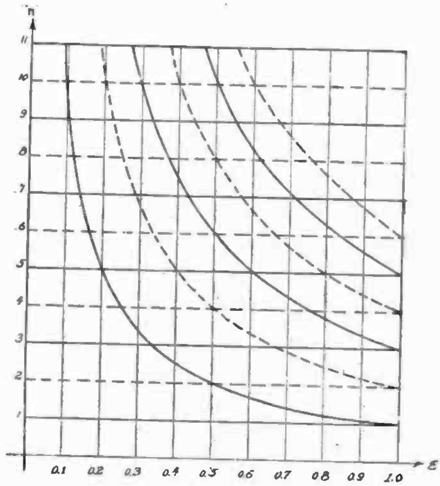

Fig. 7—Theoretical Relation between n and ϵ for Case of Neglected Transformer Inductance.
Intersections of full lines give rise to maxima, those of dotted lines to minima.

of the half-period we shall call the "period of free oscillation," because here the secondary circuit is virtually free to oscillate by itself due to the fact that the transformer is above saturation and its mutual inductance zero (according to our idealization of the B-H curve).

The steady state will be a succession of damped sinusoids whose amplitudes and phases change discontinuously at the boundaries of the coupling period and the period of free oscillation. Let us take the beginning of a positive shock as a time reference, i.e., as a reference for amplitudes and phases of the secondary current. Then if we denote the coupling period by the subscript k, and the period of free oscillation by the subscript σ, the periodicity of the steady state demands that at the beginning of the shock

$$-I_\sigma e^{m\theta}+\frac{E}{L_2q_2}=I_k, \tag{23}$$

and at the end of the shock

$$I_k e^{mk}-\frac{E}{L_2q_2}=I_\sigma e^{mk} \tag{24}$$

where the quantities are as defined by (17) or illustrated by Fig. 5. This pair of equations is easily solved for the amplitudes giving:

$$\left.\begin{aligned} I_k &= \frac{E}{L_2q_2}\,\frac{1+e^{m\sigma}}{1+e^{m\theta}} \\ I_\sigma &= \frac{E}{L_2q_2}\,\frac{1-e^{-mk}}{1+e^{m\theta}} \end{aligned}\right\} \tag{25}$$

These amplitudes are of course complex and can be written

$$\left.\begin{aligned} I_k &= |I_k|\, e^{jq_2\tau} \\ I_\sigma &= |I_\sigma|\, e^{jq_2\delta} \end{aligned}\right\} \tag{26}$$

where τ and δ are the time phases.

The pair of equations (25) completely solve the case for the type of voltage illustrated by Fig. 5 applied to an oscillatory circuit with constants R_2, L_2, and C_2. Fig. 6 serves to illustrate the application of these equations to an arbitrarily chosen case with a breadth factor $\epsilon=\frac{k}{\theta}=0.075$ and a frequency multiplication $n=\frac{q_2}{\omega}=7$. The scale is immaterial since we are not interested in quantitative values here. The quantities Δ_1 and Δ_2 are the sudden changes in current amplitude. At the left we show the diagram of rotating and decaying vectors; at the right, the corresponding time plot.

It is interesting to note that the sudden drop in voltage is as effective in increasing the amplitude as the sudden rise, and that the discontinuities in the applied voltage alone produce an increase in amplitude. Whether the discontinuities produce amplitude changes (Δ_1 and Δ_2) which increase or decrease the absolute magnitude of the current depends upon the phases $q_2\tau$ and $q_2\delta$ as can readily be seen from the figure. These phases naturally depend upon the values of n and ϵ, so that it is evident that the question of a favorable or unfavorable result will depend upon these two parameters. Fig. 7 illustrates this dependence.[2] The full lines indicate favorable and the dotted lines unfavorable conditions. The combinations of n and ϵ, corresponding

[2] For the derivation of the equations involved here and for further illustrative examples we refer the reader to the reference given above from which the figures given here are taken.

to the intersection of two full lines, give rise to maximum amplitudes, and those corresponding to the intersection of two dotted lines give rise

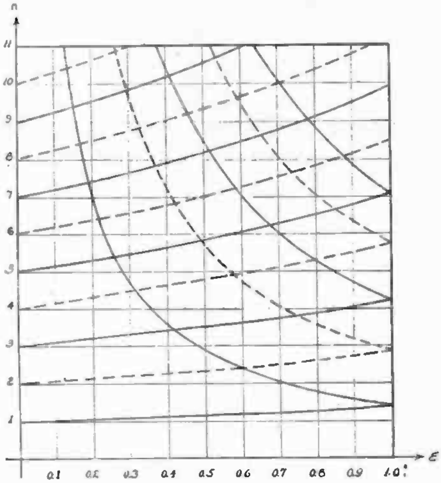

Fig. 8—Theoretical Relation between n and ϵ Taking into Consideration Transformer Inductance.
Full and dotted lines have same significance as for Fig. 7.

to minimum amplitudes. If it were possible to show a third (vertical) co-ordinate for secondary r.m.s. current, then we should have a relief

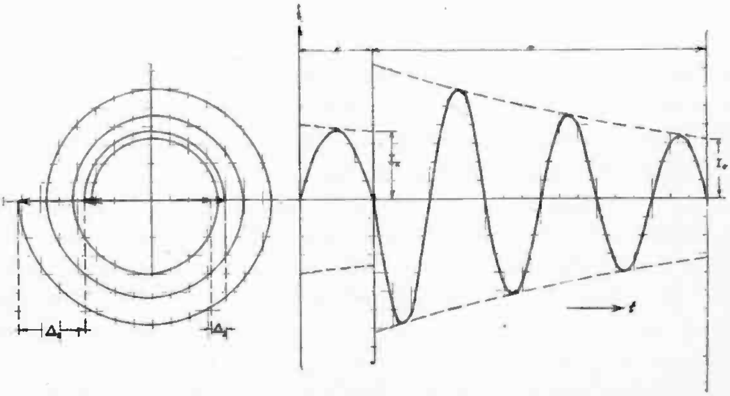

Fig. 9—Shock-Excitation Method Applied to Case in Which Transformer Inductance is Considered.
Condition for maximum secondary response with $n=7.414$ and $\epsilon=0.19$.

map, the peaks of which would correspond to current maxima, and the hollows to current minima. The intersections of a dotted with a full line would correspond to a saddle point or "ridge." Cross-sections of this relief map would show the current variation for constant n and varying ϵ or constant ϵ and varying n.

An investigation of such cross-sections is interesting from the following point of view. If we go back to the standpoint of filtering harmonics then we should expect maxima to occur whenever the secondary is tuned to an odd multiple of the primary, since the Fourier analysis of the impressed voltage contains only odd harmonics. However, although this is true, we are very likely to overlook the importance of the effect of ϵ upon the magnitude of the Fourier coefficients. For instance, with $\epsilon = 2/3$ and $n = 3$ we should get a very small secondary current. With a given circuit and fixed shock duration, therefore, we should expect to find that certain odd harmonics in the secondary are very much suppressed as a result of an unfavorable ϵ.

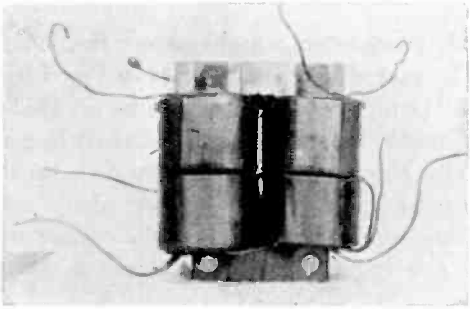

Fig. 10—Transformer Used in Experimental Work.

The picture of the actual state of affairs is still incomplete since we have thus far overlooked the fact that the total secondary inductance is *not constant*. For the shock duration it is equal to the external secondary inductance plus the secondary transformer inductance, and for the remainder of the fundamental half-period it is equal to the external secondary inductance alone. This additional feature alters the situation quite considerably and gives rise to the anomalies which the present experimental investigation proposed to illustrate.

The method of treating the problem from the standpoint of recurring transients is still applicable to the actual problem with non-constant secondary inductance, but involves—in addition to the transient set up by the sudden voltage change—also a transient due to the simultaneous inductance change. The effects of these transients superpose to form the net effect.

The application of the result is the same as for the simple case just treated, and in vector form has the same appearance as shown by Fig. 4.

The only difference is that the sudden amplitude change Δ depends not only upon ΔE but also upon the change of inductance.[3] Also the magnitude of Δ will not be the same at the beginning as at the end of the shock. We shall not take the time to go into details here, but prefer to show the theoretical results and compare them with the experimental results found in this investigation.

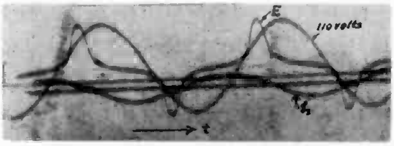

Fig. 11—Primary Current and Transformer Induced Voltage with 110 volts, 60 cycles impressed. Primary current 2 amperes.

Here again the parameters n and ϵ govern the occurrence of maxima and minima. The plot which corresponds to Fig. 7 for the simple case is given in Fig. 8. Comparing Figs. 7 and 8 we see that now the maxima do not occur for odd values of n, but for slightly larger values depending upon ϵ. Why this should be the case can be seen from Fig. 9, which illustrates the application to a case where maxima conditions are fulfilled. Here we see the effect of the higher inductance during the shock very clearly, in that the half-period there is considerably

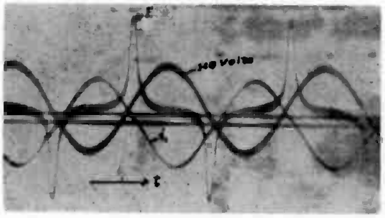

Fig. 12—Primary Current and Transformer Induced Voltage with 110 volts, 60 cycles impressed. Primary current 7 amperes.

longer than for the rest of the fundamental half-period. Also the amplitude is smaller, as it should be, considering that the same inductive countervoltage is here produced with a larger total inductance. Now, since the frequency multiplication n is taken as the ratio of the frequency during the period of free oscillation to the fundamental frequency, it is quite clear that this ratio must now be somewhat larger than an odd number of half cycles. This is evident from Fig. 9 and needs no further comment.

[3] For the derivation of this case see the reference above.

As a matter of fact it becomes rather difficult to define the frequency multiplication accurately, since there are actually three different ways in which this may be done. We could take the ratio of the frequency either during shock or during the remaining fundamental half-period to the fundamental frequency, or an average of these two values as that figure which is to define the frequency multiplication of the circuit as a whole. However, since the time of free oscillation greatly predominates over the duration of shock, and since in most practical circuits the secondary inductance of the transformer is small compared with the external inductance, it seems most reasonable to call the free oscillation frequency the secondary frequency and base the frequency multiplication upon it.

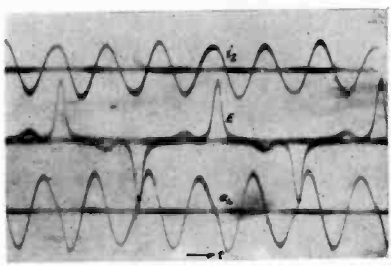

Fig. 13—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Third Harmonic.
Primary current 5.5 amperes, secondary current 2.35 amperes. Condition of oscillations of the "second kind." Efficiency 51.2 per cent.

With this brief review of the fundamental principles involved, we are prepared to consider some of the experimental results obtained and discuss them in the light of the theory.

The circuit used was exactly that shown in Fig. 1. A 60-cycle alternator was used as a primary source and the primary inductance was of the iron-core type, but so dimensioned that its core remained well below saturation throughout the cycle. The transformer, shown in Fig. 10, was of the closed magnetic circuit type, and provided with four windings with taps so that various ratios could be tried out. To all of the following figures and discussions a 1 to 1 ratio applies.

Since it was desired to vary the duration of shock or the parameter ϵ, the transformer was first provided with a variable air gap for this purpose. This method proved very ineffective, however, and ϵ was therefore varied by means of the primary current amplitude alone. That this is possible may easily be realized from the fact that with a small primary amplitude the transformer core remains below saturation for a longer time that it does for a large primary amplitude. This is illustrated by the oscillogram Figs. 11 and 12, which were taken with primary currents of two and seven amperes respectively. This method of varying the shock duration had the disadvantage of varying the induced voltage maximum at the same time. This fact could, however, easily be taken into account in interpreting the results, so that the method was still perfectly suitable for comparative work.

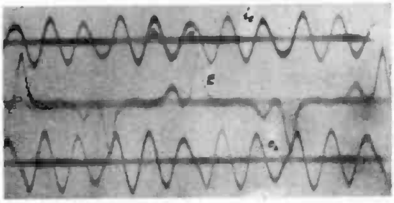

Fig. 14—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Fifth Harmonic.
Primary current 5.0 amperes, secondary current 1.47 amperes. Condition of oscillations of the "second kind."

How the secondary period is lengthened during shock, by reason of the higher total secondary inductance, is shown by oscillogram Figs. 13, 14, and 15. Fig. 13 is taken with the secondary tuned to the third harmonic. At the top is the secondary current, in the center the secondary terminal voltage, and at the bottom the voltage induced in the secondary external inductance. Here we see that the half-period

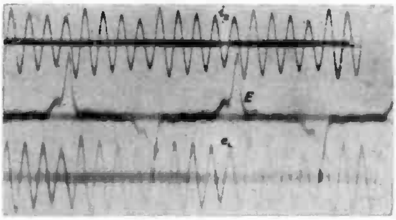

Fig. 15—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Ninth Harmonic.
Primary current 8 amperes, secondary current 0.8 ampere. Condition of oscillations of the "second kind."

corresponding to the shock duration (total secondary inductance equal to transformer secondary plus external) is appreciably longer than the following two periods which correspond to free oscillation of the secondary (total secondary inductance equal to external only). The same effect is illustrated for the fifth harmonic in Fig. 14 and for the ninth harmonic in Fig. 15. We shall refer again to these figures in connection with secondary voltage distortion.

In Fig. 9 we have shown a theoretical condition for maximum secondary response. This condition requires that the secondary half-period during shock should equal the duration of the shock, and that the remaining fundamental half-period should be just long enough to accommodate an even number of half-periods of free oscillation. In Fig. 16 we show an oscillogram which was taken with these conditions fulfilled for the seventh harmonic. The shock duration is just right, the result is smooth and regular, and the efficiency and secondary current are maxima. Incidentally the smoothness of the current can best be recognized by an inspection of the voltage induced on the secondary external inductance, since this voltage magnifies any irregularities that may be present in the current.

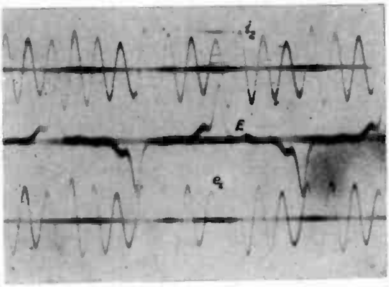

Fig. 16—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Seventh Harmonic.
Primary current 6 amperes, secondary current 0.88 ampere. Condition of oscillations of the "second kind," and shock duration adjusted for smoothest wave form. Note this in the inductance voltage.

What happens when the shock duration is too short is illustrated by Fig. 17, which was taken with the secondary tuned to the third harmonic. A shock duration which is too short need not necessarily give rise to these irregularities, however, as may be seen in Fig. 13, which is taken for exactly the same primary conditions and hence for the same shock duration on the basis of no-load conditions. These features are intimately tied up with the question of oscillations of the *first and second kind*.

In order to discuss this point adequately it is necessary to recall some well-known facts about frequency multiplication by means of the a.c. arc. The frequency multiplication circuit for the arc is identical with that shown in Fig. 1 except that the transformer is replaced by an arc. We know from our arc circuit theory that the voltage produced across the arc by an alternating current flowing through it is substantially a square wave. This square-wave voltage is then impressed upon the secondary oscillatory circuit which responds to the harmonics contained in it. This is the same situation as for the case of frequency multiplication by inductive means with the special condition of a breadth factor of unity or a shock duration of half a fundamental period, assuming of course that a breadth factor of unity were pos-

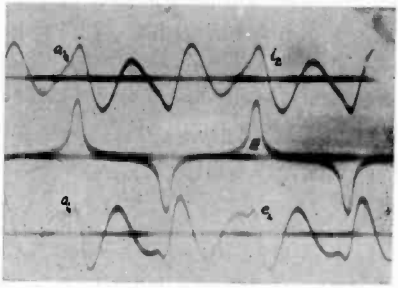

Fig. 17—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Third Harmonic.
Primary current 5.5 amperes, secondary current 0.54 ampere. Condition of oscillations of the "first kind" and shock duration too short for smoothest wave form. Note irregularities in inductance voltage. Efficiency 28 per cent. Compare with Fig. 13 for same primary conditions but operation within region of oscillations of the "second kind."

sible with the transformer method. The Fourier treatment is identical for these two problems.

From this similarity it is to be expected that other peculiarities found in the arc circuit should be recognizable in the magnetic circuit

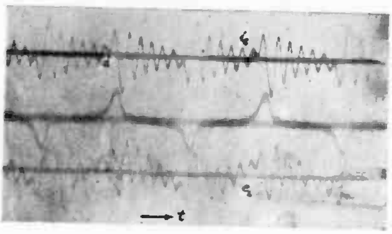

Fig. 18—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Twelfth Harmonic.
Primary current 3 amperes, secondary current too small to read. Phase reversal at point *a*. Phase reversals must occur when secondary is tuned to even harmonics.

also. Now we know that in the arc circuit we may obtain oscillations in the secondary for which the arc current reverses only once for each fundamental half-period. Such oscillations are known as oscillations of the *first kind*. If the secondary current amplitude exceeds a certain limit, then the arc current reverses *twice* in each fundamental half-period and the result is called oscillations of the *second kind*. If the arc current reverses *three times*, we get oscillations of the *third kind*, etc.

An exactly analogous phenomenon takes place in the magnetic circuit discussed in this paper. Instead of *arc current*, however, we have to consider the *transformer flux*. If n times the secondary current amplitude is less than or equal to the primary current amplitude (n being the frequency multiplication and the transformer ratio unity), then the flux in the transformer core reverses only once for each fundamental half-period and the resulting secondary oscillations should be called oscillations of the first kind. If the secondary amplitude exceeds this value, then the transformer flux reverses twice in each fundamental half-period, and the result should be termed oscillations of the *second kind* by analogy to the arc circuit conditions.

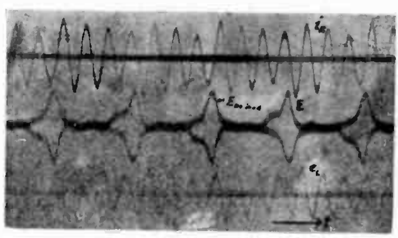

Fig. 19—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Seventh Harmonic.
Primary current 3 amperes, secondary current 0.46 ampere. Oscillations just within region of "second kind." Transformer voltage shown under load and no-load conditions to emphasize distortion.

Without going into more details we can now discuss the conditions shown in Figs. 13 and 17. Both are for a frequency multiplication of three and a primary current of 5.5 amperes. Fig. 17 is taken for a secondary current of 0.54 ampere, while Fig. 13 is taken for a secondary current of 2.35 amperes. Hence Fig. 13 represents oscillations of the second kind, and Fig. 17 oscillations of the first kind. This can also be seen directly by examining the transformer voltage. In Fig. 17 it is smooth except for the peak, while in Fig. 13 it contains an additional hump. This additional hump means an additional reversal of the transformer flux, and hence oscillations of the second kind. Oscillations of the second kind occur for low secondary resistance and low attenuation. The effect produced is to broaden the voltage shock at the base and thus avoid, to a certain extent, the appearance of secondary current irregularities. Unless the shock duration is properly adjusted, therefore, it is desirable from the standpoint of smoothness to operate in the region of oscillations of the second kind. We shall later show that maximum efficiency is also obtained at the point where oscillations of the second kind begin, so that this should be the operating condition sought after. Figs. 14 and 15 also show conditions where oscillations of the second kind were obtained.

We have one other interesting point to illustrate by oscillograms, and that is in connection with the production of even harmonics in

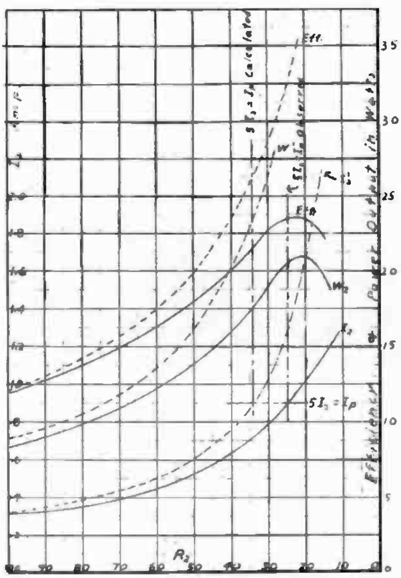

Fig. 20—$N=5$. Comparison of Observed and Calculated Results.
———— Observed
- - - - - - - Calculated
$I_p=4.5$ amperes

the secondary. In order to be able to recognize them as such, it will be necessary to recall the following: it is quite clear that the secondary current during a negative fundamental half-period must be an exact negative duplicate of the preceding positive fundamental half-period, since there is no reason for positive and negative dissymmetry. Hence it follows that in each fundamental half-period there be *exactly* an odd number of secondary half-periods—including the sudden phase shifts as illustrated by ϕ in Fig. 4. If, therefore, we tune the secondary to an even multiple of the primary, then the sum of the sudden phase shifts during any fundamental half-period must necessarily equal π radians or be equivalent to one secondary half-period. It is this phase shift which makes up the discrepancy between the number of secondary half-periods actually accomplished by the current and the necessary odd number demanded by the symmetry of the steady state.

If the shock duration is adjusted so that it is equal to an odd number of secondary half-periods below saturation, and the secondary is tuned to an even multiple of the fundamental, then this phase shift of π radians will occur at *one point* and result in a *phase reversal* of the secondary current. In the oscillogram Fig. 18 we show such a condition for the secondary tuned to the twelfth harmonic. The point

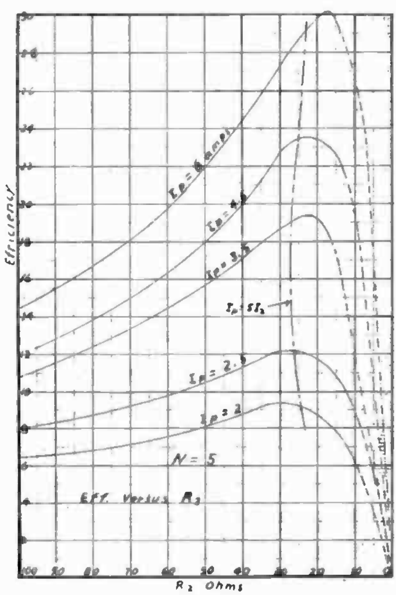

Fig. 21—Efficiency vs. Load Resistance for Various Primary Currents with $N=5$.

of reversal is marked at *a*. The shock duration is approximately equal to one-quarter of the fundamental half-period.

The question of transformer secondary terminal voltage distortion has already been discussed to some extent in connection with oscillations of the second kind. The oscillogram, Fig. 19, serves, however, to illustrate this still more clearly. Here we have a frequency multiplication of seven and a condition of operation which is just within the region of oscillations of the second kind. Both the transformer no-load and load voltage are shown. They were taken in opposite phase positions so that the picture would be clear and easy to read. The effect of the secondary current is to broaden the base of the shock and make it more pointed. The reason for this can easily be shown by drawing in the flux curve by means of the magnetization curve.

A large amount of data were also taken on power output and efficiency vs. frequency multiplication, primary, and secondary current. A few of the more interesting curves will be given here. Fig. 20, for instance, shows the secondary current, power output, and efficiency as functions of the secondary load resistance for a frequency multiplication of five and a primary current of 4.5 amperes. The broken line is drawn in at the dividing line between oscillations of the first and second kind. It is interesting to note that maximum efficiency coincides with maximum power output and that these maxima occur just within the region of oscillations of the second kind. This latter point is fortunate, since we have already seen that this operating condition also gives us the smoothest secondary wave form.

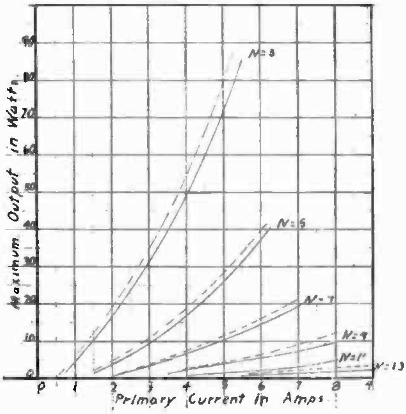

Fig. 22—Curves of Maximum Output vs. Primary Current for Constant Ratios of Frequency Multiplication.
——————— with 120 turns on secondary
- - - - - - - with 160 turns on secondary

Fig. 21 shows the efficiency as a function of the load resistance for various values of primary current and a multiplication of five. Again the broken line indicates the division between oscillations of the first and second kind. Here it is interesting to note that the maximum efficiency increases with increasing primary current. How long this would continue we were unfortunately unable to determine, since it was not possible to go to higher primary currents without danger of primary condenser and inductance insulation breakdown.

Figs. 22 and 23 show the maximum power output and efficiency as functions of the primary current for various values of frequency multiplication. They show quite conclusively that this method rapidly becomes very inefficient for high values of multiplication. Thus it is clear that a multiplication of nine by two successive stages of three would be considerably more efficient than by one single stage. This question of multistage frequency multiplication will be discussed more thoroughly in a following paper on multiphase and multistage methods of extending the present fundamental principle.[4]

In conclusion we may state that this method of frequency multiplication is useful only where perfect wave form is not essential. The wave form is improved by making the external secondary inductance large compared with that of the transformer, and by operating just within the region of oscillations of the second kind as defined in this paper. Since the efficiency increases with increased transformer inductance, it will be a matter of judgment as to how far to go in the relative magnitude of external to transformer inductance. In the

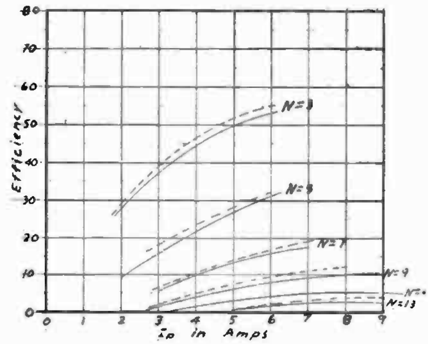

Fig. 23—Curves of Maximum Efficiency vs. Primary Current.
——————— with 120 turns on secondary
– – – – – – – with 160 turns on secondary

theoretical Fig. 9 this ratio was taken as unity, and this figure thus gives us a good idea of what such a ratio does to distort the secondary current. In the cases for which the oscillogram figures are here given the ratio of external inductance to total was about 0.7. The resulting distortion on this account is certainly not excessive. For further data as well as many interesting oscillograms and figures relative to this method of frequency multiplication we refer the reader to the complete report[5] of this investigation, which is filed in the library of the Massachusetts Institute of Technology, and of which this paper is merely an abstract.

[4] To appear in a forthcoming issue of the PROCEEDINGS.

[5] P. T. Rumsey, "Experimental Investigation of Frequency Multiplication by Means of Iron-Core Coupled Circuits." M.I.T. thesis.

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ON THE SHORT-WAVE LIMIT OF MAGNETRON OSCILLATIONS*

By
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Summary—*In this paper the theoretical considerations and experimental results pertaining to the short-wave limit of magnetron oscillations are given.*

Oscillations of only a few centimeters wavelength were successfully produced.

INTRODUCTION

FOR several years research work on the generation of very short waves has been carried on by Professor Yagi and others in our laboratory. While engaged in this work, the writer found it possible to produce very short wavelength oscillations with magnetrons of special form.

Certain of the details of this work have been published.[1] Since then the experiments have been extended and oscillations can now be produced with increased intensity and with even shorter wavelengths. Only the method and prospect of obtaining shorter wavelengths are discussed in this paper. The details regarding the means of increasing the intensity of the oscillations will be dealt with in another paper.

1. THEORETICAL CONSIDERATIONS

The distribution of the space charge in the magnetron is quite similar to that in the triode when the latter is connected to produce oscillations similar to those observed by Barkhausen and Kurz. By considering a "virtual cathode" and a "virtual grid," and an equivalent anode potential, the existence of a negative resistance can be proved in the same way as has been done by Tonks[2] in the case of the Barkhausen and Kurz type of oscillations. Such a negative resistance characteristic gives rise to a state of instability. Sometimes oscillations may be due to a dynatron characteristic and at other times oscillations may be due to the dispersion or discharge of the space charge.

It seems very likely that oscillations in the magnetron occur in the following manner:

* Dewey decimal classification: R133. Original manuscript received by the Institute, October 5, 1928. Revised copy received, November 7, 1928.

[1] Hidetsugu Yagi, "Beam Transmission of Ultra Short Waves," PROC. I. R. E., 16, 715; June, 1928.

[2] Lewi Tonks, *Phys. Rev.*, October, 1927.

When the magnetic field is greater than the critical value, the anode current is cut off and there is an accumulation of charges near the anode and cathode. The accumulation of charges near the anode is the true seat of oscillatory phenomenon.

As soon as the conditions become such that a negative resistance is possible, the accumulated charges begin to be dispersed or discharged and the state of negative resistance disappears until a further accumulation of charges occurs. The time required for the accumulation of these charges after the preceding discharge has occurred is approximately equal to the time, t, which will be required by an electron in travelling from the cathode to the region near the anode where the charge density is greatest. This process of accumulation and discharge will be repeated and cause oscillations.

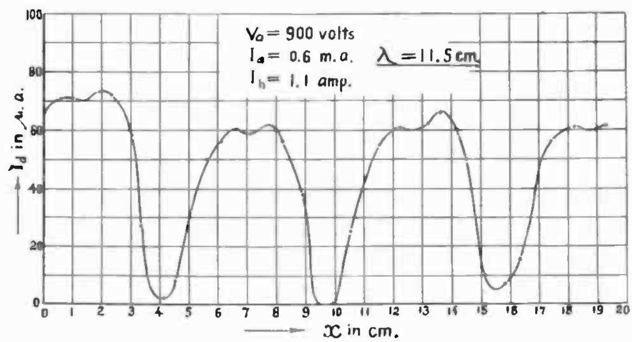

Fig. 1

The period T, which corresponds to one cycle of oscillation will be

$$T = t + t'$$

where t' represents the time required for the electronic discharge. We have no means at present for estimating the value of t'. The value t, however, may be calculated.

Since the phenomenon is alternating, it may not be unreasonable to assume that t' is approximately equal to t. Then the wavelength may be calculated by the following semi-theoretical formula,

$$\lambda_0 = 2ct \tag{1}$$

where c represents the velocity of light.

The time t can be obtained from a graphical solution of the following equation:

$$\frac{dt}{dr}=\frac{1}{\sqrt{\frac{2eV_a}{m}\frac{\log\frac{r}{r_o}}{\log\frac{r_a}{r_o}}-\frac{H^2e^2r^2}{4m^2c^2}\left(1-\frac{r_o^2}{r^2}\right)^2}} \tag{2}$$

where t, r, r_o, r_a, V_a, and H represent the time, the distance from the central axis, the radius of the cathode, the radius of the anode cylinder, the anode voltage, and the intensity of the magnetic field, respectively. e and m are the charge and the mass of an electron. c denotes the velocity of light as in the case of (1).

For the sake of simplicity, let us now assume that r_o is extremely small; then (2) becomes the following:

$$\frac{dt}{dr}=\frac{1}{\sqrt{\frac{2eV_a}{m}-\frac{H^2e^2}{4m^2c^2}r^2}}. \tag{3}$$

The graphical solution of (3) shows that the assumption introduces only a small error, and this error is not serious since it only results in a calculated value of wavelength which is somewhat too small.

By integration of (3), we obtain

$$t \fallingdotseq \frac{2mc}{He}\sin^{-1}\frac{r'}{\sqrt{\frac{8V_amc^2}{H^2e}}} \tag{4}$$

where r' is the value of r which satisfies the condition,

$$\frac{dr}{dt}=0$$

or

$$r'=\sqrt{\frac{8V_amc^2}{H^2e}}. \tag{5}$$

From (4) and (5) we obtain

$$t \fallingdotseq \frac{\pi cm}{eH}. \tag{6}$$

Assuming on the other hand r_o extremely large, we have

$$\frac{dt}{dr}=\frac{1}{\sqrt{\frac{2eV_a}{md}r-\frac{H^2e^2}{m^2c^2}r^2}} \tag{7}$$

in which d represents the normal distance between two electrodes which now form two parallel plates, and r represents the distance from the cathode surface.

The following equations (8), (9), and (10) correspond to the equations (4), (5), and (6).

$$t=\frac{2mc}{He}\sin^{-1}\sqrt{\frac{deH^2}{2V_amc^2}r'} \tag{8}$$

$$r'=\frac{2V_amc^2}{deH^2} \tag{9}$$

$$t=\frac{\pi cm}{eH} \tag{10}$$

Equations (6) and (10) are exactly the same, so that in both cases the wavelengths can be calculated by the following equation,

$$\lambda_0'=\frac{2\pi c^2m}{eH}$$

or

$$=\frac{10650}{H}\text{cm} \tag{11}$$

This equation shows that the wavelength depends only upon the intensity of the applied magnetic field, and is independent of the value of the applied anode voltage. These characteristic features are in exact conformity with what have been observed in actual experiments.[3]

The corresponding equation representing the experimental results has the form

$$\lambda_0''=\frac{13000}{H} \tag{12}$$

so that the only difference is in its numerical value.

The wavelengths calculated from (11) should naturally be smaller than those obtained from (2), and hence the proportional difference of about 20 per cent between (11) and (12) can be accounted for.

[3] Note: Fig. 28 of Yagi's paper, loc. cit., does not show the relation between the applied anode voltage and the wavelength with the meaning as here referred to. In that case the intensity of the magnetic field was always so adjusted that the oscillation becomes maximum at every setting of the applied anode voltage.

The measured wavelength and the values calculated from (12) are given in Table I, where the "difference" in the third column is the percentage difference as follows:

$$\text{"difference"} = \frac{\lambda_0'' - \lambda}{\lambda} 100 \text{ per cent.}$$

It may be shown that, under the condition of the maximum oscillation, the radius of the anode cylinder should be related definitely to the square root of the applied anode voltage. The maximum os-

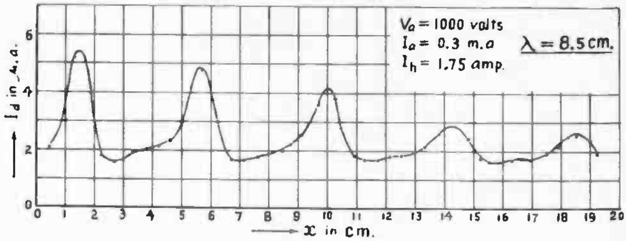

Fig. 2

cillation generally occurs at the neighborhood of the critical field strength and consequently, when H is given, the following relation should exist, which is deduced by putting $r' = r_a$ in equation (5).

$$\left.\begin{aligned} r_a &= \sqrt{\frac{8mc^2}{H^2 e} V_a} \\ \text{or} \qquad &= \frac{6.7}{H}\sqrt{V_a} \text{ cm} \end{aligned}\right\} \tag{13}$$

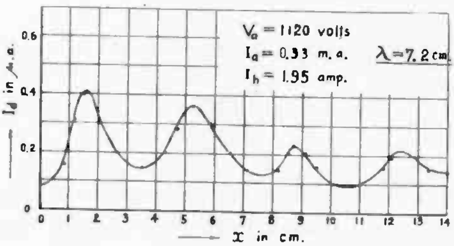

Fig. 3

All of the above equations are applicable to most of the cases of the magnetron oscillations, except the cases where the external circuits are important.

2. Experimental Results

Experiments have been carried out with the small split-anode magnetrons as shown in Table II.

TABLE I

Wavelengths Obtained by Experiments (cm)	Wavelengths Calculated from Eq. (12) (cm)	Difference (%)	
38	41.2	+ 8.4	Table 4 in *Jour.* I.E.E. of Japan, No. 467
36	39.3	+ 9.2	
32.3	36.5	+13.3	
42	43.5	+ 3.6	Table V in the above paper.
38	41.3	+ 8.7	
31	35.3	+13.9	
28.5	32.4	+13.7	
52	52.3	+ 0.6	Table I in *Jour.* I.E.E. of Japan, No. 469.
50	49.1	− 1.8	
49	44.8	− 8.6	
39	39.4	+ 1	
36.5	36.3	− 0.6	
33	34.2	+ 3.6	
32	33.3	+ 4	
70	62	−11.4	Table II in the above paper.
51	52	+ 2	
44	44.8	+ 0.9	
39.5	40.8	+ 3.3	
37	36.1	− 2.4	
34	34.2	+ 0.6	
47	41.5	−11.7	Table 7 in the Technology Report of the Tohoku Imp. Univ. Vol. 7, No. 4.
42	42	0	
45	46.4	+ 3.1	
54	55.2	+ 2.2	
52	52.7	+ 1.4	
48	48.2	+ 0.4	
11.5	11.2	− 2.6	Figs. 1, 2, 3, 4, and 5 in the present paper
8.5	7.1	−16.5	
7.2	6.4	−11.1	
6.8	5.7	−16.1	
5.6	5.4	− 3.6	

TABLE II

No. of Vacuum Tube	Diameter of Anode Cylinder (cm)	Length of Anode Cylinder (cm)	Diameter of Cathode Wire (cm)
I	0.5	0.8	0.01
II	0.35	0.5	0.01
III	0.25	0.5	0.01

The experimental results obtained using tube No. 2 are shown in Figs. 1, 2, 3, and 4. In these figures, V_a, I_a, and I_h denote anode voltage, the anode current, and the field current, respectively. In this case the field intensity H is 1050 gauss per ampere.

Throughout the measurements, the crystal detector bridging the parallel wires was kept at a fixed position and the short circuiting condenser was moved back and forth to obtain the wavelength measurements. I_d and x represent the detector current and distance through which the short circuiting condenser was shifted.

From the figures, it can be seen that the equations given do apply to the case where the wavelengths are extremely short. (Refer to Tables I and II and to equations (12) and (13)).

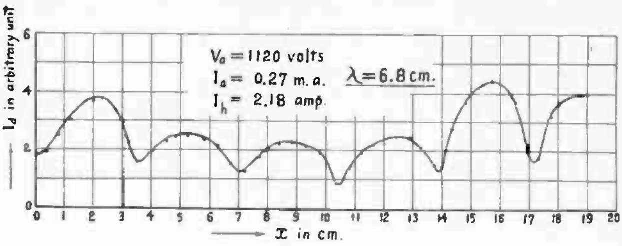

Fig. 4

The shortest wave that has ever been obtained by this method was produced with tube No. 3. The oscillations in this case had a wavelength of 5.6 cm. The potential distribution on the parallel wires is shown in Fig. 5.

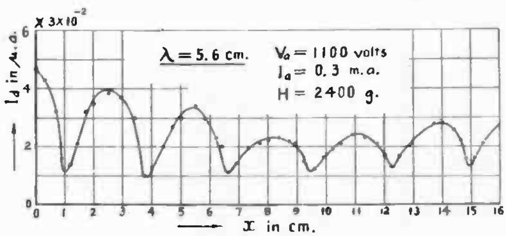

Fig. 5

3. Conclusion

The theory proposed in the present paper seems to be a very reasonable one, though not very exact, since there are many details left out of consideration. These details are to be studied further.

The author is of the opinion that this theory is also applicable to the type of oscillations observed by Barkhausen and Kurz.

The following is an approximate example of the limit of the wavelength, which can be obtained with a magnetron oscillator. These values are estimated from the summary of theoretical considerations given in an early part of the paper.

Suppose that the intensity of the applied magnetic field can be raised to 20,000 gauss—equation (12) gives the wavelength to be 6.5 millimeters. Then, if we desire to obtain maximum oscillation the anode radius must be made 0.34 millimeter, (13), and at the same time the anode potential must be raised to 10,000 volts. The practicability and even the possibility of such a small scale experiment is questionable. We may say, however, that the practical lower limit of the wavelength of oscillations produced by the magnetron under the condition of maximum oscillation is of the order of a few millimeters. If the intensity of oscillation be disregarded, it may be possible to reduce the wavelength to a slightly lower value.

The shortest wavelength actually produced so far (5.6 cm) is less than half of that previously reported,[1] and about one fourth of the lowest wavelength obtained by the Barkhausen and Kurz method. The intensity of the oscillations at this wavelength might be further increased if the anode voltage could be raised to the point indicated by (13).

The harmonic waves are not considered for the time being, though they are by no means unimportant, since the oscillations obtained from the magnetrons are apt to be distorted waves.

We can produce very short wavelength oscillations of two different types with a special form of magnetron. These two types were (a) the oscillation whose frequency is approximately independent of the external circuit, and (b) the one whose frequency depends principally on the external circuit. Most of the intense oscillations which we were able to produce were combinations of types (a) and (b).

The oscillations obtained by Professor Zacek, of which we learned recently, seem to correspond to type (a) just as do these described in the present paper.

The author wishes to express his thanks to Professor Yagi for his valuable instructions and suggestions.

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THE USE OF THE ELECTRON TUBE PEAK VOLTMETER FOR THE MEASUREMENT OF MODULATION*

By
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METHODS of measuring the degree of modulation of a radio-frequency generating set have been described[1] by Nelson and Conrad, but both systems have the disadvantage that they measure the amplitude of the audio-frequency current before it is impressed on the radio-frequency current. The oscillograph has been used extensively and provides an ideal method of measuring the degree of modulation, but it is an expensive instrument and not readily portable.

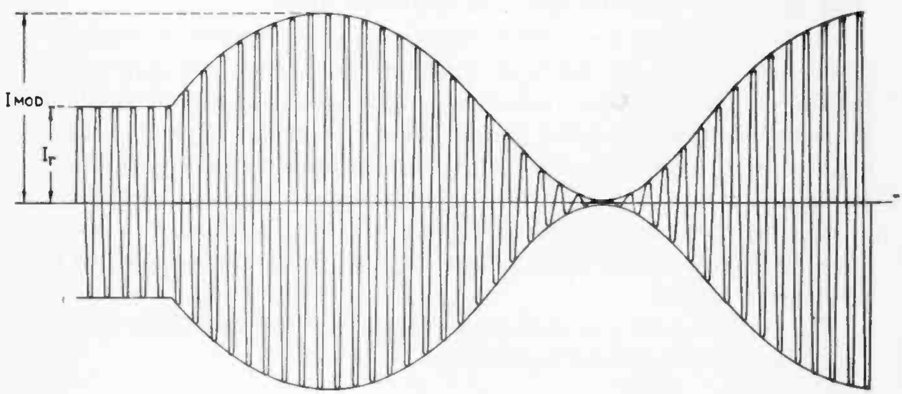

Fig. 1—Sine Modulated Radio-Frequency Current.

The electron tube peak voltmeter described by van der Bijl[2] may be used as a means of measuring the modulation of a radio-frequency current whose maximum value for complete modulation varies between the limits zero and two times the maximum value of the unmodulated radio-frequency current (see Fig. 1). Such a current is produced by the so-called "Heising" or "constant current" system of modulation. In brief, if the peak value of the radio-frequency current

* Dewey decimal classification: R275. Reprinted at the request of The Committee on Standardization from *Jour.* Opt. Soc. of Amer., 9, December, 1924. Published by permission of the Director of the Bureau of Standards of the U. S. Department of Commerce.

[1] Nelson, U. S. Patent 1,478,050. Conrad, U. S. Patent 1,477,316.

[2] van der Bijl. Thermionic Vacuum Tubes, p. 367.

is measured without modulation and then the modulation is applied and the peak value again measured the ratio

$$\frac{(I_{MOD}-I_r)\times 100}{I_r}=\text{per cent modulation} \tag{1}$$

I_{MOD} = peak value of the modulated current.

I_r = peak value of the unmodulated radio-frequency current.

The actual apparatus needed is shown in Fig. 2. The current I flows through the resistance R (1000Ω), so that measurements on the

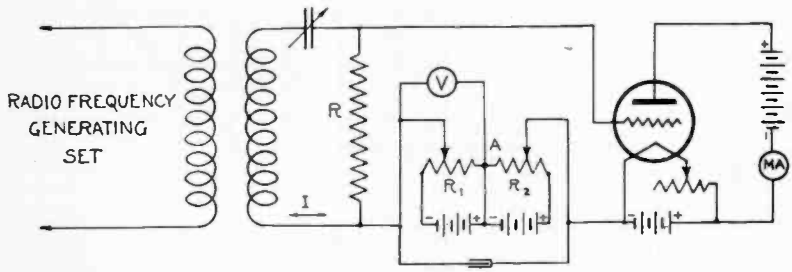

Fig. 2—Electron Tube Peak Voltmeter Used to Measure Percentage Modulation.

variation of the voltage across the fixed resistance R give values which are proportional to the variation in that current. The current I is proportional to the output current of the radio-frequency generating set.

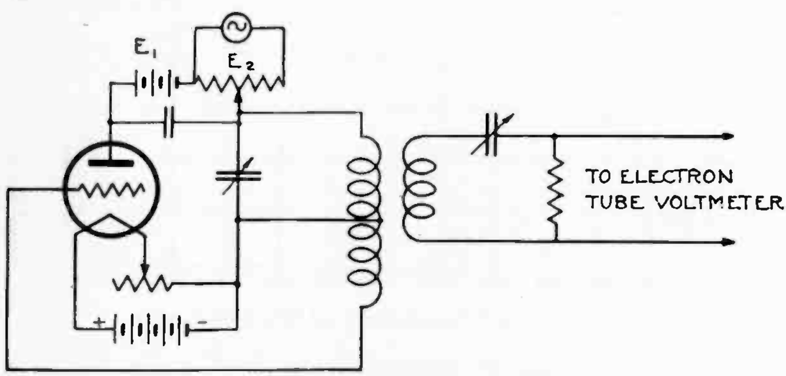

Fig. 3—Generating Circuit Used to Test Method.

The principle of the voltmeter is that the voltage of the grid is adjusted until the milliammeter just fails to show a plate current. The resistance R_2(0−2000Ω) with R_1(0−100Ω) set at zero (A) is adjusted until the plate milliammeter reads zero. The unmodulated radio frequency is then applied and R_1 adjusted until the milliammeter again reads zero. The voltmeter V gives then the peak value of the radio-frequency voltage which is proportional to the current I_r. The radio frequency is then modulated and R_1 further increased until the plate milliammeter again reads zero. The voltmeter will then read the peak value of the modulated radio-frequency voltage which is proportional to I_{MOD}. The percentage modulation then follows from (1).

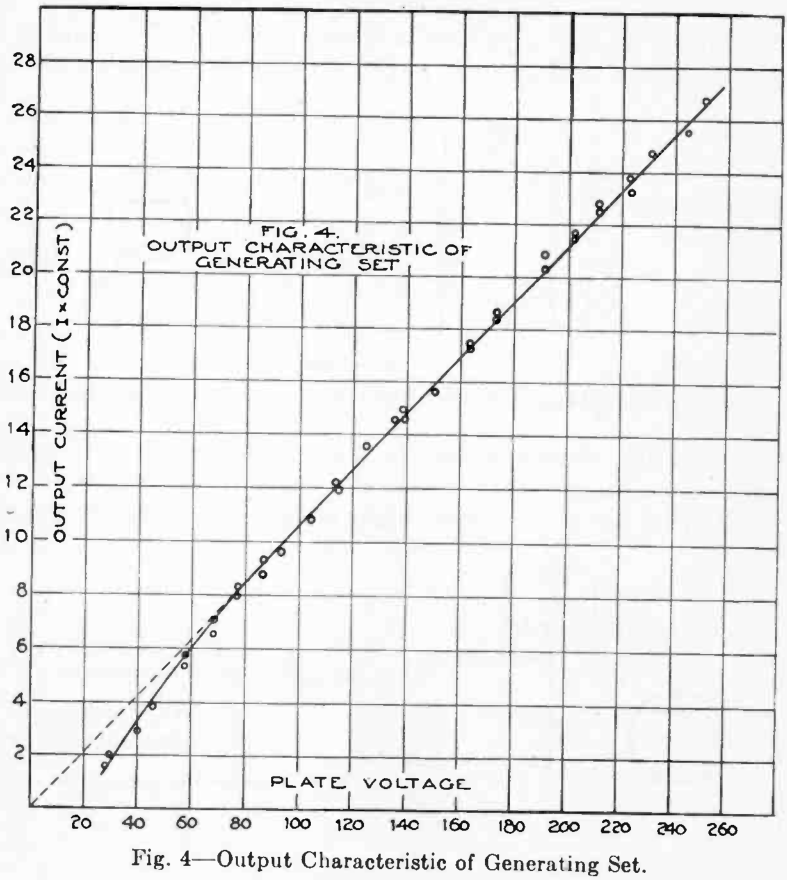

Fig. 4—Output Characteristic of Generating Set.

To test this method the generating circuit shown in Fig. 3 was used. The variation of output current with plate voltage for this circuit is shown in Fig. 4. Since, for the upper portion of the curve the output current is proportional to the plate voltage, the degree of upward modulation may be calculated from the measured values of E_1 and E_2. E_2 was either a 500 or 60-cycle sinusoidal a.c. source having an effective voltage of 120 volts. E_1 is a d.c. source and may be made any value.

The following table gives sample data on the values of the per cent modulation calculated and measured. This table shows that the accuracy of the measurement as made was not greater than about 4 per cent. The method, however, is sufficiently accurate for many purposes.

It has proved very useful in making quantitative measurements of the percentage modulation in connection with measurements on electron tube properties which involve the degree of modulation. It could be used in a system to indicate continuously the limits of modulation in a radio telephone transmitting station. The circuit could be coupled to the transmitting set and the resistance R_1 adjusted so that with no modulation the plate milliammeter reads zero. When the current is modulated then the milliammeter would show a deflection

TABLE I

E_1	E_1 Eff.	I_r	I_{MOD}	$I_{MOD}-I_r$	Measured Value $\frac{I_{MOD}-I_r}{I_r}$	Calculated Value $\frac{E_1 \times 1.41}{E_1}$
Modulation frequency = 500 cycles						
120	85	14.3	29.1	14.9	1.05	1.00
147	85	17.8	32.2	14.4	0.81	0.82
185	85	23.8	38.1	14.3	0.60	0.64
Modulation frequency = 60 cycles						
135	96	15.5	31.0	15.5	1.00	1.00
135	90	15.9	31.1	15.2	0.96	0.94
	80	15.8	28.8	13.	0.82	0.83
	60	15.9	25.7	9.8	0.62	0.62
	50	15.9	24.	8.1	0.51	0.52
	40	15.9	22.1	6.2	0.39	0.42
154	60	18.2	27.8	9.6	0.53	0.55
142	50	16.4	24.3	7.9	0.48	0.50
126	85	14.3	28.3	14.0	0.98	0.96
Radio frequency = 470 kilocycles						

which could be calibrated in terms of per cent modulation. The modulation then could be kept adjusted so that the milliammeter reads the degree desired. The instrument could be used at any time to measure the exact degree of modulation.

This method gives a fairly accurate measure of the degree of upward modulation. It cannot, however, give any indication of the characteristic of the current when the modulation is downward. It is believed that it is superior to other simple methods of measuring modulation.

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RADIO RECEIVER TESTING EQUIPMENT*

BY
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Summary—*The difficulties of measuring radio receivers and the lack of standardized equipment has made necessary the development and construction of special apparatus for this purpose. Generally speaking, determining the characteristics of the input signal to the receiver, the conditions of receiver operation, and the character of the resulting output will serve as a proper measure of performance. The "transmitter" used in the measuring setup has associated equipment in the way of attenuators, vacuum-tube voltmeters, modulation meter, and monitor and frequency supervision, so as to control rigidly and to know with great accuracy the character of the input signal. The design and usage of these various units are described in detail. Emphasis is placed on the proper design and use of vacuum-tube voltmeters. A novel modulation meter is described, being simple and accurate.*

Power measuring panels, screening, and other precautions are noted for maintaining the receiver under test in proper operating condition. The output power may be measured under different conditions of usage. A "distortometer" of new design is described, enabling the distortion and overload of the output to be directly measured.

Proper operation of this test equipment will measure radio-frequency amplification, audio-frequency amplification, and detector sensitivity. It will show the radio-frequency selectivity, the audio-frequency fidelity, the overall fidelity, and the overall selectivity. The frequency range of the receiver may be measured. The "hum" voltage of the output may be noted. The maximum undistorted output of the receiver and the percentage of overload at any output can be found. The methods of obtaining these data are given in the paper.

I. Introduction

THE performance characteristics of radio broadcast receivers are measured to aid in development, to check the standards of production and to compare competitive receivers. As these performance characteristics are new conceptions in the electrical industry, new definitions must be agreed upon, and new methods of measurement must be devised to evaluate properly these characteristics. Although the radio industry is still moving rapidly forward, ten years of experience have shown the fundamental basis on which radio receiver performance must rest. It was therefore possible for a group of representative engineers in the radio world to meet, compare experiences, and agree upon certain definitions of performance characteristics. The work of the Standardization Committee of the Institute

* Dewey decimal classification: R343. Original manuscript received by the Institute, December 8, 1928.

of Radio Engineers has almost been brought up to date, and it is expected that their report will shortly be released.

In addition to defining the performance characteristics of a radio receiver, the Standardization Committee has indicated in a general way how these characteristics shall be measured. There are two ways of measuring such characteristics. The first is relative and consists of comparing the output with the input or with some other unit maintained as a standard. The second way is absolute and consists of determining the absolute values of voltages, currents and power involved, and making such comparisons as are desired from the magnitude of the quantities determined. In general, an absolute method of measurement is best, although usually the most difficult.

The Standardization Committee, realizing the dearth of information on such radio-frequency measurements as they have deemed necessary, have not made detailed provision for the actual measuring equipment. The purpose of this paper is to describe in some detail testing equipment capable of measuring the performance characteristics of modern radio receivers, with design conforming to the restrictions of the Standardization Committee.

In addition to the purely technical problems of measuring extreme values of sensitivity, selectivity, etc., each set provides problems of connection, power supply, operation, and subsidiary equipment. The design of measuring equipment had to include such features as would simplify the actual testing of the receiver. Accuracy was, of course, paramount in the design, although simplicity, reliability, freedom from personal equation, and cost were also considered. The construction of this equipment was begun in June and finished in December, 1927, the installation being made in the Receiver Development Laboratories of the Crosley Radio Corporation at Cincinnati. The progress was necessarily slow because of lack of outside information and because each unit had to be tested as built. In June, 1928, the complete equipment was retested and recalibrated, and some changes, which six months' use had indicated were desirable, were made. The equipment has answered every demand made upon it, and is considered by engineers who have seen it to be the very latest in measuring equipment.

II. General

The equipment is divided into two parts, a transmitter and a receiver measuring unit. The transmitter is enclosed in one copper screened room and the receiver equipment in another screened room. Both of these small shielded rooms are enclosed in a third screened room. These three screens are conductively connected at only one point, thus insuring freedom from circulating currents that might otherwise decrease the effectiveness of the screening. When the door is shut, a 500-watt transmitter 300 feet away cannot be detected even when a sensitive receiver is tuned directly on the carrier.

A view of this screened room, or measuring booth, is shown in Fig. 1. All screens are soldered together and across the ends to keep the resistance in any direction low. Portions of both the inner and outer screens are carried on the door, the sides of which are tapered as on an ice box. When the door is closed, the screens wedge tight and

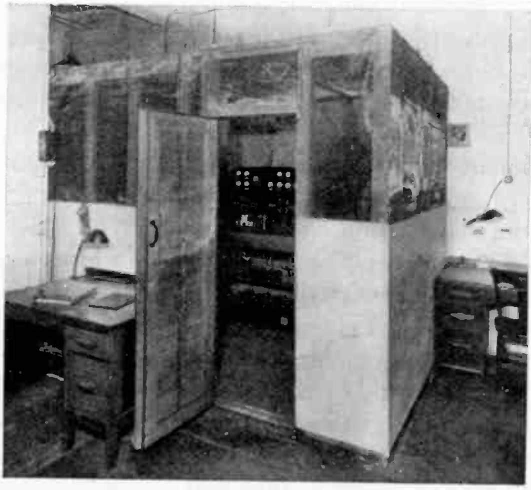


Fig. 1—External View of the Measuring Booth.

form a complete shielding surface with almost no crack. "Leaks" which are quite apparent when the door is slightly ajar disappear when pulled tight shut. A similar door is placed between the two inner rooms to insure complete shielding between rooms, while allowing passageway. Due to other shielding on the transmitter, it is seldom necessary to close this door, although for extremely sensitive sets this too has been found necessary.

A view of the transmitter unit is shown in Fig. 2. This is essentially a modulated transmitter with proper controls and means for measuring exactly what is being sent to the receiver. The use and construction of the units will be detailed later. Beginning at the upper left these are: crystal oscillator, modulation meter, monitor, radio-frequency oscillator, radio-frequency voltmeter, radio-frequency attenuator, audio-frequency oscillator, audio-frequency voltmeter, audio-frequency attenuator.

The meters are placed in a horizontal panel in front of the units for convenience in adjusting the transmitter. Any unit, or this meter panel, can be removed independently for test or recalibration.

The power supply for the transmitter comes from storage and dry batteries underneath the units. A small 100-watt motor generator outside the booth furnishes plate supply to those units where the

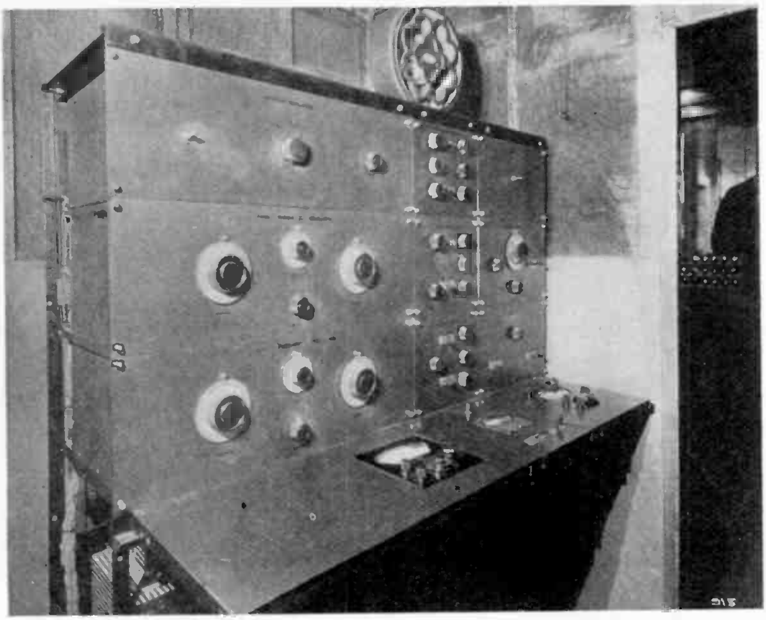


Fig. 2—Transmitter Unit.

plate voltage is not critical. The d.c. output of this generator, properly filtered with respect to commutator ripple and radio-frequency pickup, enters the screen booth through the center of the tube used to connect the screens together and to the ground system. All other external power or control leads, properly filtered, also enter through this pipe, so that the shielding is not impaired.

Fig. 3 shows the receiver testing equipment. The large center panel is for connections and power supply. The units at the ends are a detector calibrating voltmeter, a detector voltmeter, a.c. hum voltage voltmeter, an interference output voltmeter, and a power output voltmeter. The unit at the extreme right is the distortometer, a device for measuring the overload and maximum power outputs of a receiver. The unit on the extreme left is a power panel having the proper meters to give 110 and 220 volts at 25 and 60 cycles and 110 and 220 volts direct current.

Batteries to supply the test equipment and to operate battery operated sets are located beneath the bench. Tube racks, tool drawer, desk, and file complete the layout.

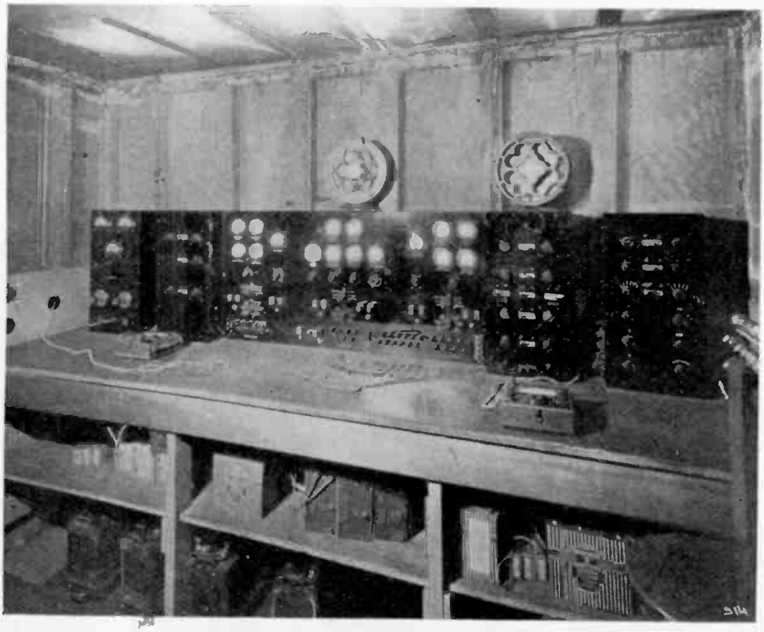


Fig. 3—The Receiver Testing Unit.

III. Radio-Frequency Generation and Measurements

In describing the equipment in detail, the plan will be followed of describing an operation, the reasons for providing a certain unit as constructed, and the means of calibrating and operating. The first and probably the most important operation is that of generating and measuring the radio-frequency voltage applied to the receiver. For some receivers and some measurements this voltage is as low as one millionth of a volt! Considering the fact that this is at radio frequency, it is obvious that straightforward methods of measurement are almost impossible. One method of measurement is to amplify this voltage with a radio-frequency amplifier of known characteristics to a point where direct measurement is possible. This method is often used when the source of voltage is not under control, as in field strength measurements.

A second method is a substitution method, but as this essentially means determining the magnitude of the substituting signal, this method is not independent except as applied to field strength measurements. A third method is to obtain this small voltage by the potential drop of a known current through a known impedance. A fourth method is to generate a small voltage by means of a mutual inductance; the frequency, current, and mutual inductance determining the voltage. A fifth method is to measure the voltage at as low a value as convenient and use some form of attenuator of known characteristics to reduce the measured voltage to the desired value. Combinations of two or more of the above methods may be used in some cases.

All of these methods have been tried at various times during the past five years in an endeavor to simplify and increase the accuracy of measurement. The method of measurement obviously influences the methods of power generation, so that the method of voltage measurement must be decided upon first.

The major objections to the first method are that the calibration is not constant as a result of the number of vacuum tubes used and their variations; the calibration varies greatly with frequency; and it is not convenient to change the ratio to conform with widely varying voltages. The principal objection to the third method is its inaccuracy for small voltages. It is impractical to measure currents much below 1 milliampere and inaccurate to use resistances much smaller than 0.1 ohm. This gives a minimum voltage of about 100 microvolts, a figure far too large for very sensitive receivers. Further reduction in the ohmic resistance leads to comparatively large errors due to the difficulty of accurately determining the high-frequency resistance, skin effect, and effective impedance due to reactance. Reactances used as coupling units have the same objections due to resistance. The objection to the fourth method is largely the same as for the third, namely, inaccuracy at low voltages although the large frequency shift is also undesirable. Without resorting to mathematical and form factor calculations (which in themselves are quite susceptible to unknown errors) it is very difficult to obtain a mutual inductance below 5 μh. This gives a minimum voltage of about 1500 microvolts, obviously too high a figure. A major objection to either the third or fourth method lies in the fact that a moment's carelessness on the part of the operator will burn out the measuring thermocouple, and in addition to the cost of a new element, it means delay in replacement and recalibration. The principal objection to the fifth method of voltage determination lies in the necessity for providing a sensitive vacuum-tube voltmeter and in the fact that there is a shift in attenuator calibration with frequency unless care is taken in the design.

The choice of a method is therefore determined by the ability to overcome the various difficulties. The design of a stable, sensitive vacuum-tube voltmeter, together with an attenuator independent of frequency, has placed the advantages of the fifth method far in advance of all others. In reliability, foolproofness, and accuracy this method seems the best possible at the present time. Voltages applied to a receiver can be varied between about 1 volt and one half of one millionth of a volt with a known accuracy of better than 3 per cent.

The Radio-Frequency Generator

The radio-frequency generator can next be considered. Due to coupling, both conductive and that resulting from radiated fields, it is desirable never to generate much more radio-frequency voltage than required. For reasons of stability and to conform with the method of voltage measurement chosen, it is necessary to generate a far larger amount of power than is normally necessary. This means that the shielding for the generator and the filter for the power leads must be the best possible. The generator was therefore built inside a copper box with a 1/16 in. wall. All the corners, except the lid, are completely soldered. The lid, with about 1/2 in. of overlapping edge, is screwed to the box with twenty-four machine screws, thus insuring a tight contact. Magnetic flux will leak through cracks that water will not flow through, and as this shielding almost determines the lower limit of voltage measurement, no precaution is too great. This copper box was placed inside a second brass box, forming the unit placed in the frame as shown in Fig. 2. The copper box and the brass box were conductively connected at only one point, thus minimizing the tendency for circulating currents to cause coupling. All ground wires and bypass to ground returns in the circuit are connected as near this ground point as possible, to decrease circulating currents. This practice was followed uniformly throughout the construction of the equipment. The object and usefulness of a shielding metal box is greatly impaired when the box is used as a ground return circuit, because of differences in potentials thus created. Such precautions may seem unnecessary, but the difficulties in removing a few unwanted microvolts resulting from circulating current coupling are otherwise almost insurmountable.

The Standardization Committee has recommended that the frequencies of 600, 800, 1000, 1200, and 1400 kc be adopted as standard, and performance characteristics be measured at these points. For some measurements, particularly selectivity and frequency range, higher and lower frequencies must be obtained. The oscillator used has a range from about 500 to 1600 kc.

The circuit is shown in Fig. 4. The approximate values used are indicated on the circuit. The Hartley circuit was used for the oscillator as it is quite stable at all frequencies and is not too rich in unwanted harmonics. A small vernier condenser is provided to give small changes in frequency. There is nothing unusual about the oscillator except

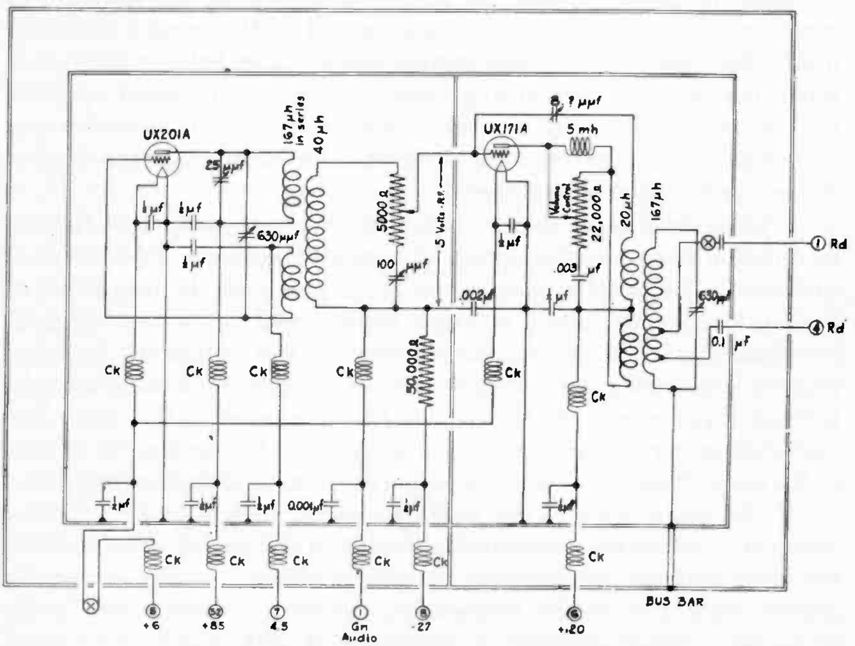


Fig. 4—Radio-Frequency Generator Circuit.

the double filter in each lead but the ground lead. As previously noted, all ground leads and condenser bypass return leads connect to a common ground point, even though the schematic circuit diagrams do not show this. The inductance of each choke is about one millihenry, so designed as to be effective within the range of frequencies used. The condensers used are non-inductive, an important point in using large paper condensers for radio-frequency bypass condensers.

Since the original design in June, 1927, this unit has been made single control, and the oscillator and filter condensers are driven from one shaft. This necessitated some change in the volume control, and that shown in the circuit was finally adopted. At full volume, the circuit is a normal amplifier neutrodyned with a "split plate" circuit. As the volume control is moved down, the plate current (r.f.) is "shunted" past the coupling coil. This control works smoothly and gives a wide range of adjustment. The usual volume control methods of filament or plate voltage variation could not be used as this amplifier tube is modulated. It was desired to change the output without changing the percentage modulation, and this could not be done if the operating characteristics of the 171A, used as the modulated amplifier, were changed. It was inconvenient to vary the mutual inductance between the plate coil and the second tuned circuit as this (due to capacity effects) varied the resonant period of the second circuit so much that single control was impracticable. The balance method is somewhat upset by this arrangement, but due to the small coupling back to the oscillator the frequency shift is only about 50 cycles in a million as the volume control is changed. Considering how this control is used such a shift is immaterial.

Grid modulation is used. This is more economical than Heising modulation from the standpoint of available voltage. Tests showed that with a low μ tube, such as the 171A, and with the magnitude of both the radio and audio voltages applied well below the overload point, no distortion due to grid modulation was obtained. In every respect, considering the voltages involved, the grid modulation method is equal or superior to the Heising method as applied in this unit. The variation in percentage modulation is obtained by varying the output of the audio-frequency generator supplying the modulating frequency.

If the magnitudes of the radio-frequency voltage and the audio-frequency voltage are maintained constant on the grid of the amplifier, the percentage of modulation remains constant (within reasonable limits) regardless of the frequencies applied. To keep the radio frequency voltage constant, a resistance of 5000 ohms and a small variable condenser of about 100 $\mu\mu f$ were used as shown. The variable condenser was adjusted to give as near a constant value of radio-frequency voltage as possible. This voltage is about 5 volts and varies only about 10 per cent throughout the entire frequency range.

There are two output coils, the values of which are roughly determined by the impedance to which they are supplying power. The lead marked $1Rd$ supplies power to the modulation meter and to the monitor. The lead marked $4Rd$ supplies power to the radio-frequency voltmeter and attenuator and receiver under test.[1] A switch is pro-

[1] The designating numbers used are those on the original layout of the equipment and are here used for convenience in detailing the circuits. All leads with the same numbers are connected. This will aid in tying the various units of equipment together.

vided inside the inner copper box to open the lead $1Rd$. In the case of extremely sensitive sets, this lead can be opened after the modulation is set, thus reducing unwanted pickup and coupling.

In building this unit, especial attention was paid to loose contacts, vibrating parts, choice of individual tubes used, with the idea of reducing frequency shifting and "wobbulation" with mechanical vibration. The frequency shift is largely determined by the plate voltage, so resistances and voltage measuring posts are provided in the power supply equipment to maintain the oscillator plate voltage absolutely constant. The amplifier plate voltage, not being so critical and requiring more current, is supplied by the motor generator.

The details of calibration and use will be reserved until later under the head of "Frequency Determination."

Vacuum-Tube Voltmeter

Vacuum-tube voltmeters as used throughout in the measuring booth are used so extensively in laboratory measurement equipment that it seems best to include a rather complete description of their theory, use, and misuse. Vacuum-tube voltmeters are roughly divided into three classes. First, the "peak" voltmeter, where the tube and associated equipment are used to indicate the maximum value of the voltage applied. The r.m.s. value, assuming sine wave shape, is obtained by dividing the peak voltage by $\sqrt{2}$. The advantage of the peak voltmeter lies in the fact that the tube used is not calibrated, and in the enormous voltage ranges over which the device will operate. The limits for a high vacuum tube are from about 0.5 volt to 300 volts. When properly used, the peak voltmeter takes no power from the circuit. The disadvantages of this form of voltmeter are its inaccuracy on low voltages, and the large error with bad wave form. This type of voltmeter was used extensively in the development of this equipment and for such work is almost indispensable.

A second type of voltmeter is the simple rectifier type with the tube plate circuit, meter, and source of voltage in series. This has no particular advantage over either the first or third types. It must be calibrated for the particular tube used and forms a load on the circuit voltage of which is to be measured. This load, in the case of high impedance tuned circuits, may destroy the usefulness of the voltmeter.

The third type, ordinarily known as the Moullin type, uses the ordinary three-element tube as a detector, with the grid biased negatively to prevent drawing current and consequently loading the circuit being measured. This type has been extensively described, but an analysis is included here to form the basis for further work on the modulation meter later. Also many constructional and circuit changes make it desirable to have the complete development data in one place.

In using the normal three-element tube as a voltmeter several conditions are necessary. The tube is maintained in an operating condition through the proper applied voltages. The grid is maintained negative at all times, grid current never flowing. The entire plate current is due to filament emission, no "gas" effects being present.

The work of Van der Bijl and others has shown that the plate current, I_p, of a vacuum tube is dependent on the grid and plate voltages, and that a grid voltage is μ times as effective as the same plate voltage. μ is approximately a constant and is consequently termed the amplification constant. This relationship can be written as

$$I_p = f\left(\frac{E_p}{\mu} + E_g + \epsilon\right) \tag{1}$$

where I_p is some function of the plate voltage E_p, the grid voltage E_g and a small voltage ϵ determined by the operating condition of the filament. In general, the interest lies not in I_p but in the change i_p when a small voltage e is applied to the grid. Let the term in parenthesis on the right of (1) be termed E_t. Then

$$I_p = f(E_t). \tag{2}$$

Now if a small voltage be impressed on the grid of the tube, (2) may be written

$$I_p + i_p = f(E_t + e). \tag{3}$$

This function may be expanded by Taylor's theorem as follows:

$$I_p + i_p = f(E_t) + \frac{e}{\underline{|1}} f'(E_t) + \frac{e^2}{\underline{|2}} f''(E_t) + \frac{e^3}{\underline{|3}} f'''(E_t) + + + + \frac{e^n}{\underline{|n}} f^n(E_t). \tag{4}$$

In this equation $\underline{|n} = 1, 2, 3, \cdots n$, and $f', f'', f''', \cdots f^n$ are the first, second, third, etc., differential coefficients of I_p with respect to E_t.* These derivatives may be written in terms of I_p and E_t as follows:

$$f(E_t) = I_p\,;\; f'(E_t) = \frac{dI_p}{dE_t};\quad f''(E_t) = \frac{d^2I_p}{dE_t^2};\quad f'''(E_t) = \frac{d^3I_p}{dE_t^3}.$$

* The differential coefficients of I_p should be taken with respect to e, but due to the symmetry of the function the derivative obtained is the same as that with respect to E_t.

In substituting these values in (4) the initial value I_p may be cancelled and only the change in current i_p retained. Thus

$$i_p = e\frac{dI_p}{dE_t} + \frac{e^2}{2}\frac{d^2I_p}{dE_t^2} + \frac{e^3}{6}\frac{d^3I_p}{dE_t^3} + + \frac{e^n}{\underline{|n}}\frac{d^nI_p}{dE_t^n} + . \qquad (5)$$

This is the fundamental equation used in most vacuum-tube theory. To evaluate this equation, the successive derivatives must be determined under the operating conditions. That is, the value of

$\frac{dI_p}{dE_t}$ will not be the same with an impedance in the plate circuit as when this impedance is removed.[2]

When used as a vacuum-tube voltmeter, the voltage applied to the grid of the tube may be written as

$$e = E_m \sin \omega t \qquad (6)$$

where E_m is the peak voltage and $\omega/2\pi$ is the frequency of supply. In the operating range of the tube, there is usually a considerable portion where only the first and second derivatives have a magnitude great enough to be considered. If an operating point be chosen within this range, all terms beyond that containing $\frac{d^2I_p}{dE_t^2}$ may be dropped from (5) without error. Operating at another point will be considered later. With this fact in mind, substituting (6) in (5) gives

$$i_p = E_m \sin \omega t\frac{dI_p}{dE_t} + \frac{E_m^2 \sin^2 \omega t}{2}\frac{d^2I_p}{dE_t^2}. \qquad (7)$$

The average current is the integral of (7) over one complete period divided by the time for that period. Thus

$$i_{p\text{ave.}} = \frac{E_m}{2\pi}\frac{dI_p}{dE_t}\int_0^{2\pi} \sin \omega t d(\omega t) + \frac{E_m^2}{4\pi}\frac{d^2I_p}{dE_t^2}\int_0^{2\pi} \sin^2 \omega t d(\omega t). \qquad (8)$$

Performing this integration and substituting limits gives

$$i_{p\text{ave.}} = \frac{E_m^2}{4}\frac{d^2I_p}{dE_t^2}. \qquad (9)$$

For a given operating point $\frac{d^2I_p}{dE_t^2}$ is constant and therefore the plate

[2] A complete discussion of this point, and the means taken to determine the operation of the tube in terms of the constants of the tube and circuit may be found in an article by John R. Carson, Proc. I.R.E., 7, 187; April, 1919.

current change indicated by a d.c. meter is proportional to the square of the applied voltage. Experimental evidence in a well designed voltmeter supports this conclusion.

In order to show the desirability of proper design, especially relating to wave form errors, the equations can be enlarged somewhat. Suppose that the voltage e is no longer a pure sine wave, but the sum of a fundamental frequency $\omega/2\pi$ and some harmonic X of this frequency. Then

$$e = E_1 \sin \omega t + E_2 \sin X\omega t. \tag{10}$$

Suppose an operating point is chosen with the third and fourth derivatives of (5) having values which are not negligible. Then substituting (10) in (5) gives

$$\begin{aligned} i_p = [E_1 \sin \omega t + E_2 \sin X\omega t]\frac{dI_p}{dE_t} + \frac{[E_1 \sin \omega t + E_2 \sin X\omega t]^2}{2}\frac{d^2I_p}{dE_t^2} \\ + \frac{[E_1 \sin \omega t + E_2 \sin X\omega t]^3}{6}\frac{d^3I_p}{dE_t^3} \\ + \frac{[E_1 \sin \omega t + E_2 \sin X\omega t]^4}{24}\frac{d^4I_p}{dE_t^4}. \end{aligned} \tag{11}$$

The average value of (11) is obtained by integrating with respect to ωt over one complete period (which corresponds to X complete periods for the harmonic voltage) and dividing by the time per cycle. The actual integration will not be given here, since the method is outlined in any table of integrals. Expanding, integrating, and substituting the correct limits results in the dropping of the first and third terms, and simplifies into

$$i_p = \frac{E_1^2 + E_2^2}{4}\frac{d^2I_p}{dE_t^2} + \frac{(E_1^2 + E_2^2)^2 + 2E_1^2E_2^2}{64}\frac{d^4I_p}{dE_t^4}. \tag{12}$$

In calibrating a vacuum-tube voltmeter, a sine wave source presumably is used. Comparing (12) and (9) shows that there is some question regarding what the voltmeter in the second case actually would read. Assuming for a moment the $\frac{d^4I_p}{dE_t^4} = 0$,

$$i_p = \frac{E_1^2 + E_2^2}{4}\frac{d^2I_p}{dE_t^2}. \tag{13}$$

As shown from (9) i_p varies as the voltage squared. Therefore the equivalent voltage of (13) is $\sqrt{E_1^2+E_2^2}$. Both E_1 and E_2 should be reduced to effective values to express correctly, but are left as peak voltages to avoid confusion later.

The effective value of the sum of two alternating voltages as read by the usual type of a.c. voltmeter is $\sqrt{E_1^2+E_2^2}$. It is therefore obvious from (13) that if $\frac{d^2I_p}{dE_t^2}$ is very large compared with the value of succeeding derivatives, the voltmeter will read true r.m.s. values of voltage regardless of wave form. Inspection of (12) shows that as $\frac{d^4I_p}{dE_t^4}$ increases (and presumably other derivatives also) the average current ($i_{pave.}$) will correspond in calibration to some r.m.s. voltage far different from that actually applied. Correct design is therefore very important, especially where the wave form is not closely sine wave. Voltages have been measured differing by 50 per cent from the true r.m.s. value, when using an incorrectly designed vacuum-tube voltmeter.

Some of these points in design are to use a type of tube and such operating voltages as to provide an operating range well in excess of the voltage to be measured. Preliminary characteristics are greatly to be desired, but in all cases, a check of the voltmeter reading against the true r.m.s. value should be made.

In general, the normal Moullin type of voltmeter is not sufficiently sensitive. The normal range of this type, using a 201A tube is between 0.7 and 3 volts. By using a more sensitive meter, or neglecting wave form errors, this range can be increased somewhat. This difficulty is corrected in the voltmeters used in this measuring equipment by using a d.c. amplifier. The plate battery for the voltmeter tube is connected between its plate and the coupling resistance of the amplifier tube. As the a.c. voltage is applied to the first tube the plate current increases, raising the grid bias on the second tube and lowering its plate current. A meter in the plate circuit of the amplifier tube serves to measure the applied voltage. A schematic circuit is shown in Fig. 5, where the first tube shown is the detector, and the second tube is the d.c. amplifier. This simple circuit is modified somewhat by the controls necessary to obtain the correct initial setting and to control the voltmeter in operation. The sensitivity is determined by the tubes used, battery voltages, coupling resistances, etc. Details are given later in discussing the individual units.

The usefulness of such a voltmeter is absolutely dependent on its accuracy. This accuracy is determined by the accuracy of calibration and the faithfulness with which this calibration is maintained. It has been found that so long as the operating point does not shift materially from its initial setting, the *change* in plate current with applied voltage is correctly given by the calibration. This means that even though the plate and grid voltages change as much as 10 per cent, the calibration will still be correct, provided that only the *change* in plate current with applied voltage be measured. To accomplish this a reverse current equal to the initial plate current of the second tube is sent through the meter. Consequently this meter reads only the change in plate current. This has an additional advantage in that it enables a very sensitive meter to be used and lowers the minimum voltage measurement limit.

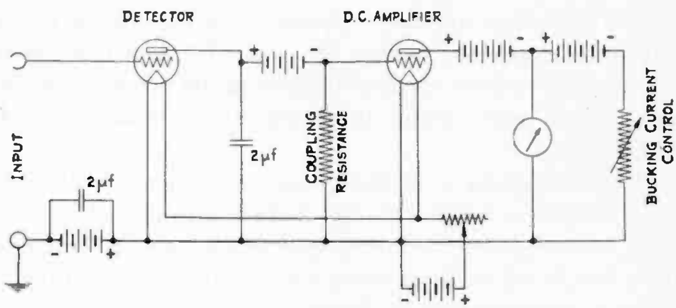


Fig. 5—Vacuum-Tube Voltmeter Circuit.

The biggest variation in calibration is caused by the initial setting, particularly with respect to the filament. Also, as the tube ages its emission characteristics change and destroy the accuracy of calibration. The usual methods of measuring filament voltage or filament current were tried without success. A simple method is to tie the grid and plate together and back to the positive filament terminal through the plate meter. The filament current is then adjusted to give some predetermined emission current, thus setting the tube in the same condition of operation every time it is set. An indication of how reliable this method of adjustment can be is shown in one voltmeter which has been in almost continual use for over three years. Repeated calibrations have shown a steady shift in calibration totaling slightly less than 3 per cent for this entire period. Other voltmeters built since show calibration characteristics indicating even greater constancy.

In determining the emission current initially, the filament current is set about 10 per cent below the normal value and the emission current measured. Under conditions of use, a shunt is connected across the indicating meter of such a value as to give full scale reading on the lowest scale (when a multi-scale reading meter is used) when the filament current is properly adjusted. A copper wire wound rheostat is used as filament current control in order to obtain small current changes and consequently accuracy in initial setting.

As the same meter is used to measure the emission and also indicate the voltage, switches marked "start" and "run" are provided to make

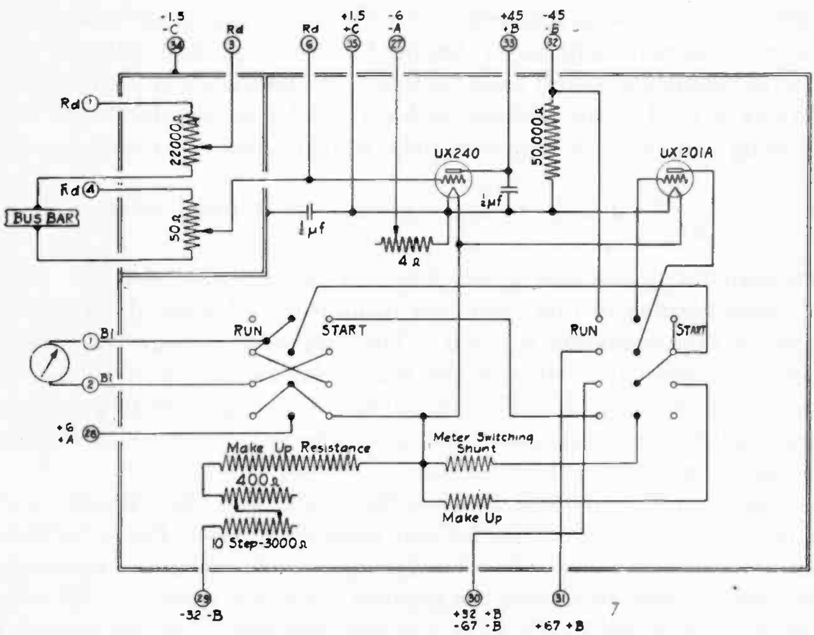


Fig. 6—Radio-Frequency Voltmeter Circuit.

the proper circuit changes. In the case of a two tube voltmeter (rectifier and d.c. amplifier) the second tube only is used to set the emission. To a reasonable approximation, both tubes change in emission characteristics alike and setting one serves to set both.

A voltage about equal to that of the plate battery of the second tube is used for the reverse current battery for several reasons. With a high voltage "bucking" battery, a high series resistance is necessary, and comparatively large changes in resistance only produce small current changes in the reverse current. It is consequently easy to adjust the indicating meter to zero reading, a particularly desirable feature if the meter is very sensitive. Another reason for using similar voltage for plate and bucking batteries is that the same type of batteries, having the same current drain, may be used. As a result, the initial zero adjustment of the meter does not have to be greatly changed during the useful life of the batteries. When replacing batteries both units are replaced at once.

The way these various points are taken care of is shown by the circuit diagram and indicated values in Fig. 6. This is the vacuum-tube voltmeter used for the radio-frequency voltages. The sensitivity of this type of voltmeter is determined by the type of tubes used, the value of the coupling resistance, and the sensitivity of the indicating meter. A Rawson multimeter, having four scales of 20.0, 2.0, 0.2, and 0.02 milliamperes at full scale, is used. As indicated, a high-μ tube, UX240, is used as the rectifier with a UX201A as the amplifier. The coupling resistance is approximately 50,000 ohms. To increase the value of $\frac{d^2I_p}{dE_t^2}$, a large bypass condenser is shunted between plate and filament of the first tube, thus increasing the sensitivity. The full scale reading of this voltmeter using the 0.02 scale of the multimeter is approximately 0.1 volt. The minimum voltage readable is approximately 0.02 volt and the maximum (using the 2.0 scale) is approximately 1.0 volt. The advantage of having a 50 to 1 voltage range will be best appreciated by those who have tried to cover large voltage ranges with a thermocouple.

The voltmeter is calibrated on 60 cycles from the 110-volt line, using the proper thermocouples and resistance units. Before calibration of the voltmeter, the thermocouple and resistance units are checked in order to obtain the greatest accuracy possible. All voltmeters used in this booth have a known accuracy of better than ± 1 per cent at any scale reading above 10 per cent of full scale. Proper construction, with repeated checks, have shown these voltmeters to be independent of frequency, being accurate at either radio or audio frequencies. For this to be true, the bypass across the coupling impedance must be effective at the frequencies at which the meter will be used. For calibration purposes, a temporary filter is placed between the rectifier and amplifier to prevent the second tube also from rectifying the applied a.c. voltage and thus giving a false reading. In the case of the audio and 60-cycle voltmeters, this filter is made permanent for the above reason.

A few more notes will be of value to anyone constructing this type of meter. Another stage can be added to increase the sensitivity, but experience has shown that the troubles with drift of the meter, leakage, and increasing shift of calibration make this step generally inadvisable. Work with the new screen-grid tubes indicates that a voltmeter built on the same general lines as this type can be made from five to ten times as sensitive if necessary. In the construction of this meter it is extremely important that loose contacts and leakage be avoided. Either of these faults will be characterized by the meter needle jumping very erratically, refusing to stay on zero and otherwise misbehaving. Old "B" batteries, with high and variable resistance, will show the same effect. As a general point in design, too high a plate voltage on either tube will cause "spotty" emission from the filament and the same "jumpy" tendency on the part of the meter needle. Ninety volts is about the usual limit with the 201A type filament, and 67 volts is even more satisfactory. The use of 112 tubes is discouraged; they are not sufficiently stable and constant in characteristics for any form of precise work. A properly operating vacuum-tube voltmeter is a "joy to the eye"; this particular voltmeter will only drift about 2 microamperes in 10 hours.

It usually takes about 5 minutes when first turned on for everything to become settled and for the zero to stay fixed. After this time the voltmeter is in every respect no more trouble than any other type of voltmeter, and is more convenient because of its greater voltage range.

In the circuit of Fig. 5, two potentiometers are shown shielded from the rest of the circuit. The upper one is the supply control to the monitor and the lower one is the supply to the voltmeter and the attenuator. This is the volume control generally used, and in order to cut to zero especial care is taken to ground the lower end. A large bus bar 2 in. $\times 1/4$ in. runs along the back of the transmitter, and all units are grounded to this bus bar with copper strips. In the case of the voltmeter, the ground returns of these potentiometers connect to the bus bar at the point of lowest potential. Thus in the "off" position, the "ground" currents (due to the capacity between the transmitter and the screen) flowing through the bus bar do not produce a voltage effective in the input circuit of the voltmeter.

The monitor control was placed in this compartment merely to preserve a neat appearance in the combined unit, and to be handy to the radio-frequency oscillator so that a short high potential lead could be used.

The Radio-Frequency Attenuator

Considering the sensitivity of the voltmeter, it was apparent that an attenuation ratio of 25,000 to 1 was desirable. Assuming a minimum coupling resistance of 0.1 ohm, this meant a series resistance of 2500 ohms. Difficulties resulting from capacity coupling, skin effect, and changing ratio are almost insurmountable in maintaining the accuracy desired. A multiple stage attenuator was therefore necessary. With two stages a ratio of about 160 was needed, with three stages about 30, and with four stages about 13. For various reasons the three-stage unit was built. As a frequency shift is extremely undesirable, capacitive and inductive reactance must be negligible. A resistance unit, comprising a short length of No. 40 bare advance resistance wire of about 40 ohms resistance, is held between two heavy copper studs. This fine resistance wire has a high value of R/L per

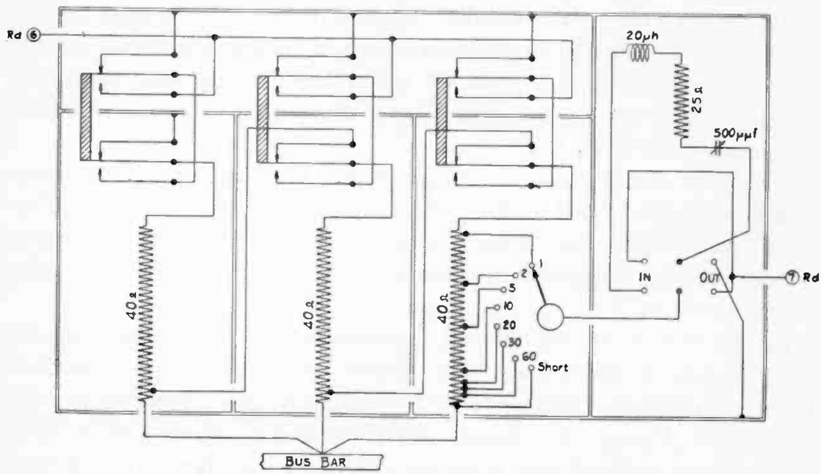


Fig. 7—Radio Attenuator and Dummy Antenna Circuit.

unit of length and a minimum of skin effect. Smaller wire than No. 40 would not have had sufficient mechanical strength. Larger wire would have had a resistance so low that connecting leads would introduce large errors. Heavy copper wedges, with the narrow edge just soldered in contact with the resistance wire, form the taps. Two of the stages have only one tap, at a 30–1 ratio, and the last stage is tapped and connected to a switch giving ratios of 1, 2, 5, 10, 20, 30, 60, and "short."

In a separate compartment in the same unit is included a dummy antenna system. The "standard dummy antenna" is used, consisting of 20 μh, 25 ohms, and 200 $\mu\mu$f connected in series. The condenser is made variable to aid in certain types of development. A switch throws this dummy antenna in or out of the circuit at will.

The complete circuit is illustrated in Fig. 7 where the stage control switches are schematically indicated. There are five separated compartments in this unit. One holds the dummy antenna, each of three holds an attenuator resistance, and the fifth holds the input lead and switch control shaft. The switches are all normally in the position indicated, and the particular number of stages desired are cut in by operating a cam shaft, depressing the proper switch. The lead which connects the high voltage (comparatively—it may only be 0.1 volt) input to the high side of the attenuating potentiometer is normally disconnected at both ends and grounded. This eliminates any capacity coupling between the input and output. Repeated tests have shown no trace of capacity coupling between input and output or between stages. This is, of course, necessary, or the total attenuation will not be the product of the individual stages, and sufficient range will not be available.

The ground potential ends of the attenuating resistances do not go to the case, but are carried through insulating bushings to a common ground connection on the bus bar. Incidentally, the common ground point is determined by these ground returns; these leads must be short to prevent inductive reactance error.

The attenuator was calibrated at radio frequencies of 500, 1000, and 1500 kc to determine the frequency shift, if any, and to observe capacitive and inductive errors. The proper operating conditions are maintained while calibrating; that is, the last stage is connected across the second, and the second across the first. The common ground point is used as the individual potentiometer ground, and not merely the stud terminal of the resistance wire. The calibration is checked in several ways. The d.c. resistance serves as an approximate check of the ratio, the attenuation at any point being the product of the stages. Another calibration using radio frequency and two vacuum-tube voltmeters gives about the same degree of accuracy. A method which is quite simple and is remarkably consistent and accurate is to take a radio receiver, note the output with a known input voltage and a 1 to 1 attenuator setting. Then change the attenuator to 2 to 1 and increase the input voltage to give the same output. This step by step calibration, with cross checks and reverse calibration, serves to establish the ratios within about 3 per cent error for the worse condition i.e., three stages. The accuracy for the last single stage is much better, being within approximately 1 per cent.

The attenuator will be in error by a considerable amount if a low impedance is connected across the output. If an impedance as low as 100 ohms should be placed across the output (no dummy antenna) the possible error would be about 10 per cent. (This on the 2 ratio of the last stage.) Smaller impedances would give greater error. However, the loads are seldom so low in impedance; if they are, a correspondingly lower resistance (higher ratio) may be used and all such error eliminated. The output from the attenuator (RD7) goes through a brass tube to an output post on a small panel in the receiver testing room. This is the upper disk seen on the left wall in Fig. 3.

This combination of radio-frequency generator, voltmeter, attenuator, and dummy antenna forms the input system to a receiver under test. Using a single stage with 1–1 ratio, the maximum applied voltage is of the order of a volt. With all three stages and a 20–1 ratio, the attenuation (based on the present calibration) is 18,200. The minimum voltage is therefore about 0.04/18200 or 2.2 microvolts.

In measuring a receiver, it is assumed that an antenna with an effective height of four meters will be used. The Standardization Committee has chosen such constants for the dummy antenna as will approximate the constants of an antenna with four meters effective height. Therefore the actual voltage applied is divided by 4 to rate the receiver measured. Thus the present equipment is capable of rating any receiver having a sensitivity value above 0.6 microvolt per meter. The accuracy at this extreme reading is approximately ±5 per cent. As this figure is normally far below the static level, it is thought this limit will be satisfactory for some time to come.

IV. Audio-Frequency Generation and Measurement

The generation and measurement of audio-frequency power and voltage is quite the same problem as that of the radio frequency, with the exception that less complete shielding and better filtering is needed. As more information regarding audio-frequency voltage generation is available, this portion will not be stressed, except to point out certain differences from other methods. The circuit and approximate values are shown in Fig. 8. Fundamentally, the audio frequency is produced by rectification of the heterodyne between two intermediate frequencies. One of these is fixed at about 140 kc, while the other varies between 100 kc and 140 kc, thus giving an audio beat note between 0 and 40 kc. The reason for this large range will be given later. With the fixed oscillator at the high frequency, it is easy to adjust the shape of the condenser plates of the variable frequency circuit to "spread out" the low frequencies, thus keeping about the same percentage accuracy at any dial setting. A small variable condenser, adjustable from the panel, is placed across the "fixed" oscillator. This has a double purpose. When the variable oscillator dial is set to zero, this small con-

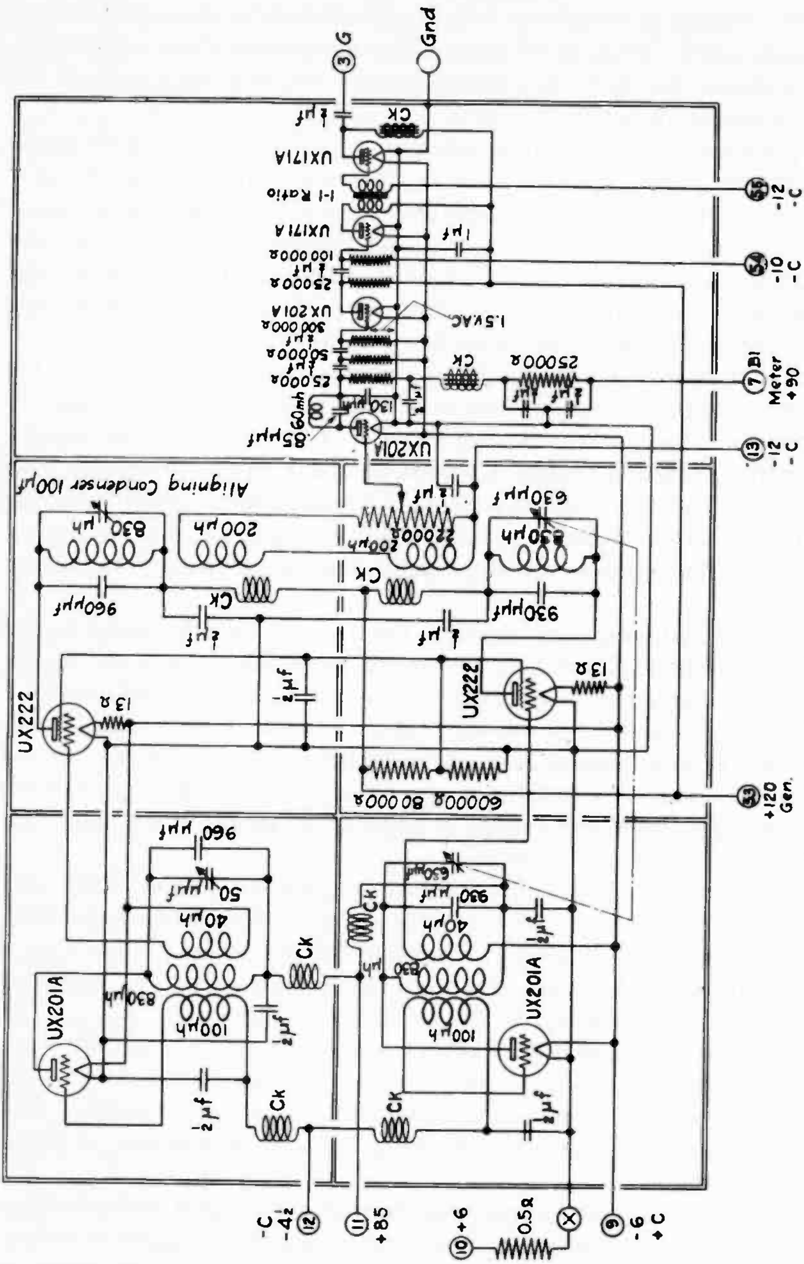


Fig. 8—Audio-Frequency Generator Circuit.

denser may be used to adjust for a true zero beat. This corrects for slight changes in frequency as a result of plate or filament voltages being in error. These small changes are immaterial above a few thousand cycles, but serve to increase greatly the frequency error if not compensated for below this point.

A second use of this small condenser (called a vernier, although it is not such in the usual sense) is to give a wide open scale at the low-frequency end. The vernier dial is set to zero and the variable oscillator dial adjusted to give zero beat. Then, having been calibrated, the small dial serves to give audio frequencies up to about 2000 cycles. The condenser is so shaped that the lower frequencies are well spread out, enabling frequencies of as low as 30 cycles to be set accurately.

As the intermediate-frequency amplifiers are not as selective as the amplifier in the radio oscillator, more care must be taken in regard to harmonics generated in the oscillators. The comparatively smaller frequency variation enables a reverse feedback circuit to be used with stability. The grid coupling coils should be adjusted to give just barely stable oscillation with all condenser adjustments. A slightly greater coupling than this is used to minimize variations resulting from voltage changes.

The oscillators are isolated from the heterodyning circuit by means of the amplifiers. There is no observable tendency to "lock in," stable frequencies as low as one cycle per second being maintained over long intervals. To avoid the necessity of neutrodyning, screen-grid tubes are used as intermediate-frequency amplifiers. Their bias is obtained by a series filament resistance giving about 1.5 volts. The value of the intermediate-frequency voltage is adjusted by means of coupling coils so as to not draw grid current.

The outputs of the amplifiers are coupled through a 22,000-ohm potentiometer, to serve as a volume control. To avoid overloading, the voltage input from each amplifier coil is set about 1 volt. This is about the limit allowable as any greater voltage raises the harmonic content of the output above the permissible value.

Following the detector, a low pass filter circuit designed for a cut off of about 70,000 cycles is placed. This prevents the amplified intermediate frequency from overloading the first audio amplifier. The values given hold only when working between impedances of about 25,000 ohms.

The detector plate current goes through a d.c. meter before returning to the supply source. At low frequencies, say 10 cycles or below, the plate filter is not effective and the meter needle will swing in accordance with the heterodyne frequency. This feature is used to set the oscillators to exact zero beat when starting the unit or checking the audio frequency.

In order to pass low frequencies in a resistance coupled amplifier, a high coupling condenser, or a high leak resistance or both, must be used. As it is almost impossible to obtain a large capacity without some leakage, this may place a positive potential on the succeeding grid. This, of course, means grid current and distortion. The leakage problem was solved by using a two-stage resistance capacity coupling, giving a product attenuation for the leakage, yet passing the low frequencies desired without much increase in loss.

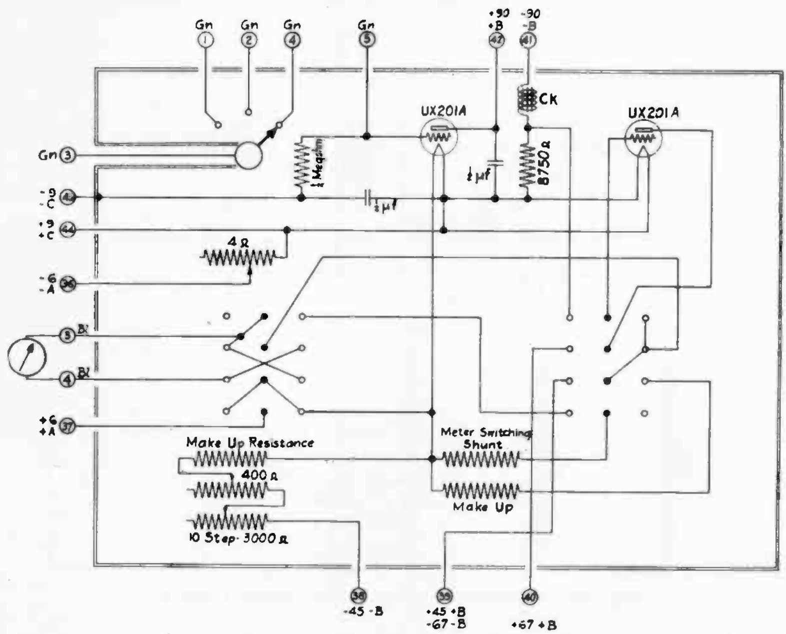


Fig. 9—Audio-Frequency Voltmeter Circuit.

An especially designed transformer is used as the input to the last stage to amplify the low frequencies. A low impedance secondary was used, thus avoiding voltage drop in case of grid current. The grid can therefore "swing" a greater amount, even positive, without causing distortion, and enabling a correspondingly greater power output to be obtained.

The variable oscillator and amplifier are uni-control. The fixed amplifier is broad enough to need no further tuning as the "fixed" oscillator is slightly changed with the vernier. The oscillator tubes are supplied with plate power from batteries with series regulating resistances to maintain the voltages constant. The remainder of the unit is supplied from the external d.c. generator.

The Audio-Frequency Voltmeter

The same method used for measuring the radio-frequency voltage is used for the audio-frequency voltage. The circuit is shown in Fig. 9. It is essentially the same as that of the radio voltmeter, and is included only for the purpose of providing a composite wiring diagram if such should be desired.

The voltmeter needs to be less sensitive and therefore a 201A is used as the first tube and only 8800 ohms for a coupling resistance.

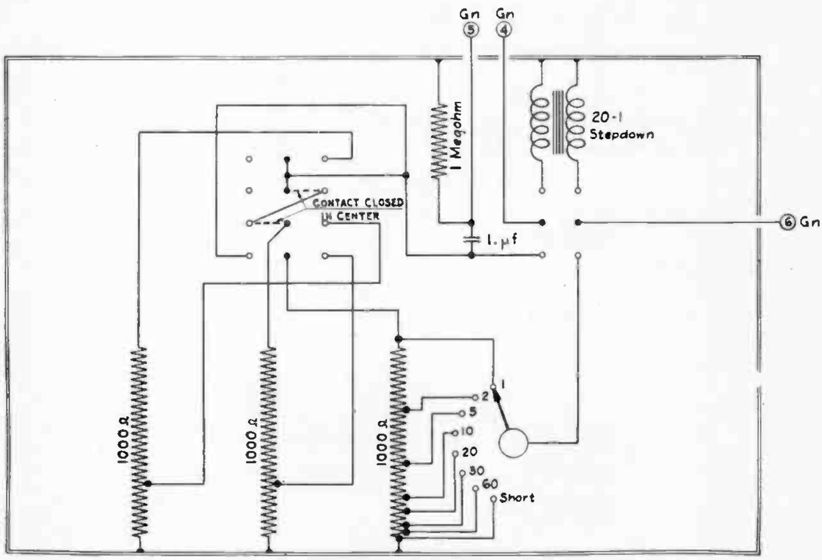


Fig. 10—Audio-Frequency Attenuator Circuit.

The voltage range is from a minimum of about 0.1 volt to a maximum of 2.6 volts. As previously noted, the audio filter is left in the circuit to prevent the audio frequencies acting on the second tube.

The output of the audio-frequency generator goes to a three-point switch located in this unit. These points go to the radio oscillator for modulation purposes, to the monitor, and to the audio attenuator. The only time the audio voltage is measured is when using the attenuator, and therefore only this connection is provided.

The Audio-Frequency Attenuator

The audio-frequency attenuator, shown in Fig. 10, is modeled after the radio-frequency attenuator, having three stages, each tapped the same as in the radio-frequency unit. The single stage resistance is 1000 ohms and as capacity effects are relatively unimportant, a Federal anti-capacity switch with properly bent contacts serves as the three-stage switch. A second switch serves to send the audio-frequency generator output through a high ratio step down transformer giving considerable power at a low voltage, as is occasionally desired. The attenuator output lead goes through a brass tube to a contact post similar to that of the radio-frequency attenuator.

V. Frequency Determination

The determination of frequency is a problem second only to that of voltage measurement. An absolute method of frequency determination

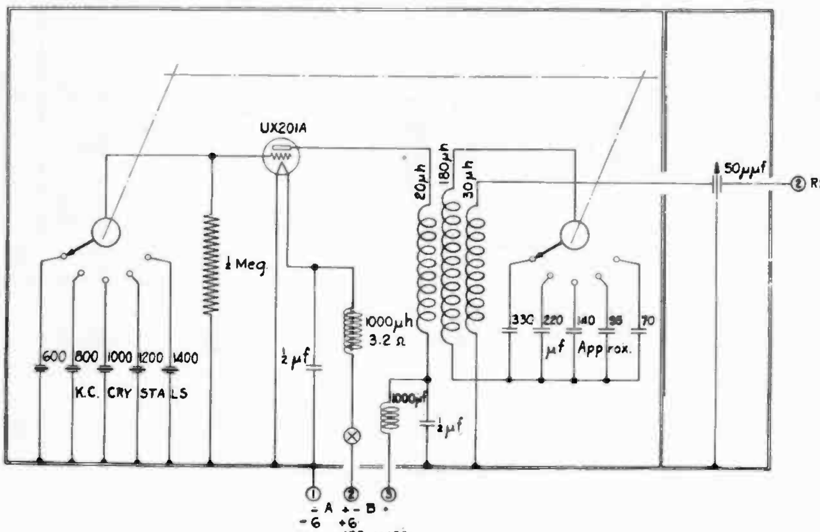


Fig. 11—Crystal Oscillator Circuit.

may be the use of a time and counter checked oscillator with a multivibrator method of frequency multiplication. Such exactness is not required in commercial practice, and therefore no fundamental standard is provided in this measuring equipment. In brief, piezo-electric crystals are used for determining the radio frequencies, and tuning forks are used for determining the audio frequencies.

Two units are provided to aid in this determination. The first is a crystal oscillator shown in Fig. 11. Five crystals with natural frequencies corresponding to 600, 800, 1000, 1200, and 1400 kc are provided. A five throw double-pole switch serves to connect the crystal and the corresponding tuned circuit desired. The output is regulated by swinging a grounded rotor between the two plates of a condenser.

There is no temperature control on the crystals. It was not felt necessary to try to maintain the frequency absolutely constant from day to day inasmuch as the frequency shift may only be 1000 cycles under quite wide temperature ranges. The harmonics of the crystals, beat against each other, indicate that the absolute accuracy is not better than ± 2 kc, and therefore no further refinement is necessary.

The monitor is the second unit used in determining frequency. The circuit is shown in Fig. 12. It consists of a tube biased to act as a

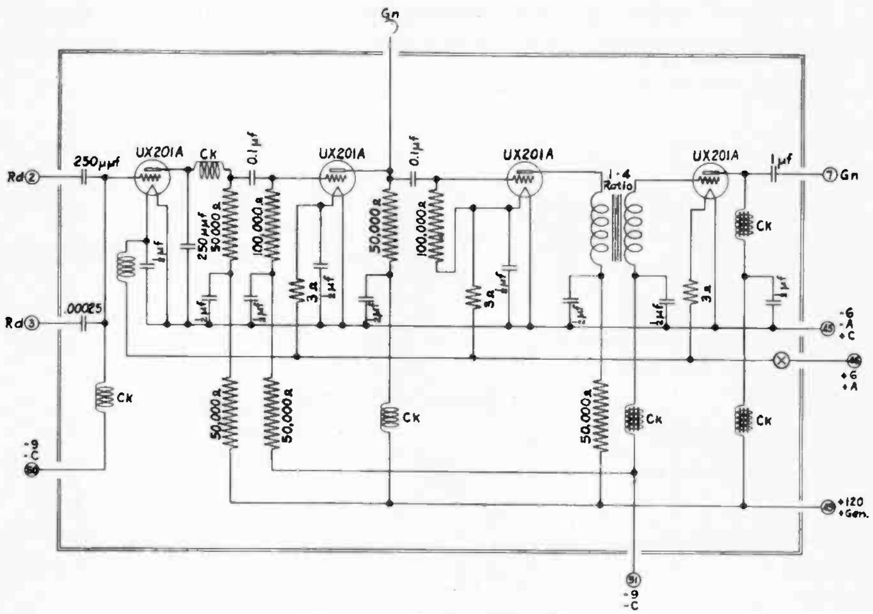


Fig. 12—Monitor Circuit.

detector, a resistance coupled amplifier stage, a second detector, and a second amplifier. The operation is as follows. The output from the crystal oscillator and a portion of the output of the radio-frequency oscillator is impressed upon the first detector of the monitor. By adjusting the radio-frequency oscillator until the resulting beat is zero, the radio frequency applied to the set is the same as that of the crystal. This method is much faster and more accurate than depending upon a calibration curve.

As a result of the detector arrangement, a frequency impressed on the first detector is multiplied and passed on to the second detector where a second beat note may result if another frequency is present.

Thus 900 kc may be obtained on the radio-frequency oscillator by setting the crystal oscillator to 600 kc and adjusting the radio-frequency oscillator to approximately 900 kc, and then carefully adjusting to make the beat frequency resulting exactly zero. This means that the third harmonic of the crystal oscillator, or 1800 kc, is beating the second harmonic of the radio-frequency oscillator, also 1800 kc, in the second detector of the monitor. This process may sometimes be carried as high as the seventh harmonic, and results in about 30 calibrating points for the radio-frequency oscillator. Any frequency desired and not given by such a possible harmonic ratio may be obtained by reference to the calibration curve. It is only a moment's work to check this calibration if its accuracy is questioned.

The determination of the selectivity curve of a receiver is a job requiring extreme accuracy. In general, a calibration curve cannot be depended upon to give accurately small frequency changes from an initial carrier frequency. Some form of mechanical verniers are often used, but are quite susceptible to error. The method used here has proved quite satisfactory. The crystal oscillator is set at the frequency at which the selectivity curve is desired, say 1000 kc. The radio-frequency oscillator is also adjusted as previously described to exactly 1000 kc. The receiver is then tuned accurately to this signal and not changed further. Suppose it is desired to obtain the receiver response at intervals of 1000 cycles difference. The audio-frequency oscillator is then set at 1000 cycles, and by means of the switch shown in Fig. 9 is connected to the second detector of the monitor. The vernier dial on the radio-frequency oscillator is then changed and a beat note results between the crystal oscillator and the radio-frequency oscillator in the first detector of the monitor. This beat note passes to the second detector and there heterodynes the 1000-cycle audio frequency, and a second beat note results. By proper adjustment of the radio-frequency oscillator this second beat note may be made zero. When this is done, the radio frequency then applied to the receiver differs from the original carrier frequency of 1000 kc by the value of the audio frequency 1000 cycles, giving 1001 kc. The receiver response is then noted on the output meter. This process can be repeated by swinging the radio frequency to the other side of the carrier, again getting zero beat and having a resulting frequency of 999 kc. Setting the audio oscillator at 2000 cycles and repeating gives two more points. Continuing this process gives the complete receiver response curve with respect to frequency, i.e., the "selectivity" curve. The frequency has the accuracy of the audio oscillator frequency setting, which is much greater than that obtained by adjusting the radio-frequency oscillator by calibration. The audio-frequency oscillator was made to go to 40,000 cycles so that the selectivity curves could be carried at least that far accurately. For greater frequency differences, the radio-frequency calibration curve can be relied upon. As a result of this accurate method of setting the frequency differences, curves can be repeated almost exactly even though they are quite sharp.

The audio-frequency oscillator is calibrated at 128, 256, 512, 1024, and 2048 cycles by means of tuning forks. A 60-cycle point can be obtained from the commercial power supply. Higher and lower frequency calibrating points are obtained by using the crystal oscillator, the radio-frequency oscillator, and the monitor in much the same way that the radio frequencies are determined. Thus the beat between the crystal oscillator and the radio-frequency oscillator may be set at 2048 cycles with the tuning fork. As a result of the harmonic frequencies generated in the monitor, a zero beat will be obtained when the audio-frequency oscillator is set at 2048, 4096, 8192, etc. By pushing the beat between the crystal oscillator and radio-frequency oscillator up to match one of these high audio frequencies, a step by step calibration is thus worked out. Cross checks, with odd harmonics, serve to increase the accuracy as much as may be desired.

Recalibration at frequent intervals is necessary as there is no fixed frequency to check against as in the case of the radio-frequency oscillator. The frequency shift observed in use has been very small, being approximately 500 cycles in 40,000 cycles after six months' use.

VI. Modulation Measurement

A radio transmitter is essentially a radiator of energy so modulated as to conform to the characteristics of the desired frequency of transmission. The amount of this modulation must vary with amplitude and may vary with frequency. In any case, the response from the receiver will vary with this modulation, and in testing a receiver this amount of modulation must be known.

It may be shown that in any conventional modulation scheme, the energy radiated consists of three parts. These parts are transmitted on three separate frequencies. The first is known as the carrier, and is usually the frequency of the radio-frequency oscillator before it was modulated. The second and the third parts are quite similar in that they usually have the same energy content and are transmitted on frequencies symmetrically placed with respect to the carrier and differing from the carrier by the frequency of the modulation. In the mathematical derivation of the formulas connected with this modulation phenomenon, the current or voltage amplitudes of these three parts are usually considered rather than the power relations. With this understood, the term "percentage of modulation" can be used without error. Suppose a carrier wave has a reference value of unity, and each other part, termed sidebands, a corresponding value of 1/2. As the relative phases and amplitudes are at all times changing, it is conceivable to have a time when the two sidebands are in phase, that is, add to a value of unity, and are out of phase with the carrier. The net result at such a time is zero. The carrier is then said to be completely or "100 per cent" modulated. There will also be a time when both sidebands and the carrier add in phase, and the net result will be an amplitude twice the value of the carrier.

The percentage of modulation, designated as K, is therefore the ratio of the sum of the sidebands to the amplitude of the carrier. As it is more convenient, both in plotting curves and in the mathematics, peak values of voltage and current are usually used.

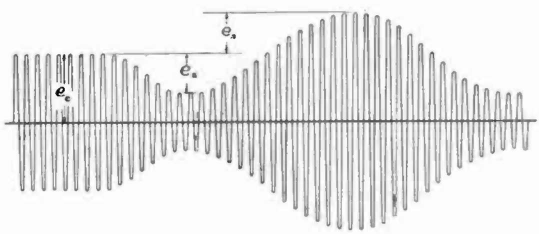


Fig. 13—Modulated Carrier Characteristic.

Fig. 13 depicts how a carrier wave looks before and after modulating. e_c represents the amplitude of the carrier and e_s the sum of the sidebands. By definition $K = e_s/e_c$.

The accurate measurement of K depends on using the curve shape as shown in Fig. 13. The so-called modulation meters on transmitters and broadcast studio control boards are merely meters indicating the relative magnitude of the audio-frequency voltage present. There are several methods of actually measuring K. The first, and by far the most common, is by the use of a cathode ray oscilloscope. An unmodulated carrier wave with an amplitude e_c is impressed on the control electrode. The length of the line on the target represents the amplitude of the carrier wave. The oscillator is then modulated and the new line observed. It will correspond to the maximum value of the envelope or $e_c + e_s$. The difference of the two readings divided by e_c gives K. This method has several disadvantages. The source of the carrier wave should be under control. The cathode ray oscilloscope is expensive. The value of K cannot be determined with much accuracy because of the necessity for scaling the length of the lines. The biggest disadvantage lies in the fact that the effective value of the carrier e_c may change when modulated. In certain methods of modulation such as that of Heising, the carrier wave may be substantially constant (if properly adjusted), but in other methods the carrier amplitude may change sufficiently to destroy the value of the determination.

If a double element oscillograph is used, the objection based on the change of carrier amplitude may be avoided. The amplitude of the carrier is shown by the distance between two reference lines, one of which is the base line for the superimposed audio frequency. When modulated, the audio frequency is observed as a sine wave rather than merely a trace. The mean of the extreme swings of this frequency is taken as the carrier amplitude, and K is figured as the percentage amplitude of the audio-frequency swing compared with the mean amplitude. If the carrier amplitude does not change when modulated, the mean of the audio swings will remain in the same line as the original base line. The enormous cost of such a unit is prohibitive to many would-be progressive laboratories. And while a visual estimation of K can be made, a scale must be provided and K calculated to obtain any degree of accuracy.

Another method is to use a peak voltmeter to determine e_c and e_c+e_s. This method is subject to practically the same disadvantages as the cathode ray oscilloscope method and is still more inaccurate because of the difficulty of determining the peak voltage accurately.

Another method is to use a peak voltmeter so arranged as to measure first e_c+e_s and then e_c-e_s. The mean of the two readings is e_c effective *while modulated*. This eliminates a large possible error, but is still comparatively inaccurate due to the use of the peak voltmeter. All of these methods have the disadvantage that the percentage of modulation must be calculated. They are in no sense "direct reading."

Considering the use of the measuring equipment, it was evident that none of these methods suited the demands of speed, reliability, accuracy, and direct reading. Based on the theory of the vacuum-tube voltmeter, a design was finally evolved which has proven very satisfactory.

If a radio-frequency carrier of frequency $p/2\pi$ is modulated by an audio frequency of $q/2\pi$ a voltage results expressed by

$$e=E\cos pt(1+K\cos qt) \tag{14}$$

where e is the voltage at a time t, and E is the peak voltage of the carrier wave. If this voltage is impressed on the grid of a vacuum-tube voltmeter with the customary plate and bias conditions, the plate current will have a form represented by substituting the value of e given in (14) into (5) as developed for the vacuum-tube voltmeter. This results in

$$i_p = E\cos pt(1+K\cos qt)\frac{dI_p}{dE_t} + \frac{E^2\cos^2 pt(1+K\cos qt)^2}{2}\frac{d^2I_p}{dE_t^2} + \frac{E^3\cos^3 pt(1+K\cos qt)^3}{6}\frac{d^3I_p}{dE_t^3} + \frac{E^n\cos^n pt(1+K\cos qt)^n}{\lfloor n}\frac{d^nI_p}{dE_t^n}. \tag{15}$$

To obtain all the components of the plate current, (15) must be expanded.[3] For the purposes of this development only the direct current and audio-frequency currents are of interest and all radio-frequency terms will be dropped. The operating point on the tube is so chosen that only the first and second derivatives have any effective value, the remainder of the series being zero. This point is easily obtained in use. With this simplification, the plate current may be shown to be

$$i_p = \frac{E^2}{2}\frac{d^2Ip}{dE_t^2}\left[\frac{1}{2}+K\cos qt+\frac{K^2}{4}+\frac{K^2}{4}\cos 2qt\right]. \tag{16}$$

The first term inside the bracket of (16) represents the increase in plate current due to the rectified carrier wave. The second term represents the audio frequency of modulation with an amplitude proportional to K. The third term represents an increase in plate current resulting from the sideband input, and is present only when the carrier is modulated. The last term is of double audio frequency, representing the two sidebands beating each other.

If the detector coefficient d^2I_p/dE_t^2 is known for the tube used, and if E and the audio current is measured, K can be calculated. Because of impedance in the plate circuit, the detector coefficient may not be constant with respect to frequency. As the voltage E is modulated, E itself cannot be read accurately. The double frequency term causes error. A more accurate method of determining K may be obtained by breaking (16) into two parts where

[3] The details of this expansion are not given here. The binomial theorem and some trigonometric equations serve to reduce this equation to first order terms of a single frequency.

$$i_p(\text{d.c.})=\frac{E^2}{2}\frac{d^2I_p}{dE_t^2}\left[\frac{1}{2}+\frac{K^2}{4}\right] \tag{17}$$

and

$$i_p(\text{a.c.})=\frac{E^2}{2}\frac{d^2I_p}{dE_t^2}\left[K\cos qt+\frac{K^2}{4}\cos 2qt\right]. \tag{18}$$

In the operation of the modulation meter, the modulated input is adjusted to give a constant d.c. change in the plate current, regardless the value of K. Let this value of current be C. Then

$$C=\frac{E^2}{2}\frac{d^2I_p}{dE_t^2}\left[\frac{1}{2}+\frac{K^2}{4}\right]. \tag{19}$$

From which

$$\frac{E^2}{2}\frac{d^2I_p}{dE_t^2}=\frac{4C}{2+K^2}. \tag{20}$$

Substituting (20) in (18) gives

$$i_p(\text{a.c.})=\frac{4C}{2+K^2}\left[K\cos qt+\frac{K^2}{4}\cos 2qt\right]. \tag{21}$$

Inspection of (21) shows that K is absolutely determined if the a.c. component of the plate current is determined. This is obtained by using a vacuum-tube voltmeter and measuring the voltage drop across a known resistance in the plate circuit.

It may be seen from (21) that the current at a frequency $q/2\pi$ is always associated with another current at a frequency $2q/2\pi$, the relation depending upon the value of K. In calibrating the vacuum-tube voltmeter a single frequency is used. Therefore, to read the effective sum of the two terms in the bracket of (21), this must be corrected to give a voltmeter reading independent of the two frequencies present. This may be done by remembering that the effective voltage sum of two voltages is equal to the square root of the sum of their squares. Thus the current is

$$i_p(\text{a.c.})=\frac{4CK}{2+K^2}\sqrt{1+\frac{K^2}{16}}. \tag{22}$$

If K is 100 per cent or 1, the error resulting from not making this correction is about 4 per cent. In general, this is unnecessary.

To complete the calibration of the modulation meter, the audio voltmeter is calibrated and a curve drawn. Some percentage of modulation, say 25 per cent, is chosen to give full scale reading on the audio voltmeter. Having chosen a value of C, in the present case 100×10^{-6}

amperes, the corresponding plate current may be found to be 48×10^{-6} amperes, peak. Knowing the sensitivity of the voltmeter, say 2 volts, peak value, full scale, the necessary plate resistance is found to be $2/48 \times 10^{-6}$ or 42,000 ohms. Therefore using this value of coupling resistance, the percentage modulation will be 25 per cent whenever the voltmeter indicating meter reads full scale, and providing the input has been adjusted to make C correct. Other percentages of modulation

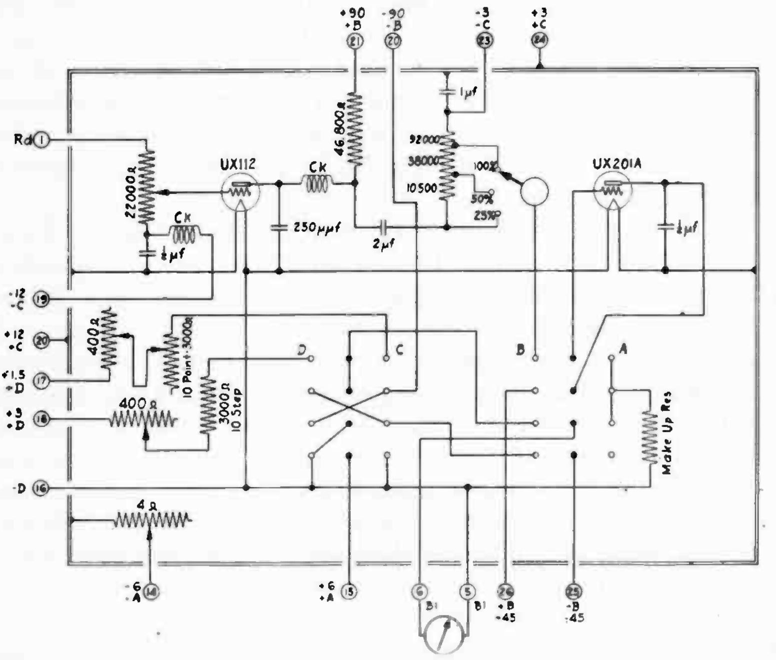


Fig. 14—Modulation Meter Circuit.
Sequence:
A closed, *C* closed, adjust for emission.
B closed, *C* closed, adjust for zero.
B closed, *D* closed, adjust for zero.
B closed, *C* closed, adjust input for *C*.
B closed, *D* closed, read *K*.

can be plotted by assuming K, determining the plate current (a.c.) multiplying by the plate resistance to obtain the audio voltage and then noting the corresponding meter reading.

In use, the plate circuit of the demodulator must be separated from the grid circuit of the audio voltmeter because of the potentials necessary to operate the tubes correctly. Therefore, a resistance-capacity coupling unit is used. The parallel resistance of the combination is used as the effective plate resistance. Taps on the grid resistance serve to vary the sensitivity and provide full scale readings of approximately 25 per cent, 50 per cent, and 100 per cent modulation.

It is important to notice in this development that the characteristics of the demodulator do not affect the reading in the least, these characteristics being cancelled out in the simultaneous solution of (17) and (18). A UX201A, a UX171 or a UX240 work equally well as a demodulator, providing sufficient input is available to give the required change C in the plate current. This independence in the determination of K of the characteristics of the demodulator is the surest proof of the accuracy of the method. The cathode ray and peak voltmeter methods are not in themselves sufficiently accurate to check this modulation meter except in a general way. Using the method indicated in the proof, of using d^2I/dE_t^2 as a check, establishes the accuracy of the method.

The circuit and approximate values are shown in Fig. 14. A radio-frequency choke and a small bypass condenser serve to eliminate the radio frequencies before reaching the voltmeter. A single meter, a Westinghouse 100 μa full scale, is used for several purposes. The emission is set on the voltmeter as previously described for accuracy of calibration, and the meter, properly connected and shunted, serves to indicate the correct value. The meter is then connected successively in the plate circuits of the first and second tubes, and the proper "bucking" current controls adjusted to give zero reading on the meter. All calibrations are made from this zero reading position. The meter is then switched back into the plate circuit of the first tube, and the modulated input adjusted with the input potentiometer to give the full scale reading C. The meter is then cut back into the second tube plate circuit where the reading, referred to the calibration curve, gives the percentage of modulation direct.

Through lack of other more accurate means of checking the percentage of modulation, the reading cannot be depended upon much better than within 5 per cent of the value read. The recommended value of K for broadcast receiver testing is set at 30 per cent by the Standardization Committee. The actual value of modulation used is probably between 29 per cent and 31 per cent, a truly negligible error. Recalibration after six months' use showed no change in characteristics, the original calibration checking within less than 1 per cent. Its service has been remarkably consistent and useful.

VII. Radio-Frequency Output Measurement

Having measured the input to the receiver, it is comparatively simple to measure the output. The problem is to measure the output without affecting the normal operating characteristics of the set. The only real point of difficulty is in measuring the radio-frequency amplification from the antenna to the detector. Any voltmeter connections across the tuned circuit feeding the detector will change the tuning, increase the feedback to the antenna, and give a false measurement.

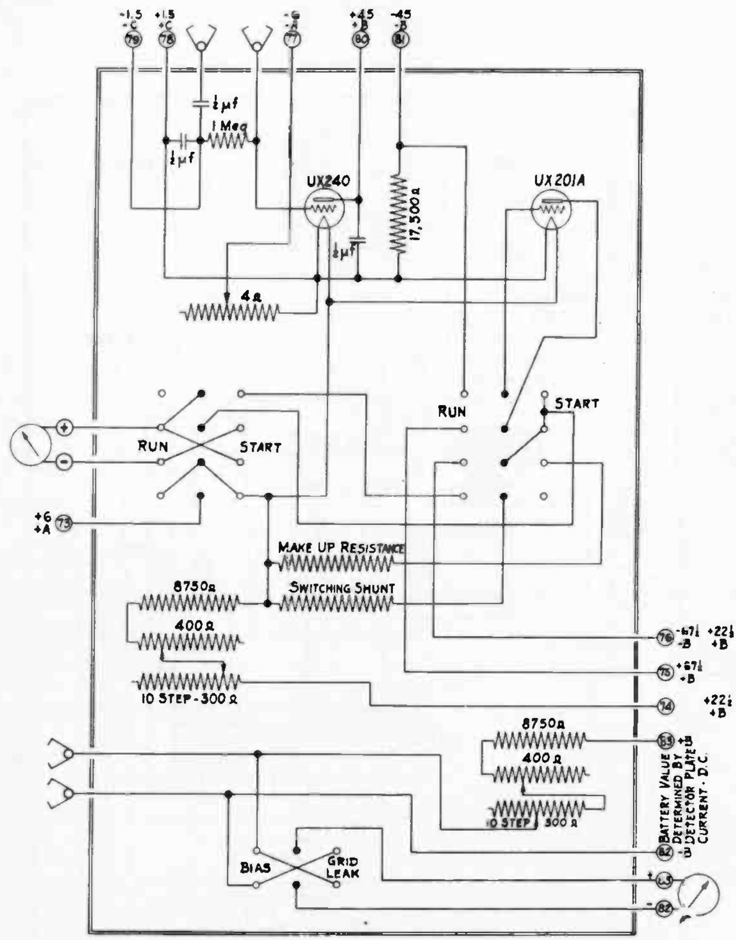


Fig. 15—Detector Calibrating Voltmeter Circuit.

These difficulties can be avoided if the detector tube in the set is used as a voltmeter. It must first be calibrated by noting the change in plate current with applied radio-frequency grid voltage. The current is measured by opening the detector plate lead below any plate impedance which might affect the operation, and inserting a meter. This meter must be fairly sensitive, and a bucking current is provided to cancel the initial plate current and bring the meter to zero reading. The radio-frequency grid voltage is measured by temporarily connecting a previously calibrated vacuum-tube voltmeter across the tuned circuit supplying the detector. After the detector is calibrated as a voltmeter, the vacuum-tube voltmeter is removed and radio-frequency

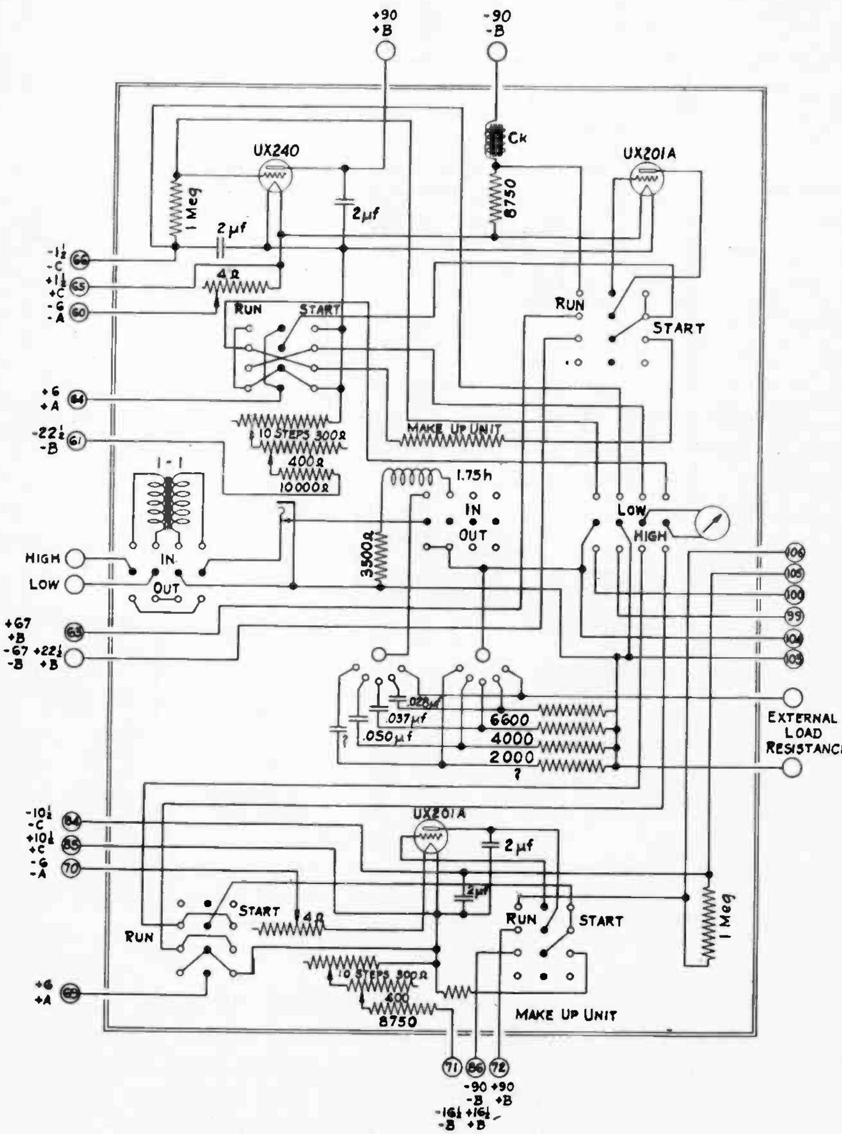


Fig. 16—Audio-Frequency Output Voltmeter Circuit.

amplification or selectivity can then be measured under operating conditions. Fig. 15 shows the circuit arrangement of the detector calibrating voltmeter and that for the detector-voltmeter. Because of wave form and overload errors with the grid leak and condenser type voltmeter, it is best to operate at low voltages. The detector voltage is usually adjusted to about 1/4 volt for the average set. Connections from the meter panel to the set are made with cords and tip jacks, with spring clips to connect at the proper points on the set.

VIII. Audio-Frequency Output Measurement

Several voltmeters are provided to read the audio-frequency output of the set. One is comparatively low reading, from 0.05 to 1.1 volts, and is used for the measurement of hum and low audio output. The other voltmeter has a range between 0.5 and 200 volts, and is used for interference and normal output voltage measurements. As these are never used simultaneously, a four pole double-throw switch serves to connect the input and output circuits properly so that only one meter is used. Except that the higher reading voltmeter uses only a single UX201A tube, they are similar to the voltmeters previously described. The complete circuit arrangement is shown in Fig. 16.

When measuring a radio receiver, care must be taken to prevent any direct current from the power tube supply system from reaching the voltmeters. If a filter circuit is provided in the receiver, this precaution is unnecessary. If there is no filter supplied, a switch cuts in a 1 to 1 output transformer built to carry the plate current of large power tubes without saturating, and having substantially a flat characteristic from 30 to 10,000 cycles. As recommended by the Standardization Committee, a load impedance equal to the internal plate filament impedance is used. This is connected across the transformer secondary, and the voltage across this load resistance measured. For convenience, five taps are provided giving 2000 ohms for a 171A, 4000 ohms for a push pull 171A, and 6600 ohms for a 120. One tap is left open for future needs to determine the value. The fifth tap connects to external posts on the panel to enable any value of load resistance to be connected.

A measurement of major importance in the development of modern a.c. operated receivers is the hum voltage present in the output. Usually the low reading voltmeter is necessary for this measurement but, unfortunately, the higher reading voltmeter has been necessary on some receivers. Because of the non-linear frequency response of the majority of present day loud speakers, some correction of this hum voltage should be made to represent the audible value of this interference. The Standardization Committee recommended a filter system, with values approximately as shown, to make this needed correction. Readings of the hum voltage on a receiver are taken both with and without the hum filter to aid in development or compare receivers. As a different capacity is required with each load resistance, these units are controlled by the same switch that changes the load impedance.

In tuning the set, or adjusting the output, it is often undesirable to have the voltmeters connected. A single closed circuit jack is arranged to connect a loud speaker across the output of the receiver, and opening the voltmeter circuits.

Because of the nature of the output, especially as the overload point is approached, it is important that these voltmeters be carefully adjusted to avoid wave form errors. The values given are approximately right for most tubes, but are checked whenever a tube is changed. The calibration for these meters is accurate to about 1 per cent and is very constant through long periods of time.

IX. Overload Determination

An extremely important measurement in the development and comparison of modern radio receivers is that of distortion. Distortion, or lack of fidelity to the original modulation frequencies, may result from three causes. The output may not contain those frequencies and corresponding amplitudes as were present in the input. The amplitude of the output on any single frequency may not be proportional to the input amplitude. The output may contain frequencies not present in the input.

From the previous discussion it is obvious that proper use of the measuring equipment may quantitatively evaluate the distortion resulting from the first two of these forms of distortion. The measurement of spurious frequencies in the output is somewhat more difficult. Several methods have been used, all having certain advantages and disadvantages. Using a double element cathode ray oscilloscope, and comparing the output voltage with the input voltage, gives one method of measurement. With no other frequencies present, a straight line results on the target. With additional harmonics present, a curved or bent line results. Proper analysis of this line shape gives the individual magnitudes of the various harmonics present. The method is fairly accurate, but rather slow, and requires that an exact replica of the input, in proper amplitude (and preferably in proper phase, or an ellipse results) be available to apply to the oscilloscope.

A second method is to use tuned circuits to pick out the harmonics, and measuring their amplitudes individually. This requires sharp tuning circuits, often difficult to obtain at low frequencies. It also means that the entire range of possible harmonics must be measured to obtain the total harmonic power present in the output. The method is rather slow, but has an advantage over the oscilloscope method in that it is not necessary to compare the output with the input.

A third method is to use a frequency bridge to cancel the fundamental frequency and leave only the harmonics. A properly designed "square law" vacuum-tube voltmeter may be used to measure the effective value of all of the harmonic voltages, and therefore evaluate the total harmonic power component. This gives no indication of the relative amplitudes of the individual harmonics, often quite desirable, but serves to measure the total harmonic power output from a receiver.

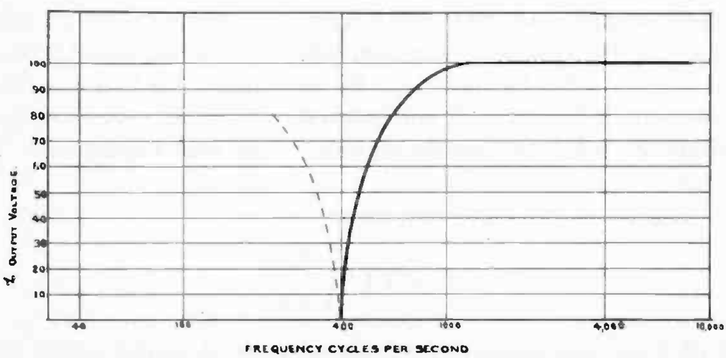


Fig. 17—Distortometer Transmission Characteristic.

The frequency bridge balance method was chosen as being the most desirable in carrying out the purpose of this measuring setup. In order to be effective, and have the accuracy desired, it was necessary that the bridge balance out the fundamental to less than 1 per cent of its original value, and that all frequencies above the second harmonic be transmitted at least 90 per cent. A Campbell frequency bridge circuit was used as the basis of this unit. The ratio between the mutual inductance and the capacity was chosen to give a sharp cutoff near the balance frequency. To accomplish this throughout the audible frequency range, five steps of inductance are used, combined with the proper capacity for balance. The frequency ranges are roughly 30 to 100 cycles, 100 to 300 cycles, 300 to 1000 cycles, 1000 to 3000 cycles, and 3000 to 10,000 cycles.

A typical transmission characteristic is shown in Fig. 17, where the bridge was tuned to 400 cycles. The transmission at 400 cycles is negligible, while that at the second harmonic, 800 cycles, is about 92 per cent. Assuming that the greater part of the harmonic power is present at frequencies of the second, third, and fourth harmonics, the average transmission characteristic is about 96 per cent. This assumption will lead to about 4 per cent error in the value measured if all the harmonic energy is concentrated in the second harmonic or in a harmonic above the fifth. The average error resulting from this unknown distribution of harmonic energy will be about 2 per cent of the value read, a truly negligible error.

To make the instrument direct reading in percentage of harmonics, the input is always adjusted to a value of 5 volts, whereupon any harmonic voltmeter reading will always correspond to some specific percentage of harmonic. Thus, if the harmonic voltage component were 5 per cent of that of the fundamental frequency, the vacuum-tube voltmeter would read 0.24 volt. This may be obtained as follows. Assume that the power is concentrated on the fundamental frequency f_1 and any one other harmonic f_n. By assumption (as based on its use) the total applied voltage, 5 volts, equals $\sqrt{E_1^2+E_n^2}$. It is desired to determine E_n when the output contains δ per cent harmonics.[4] Then $E_n=\delta_n E_1$

Solving these two equations gives

$$E_n=5\sqrt{\frac{\delta_n^2}{1+\delta_n^2}}. \tag{23}$$

With 5 per cent harmonic, as assumed above, δ_n is 0.05, and $E_n=0.25$. As a result of the transmission characteristic of the frequency bridge, averaging 96 per cent, the actual voltage reading will be 0.24 volt.

The calculation of sufficient points in this manner and the translation of the actual voltage-meter readings to the distortion factor-voltage calculation enables a direct reading curve to be drawn. The instrument so designed and directly calibrated has been named a "distortometer."

As may be seen from the development of (23), the harmonic voltage E_n is always used as E_n^2. Therefore E_n^2 may be considered as the effective voltage squared, resulting from the summation of the power in all the harmonics, and the method is general in application. That is, the individual harmonics may have any frequency and amplitude relation and the distortometer will still give the correct percentage of effective voltage, referred to the fundamental, in the harmonics.

[4] The symbol δ is here used for the distortion factor and in general use signifies the percentage of voltage, referred to the fundamental, in any harmonic denoted by the subscript. Without the subscript, δ signifies the total effective voltage component of all harmonics in percentage of the fundamental.

The voltage-meter reading calibration is obtained by impressing known voltages on the input of the distortometer and unbalancing the bridge sufficiently to give practically 100 per cent transmission. The calibration, as may be seen from the circuit diagram, (see Fig. 18) is one which includes the input tube and the bridge circuit as well as the voltmeter tube itself.

The output voltmeter is used in adjusting the input to the distortometer to exactly 5 volts. As a result of this setup, the output attenuator was placed in the distortometer cabinet. The complete circuit of

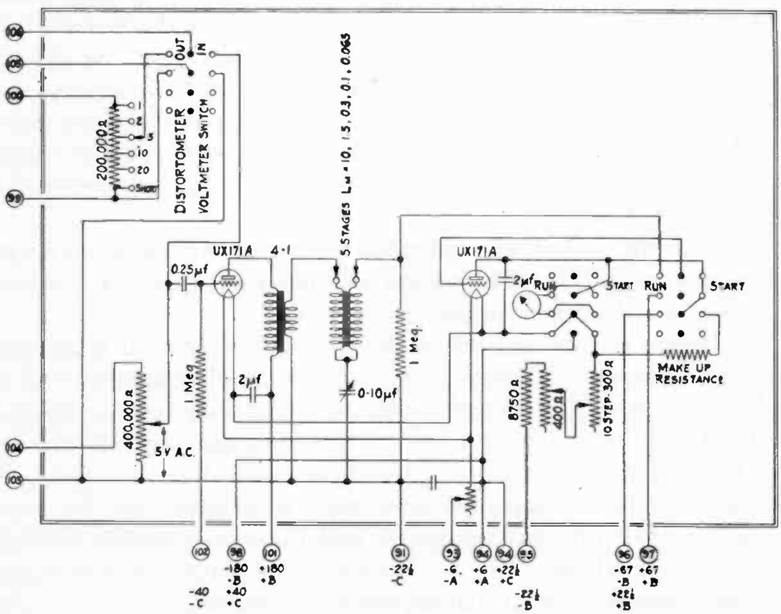


Fig. 18—Distortometer Circuit.

the connections of the output voltmeter and distortometer is shown in Fig. 18. The operation is briefly as follows. The output of the set, at some specified audio frequency, is cut into the distortometer by raising the arm on the 400,000-ohm input potentiometer. The bridge is then tuned by adjusting the mutual inductance and the variable capacity to give a minimum reading in the distortometer output meter. If the output of the set is low, i.e., no harmonics present, this minimum will be practically zero.

The output voltmeter is connected across the input terminals of the unit by throwing the proper switch to the "in" position. The input to the distortometer is then set at 5 volts by proper adjustment of the

400,000 potentiometer. The reading of the output meter on the distortometer gives the percentage harmonics. If it is desired to determine the power output of the set at some particular value of harmonic, say 5 per cent, the set output may have to be increased. As the set output is increased by proper use of the volume control or by increasing the radio input, the input potentiometer to the distortometer is gradually backed off, keeping the input as indicated by the output voltmeter constant at 5 volts. The output meter on the distortometer will gradually rise, indicating an increasing percentage of harmonic voltage. When the distortometer indicates an output reading of 0.24 volt, the output voltmeter is taken out of the distortometer circuit and cut back across the output load of the set by throwing the switch to the "out" position. This measures the output voltage of the set, and knowing the load resistance, the power output may be calculated. As the distortometer has indicated the percentage of power in the harmonics (0.25 per cent), the set output can be corrected for this amount if desired.

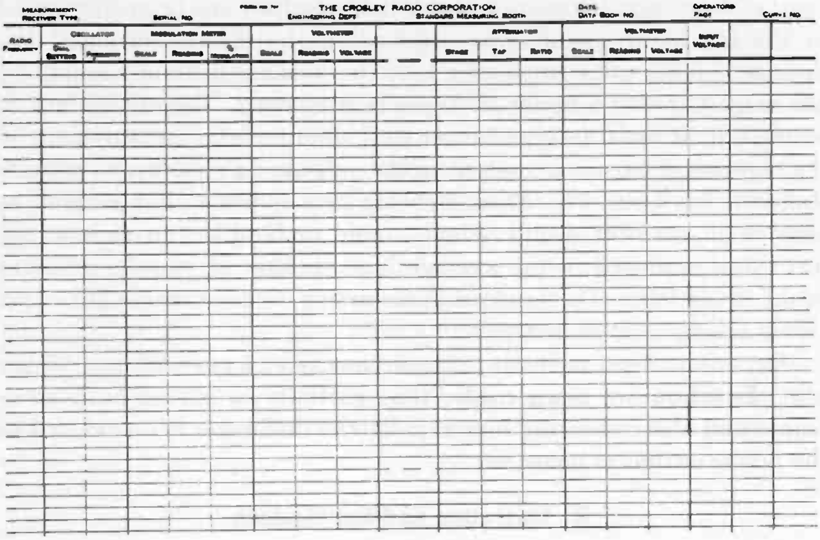
The circuit values are indicated in the drawing and no lengthy comment is necessary. The output voltmeter is adjusted for emission and set as are the other voltmeters described.

To obtain a sharp cut-off on the frequency bridge, it is necessary that the generator supplying the voltage be of comparatively low impedance. Also, this bridge should not load the circuit being measured. A UX171A, having a low plate impedance and followed by a 4 to 1 step down transformer, serves to accomplish both objects. The values of mutual inductance used were determined by the cut-off necessary, their internal resistance, and those coils readily available. The ratio is in all cases 1 to 1. The coils were placed on their cores so that no correction for mutual capacity was necessary.

It is extremely important to use a low flux density in these cores. Saturation otherwise will occur and the harmonics developed in these inductances may be greater than that from the set. This is particularly true on projecting points where crowding of the flux may saturate a small portion of the iron even though the major iron path is quite all right. Until discovered and corrected, this error proved an exasperating problem.

Having a small variable condenser, the capacity range is continuously variable from almost zero to 10 μf, and enables the bridge to be quite accurately set.

If some definite amount of harmonic content is agreed upon as the maximum permissible, it is possible to specify the "maximum undistorted power output" of a receiver. Perhaps a better way of expressing



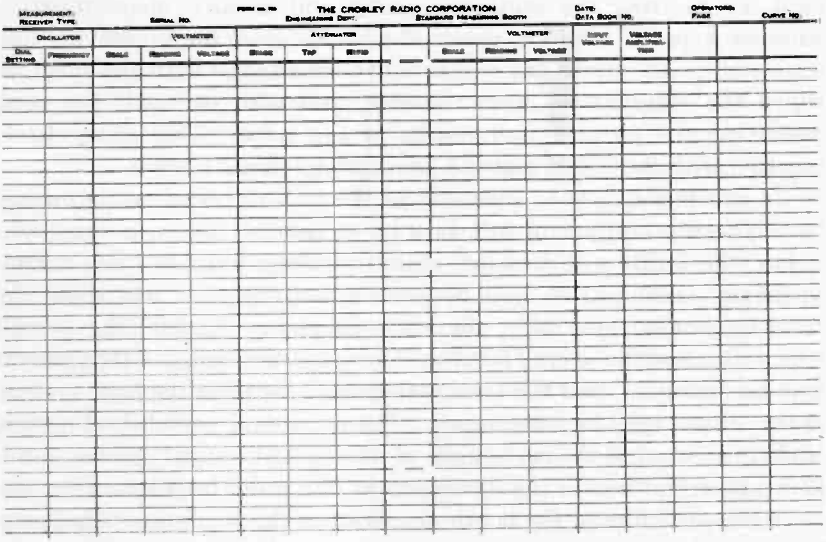


Fig. 19—Typical Data Sheets

this would be the "overload point." Among engineers[5] a tacit agreement of 5 per cent harmonic voltage in the output has been considered as the maximum permissible. Believing the discrimination of the average listener to be somewhat less, the Standardization Committee has suggested that a figure of 10 per cent be used. Because of lack of correlation of aural interpretation with simultaneous measurement of the percentage harmonic present in the output of a receiver, no definite standard has been set. Measurements here indicate that a figure as high as 20 per cent would be even more preferable. In view of the distortion produced in the average loud speaker at various resonant peaks, the setting of a standard of distortion, or permissible distortion factor δ, may well be postponed.

For comparison purposes, measurements at 5 per cent and 15 per cent distortion are being made, thus partially satisfying both viewpoints and also indicating how rapidly the distortion is introduced as the power output is increased.

X. Operation and Conclusions

For convenience in handling all of this equipment, printed data sheets are provided for the recording of the observed data. Two of these charts are reproduced in Fig. 19. Semi-log ruled curve sheets are used for plotting the audio-frequency and fidelity characteristics. Another type of semi-log curve sheets are used for sensitivity and selectivity. All curves are numbered to correspond with the sheet on which the observations were recorded, and with the daily log book describing the purpose and results of the tests. This cross check, together with the index, makes a complete reference system.

In conclusion, a brief summary of the major operations performed by this testing equipment will show its usefulness. By proper manipulation, the outfit will measure radio-frequency amplification, audio-frequency amplification and detector sensitivity. It will show the radio-frequency selectivity, the audio-frequency fidelity, the overall selectivity, and the overall fidelity. The frequency range of the receiver may be measured, and the hum voltage and the effective hum voltage of the output may be determined. The maximum undistorted output of the receiver and the percentage of overload at output can be found. It is a powerful tool in the development of a radio broadcast receiver.

Within the files of the booth described in the preceding pages may be found the performance records of practically every modern radio receiver. The vices and virtues of these receivers are revealed with a completeness that might be shocking to their respective designing

[5] This is largely based on tube distortion calculations.

engineers. But only as the vices of these receivers are known and corrected, and the virtues seen and appreciated, may the technical operation be improved. It is hoped that this paper may aid in placing the performance rating of a broadcast receiver upon the proper basis, and that from this basis, the improvement in performance may be never ending.

Addendum

In accordance with the idea of keeping this equipment up to date, several minor changes have been made to facilitate the operation since

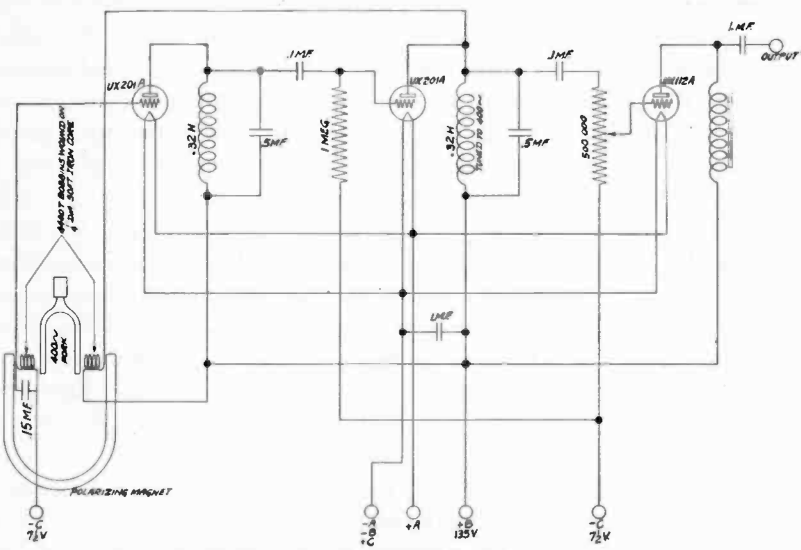


Fig. 20—400-cycle Tuning Fork Oscillator.

this description was written. The advantages demonstrated by the use of a new oscillator are so pronounced that this further addition to the paper seems justified.

Since a considerable portion of the work of measuring the characteristics of a radio receiver is done at 400 cycles, it was felt desirable to install an electrically driven tuning fork to supply power at this frequency. The circuit is shown in Fig. 20, and does not require much comment. The tuning fork used was made by the Waverly Musical Products Corporation. The oscillating circuit consists of the fork circuit, two vacuum tubes, and associated circuits tuned to approximately 400 cycles. Two tubes were found necessary to maintain the proper amplitude and constancy of oscillation. A UX112A as an output tube was found to deliver sufficient power to modulate the radio-frequency oscillator to a value of about 60 per cent modulation, and consequently no greater power output system was included. The output contains less than 1 per cent of voltage harmonics, as the wave form is extremely pure. This depends somewhat on the amplitude of oscillation; with the characteristics of the tuned circuits used no amplitude control was found necessary.

The system is rather susceptible to mechanical vibration, and to insure a good wave form the entire unit was mounted on a heavy brass plate and suspended on a soft cushion support. This unit was placed in the same cabinet as the crystal oscillator, some re-arrangement of the parts and controls on the panel being necessary. An output switch (not shown on the diagram) serves to connect the output to the radio-frequency oscillator for modulation or to the monitor for calibration uses, or to the audio attenuator for audio amplification measurements.

The use of this unit, in conjunction with the other apparatus, greatly simplifies the measurement of selectivity. The operation is exactly as previously described, except that the audio oscillator does not have to be re-set on 400 cycles after the radio-frequency has been determined. Extreme accuracy is possible. While valuable perhaps only as a trick, a selectivity curve can be obtained with the radio frequency at the points observed, differing by less than 100 cycles.

Another very valuable use for this oscillator is in the check and calibration of the audio oscillator. As a result of deliberate distortion in the monitor, many harmonics are produced from any applied input. Consequently when the 400-cycle oscillator and the audio oscillator are both connected to the monitor, a great number of "beat" calibrating points are obtained. The detail of calibration is limited almost solely by the ability of the operator to determine the order of the harmonic and thereby the absolute frequency.

It is certain that this oscillator is a valuable addition to the testing equipment, resulting in both a saving of time and an increase in the accuracy of its application.

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MEASUREMENTS OF THE HEIGHT OF THE KENNELLY-HEAVISIDE LAYER*

BY

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Summary—*In this paper we have sought to offer some further contribution to the Kennelly-Heaviside layer problem; first, in the form of experimental data showing clearly evidence of the diurnal cycle in layer height, and secondly, in the form of a discussion of methods for the interpretation of group time and phase retardation experiments and the problem of determining the relationship between the "virtual" and "true" heights. Methods of successive approximation for arriving at the "true" height, from group time or phase retardation measurements, are also discussed and applied. Close accord is found between the results of these methods and the approximation used by Schelleng in a recent paper. The results shown in Fig. 16 also indicate the necessity for further experiments in the important frequency range from 1 to 4 megacycles where no data are available.*

THE past few years have contributed much theoretical work and varied experimental data of importance to the problem of determining the properties of the Kennelly-Heaviside Layer and of predicting quantitatively therefrom radio transmission phenomena. It has become evident, however, that the phenomena involved are so many and varied that a considerable mass of experimental data will be necessary before even a semi-quantitative theory of radio transmission will be realized.

Our knowledge of the constitution of the upper atmosphere is so limited that it seems probable that many of its properties will have to be determined by suitably chosen radio transmission experiments before sufficient information will be available for constructing a quantitative transmission theory. Important in this group are experiments dealing with the height of the layer.

One of the objects of this paper is to present a small further contribution to this store of data. However, a brief summary and comparison of accumulated data in the light of recent theoretical contributions, and results readily deducible therefrom, seem appropriate at this time.

As has been frequently pointed out, the phenomena of short-wave transmission are more in the nature of a diffraction than a reflection,

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and the region of the atmosphere of importance in the turning back of waves may be many kilometers in depth. Under these conditions, the terms "layer" and "height of layer" may be misleading unless carefully defined. By "true height of layer" to a ray we will understand the highest point reached by the ray during its trajectory. By "thickness of layer" we will understand the difference between the

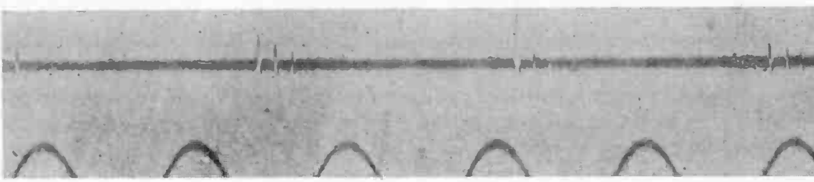

Fig. 1—Transmission from NKF on 4435 kc as Received at Philadelphia on the Evening of September 30, 1928.

height defined above and the lowest height from ground at which the electrical properties of the atmosphere produce an appreciable alteration of the wave trajectory (see Fig. 11). Both of the quantities defined above are, of course, functions of frequency and also the initial angle of the ray.

In one of the major types of experiments which have been used to measure the height of the Kennelly-Heaviside layer, the difference

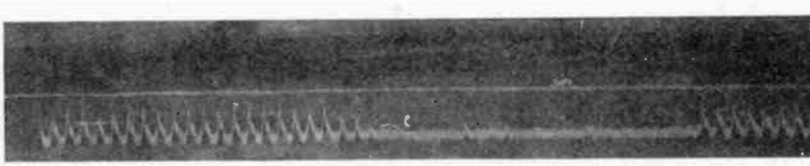

Fig. 2—Transmission with 500-cycle Modulation on October 17, 1928 at 1:45 P.M. Eastern Standard Time.
Note effects of multiple transmission paths.

in group retardation time for a sharp jab or pulse transmitted over two different transmission paths (as, for instance, the ground and first refracted wave paths) is measured.[1] Another of the major types of experiments measures the difference in wave number over several transmission paths.[2]

[1] Breit and Tuve, *Phys. Rev.*, 28, 554–575; 1926.
[2] E. V. Appleton and M. A. F. Barnett, *Proc. Roy. Soc.* A, 113, 1926. J. Hollingworth, *Jour.* I. E. E., 64, 579–595; 1926.

In each case computations of height have almost universally been arrived at by picturing the transmission to take place between two perfectly conducting, sharply defined planes or spheres at the surface of which the waves were reflected without phase change, although the investigators have in all cases been careful to emphasize that the heights so obtained were not the true heights as defined above. We will term the apparent height thus computed the "virtual height."

It is not a priori evident, for a given layer and frequency, that the results which would be obtained from these two types of experiments

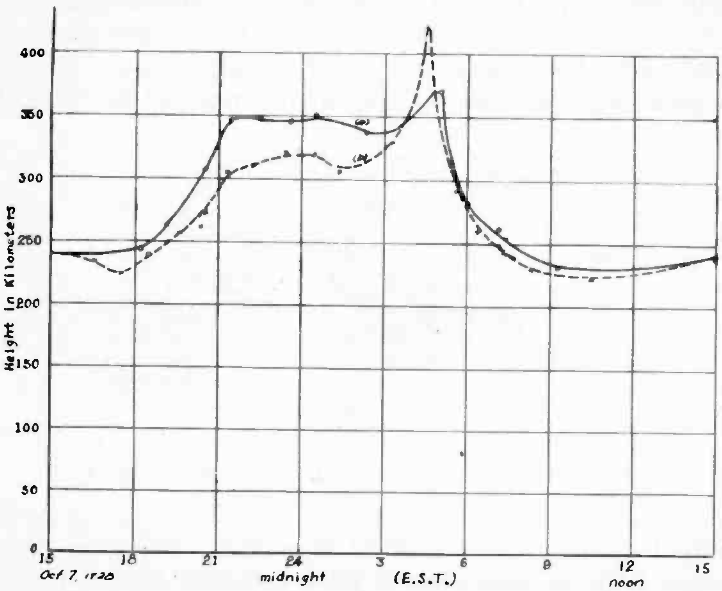

Fig. 3—"Virtual Heights" Obtained during 24-hour Observations, October 7th and 8th, 1928. (a) Observed at Moore School of Electrical Engineering, University of Pennsylvania. (b) Observed at Department of Terrestrial Magnetism, Washington, D. C. Transmission from Naval Research Laboratory, Washington, D. C. on 4435 kc.

would yield the same value for the virtual height, nor is it evident how great a discrepancy exists between the values thus obtained and the true height.

I. Recent Experimental Results

Recent experiments carried on at the Moore School of Electrical Engineering of the University of Pennsylvania in collaboration with the Department of Terrestrial Magnetism and Naval Research Laboratory have furnished further experimental data using the pulse method.

The transmission was in the form of single sharp, widely separated pulses produced by means of a multivibrator circuit already described in the I.R.E. PROCEEDINGS. The form of the groups of received pulses at Philadelphia (showing clearly the effects of multiple transmission paths) is shown in Fig. 1. In some tests, where 500-cycle modulation was employed, a received signal of the form shown in Fig. 2 was obtained.

In the experiments described here transmission was carried on from the Naval Research Laboratory at Washington (NKF) during alternate half-hour intervals. Simultaneous transmission was furnished from two separate transmitters on the frequencies 4435 kc and 8870 kc. Presumably as a result of skip distance phenomena the higher frequency was not consistently received in Philadelphia

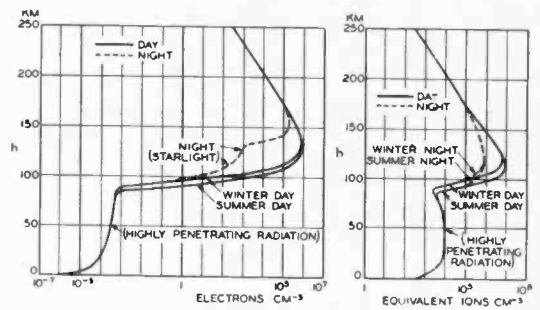

Fig. 4—Electronic and Ionic Densities as Functions of Height and Time (Pedersen).

or at the Department of Terrestrial Magnetism. Faint signals on the frequency 8870 kc were secured when 500-cycle modulation of the form shown in Fig. 2 was employed, but the pulse transmission was not audible or oscillographable at the Philadelphia station.

In Fig. 3 the virtual heights obtained from a 24-hour run on October 7 and 8, 1928, are shown. The full curve shows the results computed from data obtained at Philadelphia, while the dotted curve shows the results reported by the Department of Terrestrial Magnetism. It will be noted that the form of the curves are in essential agreement; i.e., each shows a consistent daytime height rising rapidly during the early evening, then more slowly and finally rising again rather abruptly to a maximum value just before dawn. After sunrise, the curves fall rapidly to the daytime value. A slight tendency for the night values at Washington to lie below those at Philadelphia is indicated by these curves, but the precision of measurement is not at present sufficient to render this difference conclusive for, due to oscillographic difficulties, the speed attainable is, at present, limited at Philadelphia to about that shown in Fig. 1. Under these conditions a 10-km difference in height is about at the limit of experimental precision.

II. Methods for the Reduction of Virtual Height Data

With the rapidly increasing store of group time data of the form presented above, and of data involving phase retardation measurements, some further quantitative consideration of the ratio of the virtual heights obtained by means of such measurements to the true height seems appropriate, for such a reduction is an essential pre-

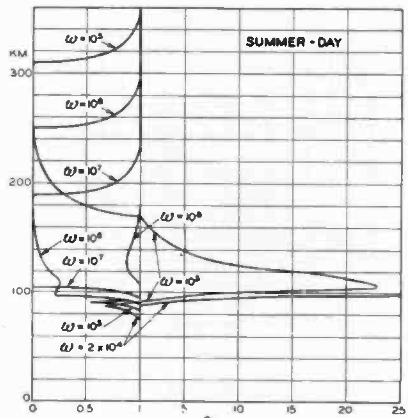

Fig. 5—Refractive Index as Functions of Height and Frequency (Pedersen).

requisite to a determination of the distribution of electronic and ionic densities in the upper atmosphere and its diurnal variation to which the shape of the curves shown in Fig. 3 are supposedly attributable.

Were this distribution known analytically, the problem of determining the ratio of "true" to "virtual" height (which we will denote by h/h') would be analytically formulatable, although the apparent complexity of the problem supposedly renders the chance of the exact integration of the differential equations involved somewhat dubious. In the absence of such data, it is necessary to secure from other sources some tentative evidence as to this distribution upon which preliminary calculations may be based. From these calculations it is possible, given sufficient data from radio experiments, to proceed by methods of successive approximations to an experimental determination of the actual distribution. These methods will be indicated in the course of this discussion.

After a comprehensive survey of the evidence available, Dr. P. O. Pedersen in a recent book[3] has evolved curves, giving what he considers the best estimates for the variation of electronic and ionic density in the atmosphere as functions of height and time. Some of the curves thus obtained are shown in Fig. 4. While the extremely limited available data upon which these curves are founded necessarily renders them controversial,[4] some of the general properties exhibited are in all probability in accord with the truth.

A substitution of the data on electronic and ionic densities into the well known equations for effective dielectric constant and conductivity as obtained and applied by Nichols and Schelleng, Taylor and Hulburt, Breit and Tuve, and others,[5] gives the index of re-

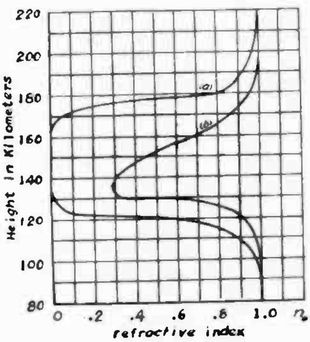

Fig. 6—Refractive Index vs. Height. (a) 4435 kc. (b) 8870 kc.

fraction curves shown in Fig. 5. A curve for the frequencies 4435 kc and 8870 kc is shown in Fig. 6.

As has been frequently pointed out, the elementary considerations of geometrical optics[6] show that, neglecting dissipation and magnetic resonance phenomena, the maximum depth of penetration of a ray does not exceed that at which the refractive index equals the sine of the initial angle of departure of the ray with the vertical. From Fig. 5 it is evident that according to Pedersen's curves for the ionic and electric distribution with height, for $\omega \backsimeq 3 \times 10^7$, the actual penetration

[3] See "The Propagation of Radio Waves," by P. O. Pedersen. "Danmarks Naturvidenskabelige Samfund," Copenhagen, 1928.

[4] Rather different estimates have been recently arrived at by Hulburt which are not in accord with Pedersen's estimates of temperature or with his assumption of the lack of diffusion in the upper atmosphere. Hulburt's results give much higher values for the height of the layer. See *Phys. Rev.*, 31, 1928.

[5] See Pedersen, page 122. He carefully reviews this material, and extends it to take into account dissipation terms, in many cases omitted in the original treatment.

[6] See p. 175, Pedersen.

of rays and hence the length of curved path in general is small. It appears possible, therefore, that the curvature of the wave trajectory and errors introduced by the assumption of a triangular trajectory may, in some cases, have received somewhat exaggerated emphasis. A curve showing the height of layer as a function of frequency and ϕ_0 as obtained from Pedersen's n curves is given in Fig. 7.

Let us consider this question in greater detail with a view to estimating the magnitude of departures between real heights and virtual heights estimated on the assumption of a triangular trajectory and

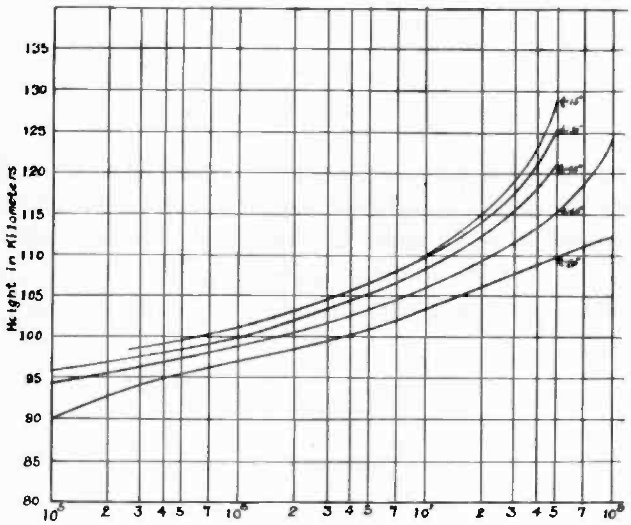

Fig. 7—Height of Layer as Functions of Frequency and ϕ_0 (the initial angle of departure from vertical) as Calculated from Pedersen's n curves (Fig. 5—summer day).

Note that the height thus obtained is probably low.

a velocity of propagation equal to that of light. Breit and Tuve have proved the interesting and important theorem that, in the absence of dissipation or important magnetic effects, the time required for the group signal to traverse the actual curved trajectory is the same as would be required for the group to traverse with the velocity of light a triangular path with the same initial angle of departure as the actual trajectory (see Fig. 8). It is only necessary, then, to compute the equation of the actual path, and to determine the ratio of its maximum ordinate to that of the apex of the triangle in order to compute the value of $\frac{h}{h'}$ for the case of the group time type of experiments.

III. Differential Equation for the Ray Path[7]

Let us compute, on the diffraction theory, the differential equation of a ray starting at an initial angle ϕ_0 in a medium of gradually varying index of refraction n (which we will subsequently consider as a function of y). We will assume the ray to start at a point where the initial

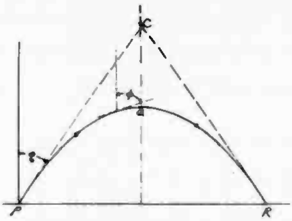

Fig. 8—Equivalent Wave Triangle.

index of refraction is unity. Our fundamental formula of refraction gives us for the angles of incidence and refraction between two media $n_0 \sin \phi_0 = n_1 \sin \phi_1$; when applied to a gradual boundary this becomes[8]

$$n \sin \phi = n_0 \sin \phi_0 = \text{Constant} \qquad \text{(III-1)}$$

Since $n_0 = 1$ in vacuo, (III-1) becomes $n \sin \phi = \sin \phi_0$

$$\frac{n^2}{\sin^2 \phi_0} = \frac{1}{\sin^2 \phi} = 1 + \cot^2 \phi$$

but
$$\frac{dy}{dx} = \cot \phi$$

$$\therefore \frac{dy}{dx} = \sqrt{\frac{n^2}{\sin^2 \phi_0} - 1}. \qquad \text{(III-2)}$$

Which is the required differential equation of our path.

IV. The Variation of n as a Function of y

In order to determine the equation of the ray trajectory in integrated form, we require the analytical equation for n as a function of y. Pedersen's computations of ionic and electronic density from an assumed atmospheric constitution (the results of which are shown in Fig. 4), and the equations which connect these estimates with the refractive index n are sufficiently involved to render small the chance of the determination of a general equation for n as a function of y

[7] See Pedersen, p. 171.

[8] In this analysis we are neglecting earth curvature, an assumption which greatly simplifies the subsequent integrations. Most of the experiments so far carried out are not over sufficient distances to render earth curvature important.

which will satisfactorily express the entire family of curves for n shown in part in Fig. 5, and at the same time when substituted in equation (III-2) result in integrals which can be evaluated in terms of known functions.

Fortunately, however, the portions of the n curves which are of importance in our calculations of ray paths as a function of y (i.e., the portion corresponding to a decreasing n) are approximated rather satisfactorily by several forms of empirical equations which at the

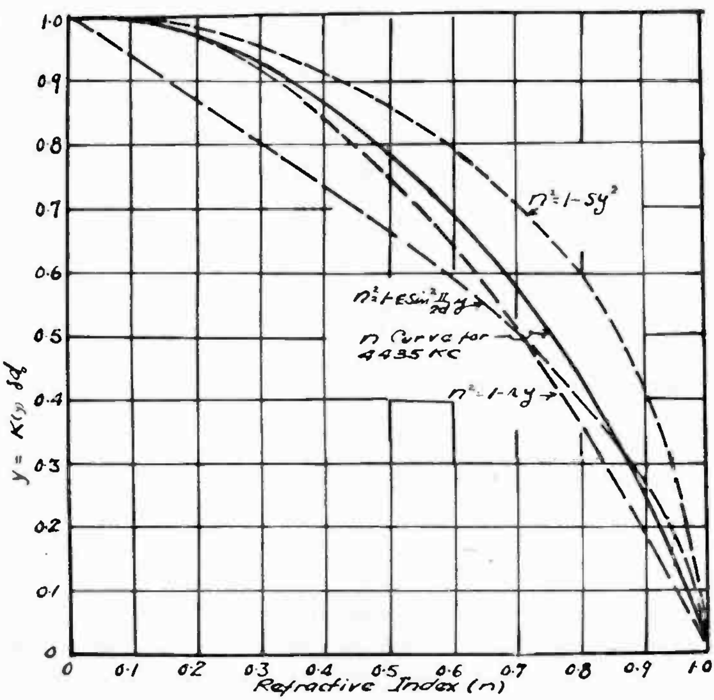

Fig. 9—Showing $K(y)$ (equals $y/\delta d_0$) as a kenetron of n in cases (as (b) and (c) when the constants s, r, E and d, are adjusted to give the same values of $\delta d/_0$ in each case.

same time lead to relatively simple integrals. The three types of equations which we will consider in this paper are

(a) $$n = \sqrt{1 - ry}$$

where r = a constant

y = height above the altitude where n starts to vary appreciably.

(b) $$n = \sqrt{1 - sy^2}$$

where s is another constant and y is the same as in (a).

(c) $$n=\sqrt{1-E\sin^2\frac{\pi y}{2d}}$$

where E and d are constants.

The form of these empirical curves is shown in Fig. 9.

V. Trajectories Obtained from the Empirical Equations for n

(a) $$n=\sqrt{1-ry}. \qquad \text{(V–1a)}$$

Substitution in (III-2) gives

$$\int_0^y \frac{dy}{\sqrt{\cot^2\phi_0-ry/\sin^2\phi_0}}=x.$$

Which yields on integration the parabolic path

$$\frac{2\sin^2\phi_0}{r}\left\{\cot\phi_0-\sqrt{\cot^2\phi_0-\frac{r}{\sin^2\phi_0}y}\right\}=x.$$

This may be written

$$\frac{\cos^2\phi_0}{r}-y=\frac{r}{4\sin^2\phi_0}\left[\frac{\sin^2\phi_0}{r}-x\right]^2$$

but

$$y_{\max}=\frac{\cos^2\phi_0}{r} \quad \text{or} \quad n_{\min}=\sqrt{1-\cos^2\phi_0}=\sin\phi_0.$$

This checks the condition pointed out in III, i.e., the ray attains a maximum height h for which $n_h=\sin\phi_0=n_{\min}$. The x coordinate corresponding to the maximum value of y is also obtained as

$$x_0=\frac{2\sin^2\phi_0\cot\phi_0}{r}=\frac{2\sin\phi_0\cos\phi_0}{r}=2y_{\max}\tan\phi_0 \qquad \text{(V–2a)}$$

or

$$y_{\max}=\frac{x_0}{2\tan\phi_0}. \qquad \text{(V–3a)}$$

In other words the maximum height attained by the ray for this type of variation of n with height (which yields a parabolic path) is just one half of the maximum height attained in a triangular ray trajectory having the same angle of departure.

(b) $$n=\sqrt{1-sy^2} \qquad \text{(V–1b)}$$

As in (a), substitution in (III-2) gives

$$\int_0^y \frac{dy}{\sqrt{\dfrac{1-sy^2}{\sin^2\phi_0}-1}}=x$$

or

$$\frac{\sin\phi_0}{\sqrt{s}}\sin^{-1}\left(\frac{sy}{\cos\phi_0}\right)=x$$

i.e.,

$$y=\frac{\cos\phi_0}{\sqrt{s}}\sin\frac{\sqrt{s}}{\sin\phi_0}x \qquad \text{(V–2b)}$$

The maximum height attained during the trajectory is again

$$y_{\max}=\frac{\cos\phi_0}{\sqrt{s}} \qquad \text{(V–3b)}$$

the substitution of (V-3b) in (V-1b) gives

$$n_{\min}=\sin\phi_0$$

which again checks our fundamental theorem in III.

The ratio of the actual maximum height attained to the height attained by a triangular trajectory of the same initial angle of departure is from (V-3b)

$$\frac{y_{\max}}{y_1}=\frac{\dfrac{\cos\phi_0}{\sqrt{s}}}{\dfrac{\pi}{2}\dfrac{\sin\phi_0}{\sqrt{s}}\cot\phi_0}=\frac{2}{\pi}. \qquad \text{(V–4b)}$$

In other words for this type of trajectory the maximum height attained by the ray is $\frac{2}{\pi}$ times the maximum height of a triangular trajectory having the same angle of departure.

(c) We still have to consider the interesting form

$$n=\sqrt{1-E\sin^2\frac{\pi}{d}y}. \qquad \text{(V–1c)}$$

Proceeding as in cases (a) and (b), we are led upon substitution in equation (III-2) to

$$\int_0^y \frac{dy}{\sqrt{\cot^2\phi_0-\dfrac{E}{\sin^2\phi_0}\sin^2\dfrac{\pi y}{2d}}}=x.$$

Letting $\frac{\pi y}{2d} = y'$ and performing a few obvious simplifications we are led to

$$x = \frac{1}{\cot \phi_0} \int_{y'=0}^{y'=\sqrt{\frac{1-\sin^2\phi_0}{E}}} \frac{dy^1}{\sqrt{1 - \frac{E}{\cos^2 \phi_0} \sin^2 y^1}}.$$

Letting

$$\frac{E}{\sin^2 \phi_0} \sin^2 y^1 = \sin^2 Z$$

we have, changing the variable again,

$$x = \frac{(\tan \phi_0)(2d)}{\frac{\pi}{\sqrt{E}} \cos \phi_0} \int_0^{\sin^{-1}\sqrt{\frac{E}{\cos^2\phi_0}} \sin \frac{\pi y}{2d}} \frac{dZ}{\sqrt{1 - \frac{\cos \phi_0}{c} \sin^2 Z}} \tag{V–2c}$$

Letting the limit of the integral become as large as possible (excluding imaginaries) we have

$$\sin^{-1} \sqrt{\frac{E}{\cos^2 \phi_0}} \sin \frac{\pi y}{2d} = 1$$

or

$$\sin^2 \frac{\pi y_{\max}}{2d} = \frac{\cos^2 \phi_0}{E} = \frac{1 - \sin^2 \phi_0}{E}. \tag{V–3c}$$

This gives on substitution in (V-1c)

$$n_{\min} = \sqrt{1 - (1 - \sin^2 \phi_0)} = \sin \phi_0,$$

again our familiar criterion. Our integral then becomes

$$x = \frac{\tan \phi_0 \left(\frac{2d}{\pi}\right)}{\sqrt{\frac{1 - \sin^2 \phi_0}{E}}} \int_0^{\pi/2} \frac{dZ}{\sqrt{1 - \frac{(1 - \sin^2 \phi_0)}{E} \sin^2 Z}}. \tag{V–4c}$$

The tabulation of this complete elliptic integral of the first kind $F(\beta, \alpha)$ as a function of $\beta \left(\text{equals } \frac{\pi}{2}\right)$ and $\alpha = \sin^{-1} \frac{\cos \phi_0}{\sqrt{E}}$ is available in tables or graphs.[3] In this case we have, for our ratio of actual

[3] See "Jahnke Einde Funktiontafeln," p. 53.

maximum height to the maximum height attained in a triangular trajectory of the same initial angle,

$$\frac{y_{\max}}{y^1}=\frac{\dfrac{d}{\pi}\sin^{-1}\dfrac{\cos\phi_0}{E}}{\dfrac{d}{\pi}\dfrac{\sin\phi_0}{\cos^2\phi_0}\sqrt{E}\cot\phi_0 F\left(\dfrac{\pi}{2};\ \dfrac{\cos\phi_0}{\sqrt{E}}\right)}$$

$$\frac{y_{\max}}{y^1}=\frac{\cos\phi_0\sin^{-1}\left(\dfrac{\cos\phi_0}{\sqrt{E}}\right)}{\sqrt{E}\,F\left(\dfrac{\pi}{2};\ \dfrac{\cos\phi_0}{\sqrt{E}}\right)} \qquad \text{(V-5c)}$$

Values of the ratio $y_{\max}/y'$ for all three cases are shown in Fig. 10 (for case (c) computed for $\phi_0=45°$).

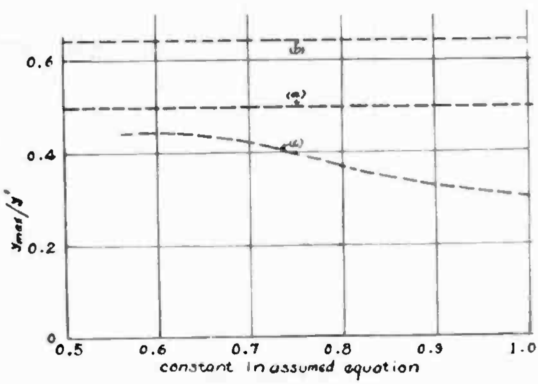

Fig. 10—Showing $y_{\max}/y'$
(a) $n=\sqrt{1-ry}$
(b) $n=\sqrt{1-sy^2}$
(c) $n=\sqrt{1-E\sin\pi/2dy}$

It will be noted that even in case *c* this ratio is surprisingly constant except near the critical angle for which $\sin\phi_0=\sqrt{1-E}$ beyond which rays are not turned back. It will be noted that this form of solution is important in that it gives a minimum value for n and hence exhibits skip distance phenomena which are absent in the other cases since the equation chosen for n goes to zero.

* This is perhaps not the best form of empirical equation for the case chosen in that the minimum given is cuspidal. (for another possibility, see Pedersen p. 160, eq. 24.)

VI. Computation of the Ratio of True to Virtual Height $\frac{h}{h'}$

The results of the previous section lead us directly to the ratio of true to virtual heights.

It will be noted that we have in each case obtained the ratio $\frac{y_{\max}}{y'}$ of the highest point reached in the wave trajectory above the height at which n starts to vary to the distance y' of the virtual triangle (for the group time case) above this same plane.

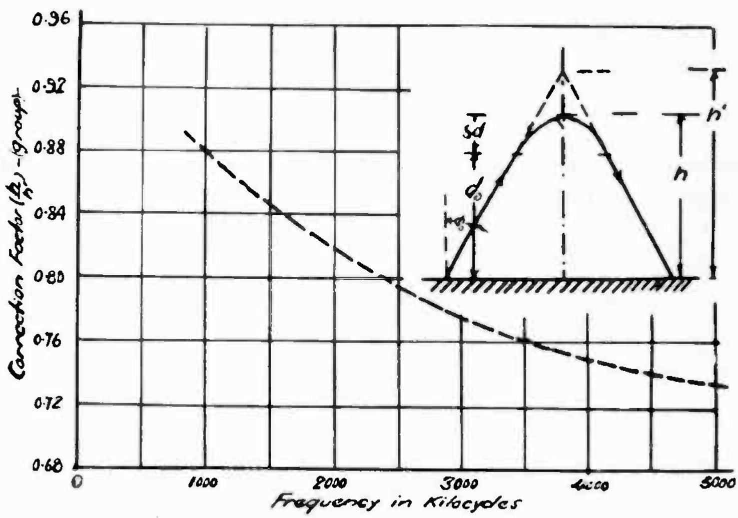

Fig. 11—Correction Factor ($h/h' = 1 - \delta d/h'$) for "Group Time" Measurements of Heights. Figs. 14 and 15 Are Used for Estimating δd and h', respectively.

As shown in Fig. 12 the total trajectory consists of this path plus two straight curves representing the path of the rays in the lower atmosphere.

Letting $\frac{y_{\max}}{y'} = \eta$ it is evident that we have from Fig. 12 by simple geometry

$$\frac{h}{h'} = \frac{d + y_{\max}}{d + y'} = \frac{d + y_{\max}}{h'} \tag{VI-1}$$

or

$$\frac{h}{h'} = \frac{d + y'}{h'} - \frac{y' - y_{\max}}{h'}$$

that is

$$\frac{h}{h'} = 1 - \frac{(1-\eta)y'}{h'} \tag{VI-2}$$

Noting that

$$\eta y' = \delta d \tag{VI–3}$$

this gives us our required result, i.e.,

$$\frac{h}{h'} = 1 - \frac{(1-\eta)}{\eta} \frac{\delta d}{h'} \cdot \tag{VI–4}$$

The factor $\frac{1-\eta}{\eta}$ is readily computable for cases (a), (b), and (c) of Section V by reference to Fig. 10.

It will be noted that the values of $\frac{\eta}{1-\eta}$ in most cases differ but little from unity so that rather closely (exactly in case (a))

$$\frac{h}{h'} = 1 - \frac{\delta d}{h'} \cdot \tag{VI–5}$$

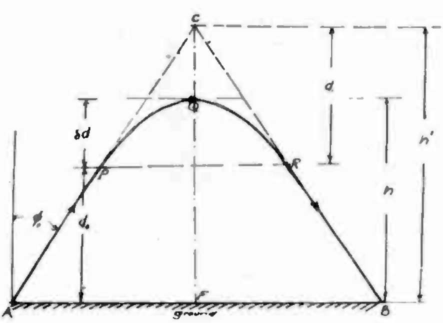

Fig. 12—Showing the Entire Upper Path of Wave Trajectory *PQR* is the Curved Portion of the Path When the Vaying Refractive Index Differs from Unity.

A simple result useful for a rapid visualization of the order of magnitude of the differences between real and virtual heights in any particular case. It will be noted in particular that when δd is small (i.e., when the layer is sharply defined) these heights are in close accord.

The results thus obtained by using the values of δd computed from Fig. 14 are shown in Fig. 11. (See Section VIII.) These curves are necessarily tentative by virtue of the inadequacy of the available experimental data.

As used in this paper, it is evident that the quantity δd is an important one, and careful consideration should be given as to its choice. From this discussion it would appear that δd should be the distance from the lowest point at which the variation in n produces "an appreciable alteration of the wave trajectory." It is evident that a more quantitative definition is highly essential. The most satisfactory quantitative definition is to be found in the relation of δd to the constants in the empirical equations used to compute the ray path, thus for equation (V–1a) $\delta d = \frac{\cos^2\phi_0}{r}$ etc., In other words the quantity δd (see Fig. 12) should be so chosen that the empirical equation used will give the best possible fit to the n curve. The fit obtained for a particular case is shown in Fig. 9. We denote by δd_0 the value of δd when $\phi_0 = 0$.

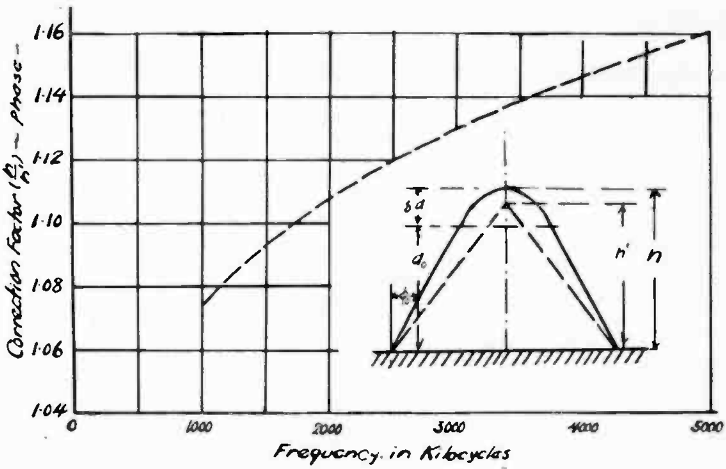

Fig. 13—Correction Factor ($h/h' = 1 - d/3h'$) for "Phase Retardation" Measurements of Heights.

VII. Phase Retardation Experiments

In the previous sections we have shown how the ratio of true to virtual heights of the wave trajectory is computable in the case of group velocity experiments; let us now proceed to a comparison of such results with those obtainable from phase retardation measurements. The procedure in such experiments is to vary the frequency over a small but finite range and to measure

$$\frac{\Delta(N_1 - N_2)}{\Delta f} \qquad \text{(VII–1)}$$

Where N_1 = The number of wavelengths in path a,* let us say the sky path.

N_2 = The number of wavelengths in path b,* let us say the ground path.

* a and b may, of course, be any two transmission paths.

$$f=\frac{\omega}{2\pi}=\text{transmitted frequency}$$

$\Delta f=$ Change of transmitted frequency.

$\Delta(N_1-N_2)=$ Change in N_1-N_2 due to the change f.

If we define the wave number N by the equation[10, 11]

$$N=fT_p=\int\frac{ds}{\lambda}. \tag{VII–2}$$

Where $ds=$ the element of arc along the path under consideration.

$\lambda=$ wavelength at the point on the path at which ds is measured.

then

$$N=fT_p \tag{VII–3}$$

and by substitution in (VII-1)

$$\frac{\Delta N}{\Delta f}=\frac{f\Delta T_p+T_p\Delta f+\Delta f\Delta T_p}{\Delta f}. \tag{VII–4}$$

Admitting the approximation

$$\frac{\Delta N}{\Delta f}\backsimeq\frac{dN}{df}. \tag{VII–5}$$

We may show that to this order of approximation the value of $\frac{\Delta N}{\Delta f}\left(\text{equals}\,\frac{\Delta(fT_p)}{\Delta f}\right)$ approaches the value of T_g, i.e., the group time as measured in group time experiments.

This may be proved as follows:

We have[12]

$$T_g=\int\frac{ds}{u} \tag{VII–6}$$

where: $T_g=$ group time; but also

$$u=\frac{df}{d(1/\lambda)}=\text{group velocity}$$

have

$$T_g=\int\frac{d}{df}\left(\frac{ds}{\lambda}\right)=\frac{d}{df}\int\frac{ds}{\lambda}=\frac{dN}{df}=\frac{d(fT_p)}{df} \tag{VII–7}$$

[10] In the subsequent analysis we will write for brevity N for a certain component (either N_1 or N_2). We will proceed with the analysis for this component. Evidently no generalization is lost and we may proceed to obtain expressions for N_1-N_2 etc. by subtraction of two such results wherever these are pertinent to the discussion.

[11] See Pedersen, p. 175.

[12] Breit and Tuve, *Phys. Rev.*, 28, 571, 1926; equation 10. Also Pedersen, p. 170, equation 46.

It should be carefully noted, however, that this does *not* imply that T_p is equal to T_g. Actually, if we neglect dissipation it follows[13] that

$$T_p = 1/f \int \frac{ds}{\lambda} = \int \frac{n\,ds}{c} \qquad \text{(VII-8)}$$

while for this case since[14]

$$n^2 = 1 - \frac{4\pi N e^2}{m\omega^2} \qquad \text{(VII-9)}$$

$$u = cn \qquad \text{(VII-10)}$$

and

$$T_g = \int \frac{ds}{cn} \,. \qquad \text{(VII-11)}$$

It should be noted that, insofar as equation (VII-5) is justified, phase retardation experiments may be used, by virtue of equation (VII-7), to compute T_g which in turn may be used to compute virtual heights which will, for a given layer, be in accord with those computed from group time measurements. The correction factors deduced in Section VI are evidently applicable to virtual heights thus computed.

In his recent paper Mr. J. C. Schelleng proceeds to use equation (VII-11) in a somewhat different manner: he writes

$$(T_{g_1} - T_{g_2}) = \frac{d(N_1 - N_2)}{df} \qquad \text{(VII-12)}$$

$$(N_1 - N_2)_{f_0} = \frac{1}{f_0} \int_0^{f_0} (T_{g_1} - T_{g_2})\,df \qquad \text{(VII-13)}$$

He now identifies the sides of a triangle with the difference of wave numbers $N_1 - N_2$. See Fig. 13.

It should be carefully noted that the altitude of this triangle is in no sense the altitude of the triangle computed from group time data and giving the group time virtual height. This fact is, of course, well appreciated by Mr. Schelleng who presents this triangle as a much closer approximation to the true height of the wave trajectory h. Pedersen has in fact proved that this triangle is indeed a very good approximation; our analysis is in accord with this statement but enables us to find for certain given assumed equations for $n = f(y)$ the exact closeness of accord. Thus we have only to compute the $N = \int \frac{ds}{\lambda}$

[13] See p. 170, Pedersen.
[14] See pp. 174–75, Pedersen.

over the actual paths given by equations (V-1a), (V-1b), and (V-1c) and then compute the height of a triangle having the same wave number N.

Let us do this for the simplest case; i.e., when $n=\sqrt{1-ry}$ we have the curve represented by Fig. 12, considering the *curved* path of the wave projectile first (i.e., curve PQR (see Fig. 12)) that the number of wavelengths for infinitesimal length ds is

$$dN=\frac{ds}{\lambda}\cdot$$

The number of wavelengths for PQR is hence

$$N=2\int_P^Q\frac{ds}{\lambda}$$

but since

$$n=\frac{\sin\phi_0}{\sin\phi}$$

$$ds=\sqrt{1+\left(\frac{dy}{dx}\right)^2}\,dx=\frac{n}{\sin\phi_0}dx$$

or

$$ds=\frac{n\dfrac{dy}{dx}dx}{\sin\phi_0\dfrac{dy}{dx}}$$

assuming

$$n^2=1-ry$$

where $r=$constant
$y=$height (vertical) from P

$$dy=-\frac{d(n)^2}{r}$$

also

$$\frac{dy}{dx}=\sqrt{\frac{n^2}{\sin^2\phi_0}-1}$$

$$\therefore \qquad ds=\frac{-\dfrac{n}{r}d(n)^2}{\sin\phi_0\sqrt{\dfrac{n^2}{\sin\phi_0}-1}}=-\frac{nd(n)^2}{r\sqrt{n^2-\sin^2\phi_0}}$$

Noting that

$$\frac{1}{\lambda}=\frac{n}{\lambda_0}$$

when

$$\frac{\sin\phi}{\sin\phi_0}=\frac{1}{n}=\frac{\lambda}{\lambda_0}$$

$$\frac{N}{2}=\int_P^Q\frac{ds}{\lambda}=-\int\frac{ndn^2}{r\sqrt{n^2-\sin^2\phi_0}}\cdot\frac{1}{\lambda}=-\int\frac{n^2}{r\lambda_0}\,\frac{dn^2}{\sqrt{n^2-\sin^2\phi_0}}.$$

Let

$$Z=n^2$$

then $Z=1$ at P, and $Z=\sin^2\phi_0$ at Q. Hence

$$\frac{N}{2}=\int_P^Q\frac{dS}{\lambda}=-\frac{1}{r\lambda_0}\int_1^{\sin^2\phi_0}\frac{ZdZ}{\sqrt{Z-\sin^2\phi_0}}$$

$$=-\frac{1}{r\lambda_0}\Big[\tfrac{2}{3}(Z-\sin^2\phi_0)^{3/2}+2\sin^2\phi_0\sqrt{Z-\sin^2\phi_0}\Big]_1^{\sin^2\phi_0}$$

$$\frac{N}{2}=\frac{2\cos\phi_0}{r\lambda_0}\left[1-\frac{2}{3}\cos^2\phi_0\right]$$

But

$$r=\frac{\cos^2\phi_0}{\delta d}$$

from which it follows that

$$\frac{N}{2}=n\int_P^Q\frac{dS}{\lambda}=\frac{2\delta d}{\cos\phi_0\lambda_0}\left[1-\frac{2}{3}\cos^2\phi_0\right].$$

The above expression gives the number of waves for the curved path; the actual path has, by referring to Fig. 12

$$\frac{N_{\text{total}}}{2}=\frac{(\sec\phi_0)(h-\delta d)}{\lambda_0}+\frac{2\delta d}{\lambda_0}(\sec\phi_0)\left[1-\frac{2}{3}\cos^2\phi_0\right].$$

We wish to find a triangle with the same wave number and same base. The altitude of such a triangle is

$$h' = \sqrt{(\sec^2 \phi_0)\left[h + \left(1 - \tfrac{2}{3}\cos^2\phi_0\right)(\delta d)\right]^2 - (\tan^2\phi_0)(h+\delta d)^2}$$

By solving for the correction factor $\left(\dfrac{h}{h'}\right)$ from the above, we get

$$\frac{h}{h'} = \frac{1}{3}\left(\frac{\delta d}{h'}\right) + \sqrt{1 + \frac{16}{9}(\sin^2\phi_0)\left(\frac{\delta d}{h'}\right)^2}$$

or by expanding

$$\frac{h}{h'} = 1 + \frac{1}{3}\left(\frac{\delta d}{h'}\right) + \frac{8}{9}(\sin^2\phi_0)\left(\frac{\delta d}{h'}\right)^2 + \frac{32}{81}(\sin^4\phi_0)\left(\frac{\delta d}{h'}\right)^4 + \cdots.$$

It will be noted that $\dfrac{\delta d}{h'}$ is a small quantity (usually less than ten per cent) and $\sin\phi_0$ is also small for most experiments. Thus, the product of the squares of two small quantities will be negligible. Hence

$$\frac{h}{h'} \doteq 1 + \frac{1}{3}\left(\frac{\delta d}{h'}\right).$$

This, again, is our familiar expression for $\dfrac{h}{h'}$ (for group case) except for the reversal of sign and the new factor 1/3[15]. In other words the equivalent triangle, which has the same number of waves

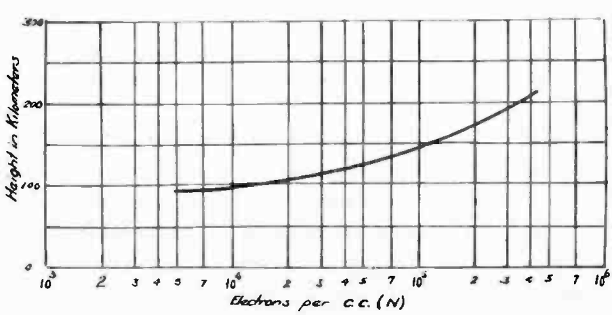

Fig. 14—Electron Density as a Function of Height, as Deduced by the Method of Successive Approximations. Compare with Fig. 4.

as the actual path, has a lower height than the "true" height of the layer. This explains the too large wave number obtained when the triangular height is taken to be the "true" height.[16]

[15] See Section VI.
[16] See Pedersen, p. 176.

The correction factors obtained from the wave number considerations (using values of δd, estimated from the N and n curves of Figs. 14 and 15) are given in Fig. 13. These factors are evidently highly approximate because of the very limited experimental data on which the N and n curves are based.

VIII. Resumé of Available Data and Method of Successive Approximations

In Sections IV to VI of this paper we have sought to outline methods by which the ratio of "true" to "virtual" heights may be computed in any case in which the equation for the refractive index n is a known function of y; we have investigated the results obtainable

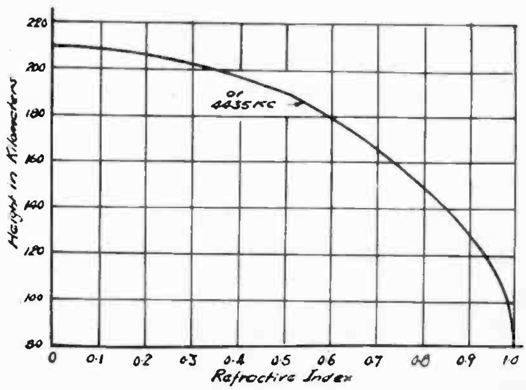

Fig. 15—n as a Function of Height for $f = 4435$ kc.
Note that the curve goes to a higher level than that given in Fig. 6.

by the use of these methods and several convenient empirical equations for n. It is interesting to note that the results of computations on several rather different assumptions give relatively constant and simple relations for the ratio of true to effective height h/h' in terms of the thickness of the n layer δd to which the wave penetrates.

In a recent paper[17] Mr. J. C. Schelleng has presented a brief summary of most of the virtual heights thus far determined by experiment. His estimate of wave number height shows a slow linear variation of this height with frequency. As is emphasized in his paper, the estimate is based on an assumed linearity of the experimental virtual height-frequency curve which is probably not the exact form which will develop when more data becomes available. However, it is interesting to compare value for the corrected height obtained by the

[17] Proc. I. R. E., 16, 1471; November, 1928.

rather different methods of Sections VI and VII with Mr. Schelleng's estimates. It is interesting to note that the mean height obtained in the experiments described here falls closely on the straight line approximation for the virtual heights as obtained by Schelleng (and used for the virtual height curve in our estimates).

As our first approximation we apply the correction factor $\left(h/h' = 1 - \frac{\delta d}{h'}\right)$ as obtained from N and n curves derived from this uncorrected (virtual) height curve. From the new height so calculated we can make a second approximation for the electron density N in

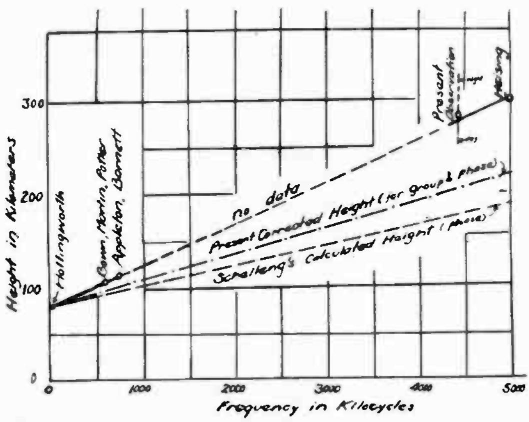

Fig. 16—Showing a Comparison of the Experimental Data, Schelleng's Calculated Height, and the Present Estimate of the "True Height" by Successive Approximations.

the upper atmosphere at different heights. The new N curve enables us to calculate again the curves of index of refraction n for various frequencies and so on. The final curves obtained in this manner (for $f = 4435$ kc) are shown in Figs. 14 and 15. It will be noted that this gives a more gradual variation of n with height, and hence a larger value of the correction factor $\left(1 - \frac{\delta d}{h'}\right)$ than indicated by Pedersen's estimates (see Figs. 4 and 5). The corrected height obtained by this method is shown in Fig. 16. This curve approaches reasonably close to the estimate made by Schelleng; in fact the corrected height obtained from his wave number heights by the application of the appropriate correction factor for this case (see Fig. 13) gives a curve which coincides with the new true height curve to a precision at least as great as warranted by the limited data on which the virtual height curve is based.

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MEASUREMENT OF THE FREQUENCIES OF DISTANT RADIO TRANSMITTING STATIONS*

By
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Summary—*The equipment installed in the Italian Royal Experimental Institute of the Communications for radio-frequency measurements is described. The measurements are based upon the utilization of harmonics generated by the Abraham multivibrator controlled by a 1024-cycle tuning fork. The factors which can influence the variation of the tuning fork, with special regard to the action of temperature, are examined. A simple and rapid method is described, which permits of control of the frequency of the tuning fork at the beginning and at the end of each measurement, and determines, with great accuracy, the frequency of the harmonics between which the wavelength to be measured is included. Such a method consists in comparing the frequency of the tuning fork with that generated by an alternator, and in recording, through a chronograph, the beats included between the two frequencies, the number of revolutions, and frequency of the alternator together with the time furnished by a pendulum of the Astronomical observatory.*

Some examples of calibrations are reported, showing that the degree of precision assured varies from 1 to 5 per 100,000.

THE problem of measuring the frequencies of distant radio transmitting stations with the best possible accuracy has recently seriously attracted the attention of workers in the radio field. This problem was not fully appreciated in the past. It was not a rare thing to find wavemeters giving uncertainties of the order of several per cent even in some of the principal radio stations. At the present time the question of precise measurement of the frequencies of radio transmitting stations has become of utmost importance because of the great increase in the number of radio stations, many of which, as for example, broadcasting stations, must be operated simultaneously. As a result, all the most important laboratories are now equipped for radio-frequency measurements of the highest precision.

In Italy the necessity of undertaking special studies and researches in the various government laboratories for the purpose of constructing radio-frequency standards was pointed out last June by Prof. Corbino, President of the Italian Committee of the International Union of Scientific Radiotelegraphy (U.R.S.I.)

Recently the Superior Committee for Vigilance on Italian Broadcasting, presided over by Senator Tittoni, recognizing the necessity

* Dewey decimal classification: R 210. Original manuscript received by the Institute, October 15, 1928. Revised manuscript received, November 14, 1928.

of checking the frequencies of the emitted waves of the Italian broadcasting stations, charged the Ministry of Communications with the execution of these high precision measurements.

The object of this paper is to describe the apparatus used in the Royal Experimental Institute of Communications for the measurement of long and medium waves between 100 and 30,000 meters (3,000 and 10 kc). The system is based on the well known principle of harmonics produced by means of the Abraham multivibrator. The set employed can also be utilized for the measurement of shorter waves, but we believe that measurements of this kind should preferably be made with other methods, the description of which will be contained in a later paper.

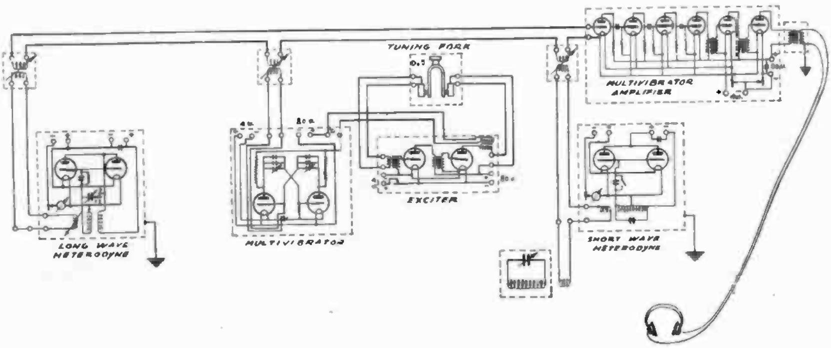

Fig. 1—Circuit Used in Measuring the Frequencies of Distant Radio Stations.

The multivibrator fundamental frequency is synchronized by means of a tuning fork giving approximately 1024 cycles per second. The tuning fork is operated and maintained in oscillation by the Eccles triode circuit, in the plate circuit of which is inserted the primary of a transformer whose secondary is in series in the multivibrator plate circuit (see Fig. 1).

Using this arrangement, when the multivibrator is approximately tuned to the tuning fork frequency a synchronizing effect takes place over a rather broad zone in the neighborhood of the tuning fork frequency. Thus the difficulty of a continuous control of the multivibrator frequency during the measurements is completely avoided.

Two heterodyne generators complete the setup. Fig. 2 illustrates the arrangement of the measuring apparatus.

Normally only one heterodyne generator is used. When it is necessary to standardize the shorter waves of the above-mentioned scale of frequencies the other generator is adjusted to the harmonics of the first one. Both heterodynes can be used for determining the order of the harmonics. For instance, the 96th harmonic of the multivibrator can be more easily found as the 8th harmonic of the 12th, or the 6th harmonic of the 16th, than by picking it up directly. Moreover, uncertainty is removed when there is the possibility of determining a particular harmonic by two different methods, especially when there is no possibility of obtaining adjacent harmonics, as in the example above. After the harmonics are identified, the calibration of the wavemeter is easily obtained by measuring with it

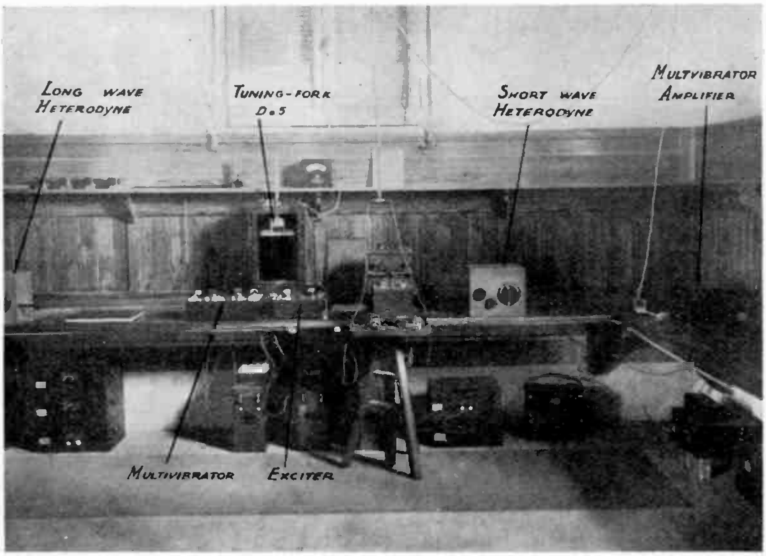

Fig. 2—Interior View of the Radio-Frequency and Calibratiom Set.

the frequencies of the electron-tube generators, successively adjusted to the different harmonics.

The measurement of a transmitting station frequency is obtained, first by synchronizing one of the generators on the unknown frequency, by means of a non-oscillating receiver, and then measuring that generator frequency with the calibrated wavemeter. If a very high precision is required the heterodyne oscillator may be used as a frequency meter. At the time of the measurement the multivibrator is adjusted to frequencies very near those to be measured and the corresponding reading of the heterodyne condenser observed. By successive interpolation the value of the frequency is obtained, which corresponds to the condenser setting for which the condition of zero beat exists between the heterodyne and the incoming signal. Every possible precaution should be taken during this procedure for maintaining the frequency of the heterodyne constant.

The piezo-electric crystals, now widely used in radio, can also be considered as high-frequency tuning forks, having a low temperature coefficient and sufficient stability to permit their use in the construction of very good secondary standards of frequency. The calibration of a piezo-electric crystal can be obtained by connecting it in the usual circuit, and then measuring its frequency by means of a calibrated electron tube generator. The same procedure described above for the measurement of a transmitting station frequency is used.

The tuning fork calibration, which is undoubtedly the principal problem connected with this kind of measurement, has required the development of a rapid and easy method, capable of giving satisfactory precision. This is obtainable in the usual electrotechnical laboratory practice.

First, it is necessary to consider how far it is possible to rely on the constancy of the tuning fork frequency.

It is known that for a tuning fork the number of vibrations of the fundamental frequency is directly proportional to the prong thickness, inversely proportional to the square of its length, and directly proportional to the square root of the fraction E/D, in which E is the modulus of elasticity and D the density of the steel of the tuning fork. Temperature, which affects directly the modulus of elasticity, is the principal factor influencing the frequency of the tuning fork (frequency).

It is known that between N, which represents the tuning fork frequency at temperature t, and N_o, which is the same frequency at zero degrees C, the following relation exists:

$$N = N_o(1 - \alpha t - \beta t^2)$$

Normally, β is very small and t^2 can be neglected, but the coefficient α in the common steel tuning forks varies between 0.000088 and 0.00012. Consequently, in the ordinary tuning forks there are frequency variations due to temperature changes which may be in some cases of the order of several parts per thousand.

In the electrically-driven tuning forks, other factors besides the temperature have an effect on the frequency stability, but their effects are not marked. It is known that the tuning fork frequency decreases when the amplitude of the tuning fork vibrations increases. Consequently, it is clear that any variation in the excitation current (that is, the filament current of the vacuum tube), or of the capacity shunting the exciting coils, or of the load applied to the secondary of the transformer, is accompanied by corresponding variations in the tuning fork frequency. A very accurate study of these subjects has been made by Dr. D. Dye.[1]

As a result of the above considerations we reached the conclusion that in measurements requiring a high degree of precision (1 to 5 parts in 100,000) it is necessary to calibrate the tuning fork each time before the measurement and to maintain an absolute constancy of temperature. When less precision (5 parts in 10,000) is required a

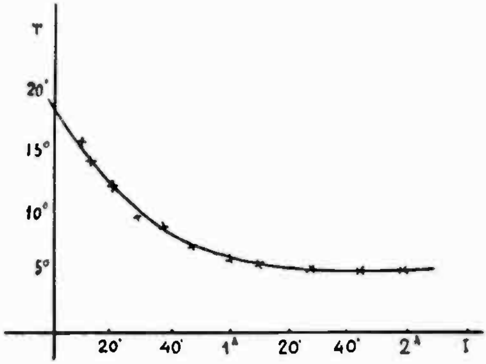

Fig. 3

former calibration can be used, but in this case the temperature coefficient must be known and the other conditions (relating to the tubes, plate current, etc.) must be exactly those which were used during the former calibration.

We believe that the exact determination of the tuning fork temperature, which lags behind the temperature of the surrounding air, constitutes the greatest difficulty of this measurement. For example, Prof. Gianfranceschi, who conducted investigations on the 870-cycle standard tuning fork of the Physical Institute of Rome, found a tuning fork cooling curve similar to that shown in Fig. 3. He has also found by calculation that in the case of the above mentioned standard tuning fork the coefficient α' of the equation of the curve

$$T = T_a - T_e \cdot e^{\alpha' T}$$

was given by:

$$\alpha' = 0.02339$$

[1] *Proc.* Roy. Soc. of London, p. 240, May, 1923.

where T and T_a were, respectively, the tuning fork temperature at the instant t and the temperature of the air, and T_e represented the excess of the initial temperature of the tuning fork over that of the temperature of the air.

By the method we have developed we can obtain the calibration of a tuning fork very simply and quickly. In Fig. 4 is shown the stator winding of a small Siemens 100-toothed alternator of the induction type connected in series with the secondary of the transformer in the output circuit of the multivibrator amplifier. This alternator is driven by a d.c. motor which carries on the same shaft a small d.c.

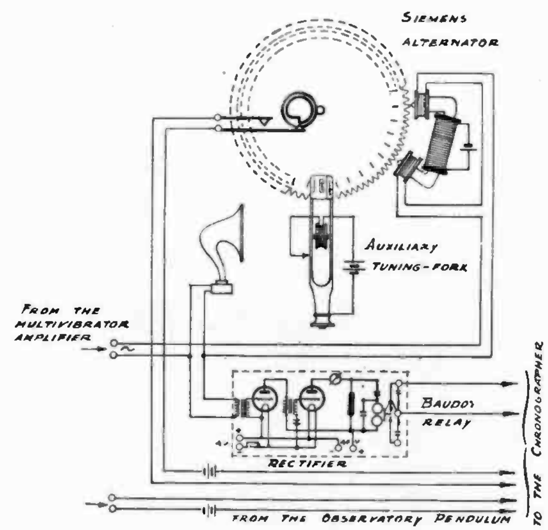

Fig. 4—Circuit Diagram of the Method Adopted for the Calibration of the Multivibrator Tuning Fork.

generator. In order to obtain a steadier speed of the revolving gear this generator can be loaded with a resistance coil of a suitable value. Moreover, a centrifugal speed regulator of the type used in the Hughes telegraphic apparatus is used on the revolving shaft for the same purpose. In order to obtain a frequency from the alternator as near as possible to the tuning fork frequency, a stroboscopic figure composed of 39 sectors is drawn on the alternator rotor and is observed through a small window, open at each semivibration, by an auxiliary electrically-driven tuning fork having a frequency of 200 cycles per second.

When the stroboscopic figure appears still as viewed through the small window, the number of revolutions per second n attained by the alternator is such that the following relation exists:

$$\frac{1''}{39\times n}=\frac{1''}{200\times 2}.$$

As the rotor has 100 teeth the corresponding frequency of the above conditions is given by:

$$f=\frac{400}{39}\times 100=1025.6$$

which is very near to the frequency of the tuning fork.

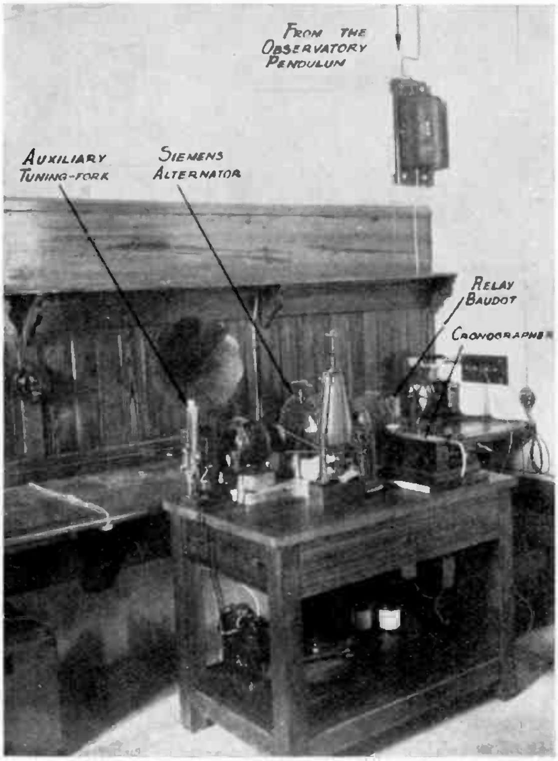

Fig. 5—Interior View of the Apparatus Used in the Tuning Fork Oscillation.

When the alternator frequency is close to the tuning fork frequency, beats will be heard in the loud speaker telephone in series with the amplifier output transformer of the multivibrator and the Siemens alternator. A suitable amplifier followed by a vacuum-tube detector shunts the loud speaker. In the plate circuit of the detector we have placed a very sensitive Baudot relay, having coils with 11,500 turns each, and a resistance of 5000 ohms per coil. This relay is operated at each beat.

The installation shown in Fig. 5 is completed by a three-pen recorder. One of the pens is actuated by the above-mentioned relay; the second is actuated by a contact which closes every 10 revolutions of the motor-alternator shaft, and the third is controlled by a contact on the standard pendulum in the Astronomic Observatory of the Campidoglio, which is received by means of a special line. Fig. 6 shows one of the records obtained. Dots (*a*) show the number of beats, the dashes (*b*) indicate the time intervals (a one-second dash each two seconds), and dots (*c*) furnish the number of tens of the alternator revolutions, that is, the number of thousand cycles of the alternator current.

Fig. 6

We have found that the alternator speed need not be kept absolutely constant during the measurement, and that it is not necessary to observe the stroboscope continuously. Conditions are satisfactory for measurements when the difference between the alternator and the tuning fork frequencies remains the same as indicated by the note in the loud speaker. This condition is ordinarily obtained when beats are not too fast to be recorded. By means of the loud speaker it is possible to control at each instant the regular working of the beat recorder relay. If small variations of the alternator speed are required, they can be obtained by means of the d.c. generator loading resistance or directly by changing the mechanical regulating mechanism.

Evidently, if n is the number of the alternator cycles in a certain time t, and a is the number of beats recorded in the same time, the tuning fork frequency will be given by $n \pm a/t$. The sign to be applied at a can be obtained by observing the stroboscopic figure corresponding to zero beat. In the case of the Institute installation, at zero beat the stroboscopic figure appeared to revolve slowly in the direction of the shaft revolution. As the same figure normally had a greater speed in the same direction for the ordinary frequencies we came to the conclusion that the sign of a was negative.

Using a temperature controlled box (thermostat) we were able to obtain the calibration of the Institute tuning forks at various temperatures, and thus it was possible to obtain the temperature coefficient, which was 0.105 part in 1000. This is very near to that of 0.12 part in 1000 stated by the tuning fork manufacturer. Such a coefficient may be used for ordinary calibration in which case the tuning fork frequency can be calculated by means of the formula:

$$N = 1026.7 \times (1 - 0.000102t)$$

where t is the tuning fork temperature (Centigrade scale) that may be deduced from the air temperature by introducing proper corrections, and 1026.7 is the number of vibrations reduced to zero obtained from a long series of tests.

Using the above mentioned formula it is necessary to keep in mind that besides the uncertainty existing in the value of t, there is another source of error by assuming the quantity $N_o = 1026.7$ as a constant. This is because the supply current in the tuning fork circuit can change in the intervals between the various measurements, causing other differences of several parts in 100,000 from one test to another. This is especially true when the tests are made on different days. Consequently, when a high precision is required the tuning fork calibration must be repeated for each measurement.

Table I, which refers to measurements made on March 15, 1928 at a mean air temperature of 19 deg. C gives a clear demonstration of the precision of the present method.

TABLE I

CALIBRATION OF THE TUNING FORK

N	Date	Temperature	Seconds	Alternator cycles n	Beats Total number a	Tuning-fork cycles $N = n - a$	Tuning-fork frequency $f = \frac{n-a}{t}$	Mean frequency	Jumps from the mean %
1	4-3-928	19°C.	104	107.000	437	106.563	1,024,644	1,024,698	−0.054
2			566	583.435	3469	579.966	1,024,674		−0.024
3			326	335.610	1551	334.059	1,024,720		+0.022
4			488	502.733	2665	500.068	1,024,720		+0.022
5			1366	1,408.716	8951	1,399.765	1,024,720		+0.022

The principal advantages of the method described are the following:

(1) It permits the tuning fork calibration both before and after the various series of measurements, that is, under conditions which are practically the same as those existing during the use of the tuning fork as an exciter of the multivibrator.

(2) It avoids the necessity of obtaining an absolutely constant speed of the alternator.

In order to demonstrate the method followed during the measurements and the degree of precision obtainable, we give the following two examples:

(1) The calibration of a piezo-electric crystal (Table II).

(2) The measurement of the frequency emitted by a tube transmitting set of the Rome (Torrenova) high power radio installation (Table III).

TABLE II

CALIBRATION OF THE PIEZO-ELECTRIC CRYSTAL NO. 1

March 15th, 1928

$t=14$ deg. C. Measured tuning fork frequency: $f=1025.20$

Number of the observations	Readings of the heterodyne condenser for the harmonics		Reading of the heterodyne condenser at zero beats with the crystal frequency	Interpolation frequency	
	71st	70th			
Means of 5 observ.	822.96	987.98	840.3	71st harmonic 70th harmonic Crystal frequency	72789.75 71764.65 72682.10

March 18th, 1928

$t=13$ deg. C. Measured tuning fork frequency: $f=1025.42$

Means of 7 observ.	807.76	973.70	827.36	71st harmonic 70th harmonic Crystal frequency	72804.8 71779.4 72683.6
Means of 16 observ.	808.41	973.79	827.76	71st harmonic 70th harmonic Crystal frequency	72804.8 71779.4 72685.2

The general mean value of the three series of measurements made on March 15 and 18, 1928, gives 72684.5 as the crystal frequency. This value, at a temperature of about 13.5 deg. C differs by +3.44 parts in 100,000 from the value 72682 determined on December 12, 1927 by the Naval Radio Laboratory in Leghorn at a temperature of 14.8 deg. C, but using a different method.

It can be seen that the sign of the frequency difference is in accord with the temperature difference, so there is a perfect agreement between Rome and Leghorn measurements.

The control installation is completed by a series of five standard piezo-electric crystals calibrated as described above.

We believe that very quick determinations over a wide range of frequencies can be easily made by utilizing the oscillations and the harmonics produced by the above mentioned crystals together with an electron tube generator.

TABLE III

Measurement of the Frequency Emitted by Torrenova (Roma) Electron Tube Transmitter

March 17th, 1928

$t = 13$ deg. C. Measured tuning fork frequency: $f = 1025.22$

Number of the observations	Readings of the heterodyne condenser for the harmonics		Readings of the heterodyne condenser at zero beats with the Torrenova frequency	Interpolation frequency	Wavelength (for $V = 299.82.10^6$ m)
	98th	97th			
I Series					
1	906.0	967.8	Mean of 8 observations: 920	97th harmonic 99446.34	m 2290.8
2	906.0	967.8		98th harmonic 100471.06	
3	906.0	967.9		Torrenova freq. 100240.78	
Mean	906.0	967.8			
II Series					
1	906.7	968.0	Mean of 11 observations: 920.84	Torrenova freq. 100235.06	m 2991.2
2	906.5	967.8			
3	906.6	968.1			
Mean	906.6	968.0			
III Series					
1	907.0	968.3	Mean of 15 observations: 920.97	Torrenova freq. 100238.06	m 2991.0
2	906.9	968.2			
3	907.0	968.4			
Mean	906.97	968.3			
IV Series					
1	907.5	969.5	Mean of 27 observations: 920.76	Torrenova freq. 100207.5	m 2990.5
2	908.0	969.6			
3	907.8	969.5			
Mean	907.8	969.5			

By means of the methods and apparatus described above the Experimental Institute of Communications is now in a position to make frequency measurements of distant radio transmitting stations with accuracy and precision and to exercise, as required, the necessary control of the waves of radio transmitting stations.

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NOTE ON EARTH REFLECTION OF ULTRA SHORT RADIO WAVES*

By

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Summary—*In analytical investigations of the resultant pattern of electric intensity about an antenna, for the longer waves, the earth in the vicinity of the antenna has generally been considered as a perfect conductor. At sufficiently high frequencies, the reflected waves may differ considerably in magnitude and phase from perfectly reflected waves. The resultant distribution of electric intensity about an ultra short antenna depends upon the nature of the reflecting surface. Some computations and curves are given for the reflection coefficients and phase angles for various surface conditions, in conjunction with a horizontal ultra short antenna. Theoretical polar diagrams have been computed for various heights of horizontal antenna above the surface.*

CERTAIN experiments with ultra short radio waves have been described by H. Yagi,[1] in which measurements were made of the field distribution about short linear oscillators of 2.6 meters wavelength. The effects of earth reflection show to a marked degree in the pattern of electric intensities about a horizontal linear oscillator, in a plane perpendicular to the oscillator, and for various heights above the ground. The experimentally determined polar diagrams mentioned suggest the inadequacy of the assumption of perfect reflection for very short waves.

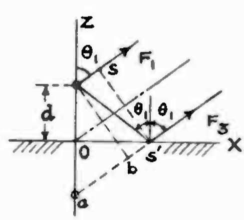

Fig. 1—Earth Reflection of Horizontal Electric Intensity.

The earth reflection from a horizontal linear oscillator is somewhat simpler to consider from the theoretical standpoint than that from a vertical oscillator[2]; also the experimental curves are more complete for the horizontal case, so that only the horizontal oscillator is here

* Dewey decimal classification: R113.6. Original manuscript received by the Institute, December 3, 1928.

[1] H. Yagi, "Beam Transmission of Ultra Short Waves," Proc. I.R.E. **16**, 715; June, 1928.

[2] Stuart Ballantine, "Review of Current Literature," Proc. I.R.E., **16**, 513; April, 1928; discussion of vertical dipol and bibliography.

considered. The exact effect of the earth as a composite dielectric, with various degrees of conductivity and permeability, is necessarily very complex, and only the simplest assumption is considered in what follows for a non-magnetic layer having a certain amount of conductance and free charge capacitance.

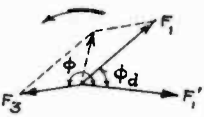

Fig. 2—Phase Relation of Direct and Reflected Intensity.

Polar diagrams for various heights of antenna above the earth have been computed for a plane perpendicular to the antenna, and are shown in Figs. 6 and 7. Points sufficiently far removed from the oscillator have been used in determining the directional intensity, so that the reflected and direct rays F_1 and F_3, Fig. 1, may be considered parallel in computing the phase difference at the wave front SS'. The resultant horizontal electric intensity thus depends upon the vector sum of the direct and reflected intensity, taking account of the magnitude and phase of each of the quantities at the wave front. Referring

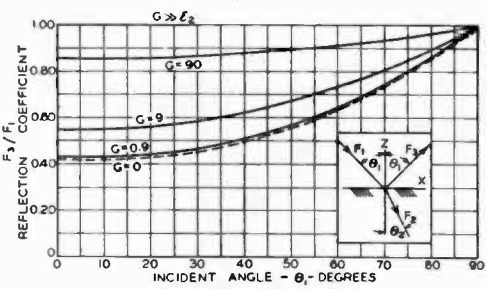

Fig. 3—Reflection Coefficients for Various Angles of Incidence, Resistivities, and Frequencies.

to Fig. 1 and Fig. 2, the excess path to the wave front for the reflected wave with reference to the direct wave is

$$ab = 2d \cos \theta_1$$

and the corresponding phase difference is

$$\phi_d = \frac{4\pi}{\lambda_1} d \cos \theta_1 \qquad (1)$$

A further phase change occurs at reflection, so that if F'_1 is the intensity before reflection, the direct wave at the wave front will be ϕ_d in advance of F_1', and the reflected intensity will have the phase angle ϕ with F_1' due to reflection. The resultant electric intensity is therefore the vector sum of F_1 and F_3.

For a perfectly reflecting surface the above computation is relatively simple, since the reflected and incident intensities are equal, and there is a phase change of 180 degrees on reflection. In this case the resultant phase difference between the direct and reflected waves at the wave front, and the resultant intensity, are determined by a simple expression. When, however, the reflecting layer has both conduction and displacement currents, both the magnitude and phase of the reflected ray vary with the angle of incidence.

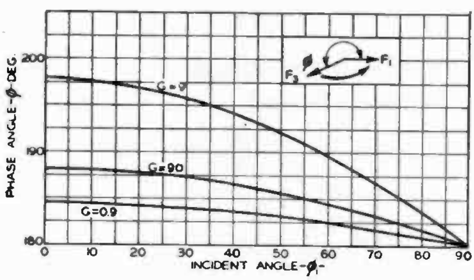

Fig. 4—Phase Angles for Various Angles of Incidence, Resistivities, and Frequencies.

From the following considerations, curves have been computed for the reflection coefficients and phase angles, for various angles of incidence and surface conditions. These curves are shown in Figs. 3 and 4. The electric intensities are f_x, f_y, and f_z, and the magnetic intensities h_x, h_y, and h_z.

The fundamental relations to be satisfied are the following Maxwell's equations in which $f_x = f_z = 0$, in accordance with the condition that the electric field is parallel with the earth's surface.

$$\frac{4\pi f_x}{\rho c} + \frac{\epsilon}{c}\dot{f}_x = \frac{\partial h_z}{\partial y} - \frac{\partial h_y}{\partial z} = 0 \tag{2}$$

$$\frac{4\pi f_y}{\rho c} + \frac{\epsilon}{c}\dot{f}_y = \frac{\partial h_x}{\partial z} - \frac{\partial h_z}{\partial x} \tag{3}$$

$$\frac{4\pi f_z}{\rho c} + \frac{\epsilon}{c}\dot{f}_z = \frac{\partial h_y}{\partial x} - \frac{\partial h_x}{\partial y} = 0 \tag{4}$$

$$-\frac{\mu}{c}\dot{h}_x = -\frac{\partial f_y}{\partial z} \tag{5}$$

$$-\frac{\mu}{c}\dot{h}_y=\frac{\partial f_x}{\partial z}-\frac{\partial f_z}{\partial x}=0 \tag{6}$$

$$-\frac{\mu}{c}\dot{h}_z=\frac{\partial f_y}{\partial x}\cdot \tag{7}$$

Therefore, the magnetic intensity $h_y=0$. (8)

Solutions of these equations are of the form

$$f=Fe^{-\alpha r}\times e^{j(\omega t-\beta r)} \tag{9}$$

where α and β are constants, $\omega=2\pi n$, and r the distance in the X-Z plane to any point from the earth's surface, along the beam.

$$r=x\sin\theta+z\cos\theta \tag{10}$$

Using subscripts 1, 2, and 3 to designate the incident, refracted and reflected beams, respectively, and noting that the value of α_1 for air is

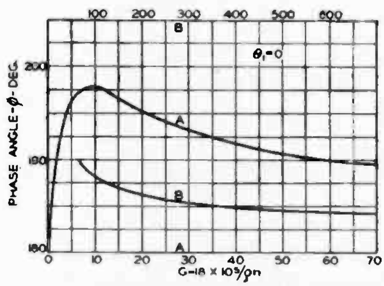

Fig. 5—Phase Angles at Normal Incidence for Various Resistivities and Frequencies.

negligible in comparison with α_2 for the earth, the electric intensities are

$$f_{y1}=F_{y1}e^{j(\omega t-\beta_1 r_1)} \tag{11}$$

$$f_{y3}=F_{y3}e^{j(\omega t+\beta_3 r_3)} \tag{12}$$

$$f_{y2}=F_{y2}e^{-\alpha_2 r_2}\times e^{j(\omega t-\beta_2 r_2)} \tag{13}$$

Considering the permeabilities $\mu_1=\mu_2=1$ for the air and the earth, also $\epsilon_1=1$ for the air, (3), (5), (7), and (9)

give
$$\frac{\omega^2}{c^2}\left(\frac{4\pi}{\rho\omega}j-\epsilon_2\right)=(\alpha_2+j\beta_2)^2 \tag{14}$$

for the earth, and
$$\frac{\omega}{c}=\beta_1=\beta_3 \tag{15}$$

for the air. From (5), (11), (12), and (13) the tangential magnetic intensities are:

$$h_{x1} = -f_{y1}\frac{c\beta_1}{\omega}\cos\theta_1 \tag{16}$$

$$h_{x3} = f_{y3}\frac{c\beta_3}{\omega}\cos\theta_3 \tag{17}$$

$$h_{x2} = -f_{y2}\frac{cj(\alpha_2+j\beta_2)}{\omega}\cos\theta_2 \tag{18}$$

Equality of tangential components at the earth-air boundary requires that for any point at the boundary and for all instants of time,

$$f_{y1} + f_{y3} = f_{y2} \tag{19}$$

and

$$h_{x1} + h_{x3} = h_{x2} \tag{20}$$

From (19) and the values of (11), (12), (13), at the earth-air boundary, $\sin\theta_1 = -\sin\theta_3$ (21), that is, the angle of incidence equals the angle of reflection, and

$$\frac{\sin\theta_2}{\sin\theta_1} = \frac{j\beta_1}{(\alpha_2+j\beta_2)} \tag{22}$$

also

$$F_{y1} + F_{y3} = F_{y2} \tag{23}$$

and from (16), (17), (18), and (20)

$$F_{y1} - F_{y3} = F_{y2}\frac{j(\alpha_2+j\beta_2)}{\beta_1}\frac{\cos\theta_2}{\cos\theta_1} \tag{24}$$

Put

$$\frac{j(\alpha_2+j\beta_2)}{\beta_1}\frac{\cos\theta_2}{\cos\theta_1} = p + jq. \tag{25}$$

From (23) and (24) the vector ratio of the reflected electric intensity to the incident electric intensity is

$$\frac{F_{y3}}{F_{y1}} = \frac{1-(p+iq)}{1+(p+jq)}.$$

The ratio of the magnitudes is

$$\frac{F_{y3}}{F_{y1}} = \frac{\sqrt{(1-p)^2+q^2}}{\sqrt{(1+p)^2+q^2}} \tag{26}$$

and the phase angle ϕ between F_{y1} and F_{y3} is

$$\phi = \tan^{-1}\frac{-q}{1-p} - \tan^{-1}\frac{q}{1+p} = \tan^{-1}\frac{2q}{p^2+q^2-1}. \tag{27}$$

The values of p and q for use in determining the amount and phase of the reflected electric intensity in terms of the incident electric intensity are determined from (14), (15), (22), and (25), and involve

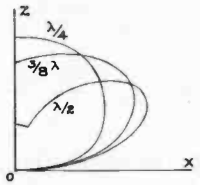

Fig. 6—Electric Intensities for Heights d of Antenna Above the Earth's Surface.

only the dielectric constant ϵ_2, the frequency n, the resistivity of the earth ρ, and the angle of incidence θ_1.

Thus, $$(p+jq)^2 = \left(\epsilon_2 - \frac{4\pi}{\rho\omega} j\right) \sec^2 \theta_1 - \tan^2 \theta_1$$

Put $$b = \epsilon_2 \sec^2 \theta_1 - \tan^2 \theta_1$$

and $$g = \frac{4\pi}{\rho\omega} \sec^2 \theta_1 = G \cdot \sec^2 \theta_1. \tag{28}$$

Then $$p = \sqrt{\frac{b + \sqrt{b^2 + g^2}}{2}} \tag{29}$$

and $$q = \sqrt{\frac{-b + \sqrt{b^2 + g^2}}{2}}. \tag{30}$$

The value of ρ in (28) is in electrostatic units. It will be more convenient to use ρ in ohms per centimeter cube, and the frequency

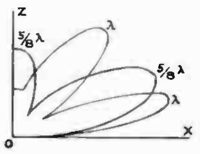

Fig. 7—Electric Intensities for Heights d of Antenna Above the Earth's Surface.

in megacycles per second. Therefore, with ρ = ohms per cm cube, and n = megacycles per sec.

$$G = \frac{18 \times 10^5}{\rho n}. \tag{31}$$

As an instance of the magnitude of G, it may be noted that for $n = 100$ megacycles per sec. (i.e., 3 meters wavelength) and $\rho = 18 \times 10^3$ ohms per cm cube, $G = 1$. Thus, taking the value of $\epsilon_2 = 6$, the values of b and g which determine for normal incidence ($\theta_1 = 0$) the coefficient of reflection and phase angle are $b = 6$ and $g = 1$.

The curves of Figs. 3, 4, and 5 have been computed from (26) and (27), taking $\epsilon_2 = 6$, and the values of G ranging from zero to very large values compared with ϵ_2. For normal incidence, the phase angle ϕ due to reflection is greatest for a value of G in the vicinity of 9, when $\epsilon_2 = 6$. For any value of b, and for sufficiently small values of n, i.e., for the longer waves, G becomes large in comparison with ϵ_2, and the earth surface may be considered a good reflecting surface.

The polar diagrams have been computed for $G = 9$. In general form, there is good agreement with the experimentally observed distributions shown by Yagi, taking into account the fact that the observations were made at such proximity to the oscillator that the converging direct and reflected rays in some instances included an angle sufficient to affect the phase computation.

Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

Discussion on
PRINCIPLES OF GRID-LEAK GRID-CONDENSER DETECTION*

(F. E. Terman)

W. J. Polydoroff[1]: The theory and methods of grid-leak grid-condenser detection as developed by Professor Terman are very valuable to radio engineers, as they deal for the first time with numerical values of the circuit constants and actual results expected.

The phenomenon of detection has been well known to us for many years, and the values computed by Professor Terman are closely in accordance with those being used for several years. Unfortunately the detector tube requirements are considerably greater than that tube can meet because of increased sensitiveness and fidelity of present-day radio receivers. The range of sensitivity of a receiver if measured on the input of the detector circuit may be determined as 0.005 volt for very weak signals (distance reception) up to 10 volts on loud local signals.

Obviously present-day detector tubes can hardly take care of efficient rectification within such large limits. If the detector is adjusted to weak signal reception (maximum sensitivity), and therefore has high values of grid-leak resistance and small capacity, loud signals of local stations are very easily distorted and high notes distortion already existing due to sideband cut-off is accentuated still more. Besides, the detector is easily overloaded by strong signals producing grid polarizing voltage and general distortion occurs. As a result a usual precaution applied to most of the receivers is that *care should be taken not to overload the detector tube.*

A compromise is reached on values of a grid leak to be about 1 megohm and grid condenser of 250 $\mu\mu$f.

The anode bent rectification in this respect is considerably more elastic within the range of signal strength and introduces less distortion, but lacks efficiency and therefore reduces overall sensitivity of a receiver.

In my belief a two-element rectifier with substantially linear rectification would be a better solution for detection, especially when it is added to the set and the detector tube converted to act as a radio-frequency amplifier to make up for the loss of the detector's amplification. A sensitive and robust two-electrode tube can be designed to act as a rectifier in itself. I had some experience with similar detectors in the past and the only objection to them is that they do not amplify as they detect. The cost of an extra inexpensive tube would be justified for the increase of working range and better fidelity. The requirements of heating energy are so small that this tube can be operated from *B* supply of all electric sets saving the greater part of troubles due to a.c. hum encountered in their design.

However, at present we have to meet the situation with the tubes we have on hand and obtain the best results. The investigation of Professor Terman shows:

"Detector is less sensitive to high modulated freqnencies, high notes are discriminated against,"

* Proc. I.R.E., 16, 1384; October, 1928.

[1] Johnson-Williamson Laboratories, Chicago, Ill.

"Grid-leak resistance affects operating grid potential,"

"Change of grid potential is simply the voltage developed across the grid-leak condenser impedance."

These quotations are taken to prove that detector characteristics are *different* at *different frequencies* and at *different incoming voltages*, and therefore for the best results detector constants should be adjustable.

In the old days when the number of controls on a panel was considered more as an asset than a liability, receivers had as many as ten variable controls, among them adjustable grid-leak and adjustable grid-condenser. Nowadays, with a tendency to simplify the operation they are out of question.

However, I would like to show an effective way to control volume, especially adaptable to the receivers using a.c. tubes. We all know that the old methods of volume control by adjusting filament rheostats or plate supply cannot satisfactorily be applied to a.c. tubes, as a very small deviation in operating conditions of these tubes will inevitably produce objectionable hum. Consequently, it is usual to control input energy to the antenna circuit; this is especially convenient when aperiodic antenna and coupling tube are employed. For the receivers not neutralized and working under oscillating conditions this method is dangerous, as the tendency to oscillate increases when signal is weakened. There are a hundred points in the circuit where a resistance can be inserted to attenuate the incoming signal, in every case either affecting selectivity or shifting the resonance point. Control of the volume on the audio side is very favorable, but does not take care of detector overload.

A few months ago while trying to design an efficient volume control for a.c. operated loop receiver I encountered some difficulties, and after many comparisons came to a conclusion that a very practical method of volume control having distortionless properties is to use old variable grid leak for volume control.

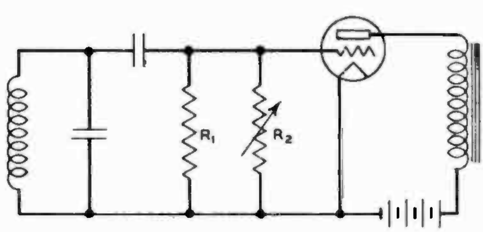

Fig. 1—The Circuit.
$R_1 = 2\omega$
$R_2 = 500{,}000 - 15{,}000$

A 2-megohm resistance is inserted in the usual way and in parallel with it a variable resistance 15,000–500,000 ohms. This variable resistance passes very small currents and therefore any plain carbon type rheostat, as used for plate supply control, can be inserted. When the resistance is in "off" position total resistance value is 2 megohms, and the detector is adjusted to maximum sensitiveness for weak signals. When 500,000 ohms are inserted in parallel it does not cause any noticeable decrease of reasonably loud signals, and for the local extra-loud signal the rheostat can be advanced to a desired volume and a signal can be brought to zero volume when total resistance is 15,000–50,000 ohms. These figures apply to any standard detector tube a.c. or d.c. type.

The system, very simple but involving several modes of operation, performs the following functions:

(1) Adjusts detector for maximum sensitiveness with undistorted rectification for any given signal strength,
(2) Regulates output volume of radio receiver from maximum undistorted signal to zero without changing characteristics of other tubes,
(3) Improves fidelity of signal when volume is tuned down,
(4) Keeps r.f. amplification all the time at its full efficiency,
(5) Broadens the selectivity of the detector tuned stage.

First four points score to the advantage of this system; the last one is objectionable. The broadening effect occurs when variable resistance values are 50,000 ohms or less, in which case the tuned circuit is short-circuited by an impedance:

$$Z=\sqrt{\left(\frac{Rr_f}{R+r_f}\right)^2+\frac{1}{C^2\omega^2}}$$

where R is variable resistance, r_f is grid-filament resistance, C is capacity of grid-condenser. According to Professor Terman's theory, the most important action which occurs in this scheme of control is the change of grid potential due to rectified current. When the signal is too great the rectified grid current changes the operated potential of the detector, which in turn causes overloading.

To avoid this, automatic means can be devised which would shift the operating grid potential of the detector in accordance with the magnitude of the carrier current receiver. This scheme is now under development and apparently offers a solution preventing detector overload. The volume control of the receiver in this case can be made in the audio parts of the apparatus, which, of course, offers no difficulties.

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Volume 17, Number 4 *April, 1929*

BOOK REVIEW

Données Numériques de Radioélectricité. EDITED BY R. MESNY. Gauthier-Villars et Cie., Paris; McGraw-Hill Book Co., Inc., New York, 1928; pp. vii+26, quarto; price bound, 30 francs, unbound, 15 francs.

The book before us is that portion of Volume VI of the "Tables Annuelles de Constantes et Données Numériques" which is devoted to data on radio. Here, for the first time in the history of the Annual Tables, space has been found for a considerable amount of attention to various matters pertaining to radio communication. Hitherto the editors have not felt that radio data had attained a sufficient degree of precision to warrant fuller inclusion.

The usefulness to radio engineers of the Section under discussion is, of course, seriously limited by the fact that it includes only data that appeared in the course of two years, and that since then five years have elapsed. Nevertheless, the years 1923–24 brought forth so considerable an amount of material that its compilation from many widely scattered sources in the present volume is well worth while.

First of all we find several pages on numerical data and characteristic curves of tubes[1] and on electron emission from filaments. Then follow a few data on detector tubes and on magnetic permeability at high frequency. Wave propagation receives considerable attention, particularly Austin's measurements of intensity of signals from various stations. Results on direction finding are summarized, together with data on wave antennas and other forms of aerial.

Among other subjects treated are atmospheric disturbances, miscellaneous data on electrical oscillations, and the tables prepared by Grover and by Curtis and Sparks for the calculation of self-inductance. It is a striking tribute to the onward march of the vacuum tube that among all these data there is no word concerning arcs, sparks or contact detectors.

It is gratifying that the Annual Tables are now giving more space to the contributions which radio is making. While recognizing fully the great amount of labor and time required in selecting and editing such data as these, still we venture to hope that Professor Mesny will find it possible in future editions to bring the material more nearly up to date.

The section on radio can be obtained separately bound in board covers. Workers in the field will find the publication a very useful addition to their libraries.

W. G. CADY†

[1] Chiefly various types of G. E. tubes from articles in the PROC. I.R.E.

† Department of Physics, Wesleyan University, Middletown, Conn.

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 various articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. Radio Communication

R007 The problem of international distribution of broadcast wavelengths: Proposals of the Polish Broadcasting Co. *Experimental Wireless and Wireless Engineer* (London), 6, pp. 3–8; January, 1929.
(Deals with European situation.)

R010 Zenneck, J. The importance of radiotelegraphy in science. Proc. I. R. E., 17, pp. 89–114; January, 1929.
(A history of the progress of the art is given.)

R020 Hund, A. Hochfrequenz-Messtechnik. (High-frequency measurements). (book). 2nd edition, 1928. Published by J. Springer, Berlin. Noted in *Experimental Wireless* and *Wireless Engineer* (London), 5, p. 22; January, 1929.
(Review of book.)

R090 Important events in radio—peaks in the waves of wireless progress—1827 to 1928. Radio Service Bulletin, No. 141; pp. 16–26; December, 1928.

R100. Radio Principles

R113 Barfield, R. H. and Munro, G. H. The attenuation of wireless waves over towns. *Experimental Wireless and Wireless Engineer* (London), 6, pp. 31–37; January, 1929.
(Experimental and theoretical discussion of attenuation of radio waves over towns and suburbs. The absorption over towns is mostly due to buildings, vertical metal conductors, etc. and the absorption due to receiving stations plays only a small part, while over suburbs the absorption is mostly due to receiving stations.)

R113 Quäck, E. and Mögel, H. Hörbarkeitsgrenzen und gungstigste Verkehrszeiten bei Hurzwellen auf den einzelnen Überseelinien. (Audibility limits and most favorable communication directions for short waves over certain transoceanic paths). *Elekt. Nach. Tech.*, 5, pp. 542–549; December, 1928.

(The charts given refer to a range of 30000 to 7500 kc. For twenty-four hours of service, one day and one night frequency are usually sufficient. For certain directions, a third wave is needed. If the rays of the sun are perpendicular to certain portions of the transmission path while the entire path is exposed to daylight, the reception is rather poor.)

R113.1 Colwell, R. C. Fading curves and weather conditions. PROC. I. R. E., 17, pp. 143–148; January, 1929.

(Sunset fading curves from KDKA made at Morgantown, W. Va. with a Shaw recorder during April and May, 1927 showed that the signal strength from KDKA during the night was sometimes below and sometimes above daylight strength. Apparent correlation with weather.)

R113.3 Duckert, P. Elektrische Schwingungen und ihre Grenzgebiete drahtlose Telegraphie—Über Fehlweisung der Funkpeilung in Abhängigkeit von Wetterlage. (Electric waves and their limited transmission: On direction shift due to weather). *Zeits. für Tech. Phys.*, 9, pp. 466–469; No. 12, 1928.

(Experiments carried on during several years (day measurements) show that the power received from a transmitter passes over a curved instead of a straight horizontal path so that the direction finder may be in error by several degrees. This is true when certain weather conditions exist, so called atmospheric border layers. During the night the direction deviations are also accompanied by fading.)

R113.4 Echos von Hertzachen Wellen (Echoes from electromagnetic waves). *Elek. Nach. Tech.* 5, p. 488; December, 1928.

(Gives the different suppositions which may explain the echo effects. One explanation (Lassen and Försterling) assumes that the rays travel along different paths in the ionized layer. The second explanation (Störmer) assumes reflections or refractions on the envelope of electron pockets of very great dimensions. The third explanation (B. van der Pol, Jr.) assumes that the group velocity of the waves through the ionized layer can be slowed down considerably for certain conditions in the layer. The fourth explanation has reference to echoes corresponding to 0.005 to 0.08 second intervals, such as observed by Taylor and Young, Hoag and Andrew, and Quick and Mögel. They are due to reflection from a layer 1500 km above the earth, from the polar nightlight zone, or from the sunset shadow wall.)

R113.4 Försterling, K. Über die Ausbreitung kurzer elektromagnetischer Wellen in der Heavisideschicht. (On the propagation of short electromagnetic waves in the Heaviside layer). *Elek. Nach. Tech.* 5, pp. 530–542; December, 1928.

(General discussion on the refraction and absorption of short waves (less than 60 meters). Assumes that the ionization is due to ultraviolet rays from the sun. The article also indicates that there can be three layers from which the atmospheric ray may descend.)

R113.4 Appleton, E. V. Some notes on wireless methods of investigating the electrical structure of the upper atmosphere. *Proc. Physical Soc.* (London), 41, pp. 43–59; December 15, 1928.

(Comparison of the wavelength change method, angle of incidence method, and the group retardation method for finding the effective height of the Kennelly-Heaviside layer. If the effect of the earth's magnetic field is neglected, the methods measure the same equivalent height which is greater than the maximum height reached by the atmospheric ray. A method for investigating the ionic concentration is given. This is partly based on an assumption by Pedersen for which the angle of incidence at the ground is greater than 30 deg.)

R113.5 Wright, C. S. Radio communication and magnetic disturbances. *Nature* (London), 122, p. 961; December 22, 1928.

(Note on the effect of magnetic disturbances on radio reception for frequencies greater than 150 kc for crystal reception.)

R113.5 Chapman, S. The ultra-violet light of the sun as the origin of aurorae and magnetic storms (letter). *Nature* (London), 122, p. 921; December 3, 1928.

(Refers to the letter by H. B. Maris and E. O. Hulburt (*Nature*, 24, 1924) on their theory of aurorae and magnetic storms. This theory assumes that occasional sudden blasts of ultra-violet light produce the aurorae and magnetic storms. Chapman expects to offer another theory of the magnetic disturbances, which will assume the cause due to a neutral ionized stream (F. A. Linderman's original suggestion) which will avoid some of the inconsistencies of the Maris and Hulburt theory.)

R114 Kenrick, G. W. The analysis of irregular motions with applications to the energy frequency spectrum of static and telegraph signals. *Philosophical Magazine* (London), **7**, pp. 176–196; January, 1929.

(The analysis shows that the energy contained within a frequency band of given width due to an atmospheric disturbance produced by a random sequence of pulses varies directly with the square of the wavelength. A numerical example shows that the energy due to pulses of the order of 10^{-3} second in duration and sharply rising exponential form can produce appreciable fields (since proportional to the square roots of the energies) on the longer wavelengths.)

R124 von Ardenne, M. Einige Messungen uber die Hochfrequenzspannungen an der Eingangsseite von Empfangern. (Some measurements on r.f. voltages induced in coil antennas). *Zeits. fur Hoch.*, **32**, pp. 199–202; December, 1928.

(An aperiodic high frequency amplifier of known amplification inserted between the coil antenna and a tube voltmeter in order to bring the sensitiveness of the tube voltmeter to the proper value. Applications to field intensity measurements on broadcast stations are given.)

R125.1 Smith-Rose, R. L. The reversibility of radio direction finders and local error at rotating loop beacon. *Jour.* I. E. E. (London), **67**, pp. 149–156; January, 1929.

(It is shown that transmission from a rotating loop beacon being received on an open antenna produces about the same kind of errors as those experienced when the antenna is employed for transmitting to a direction finder. Methods are given for eliminating local errors.)

R125.6 Meissner, A. and Rothe, H. On the determination of the optimum radiation angle for horizontal antennas. PROC. I. R. E., **17**, pp. 35–41; January, 1929.

(The most favorable radiation angle for 15 and 20-meter wavelengths was determined using horizontal multiple antennas in connection with a parabolic reflector. It was found that the most favorable radiation happened when it took place along the tangent of the surface of the earth.)

R132 Colebrook, F. M. A generalized analysis of the triode valve equivalent network. *Jour.* I. E. E., (London), **67**, pp. 157–169; January, 1929.

(Analytical treatment (graphical) of a vacuum tube amplifier. It is based on Miller's assumptions (Bureau of Standards, **15**, p. 367; 1919) with modifications due to Hartshorn (*Proc.* Physical Soc., London, **39**, p. 108, 1926–1927). One conclusion is that at very high radio frequencies the voltage amplification factor may apparently exceed the voltage factor of the tube if a pure inductive plate load is provided.)

R132 Forstmann, A. and Schramm, E. Ueber Maximalleistungen von Verstärkerröhren. (On the maximum output of electron tube amplifiers). *Zeits. für Hoch.*, **32**, pp. 195–199; December, 1928.

(Formulas are derived for optimum output of vacuum tubes which are loaded either by a resistance or by an inductance. It is shown that for a constant plate potential and a resistance load the ratio of external to internal resistance is smaller than unity.)

R132 Snavely, B. L. and Webb, J. S. Radio frequency amplifying circuits. PROC. I. R. E., **17**, pp. 118–126; January, 1929.

(An equation is developed showing relation between circuit constants and critical grid resistance in a resistance stabilized amplifier having a tuned-grid circuit and a pure inductance plate load.)

R133 Kohl, K. Über kurze ungedämpfte elektrische Wellen. (On short sustained electromagnetic waves). *Zeits. für Tech. Phys.*, **9**, pp. 472–473; No. 12, 1928.

(The author attributes the frequency of ultra short waves in the Barkhausen-Kurz circuit partially to the natural frequency of an ordinary oscillating circuit. The frequency is increased on account of the dielectric constant of the electron gas between the grid and the anode since this constant is smaller than unity. The frequency is therefore not due to a pure electronic oscillation.)

R133 Edgeworth, E. Frequency variation of the triode oscillator (letter). *Philosophical Magazine* (London), 7, pp. 200–203; January, 1929.

(Further discussion D. F. Martyn's paper (*Phil. Mag.*, Nov. 1927) with reference to the paper by E. Edgeworth in *Jour.* I. E. E., Wireless Section, Jan. 6, 1926.)

R134 Terman, F. E. and Googin, T. M. Detection characteristics of three-element vacuum tubes. Proc. I. R. E., 17, pp. 149–160; January, 1929.

(Comparative tests of different tubes using grid detection.)

R144 Butterworth, S. The high frequency resistance of toroidal coils. *Experimental Wireless and Wireless Engr.* (London). 6, pp. 13–16; January, 1929

(A discussion of the toroid coil in comparison to a single layer solenoid. It is shown that although the toroid has the advantage of being astatic it has always more high frequency resistance than a properly designed single layer coil and that the best possible toroid has more than twice as great a high frequency resistance as an equally compact single layer coil.)

R148 van der Pol, B. Some remarks on ultra short wave broadcasting. *Experimental Wireless and Wireless Engr.* (London), 6, pp. 9–12; January, 1929.

(Theory of pure amplitude modulation showing that for ordinary short wave transmitters a frequency modulation due to the resistance of the oscillator occurs in addition and that this interference can be avoided by using a piezo-electric controlled oscillator.)

R200. Radio Measurements and Standardization

R210 Hitchcock, R. C. A direct reading radio frequency meter. Proc. I. R. E., 17, pp. 24–34; January, 1929.

(A new type of direct reading r. f. meter.)

R210 Pierce, G. W. Magnetostriction oscillators. Proc. I. R. E., 17, pp. 42–88; January, 1929.

(Use of magnetostriction to produce and control electrical and mechanical frequencies of oscillations in a range of frequencies extending from a few hundred cycles per second to more than three hundred thousand cycles per second. Methods of calibration of vibrators and their use in the calibration of frequency meters and otner data pertaining to such oscillators are given.)

R214 Gerth, F. and Rachow, H. Die Temperaturabhangigkeit der Frequenz des Quarzresonators. (The temperature dependence of the frequency of quartz resonators). *Elekt. Nach. Tech.*, 5, pp. 549–551; December, 1928.

(Measurements of the absolute temperature coefficient of a quartz plate gave a value of 60 parts in one million per degree Centigrade. This figure was found for a temperature variation of 10 to 25°C. The quartz plate was silvered and determination is independent of the crystal holder.)

R214 Wright, J. W. The piezo-electric crystal oscillator. Proc. I. R. E., 17, pp. 127–142; January, 1929.

(Theory and operation of the piezo oscillator.)

R220 Wilmotte, R. M. A quick and sensitive method of measuring condenser losses at radio frequencies. *Jour. Scientific Instruments* (London), 5, pp. 369–377; December, 1928.

(The substitution method for measuring the effective condenser resistance has been improved by employing proper screening. Curves for several air condensers are given. The method can be used from up to approximately 6000 kc.)

R220 Weihe, W. Die Messung von Kapazitäten mit dem Ueberlagerungsverfahren. (The measurement of capacities with the heterodyne principle). *Zeits. für Hoch.*, 32, pp. 185–194; December, 1928.

(A beat note method using the silent zone is employed for measuring the effective grid-filament capacity for different external anode resistances (for 2000 to 10,000 kc.)

R300. Radio Apparatus and Equipment

R334 Westman, H. P. UV–861, a screen-grid tube for the high power amateur transmitter. *QST*, 13, pp. 41–43; February, 1929.

(Characteristics of tube.)

R341 Pike, O. W. and Maser, H. T. A new type of rectifier tube for amateur use. *QST*, 13, pp. 20–22; February, 1929.

(Description of UX-866 rectifier tube known as the hot cathode mercury vapor rectifier. Characteristics of tube are given.)

R351 Lampkin, G. F. An auxiliary frequency control for r.f. oscillators. Proc. I. R. E., 17, pp. 115–117; January, 1929.

(A method is described for varying the frequency of an oscillator in small amounts by the use of a control which operates on the normally fixed element in the oscillating circuit.)

R386 Plebanski, J. Filtering antennas and filter-valve circuits. Proc. I. R. E., 17, pp. 161–73; January, 1929.

(Describes methods of coupling together many circuits or antennas giving them simultaneous excitation from the same source of power. The purpose of these arrangements is the construction of practical filter circuits (filtering antennas) giving square topped resonance curves with good efficiency.)

R400. Radio Communication Systems

R401 Wollner, E. Die Fernsprechverbindung zwischen Europa und Amerika. (Long wave communication between Europe and America). *Elek. Nach. Tech.*, 5, pp. 489–522; December, 1928.

(A review of transoceanic radio communication with long waves (corresponding to a range from 17 to 60 kc).)

R470 Bodie, C. A. and Curtis, R. C. The transmission of high frequency currents for communication over existing power networks. *Journal* A. I. E. E., 48, pp. 37–41; January, 1929.

(Tuned choke coils are used to isolate the communication channel from the remainder of the power system. Such a system gives improved quality of speech, reduction in the noise level, freedom from variation in signal due to switching.)

R500. Applications of Radio

R526.4 The radio altimeter. *Science and Invention*, pp. 952–53; February, 1929.

(General Electric Company altimeter designed by E. F. W. Alexanderson.)

R536 Gleason, C. S. How radio prospecting takes the gamble out of mining. *Radio News*, 10, pp. 716–719; February, 1929.

(Methods by which geologists study the distribution of minerals under ground.)

R592 Moullin, E. B. Radiotelegraphy and radiotelephony. *Jour.* I. E. E., (London), 67, pp. 170-176; January, 1929.

(Review of many phases of radio. It includes the achievements in beam communication, electron tubes, transatlantic telephony, action of the Kennelly-Heaviside layer and other miscellaneous branches in radio.)

R800. Non-Radio Subjects

537.65 Vigoureaux, P. Development of formulae for the constants of the equivalent electrical circuit of a quartz resonator in terms of the elastic and piezo-electric constants. *Philosophical Magazine* (London), 6, pp. 1140–1153; December, 1928.

(The equivalent electric constants of a quartz resonator are derived for the case of a long bar using Lamb's equation and the equivalence between the electrical and mechanical vibrating systems.)

621.385 Herd, J. F. The transmission unit and its application to radio measurements. *Experimental Wireless and Wireless Engr.*(London), 6, pp. 17–22; January, 1929.

(Brings out the merit of the TU and compares the same with the Napier unit used in some countries.

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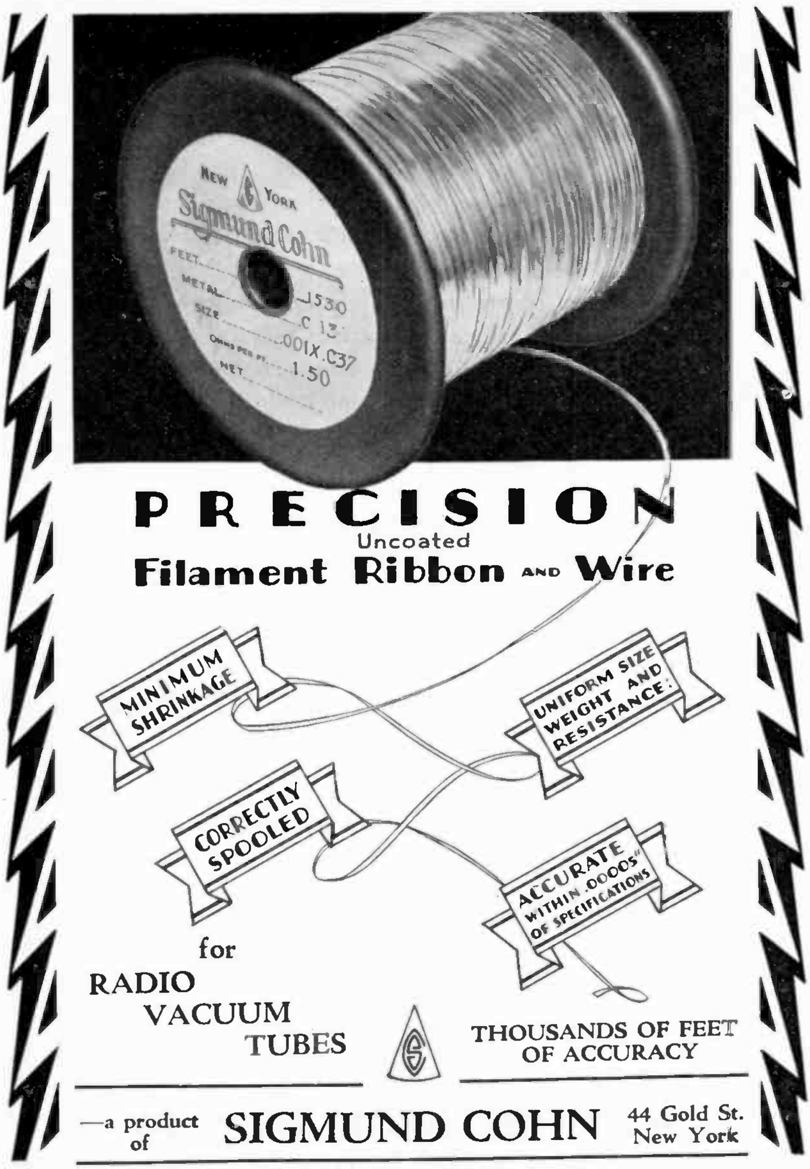
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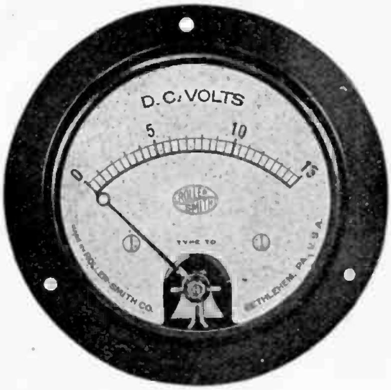

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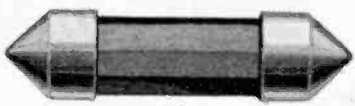

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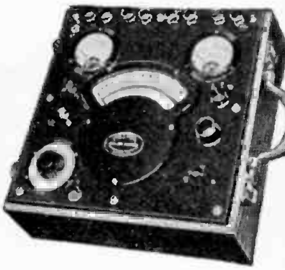

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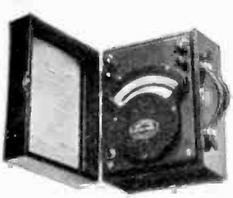

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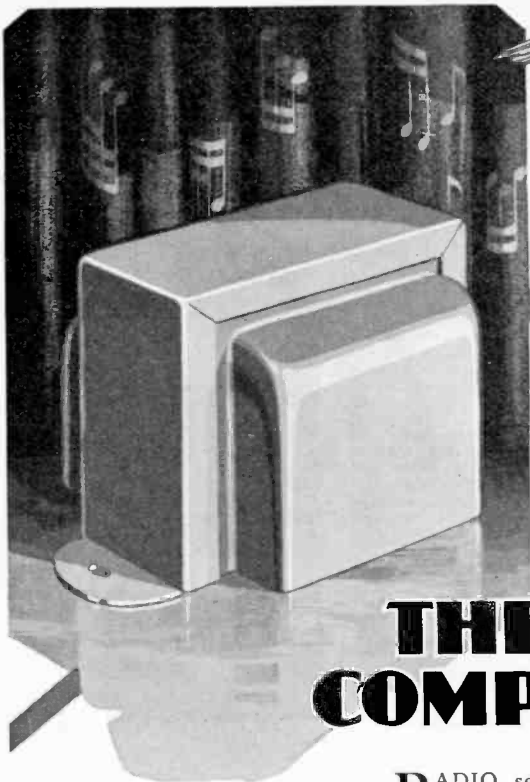

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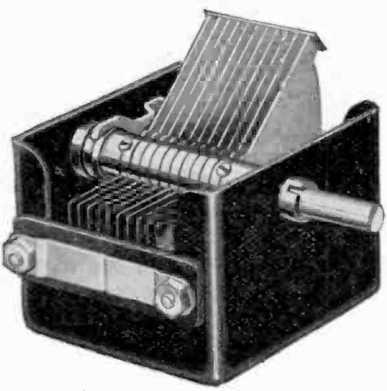

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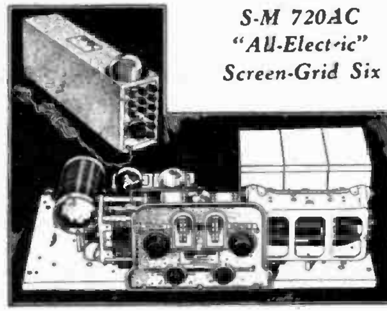

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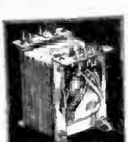

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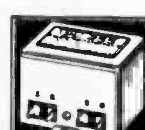

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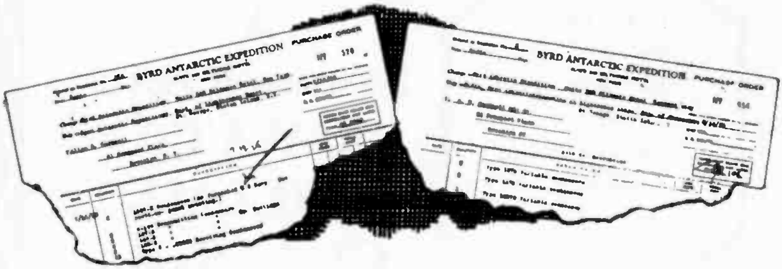

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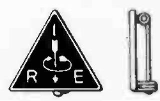

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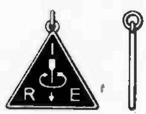

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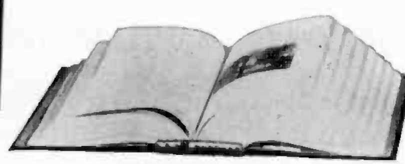

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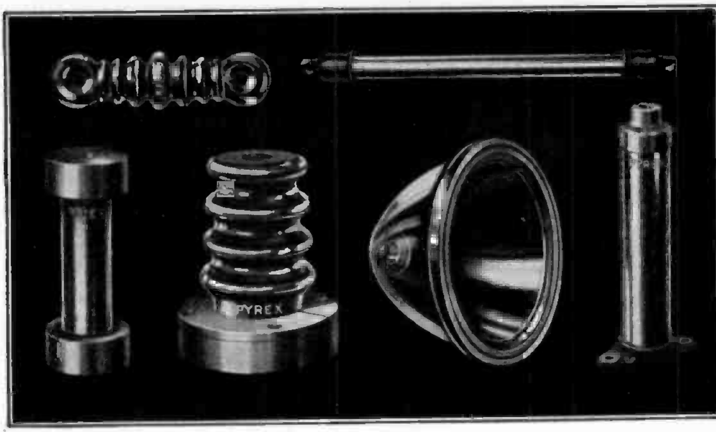

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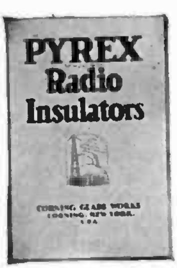

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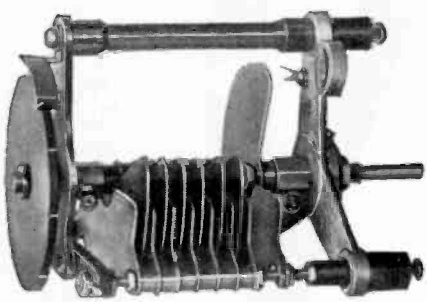

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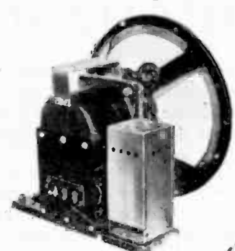

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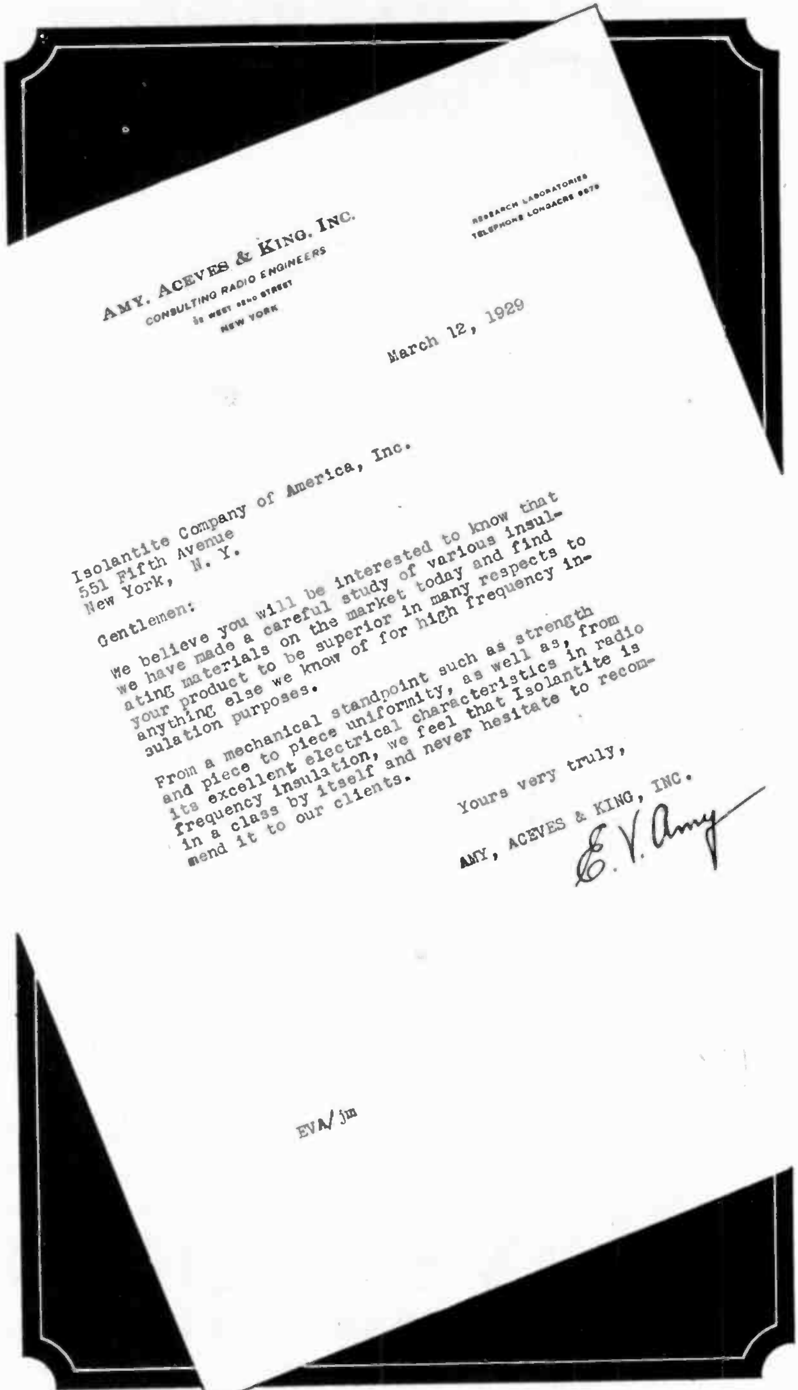

AMY, ACEVES & KING, INC.
CONSULTING RADIO ENGINEERS
NEW YORK

RESEARCH LABORATORIES
TELEPHONE LONGACRE

March 12, 1929

Isolantite Company of America, Inc.
551 Fifth Avenue
New York, N. Y.

Gentlemen:

We believe you will be interested to know that we have made a careful study of various insulating materials on the market today and find your product to be superior in many respects to anything else we know of for high frequency insulation purposes.

From a mechanical standpoint such as strength and piece to piece uniformity, as well as, from its excellent electrical characteristics in radio frequency insulation, we feel that Isolantite is in a class by itself and never hesitate to recommend it to our clients.

Yours very truly,

AMY, ACEVES & KING, INC.

E. V. Amy

EVA/jm

PROFESSIONAL ENGINEERING DIRECTORY

For Consultants in Radio and Allied Engineering Fields

(Continued)

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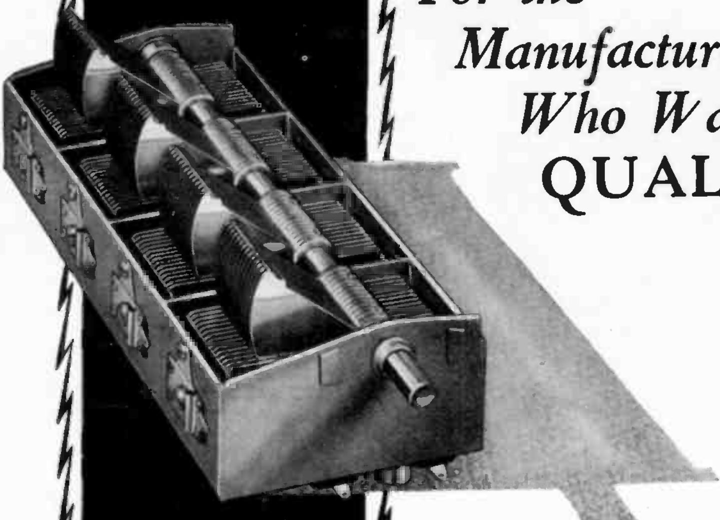

...and Polymet makes them"

The customer who said that didn't have to see them, test them, try them. He *knew*, from experience. that if Polymet makes them, they're o. k.

He was buying coils. He needed good coils, promptly delivered, at a right price. He knew that Polymet had gone into the coil business from an announcement he had read. He had used Polymet Condensers and Resistances before. Confidently, then, he placed his order for coils with Polymet.

This is the sort of good-will that is a real asset to a business. We are proud of ours and intend to keep it intact—to add to it—to continue our unvarying policy of Polymet Quality, Service, and Dependability.

Polymet supplies from the largest manufacturers to the smallest set builders with Quality electric set essentials.

Write for our catalogue.

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